5.75

$1.50

ham radio magazine

JANUARY 1978

JANUARY 1978

ham radio magazine

JANUARY 1978

volume 11, number 1

T. H. Tenney, Jr., W1NLB
publisher

James R. Fisk, W1HR
editor-in-chief

editorial staff

Charles J. Carroll, K1XX
Alfred Wilson, W6NIF
assistant editors

Patricia A. Hawes, WA1WPM
Thomas F. McMullen, Jr., W1SL
Joseph J. Schroeder, W9JUV
associate editors

Wayne T. Pierce, K3SUK
cover

publishing staff

Harold P. Kent, WA1WPP
assistant publisher

Fred D. Moller, Jr., WA1USO
advertising manager

James H. Gray, W1XU
assistant advertising manager

Therese R. Bourgault
circulation manager

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

U.S. and Canada: one year, $12.00
two years, $22.00
three years, $30.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00
two years, $45.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00
two years, $34.00

foreign subscription agents

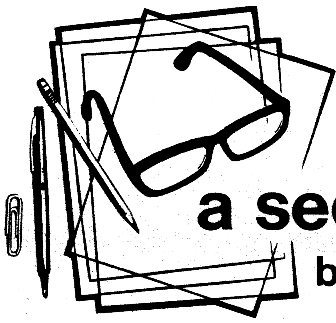
Foreign subscription agents are
listed on page 117

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 23340

contents



a second look

by Jim Fisk

Seventy-five years ago this month, on January 19th, 1903, the first two-way radio transmission was completed between the United States and England when President Theodore Roosevelt exchanged greetings with King Edward VII of England through the Marconi wireless station at South Wellfleet on Cape Cod. Only ten years earlier it was an achievement which would have seemed impossible, even to the most foresighted scientists; they were delighted when their crude laboratory apparatus could be made to span a few meters — to transmit a radio signal across 5000 kilometers of ocean was beyond their wildest imagination. Young Guglielmo Marconi, not yet 30 years old when he accomplished the feat, was not restrained by such learned skepticism, but from the viewpoint of his first attic laboratory in 1894, even he would have been surprised by his great successes of the future.

The son of a well-to-do land owner, Marconi had little formal technical training, but through his family's friendship with Professor Righi at the University of Bologna he was allowed to audit the professor's physics classes. It was there that Marconi was first exposed to the radio experiments of Heinrich Hertz, Oliver Lodge, and others. He soon set up a small lab in the attic of his father's large estate and began to experiment with electromagnetic phenomena. With Righi's spark gap mounted in the center of a short dipole and Lodge's coherer (a glass tube filled with metal filings) for his receiver, he was able to transmit radio signals that rang a bell at the other end of the attic.

Guglielmo moved his equipment out into the yard and was soon working over distances of 50 meters or more. He found that he could extend his range by increasing the height of his antenna, but with greater heights it became more and more impractical to install the spark gap at the center of the antenna. When he installed the spark gap at ground level and connected it between the antenna and ground, his range increased dramatically. What was happening, of course, was that greater antenna heights meant lower operating frequencies since the dipole was the only resonant circuit in the transmitter, and the longer wavelengths provided greater range (Marconi's earliest experiments were at about 300 MHz, which limited transmissions to line of sight).

In 1896 Marconi took his apparatus to England where some friends had arranged a meeting with William Preece, chief engineer of the British Post Office, the governmental department which had jurisdiction over all electrical communication in Great Britain. Preece was apparently impressed by the young Italian because arrangements were quickly made for a practical demonstration of the equipment for various British officials. By the autumn of 1896 Marconi had transmitted and received signals over a distance of 3 kilometers at Salisbury; in the spring of 1897 he transmitted signals across the Bristol Channel (14km); and later that year he communicated between two ships at sea 16 km apart.

Although Marconi was enjoying the full support of Preece at the Post Office, apparently the bureaucracy moved too cautiously (or released funds too slowly), for he formed his own firm, the Wireless Signal and Telegraph Company in July, 1897 (in 1900 the name of the company was changed to Marconi's Wireless Telegraph Company).

In the late 1890s Marconi continued to improve his equipment, and, in 1899, at the invitation of the French government, he bridged the English Channel, a distance of more than 50 km. By 1900 science had progressed to the point where signalling over distances of 300 km was possible, and Marconi began hinting that even the wide expanse of the Atlantic was not a barrier to wireless. In October the Marconi Company began construction of what was to be the most powerful wireless station in the world, at Poldhu Point in southwest England; by November, 1901, all was in readiness and Marconi sailed for St. John's, Newfoundland, the point in North America which was nearest to Poldhu. On December 12th, he and his assistants were successful in copying the Morse letter *S* transmitted from Poldhu — thirteen months later two-way wireless communications across the Atlantic were accomplished and the era of long-distance radio had begun.

Marconi had a great deal of respect for the radio amateurs, for it was amateurs who showed that the short waves were not a "vast wasteland," as many scientists of the day believed, but were more valuable to long-distance radio than the lower frequencies favored by the commercial interests. In his later years Marconi often referred to himself as an "amateur" — it's only fitting that the 75th anniversary of his two-way radio transmission across the Atlantic will be celebrated this month on the amateur bands by KM1CC, a special events station operated by amateurs on Cape Cod. Look for them on 160 through 10 meters the week of January 14th. During the same period amateur stations will also be operating from the original Marconi transmitting sites at Poldhu and Clifden, Ireland.

Jim Fisk, W1HR
editor-in-chief

broadcast quality television camera

How to combine a sync generator with three additional circuit boards to obtain a versatile, high-quality TV camera

This television camera's main features combine high quality and ruggedness with relatively low cost and easily obtainable components. Printed-circuit construction and detailed circuit descriptions were used to produce a project that will be operationally complete; it's not just another partially assembled unit, hastily placed under the workbench because of the lack of parts, technical knowledge, or article documentation. An attempt has been made to keep the special tools and materials stocked only by machinists separate from this project. And though I have found that compromises must be made to most ideal goals, every practical effort was made to keep these compromises to a minimum.

A number of interesting features were incorporated into this camera. Most of them allow flexibility, with applications extending beyond ATV use. Indeed, application flexibility weighs very heavily for any homebrew project, so study the following features and compare them to other cameras and your intended use.

1. Resolution capability. In excess of 500 lines. The video processor has a 3 dB response of approximately 6.5 MHz but actual resolution depends largely upon the quality of lens, Vidicon, and yoke in that order.

2. EIA interlaced scanning. Commercial broadcast quality sync is provided to further enhance stability, resolution capability, and weak signal lock-in when the picture is viewed through snow. Also, if a video tape recorder is used, exceptional frame lock stability is provided.

3. Crystal-controlled timebase. A 3.15 MHz crystal is used to derive horizontal and vertical scanning, eliminating a 60-Hz line requirement.

4. High acceleration voltages for the Vidicon. Resolution is basically increased as the G3-G4 grid voltage is raised but the main intent is to enhance the Vidicon's amplitude response, therefore boosting the performance of weak Vidicons.

5. High-video output. 1 volt p-p positive going video is available at the output connector. The video is ac coupled for 75-ohm line drive requirements providing sufficient video for even the most stubborn modulator.

6. Simple operating controls. The normal focus, target, and beam controls are rear panel mounted. No linearity adjustments are needed because of the current feedback, in the ramp generators, which provides a linear sweep.

7. Poor-man's special effects. Variable vertical and horizontal blanking is provided as an option for multi-camera superimposed image applications.

By Arthur Towslee, WA8RMC, 180 Fairdale Avenue, Westerville, Ohio 43081

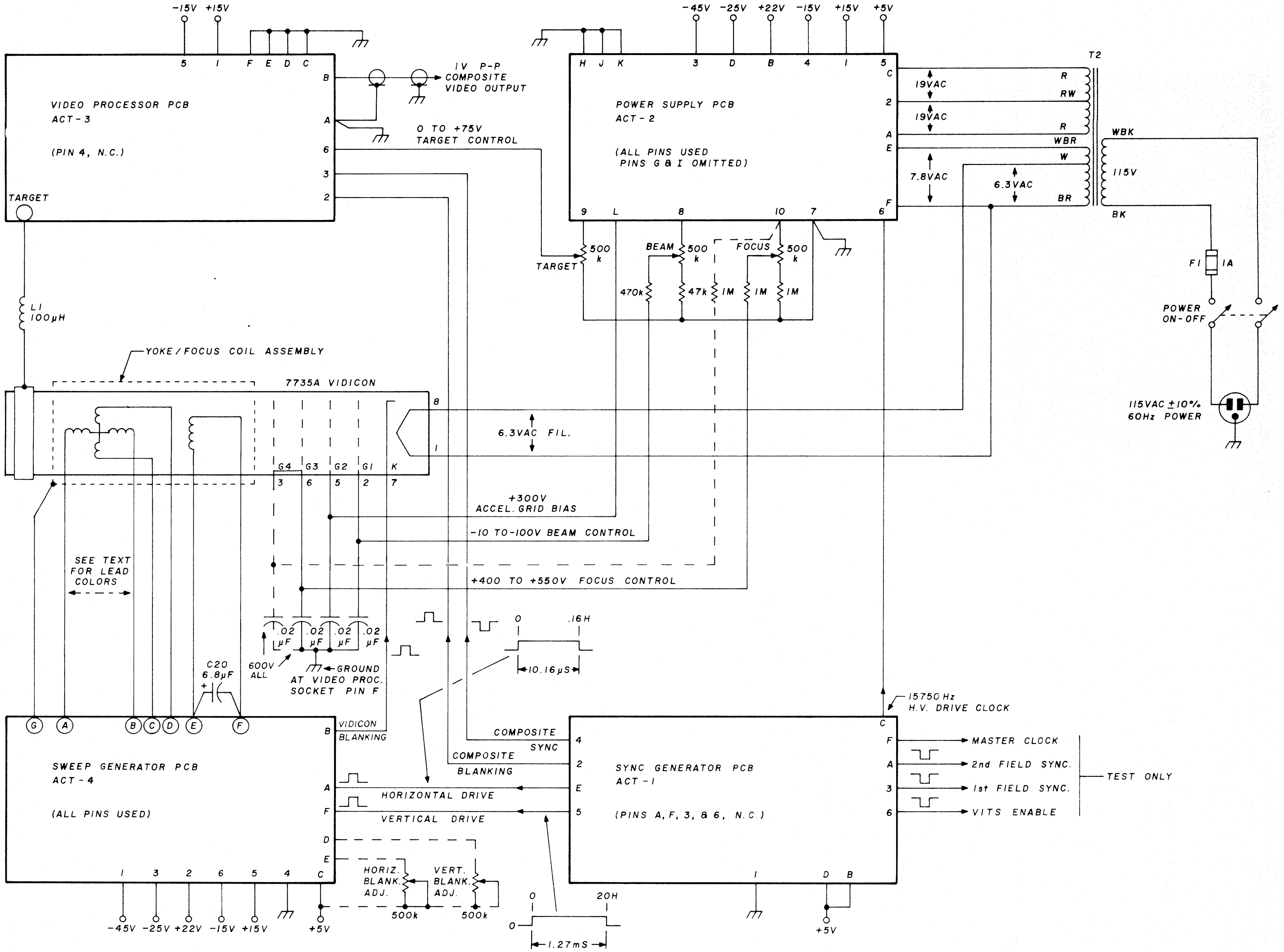


fig. 1. Wiring diagram for the frame of the camera. The connector for the power supply printed-circuit board is a Cinch 50-20A-30. The other boards use the Cinch 50-12A-30. The C-mount lens-mounting plate (3833), yoke and focus coil assembly (2045), and Vidicon socket (2100) are available from Denson Electronics. Box 85, Vernon. Connecticut 06066. A suitable power transformer (ACT-760512.0) is available from Automation Engineering for $10.00, postpaid. A set of four etched and drilled circuit boards is available for $25.00, postpaid, also from Automation Engineering.

8. Low power requirements. Only 25 watts of ac power is required; portable applications can make use of low-power inverters when 12 Vdc operation is used.

9. Printed-circuit construction. Four **3** x 6-inch (7.6x15.2cm) circuit boards are used, thereby eliminating most hand wiring.

10. Rugged and simple construction. Easy-to-obtain aluminum is used. The frame approach produces a camera that is mechanically rugged and allows easy access to all components.

11. Vidicon and yoke flexibility. The circuit design allows for many different magnetic focus/deflection Vidicons to be used. Also, a wide variety of deflection yokes can be accommodated.

12. Low cost. I estimate that the average amateur, with a well stocked junk box, can build this camera for less than 150 dollars. New parts cost (via surplus outlets) is approximately 275 dollars.

13. Optional automatic light compensation. By the addition of a 500 to 1000 megohm resistor in the video processor, reasonably good light compensation can be obtained.

14. Clamped black video level. As the average scene illumination changes, the black level position remains constant. This produces a constant reference level needed for proper setup of bias levels in video modulators and final rf amplifiers.

I've found all of these features to be important, and are lacking in many camera designs. The origin of this design grew from seeing many other camera designs; I liked some of the features of each but not all in any one camera. Thus, I undertook the task of designing a camera from scratch, the way I wanted it. Also, I was determined to finish this project before starting another, and I've got a lot of things around the house that need attention!

general description

A block wiring diagram of the camera is shown in **fig. 1.** The role of the power supply is obvious, and includes the generation of the +450 Vdc needed for Vidicon operation. All main timing and beam scan control is provided by the sync generator which was covered in detail in the September, 1977, issue of *ham radio.* This board provides the clock for the high voltage generation in the power supply, along with the horizontal and vertical pulse information to operate the sweep generator and video processor circuit boards. The sweep generator supplies the operating voltages for the yoke/focus coil and Vidicon. The

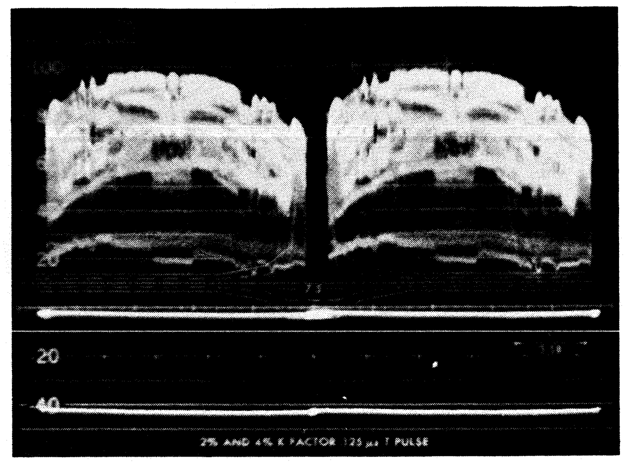


fig. 2. Photograph of the vertical video waveform. The blanking pulse can be seen as a slightly intensified portion of the base line, while the sync pulse is visible as an intensified portion on the – 40line.

video processor circuit board, as the name implies, processes the minute Vidicon target current into composite video, inserting sync and blanking on the output video. Typical horizontal and vertical output waveforms, at the video output connector, are shown in **figs. 2** and **3.**

Although complete circuit description and theory of operation becomes encyclopedic, and the resultant space would be prohibitively large, I hope that sufficient information is given here to enable you to properly troubleshoot and alter the circuitry as required to suit your own requirements. Most parts are non-critical and substitutions can be made if full knowledge of their function is understood. There are many areas which are considered "designer's choice," implying that there is more than one way to achieve the same result.

Sync generator. All of the outputs provided on this circuit board are not used, so I'll recap only the functions needed and generally describe what their role is in this camera.

First, the high-voltage power supply uses the 15-kHz square wave at pin C. The composite sync at pin 4 and the composite blanking at pin 2 are fed directly to the video processor for insertion into the raw video. The horizontal drive (pin E) and the vertical drive (pin 5) feed the sweep generator to trigger the respective ramp generators and also to provide Vidicon blanking during scanning beam retrace. The high-voltage disable input is not used so it must be connected to pin D (+5V) to enable the high-voltage drive.

Power supply. A great deal of effort was made to provide a complete power supply **(fig. 4)** on one circuit board that was as simple as possible, easy to troubleshoot, and used commonly available components. I almost made it, with the exception of the high-voltage transformer.

A +5 Vdc power supply circuit is used with the popular LM309K regulator IC as the series pass element. Particular attention should be given to the transformer secondary ac voltage. I point this out because I've seen many projects using a 6.3 Vac winding to obtain regulated 5 Vdc from a *bridge* rectifier. It will be found that most of these power supplies drop out of regulation at ***110 Vac input,*** even with very high filter capacitor values. It's possible to use a center-tapped 12 Vac transformer with a full-wave center-tap arrangement, as shown in the sync generator article, since there is only one diode voltage drop. A diode bridge has an extra voltage drop to add in, thus a higher ac voltage is required to maintain regulation.

A simple, little known formula for determining the minimum filter capacitance is presented here for those who may want to calculate the minimum value of C4.

First, the average dc voltage at the filter capacitor (C4) is roughly 1.1 times the secondary ac rms voltage. Using this information, apply the formula $CV = IT$

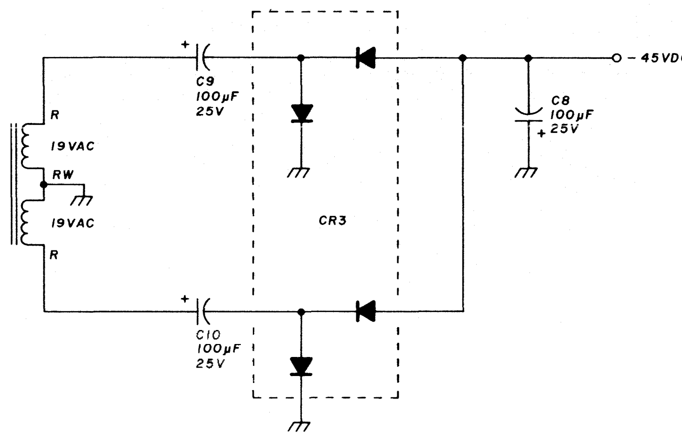
where C = Capacitor value in farads
V = Peak ripple voltage at the filter capacitor in volts
I = Load current at the output of the regulator in amperes
T = Time between charge peaks in seconds

For the LM309 to regulate, a minimum input to output differential of 1.5 volts must be maintained. Therefore, at 5 Vdc output, the input must never go below 6.5 volts. At a low line voltage of 105 Vac, the dc level will drop to 7.3 Vdc. Therefore, only 0.8 volt of ripple is allowed. These facts produce the following:

$$C = \frac{IT}{V} = \frac{0.00833 \times 0.375}{0.8} = 0.003906 \text{ farads } (3906\mu F)$$

Therefore, at least 3900µF of capacitance is required. I might have parted a bit from the main topic here, but I feel the above information is sadly lacking from typical power-supply designs. I'm a collector of rule-of-thumb formulas and this seems to be a good place to exercise this one.

The +15 Vdc regulator circuit is similar to the +5-volt circuit except a full-wave center-tap rectifier is used. The output is taken from U1. The – 15 Vdc regulator uses the other half of the CR2 bridge with the +15 Vdc output as a reference. This provides tracking between the two supplies. U3 is used as an error amplifier to compare the R1-R2 connection to ground. If +15 and – 15 volts are equal and opposite, and R1 = R2, zero volt will be present between pins 2 and 3 of U3. If not, U3 will drive Q2 to change the minus voltage until the difference is zero. This arrangement works quite well despite the fact that there are dual-tracking regulators on the market doing the same job. The most cost effective approach is used here. The last of the low-voltage rectifier circuits is the –45 Vdc supply. This is a full-wave bridge doubler composed of C8, C9, C10, and CR3. This combination is used to eliminate the need for another transformer winding. The circuit, if redrawn, will reveal that it is actually two cascade voltage doublers arranged in push-pull to provide full-wave rectification with better regulation than a half-wave arrangement.



The power transformer used to supply the voltages for the previous circuits is the size of a 6.3 Vac 3-amp filament transformer. Multiple transformers to obtain the required voltages could be used here, but to fit into the available space I have designed a single unit to fulfill these specific needs. It is also possible to wind a transformer by removing the secondary of a filament transformer and winding back on the required wire (if you ***really*** like to wind transformers).

The high voltage needed to operate the Vidicon is obtained by the use of a dc-dc converter driven by the 15750-Hz square wave from the sync generator. T1 is a ferrite cup core around a hand-wound bobbin. In operation, when the logic level drive signal at pin 6 goes positive, Q1 turns on and saturates the core of T1. When Q1 is turned off, the collapsing field in the primary of T1 is transferred to the secondary with a magnitude determined by the turns ratio. The ac voltage induced across the secondary is then rectified and filtered in a conventional manner.

The circuit feeding the focus potentiometer is a half-wave cascade voltage doubler formed by C19, CR4, and CR5. All other voltages are obtained from simple half-wave rectification circuits. The filter on the primary of T1 is used to prevent the switching spikes and the collapsing field of T1 from being fed back into the +15 volt regulated supply. The point that Q1 switches is halfway across the horizontal active scan, so if care is not taken to suppress the spikes, they will appear as a vertical line in the center of the picture.

Various chokes were tried for L1. The best commercially available unit I could find is a Miller 100-μH hash choke. However, for those who find this item

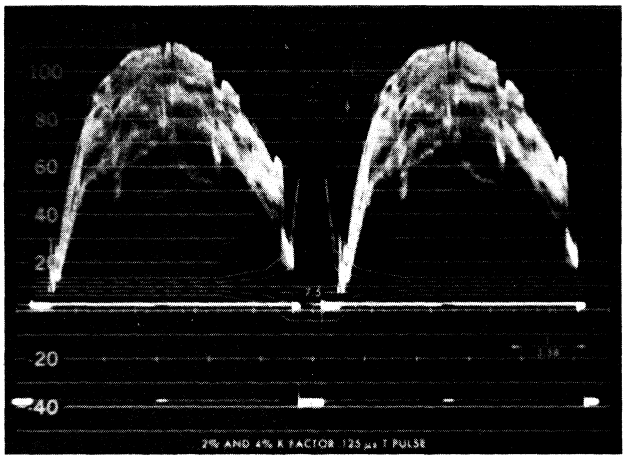


fig. 3. Horizontal video waveform as seen on an analyzer. The sync pulse is seen on the −40 line.

hard to obtain, a toroid-wound choke consisting of 24 turns of no. 22 AWG (0.6mm) enameled wire wound on an FT50-75 core is quite satisfactory.

The high-voltage transformer T1 is a homemade unit wound on a Magnetics OF42213-UG ferrite core and a PCB2213-23 bobbin. Actual winding and assembly is quite simple and total construction time should be less than 15 minutes; careful winding of the primary will save time later. See the construction section for full details.

sweep generator

The sweep generator (fig. 5) provides a number of functions; first and most important is the generation of linear ramp currents to sweep the electron beam across the target area of the Vidicon. In addition, logic is provided to blank the Vidicon during ramp retrace and also to blank the Vidicon in case of sweep failure. Finally, a constant-current source supplies the focus coil with the proper current to establish magnetic focus for the Vidicon.

Vertical ramp generator. Operational amplifier U1A acts as an integrater to produce a vertical ramp controlled by the vertical drive pulse. This circuit is different than most integraters because the integrating capacitor C1 is referenced to ground,[1] thereby simplifying the vertical ramp reset circuitry (Q1). This configuration also allows for any required linearity correction. The component values shown provide a linear ramp (R1 and R2 are equal) but the output is an exponential function in which the magnitude and sign of the exponent may be varied by changing the ratio of R1 and R2 while holding the total resistance constant. The sum of R1 plus R2 multiplied by C1 [$C1(R1+R2)$] determines the slope of the ramp. Values of C1 other than 1 μF may be used providing the RC constant remains unchanged and the value of R1 and R2 does not exceed approximately 250 kilohm. In fig. 5, C1 = 1 μF while R1 and R2 equal 100,000 ohms. Therefore, the RC constant is 200,000. If C1 was 10 μF, then R1 and R2 must each be 10,000 ohms.

The amount of positive feedback through R4 and R5 controls the exponential output. These resistors must be equal for a linear ramp but the main function is to set the gain (gain of 2). Any value of resistance for R4 and R5 between 4.7k and 100k produces no significant change in the output. Caution must be exercised when selecting values for C1. Stable, low-leakage capacitors (preferably non-electrolytic) must be used. I specify a mylar capacitor but tantalum units may be substituted if the correct polarity is observed. In operation, the positive-going vertical drive pulse resets the ramp generator by turning on Q1 and discharging C1 to ground. After the vertical pulse is completed, C1 linearly charges producing the +5 volt output at pin 12 of U1A. The next pulse resets Q1 and the cycle repeats again. The relatively large signal output (+5 volts) from U1A is intentional, to produce a large signal-to-noise ratio.

The second stage (U1B) is a voltage follower placed in the current loop to provide adjustable current gain and dc offset. The signal is reduced before pin 6 to approximately 0.6 V p-p and then compared to the 0.6 V p-p signal at pin 7. The output from pin 10 then feeds a complementary transistor pair (Q2 and Q3) which provides a current boost to drive the vertical yoke. R14 and R15 bias Q2 on slightly before Q1 turns off, eliminating crossover distortion; R16 also helps supply current at the crossover point. R17 and R18 are simply current limiters for the transistors and also provide some degree of decoupling.

Capacitor C3 shunts the small component of horizontal ramp signal around the vertical coils. R21 is used to sense the current through the vertical yoke since the voltage developed across this resistor is proportional to the current flowing through the coils. R19 and R20 form a voltage divider across R21 and

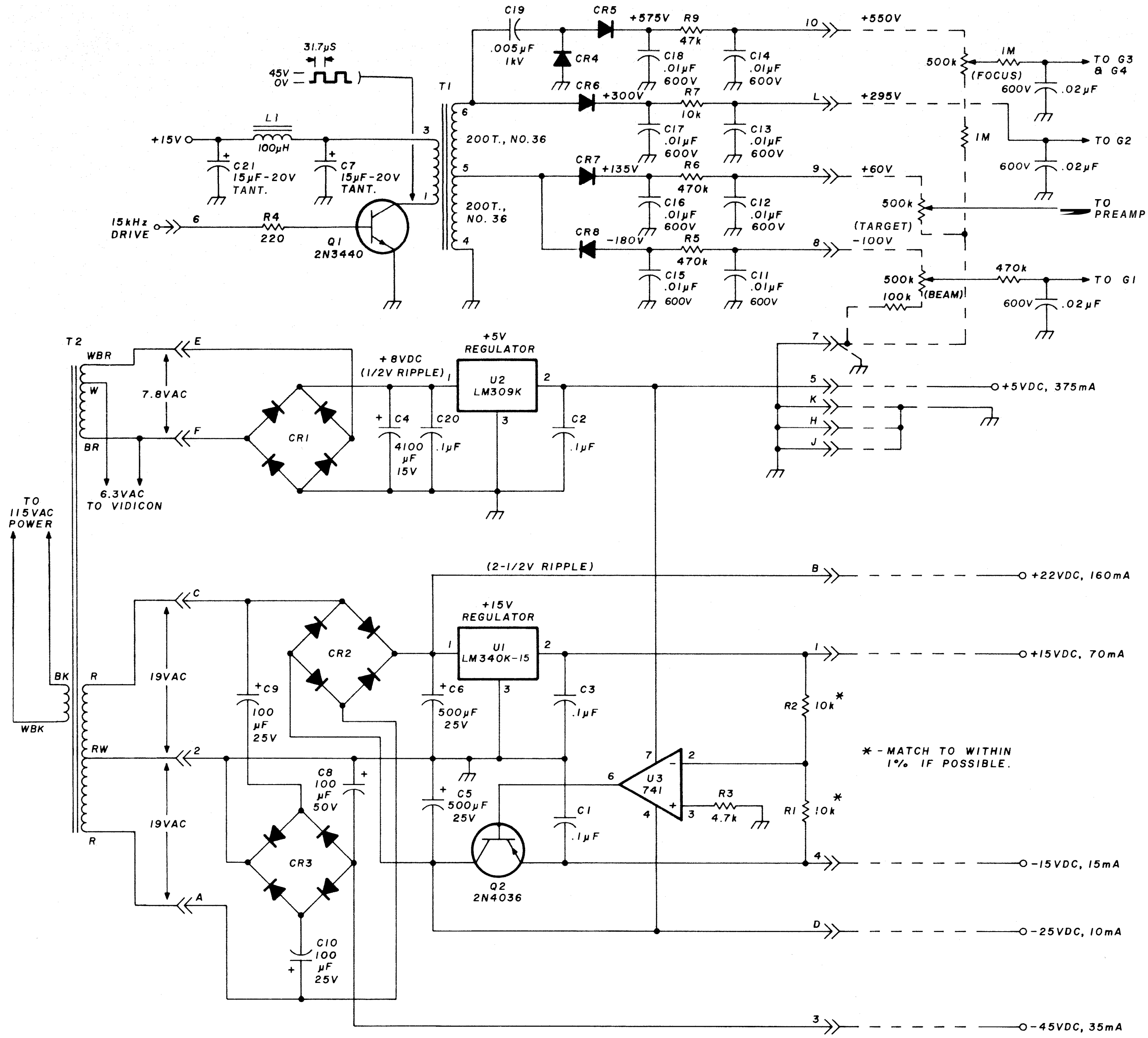


fig. 4. Schematic diagram of the power supply board. Diodes **CR1**, **CR2**, and **CR3** are Varo **VE18** diode bridges, or can be replaced with individual **1N4004** diodes. All other diodes are **1N4007s. R1** and R2 should be matched to within **1** per cent; all other resistors are **10** per cent tolerance. The 100 μ**H** choke is a J. W. Miller 5250. All the capacitor values are minimum values which can be increased if the new capacitor will fit within the available space. The high-voltage transformer is mounted by an **8-32 (M3.5)** screw through the circuit board. Flat washers should be used to space the transformer approximately **1/8** inch **(3mm)** above the board.

feed back a portion of the voltage to the inverting input of U1B. The overall gain of the loop is determined by

$$\frac{R11+R13}{R11} E_{in}=E_o$$

where E_{in} is the voltage at pin 7 and $E,$ is the voltage at R20's wiper.

By changing the vertical yoke current, the height is varied with minimum current equal to maximum height and vice versa. Yokes with different deflection factors can be accommodated by changing the value of R21 which changes the range of the height potentiometer R20. The value used here (100 ohms) will yield an R20 range of 12 to 80 mA. Decreasing this value to 50 ohms would produce a range of 25 to 160 mA. However, most deflection factors fall in the range of 12 to 80 mA. C2 compensates U1B and prevents ringing due to overshoot by reducing the gain of U1B at high frequencies. Centering pot R10

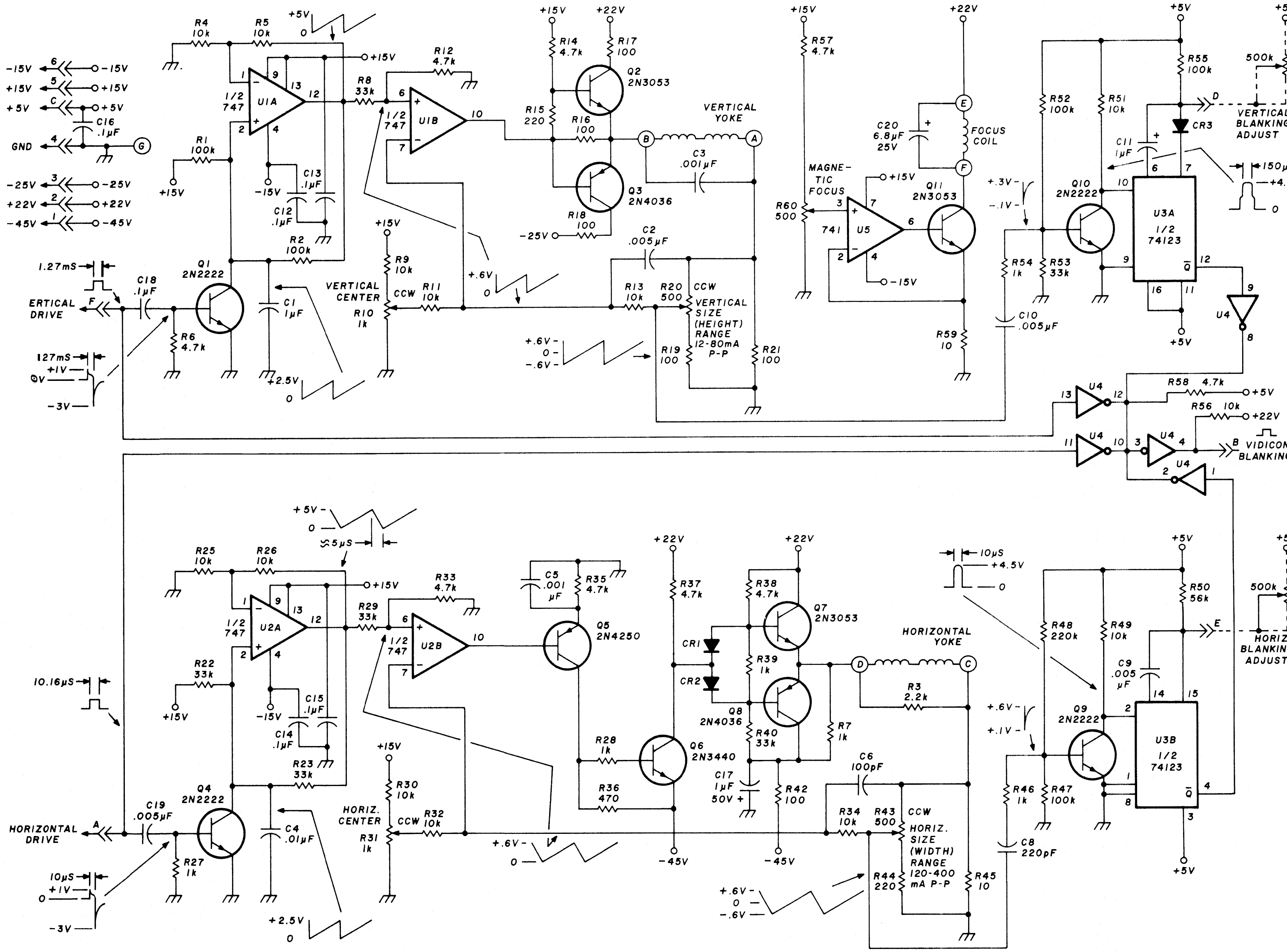
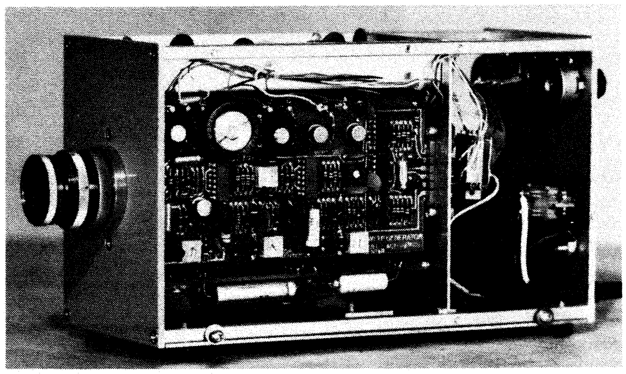


fig. 5. R7 is a ½-watt, 10 per cent, composition resistor. All other resistors, except the potentiometers, are ¼-watt, 10 per cent. The potentiometers can be either Helipot 72PM, Bourns 3386-P, or Bourns 3386-F. All diodes are 1N4151 but most other silicon diodes are acceptible. The heatsinks are Thermalloy 2212B-5 or 2227-B for Q8, and Thermalloy 2228-B for Q7 and Q11. C1 and C4 are mylar capacitors; C11 should be a tantalum capacitor.



This side view of the camera shows the sweep generator card. The additional components for the alignment coil can be seen on the socket for this circuit board.

provides a variable dc reference to center the ramp symmetrically about zero. Increasing the bias shifts the picture on the monitor from the top to bottom.

Horizontal ramp generator. This circuit is identical in concept to the vertical ramp generator discussed above. In fact, a visual inspection of the schematic will reveal only one added part, R24 in the circuit of U2A. When the horizontal pulse resets the ramp, C4 is not completely discharged due to the 10-ohm resistor, but left with a slightly positive bias which shows up at pin 6 of U2A as an offset. This small offset opposes that presented by the horizontal centering control and will provide sufficient adjustment on either side of zero.

The slew-rate upper-frequency limitation of U2A is used to an advantage here. Since slew rate is measured in volts per unit of time, the higher the ramp voltage becomes, the more time it takes to retrace. Steep slope retrace times will tend to produce ringing in the following stages; By keeping the peak ramp voltage high (+5V), the retrace takes longer due to slew-rate limiting and yoke ringing is minimized. The resistive divider (R29 and R33) then attenuates the signal to approximately 0.6 volt at pin 6 of U2A.

The output voltage and current boost circuitry, formed by transistors 05-08, provide the required gain and current drive for the horizontal yoke. Q5 and Q6 provide some gain but serve primarily as dc level shifters to drive the complementary stage (Q7 and Q8). Since the use of higher voltage levels was required, I had to fall back on discrete logic. The higher peak-to-peak voltages are due to two factors. First, the horizontal deflection factor of the yoke is roughly six times higher than the vertical circuit. This is due to the smaller number of turns of wire to minimize the inductive reactance. Second, because of the increased inductive reactance, even with fewer wire turns, more voltage must be impressed across the windings to source the same current as that required by the vertical circuit.

Obviously, there are compromises and tradeoffs while trying to maximize current with a minimum of voltage. Notice that a negative supply voltage of –45 volts is used. This is because maximum current is needed during retrace where the rise time is the greatest. An oscilloscope on the collector of Q6 would reveal a sharp negative-going pulse.

Theoretically, an infinitely steep voltage pulse, discharged into a pure inductance, will produce a current ramp. This is how some commercial cameras generate a sweep. However, no circuit achieves theoretical factors so non-linearities creep in. That is why most cameras provide linearity adjustments.

Since the circuit presented here is a true closed-loop current source, the source voltage is automatically controlled to produce a linear current ramp. Thus, no linearity adjustments are needed and the only non-linearities which do exist are in the yoke itself. Because of the ramp generator design, compensation can be made to overcome this if required. For all practical purposes, however, yoke non-linearities are not detectable to the average viewer.

The zero crossover problem, discussed in the vertical circuit, is more difficult to overcome due to the higher frequency involved. Hence diode biasing is used to overlap the turn on of Q7 before Q8 turns off. R39 provides the proper degree of overlap and also balances the base drive for Q7 and Q8. Notice at this point that R7 is placed across the emitter-collector of 08. If a yoke is used that requires more than 300 mA p-p for proper operation, base drive limiting can take place due to the low current gain of Q8. Therefore, R7 helps to source current when the ramp is maximum negative, reducing the dissipation. Even with R7 in the circuit, Q8 is driven much harder and for a longer duration of the cycle than Q7.

I would like to point out here that a large imbalance of the centering pot for a long duration will cause excessive dissipation in Q7 or Q8, depending upon which side of zero the imbalance occurs. Steps must be taken to check this when initially powering the circuit to avoid the premature replacement of the transistors.

The value of R3 depends upon the yoke inductance and is used to dampen the inductance and lower the *Q* of the circuit to prevent ringing. Use a resistor decade box and decrease the resistance to a point where any vertical bar shading irregularities at the left side of the video monitor just disappear. If an oscilloscope is handy, connect it from terminal **C** to ground and view the horizontal ramp. Adjust R3 to eliminate any ringing on the leading edge of the ramp immediately after retrace. For the yoke specified, this value is 2.2 kilohms.

The current sense element for the horizontal ramp is R45 and operates in exactly the same manner as R21 in the vertical circuit. Changing the value of this resistor will change the range of the width potentiometer R43. With an R45 value of 10 ohms, a current range of 120 to 400 mA can be accommodated. It is important to realize that when it is desired to view the ramp waveform on a scope, only the *current* waveform is meaningful, because it is the current that deflects the electron beam within the Vidicon. (Remember that we are dealing with *magnetic* deflection Vidicons.)

Don't confuse the current waveforms with voltage waveforms shown on many commercial camera schematics. In many cameras, current waveforms are difficult to make because of the lack of a grounded current reference point. However, in this case R21 (vertical) and R45 (horizontal) are in the current path. A scope from terminal **A** or ***C*** to ground will display current if the voltage on the CRT is divided by the R21 or R45 resistance.

The Vidicon must be blanked during both vertical and horizontal retrace even though proper blanking of the video waveform is taken care of farther downstream; this is because of the persistance and lag of the Vidicon — an actively scanned retrace will produce dark diagonal lines superimposed on the normal scan. Methods to blank the Vidicon vary from cathode blanking to G1 grid blanking or both. In this case, simple cathode blanking is used by raising the cathode approximately 22 volts positive during the retrace intervals. The vertical and horizontal drive signals are inverted, wire ORed and again inverted to drive the Vidicon cathode through U4. I have noticed that when using a type 4478 Vidicon, because of its higher cathode cutoff point, marginal cutoff of the beam can occur, especially at low-light levels, producing a very slight retrace line. Using a type 7735 Vidicon has not been any problem.

The Vidicon *must* be protected from accidental loss of either vertical or horizontal sweep, even for time durations as short as a fraction of a second. If sweep failures occur without protection, the scanning beam in the Vidicon would permanently burn the target material, causing a serious line or spot in the picture after normal sweep resumed. U3, a 74123 dual retriggerable one-shot, is the heart of a scheme devised to blank the Vidicon (turn off the scanning beam) if either horizontal or vertical sweep is interrupted. In operation, C10 and R54 integrate the vertical ramp, negative-going retrace into a short duration pulse at the base of Q10. Q10 turns on during this time to produce a positive pulse at pin 10 of U3 to reset the timer. The output (pin 12) is low at this point. After the input pulse, U3 starts a time delay

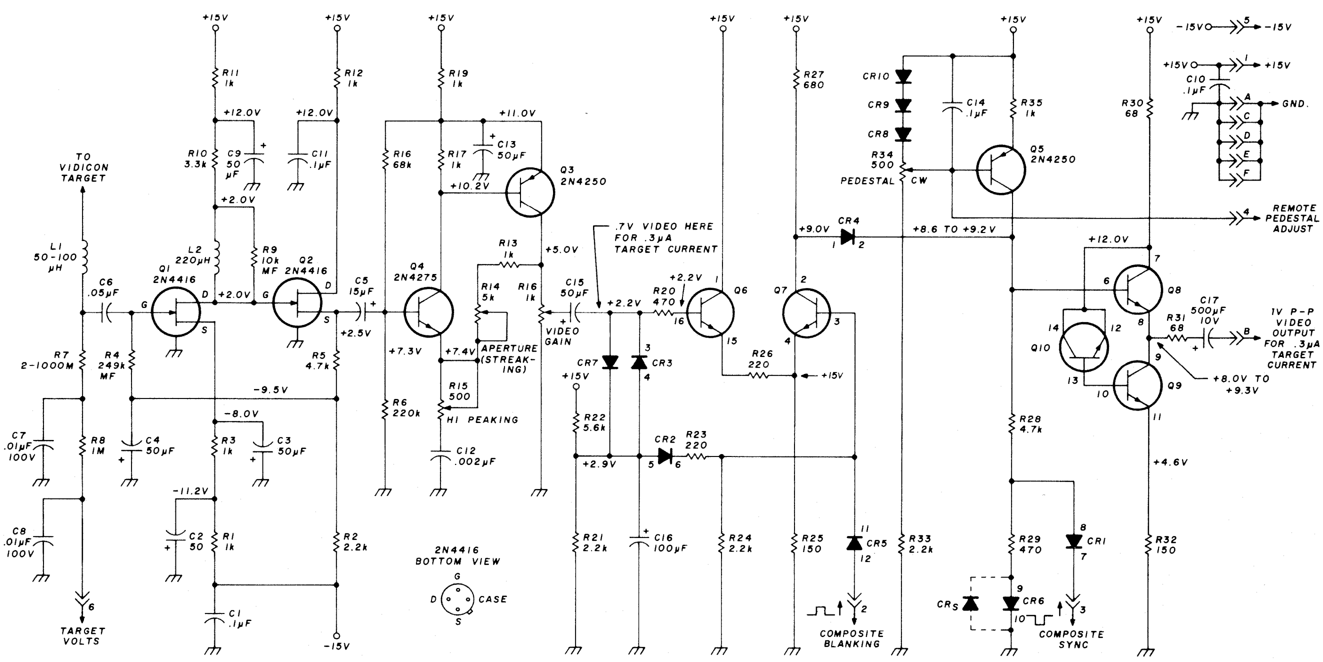
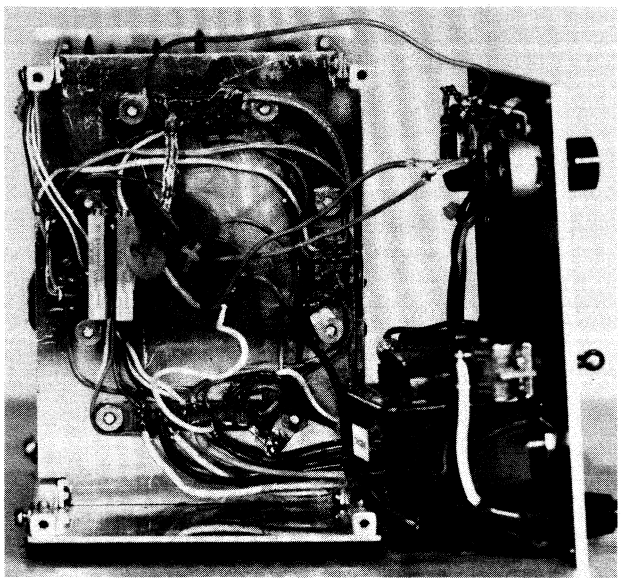


fig. 6. Schematic diagram of the video processor; all voltages are shown with no video (lens capped) and R16 turned fully counter-clockwise. R4, R7, and R9 are metal-film or deposited-carbon resistors. All other resistors are %-watt, 10 per cent carbon composition. Diodes CR7-CR10 are 1N4151 or equivalent silicon diodes; L1 is a Nytronics WEE-100 or J. W. Miller 70F755A1, and L2 is a Nytronics WEE-220 or J. W. Miller 70F224A1. U1 is an RCA CA3039 and U2 is a CA3083. The output voltage will vary depending upon the pedestal setting. R29's value will produce 40 IRE units of sync; to change to 20 units, R29 must be reduced to 220 ohms. All voltages were measured with a 10,000 ohms/volt meter.



Rear view of the camera with the back plate removed. The three capacitors which are connected to the Vidicon socket are grounded through a common piece of braid to the video processor socket.

which, if allowed to time out, will cause pin 12 to go high. For normal operation, before the timer times out, a new pulse arrives at pin 10 resetting the timer and repeating the cycle. If the sweep had failed to retrace, no pulse would occur at pin 10 and the timer would time out thus driving U4 pin 8 low and blanking the Vidicon.

The time delay is set by R55 and C11 and is slightly longer than a normal 16-millisecond vertical-scan cycle. The horizontal circuit operates in an identical manner. Note that I have brought the C11-R55 junction to pin D on the connector. Normally this pin is not used, but if the delay is shortened to time out before the end of the normal scan cycle the Vidicon will be blanked from that point until the start of the next cycle. By connecting a 500k potentiometer from pin D to +5 Vdc, a poor man's special effect can be made to "wipe" from bottom to top. An additional 500k potentiometer from pin E to +5 Vdc will produce a wipe from right to left. This combination will change the picture from full screen to a small square in the upper left of the screen. One caution must be observed. By blanking the Vidicon at a point normally in the active scan region for a long time on relatively bright scenes, a line will be produced on the screen when the blanking point is changed. This could produce a permanent burn in the Vidicon.

The Vidicon magnetic focus control circuit is also on this circuit board. It is a simple current sink to supply constant current to the focus coil independent of supply voltage and coil resistance variations. R60 is adjusted to set the current desired to focus the Vidicon, The voltage developed at the emitter of Q11 is compared to the voltage at the wiper of R60 by U5. U5 will then drive Q11 until these voltages are equal, producing a voltage drop across R59 which is proportional to the focus coil current. The focus coil current, $I, = E_{R60} wiper/R59$. The op-amp's open loop gain and excellent temperature stability are combined to produce very stable regulation of the focus current.

Most focus coils typically require approximately 40 mA if the coil resistance is about 400 ohms, producing E_{R60} equal to 0.4 volts. The lower the coil resistance, the higher the required current will be. If the focus coil used approaches 100 ohms, it will be necessary to place a small resistance in series with the coil (approximately 22 ohms) to prevent excessive dissipation in Q11. The voltage across terminals E and F should be close to 15 volts. Finally, capacitor C20 must be placed across terminals E and F to suppress horizontal spikes occurring at a rate too fast to be corrected by U5 (there was not room enough to mount this capacitor on the circuit board).

video processor

The video processor, as the name implies, processes the extremely low level video signal from the Vidicon. The preamplifier portion (**fig. 6**), composed of Q1 through 04, amplifies the signal and also provides response modification. The amplified signal (approximately 0.7 volt p-p) is then presented to the processor where blanking and sync are inserted.

The preamplifier accepts the extremely low level current, from the target of the Vidicon, which is produced by the discharge of the target by the scanning beam. This signal, in the range of 0.05 to 0.5 microamp, is ac coupled to Q1 and discharged through R4 to produce a voltage equal to:

$$I_t \left[\frac{R4 \times R7}{R4 + R7} \right]$$

at the gate of Q1. For 0.3 microamp of target current (a relatively bright scene), approximately 60 mV is developed at QI's gate; a higher voltage level than one would suspect. Wiring must be kept to a minimum and also well shielded at this point to prevent broadcast interference pickup due to the high impedance involved.

The target bias is also applied at this point. This positive voltage will charge the capacitive surface of the Vidicon that eventually will be discharged by the scanning beam. The bias, usually about +20 volts, must be varied as the average scene level changes and ideally must be a constant-current source for automatic light compensation. By making the value of R7 very large (greater than 100 megohms) a constant-current source is approximated because the Vidicon is basically a constant-current generator. Therefore, partial ALC is achieved. My original design uses a 2.2-megohm resistor for R7 with no ALC. However, if ALC is desired, use a 500- to 1000-megohm resistor for R7 and increase the value of the target pot from 500k to 1 megohm. If this is done, the recovery from abrupt light level changes will be slow, especially as R7 approaches 1000 megohrns because C6 must be charged (or discharged) to a stable level in the process.

In operation, the high impedance gate of 01 amplifies the video signal with some series peaking in the drain circuit. This technique helps to compensate for the normal high-frequency rolloff of the Vidicon. R9 shunts L2 to shape the response of the peaking by reducing the Q of L2. At low frequencies R10 is primarily the load resistor, while at high frequencies, where the reactance of L2 is higher, R2 and X_L of L2 are additive to increase the gain of the stage. Q2 is basically a source follower to isolate the signal from the previous stage and to present a lower impedance for coupling to Q4. In addition, it provides a dc stabilized feedback to Q1. 01 and Q2 therefore operate together to provide a single inversion of the signal with a total mid-frequency gain of approximately 15.

The video at Q2's source is ac coupled to Q3 and 04; Q4 is used as a bootstrapped emitter follower with a dc gain of 1. Degenerative feedback is provided to vary the high-frequency rolloff point (aperture) and the slope of the gain curve to the rolloff point (high peaking). The aperture is normally adjusted by viewing a scene with high contrast ratios and adjusting R14 for no white *tails* following a white to black transition. The high-peaking pot, R15, is adjusted by observing the maximum picture detail without noticeable oscillation effects. The corrected video is then fed to the video gain pot R16 for presentation to the sync and blanking processing stages.

The blanking and sync insertion circuitry2 is built around an RCA CA3039 high-speed diode array. The circuit restores the dc level to the incoming positive-going video signal, inserts blanking, sets up a pedestal level, and then adds composite sync. This type of processing precisely inserts the sync and blanking information, not merely adding it to the video. Linear addition is not good enough because the video information would ride on top of the sync information. Sync must only contain position information; therefore, the video is clamped to a voltage level at the positive terminal of C16 by diode CR3.

Diode CR7 protects CR3 from excessive reverse

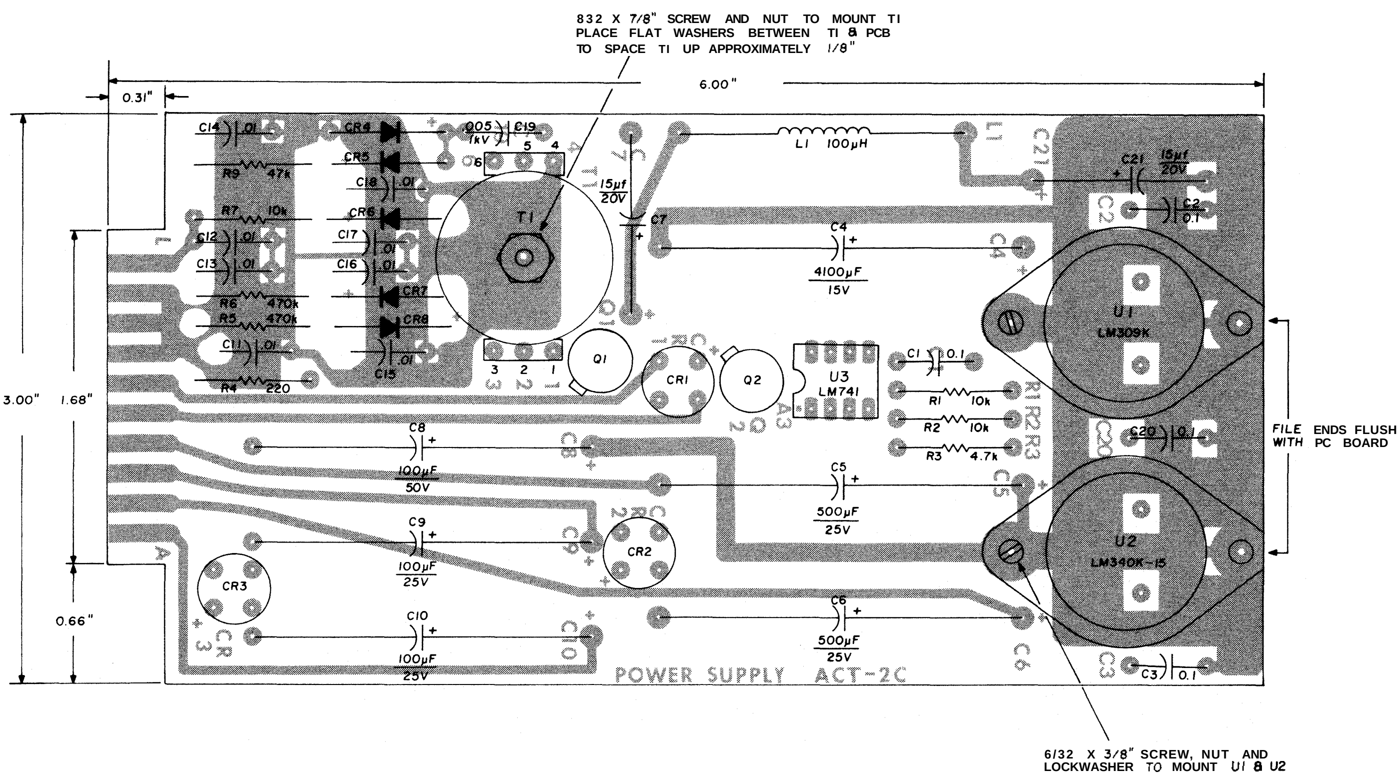
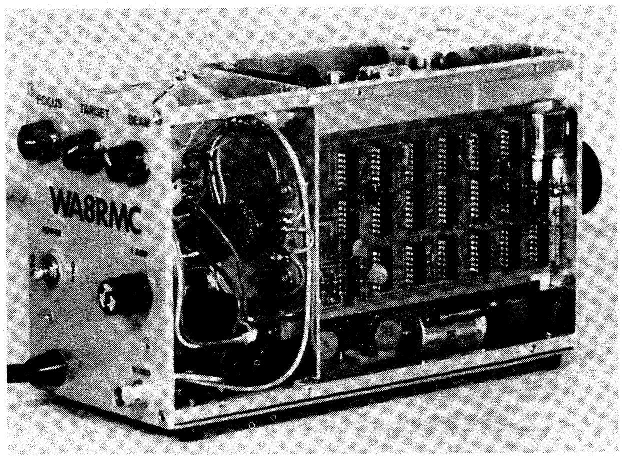


fig. 7. Component placement for the completed power supply board.



The 0.01 μF disc capacitor mentioned in the text can be seen attached to the sync generator board. Its inclusion will eliminate a vertical line in the picture.

voltage and will clamp the absolute video excursions to approximately 0.7 volt. The resulting signal is then fed into a differential amplifier consisting of Q6 and Q7, a matched pair within the CA3086 (U2). This amplifier has a low gain (2.5) between the base of Q6 and collector of Q7 but a much higher gain (6.5) between the base of Q7 and its collector. With CR5 reverse biased during scan time, amplified video will appear at the collector of 07. When a horizontal or vertical blanking pulse appears at pin 2, CR5 is forward biased and saturates Q7, thereby adding a very large pedestal to the video. This pedestal must then be clipped and sync added to complete the process.

Transistor Q5 and the associated circuitry is connected to function as a constant-current source which will set up a variable voltage at the cathode of CR4 depending upon the presence or absence of a sync pulse. Thus, whenever the signal at Q7's collector is more negative than this potential, CR4 will isolate the video signal from the base of Q8. However, if the level at Q7's collector is more positive, the video signal will appear at Q8. If the current from Q5 produces a level at Q8 base just beneath the video black level, the entire video will rest upon the blanking level. When CR1 conducts due to an incoming sync pulse, R29 will be short-circuited, and the voltage at the base of Q8 will drop a corresponding amount. The resultant composite video is then presented to the output transistors Q8 and Q9. Transistor Q10 is connected as a diode to provide negative feedback. Because of Q10's low reverse breakdown of 6 volts, it acts as a zener to clamp the feedback at 6 volts. R30 and R32 are purposely selected to be low values to keep impedances low and improve bandwidth.

The entire processor circuit, including the preamp, has been tested and found to have an overall 3 dB bandwidth in excess of 6.5 MHz. This value is more than adequate for all but the most expensive Vidicons.

There are a few comments about component selection which deserve mention. The preamp resistors R4, R7, and R9 should be metal film to keep noise to a minimum. C6 must be a non-polarized and low-leakage capacitor; most ceramic capacitors are fine. In the processor U1 is a diode array selected primarily for high-speed operation. Substitution of descrete slow-speed diodes could be disastrous. U2 is a high-current transistor array to handle the relatively high current in the output stage; it also runs warm in normal operation. In some applications, particularly for ATV operation, the sync amplitude may have to be altered. R29 is selected to produce approximately 40 IRE units of sync. To reduce this level to 20 IRE units, decrease this value to 220 ohms.

A number of prototypes were built before an arrangement was obtained that provided compact size, easily obtainable materials, a minimum of special tools for fabrication, and accessibility to all printed circuit boards while the camera was operational. Of course, there are always tradeoffs to any design, but the main criterion here lies in the ability for the maximum number of people to be able to reproduce this design with the minimum of effort.

A standard chassis approach was abandoned early because no standard size existed which was close to the dimensions needed to qualify the camera as a compact. Instead, I settled upon a basic frame approach. This may require more individual pieces but it keeps bending requirements to a minimum.,

Basically, the front and rear plates are made from 118-inch (3mm) aluminum. Thinner material could be used but mechanical rigidity will suffer. The frame studs that hold the front and rear plates are 1/4-inch (6.4mm) square aluminum bar stock. This material is not generally available in hardware stores and I had to go to an industrial aluminum supplier and buy a 12-foot (3.7m) long piece. I've got a lot left over, but the cost was less than three dollars. The back plane bracket, video PCB shield, sync generator PCB shield, and the bottom plate are all made from I/16-inch (1.6mm) aluminum. Here, substitution of thinner material down to about 0.040-inch (1mm) thick is fully satisfactory. Thicker material, however, will be difficult to bend and should be avoided.

*A copy of the complete set of mechanical drawings is available by sending a 9½ x 11 envelope with 35 cents postage to *ham* radio, Greenville, New Hampshire 03048.

Mounted directly behind and fastened to the front plate is the front mounting block. This serves as a very convenient means of securing all four circuit boards and establishes a good low-impedance electrical ground. It also spaces the yoke/focus coil back from the Vidicon to establish proper focusing. I made this block from 114-inch (6.4mm) aluminum, bur thicker pieces up to 3/8-inch i9.5 mm) can be used if care is taken to chamfer the edges where they contact the circuit boards so no shorts occur. Since the yoke/focus coil mounts to this block, check the mounting dimension requirements of the yoke you obtain before drilling these holes.

The outer cover that wraps around the entire camera is not detailed because of the variety of materials that could be used. Solid-sheet aluminum may be used if ventilation holes are drilled along the sides and top. Many types of decorative perforated aluminum are available at hardware stores and are desirable for ease of bending. I used 0.050-inch (1.3mm) perforated steel because of availability. Don't use this unless you have access to a bending brake and can do it right the *first* time!

Finally, the hole size in the front plate depends upon the method of mounting the lens but must be larger than 1-118-inch (2.9cm) in diameter in order to insert and remove the Vidicon without removing the plate. I have found that a universal lens mounting method to satisfy all situations is not obtainable. I have access to a lathe, so an adapter was machined with threads to accept a standard C-mount lens which would also screw into the front plate. Not everyone can do this so I suggest an adapter available from Denson Electronics (part 3833) to mount the lens on the front plate.

The printed-circuit board layouts may be obtained by sending a self-addressed, stamped-envelope to *ham radio,* Greenville, New Hampshire 03048 or, purchased as etched and drilled blank boards from Automation Engineering Company.* The parts placement drawings are shown in **fig.** 7, **8,** and **9.** Breadboard construction is not recommended because of the somewhat critical layout of components on some boards. In general, no special assembly of components is required on any of the boards. However, I might suggest three cautions that should not be overlooked.

1. Check for proper polarity of capacitors and diodes.

2. Check solder connections on the *component* side of the board.

3. Be careful not to leave component leads too long on the solder side; they may touch the focus coil.

Most components are commonly available and

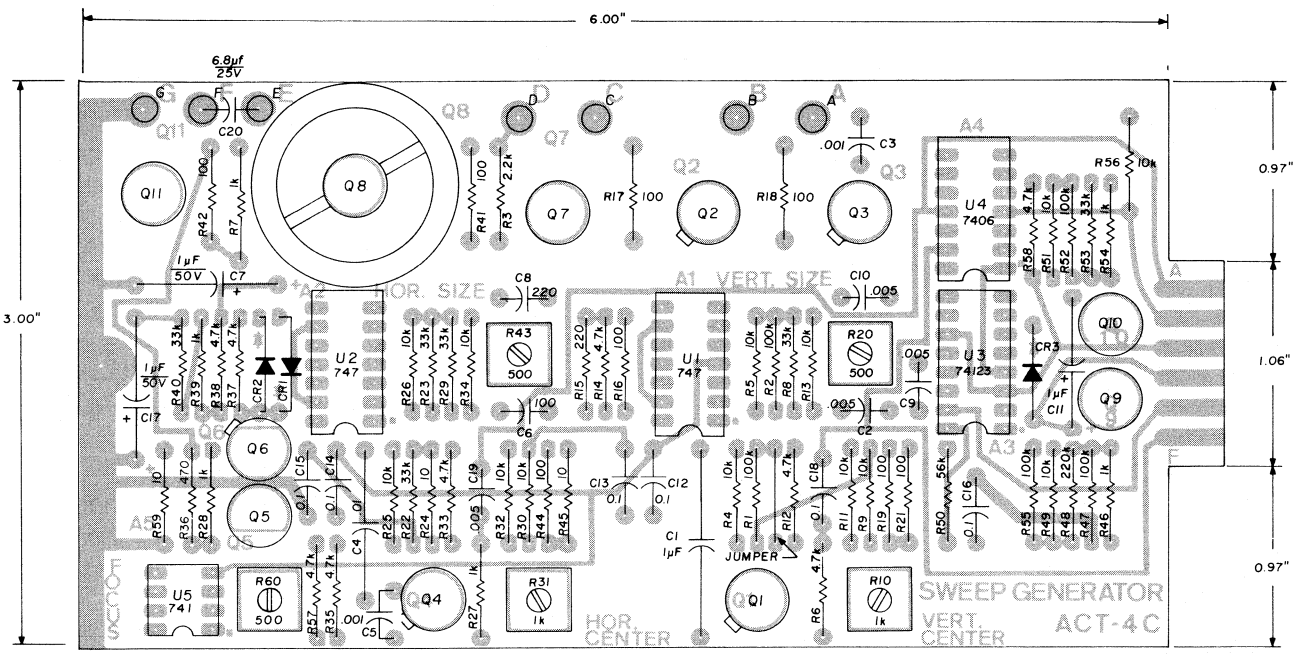


fig. 8. Component pattern diagram for the sweep generator. Plastic transistor spacers should be used under Q7 and Q11 to prevent the heatsinks from touching other components.

*Automation Engineering Company, 3621 Marine Drive, Toledo, Ohio 43609.

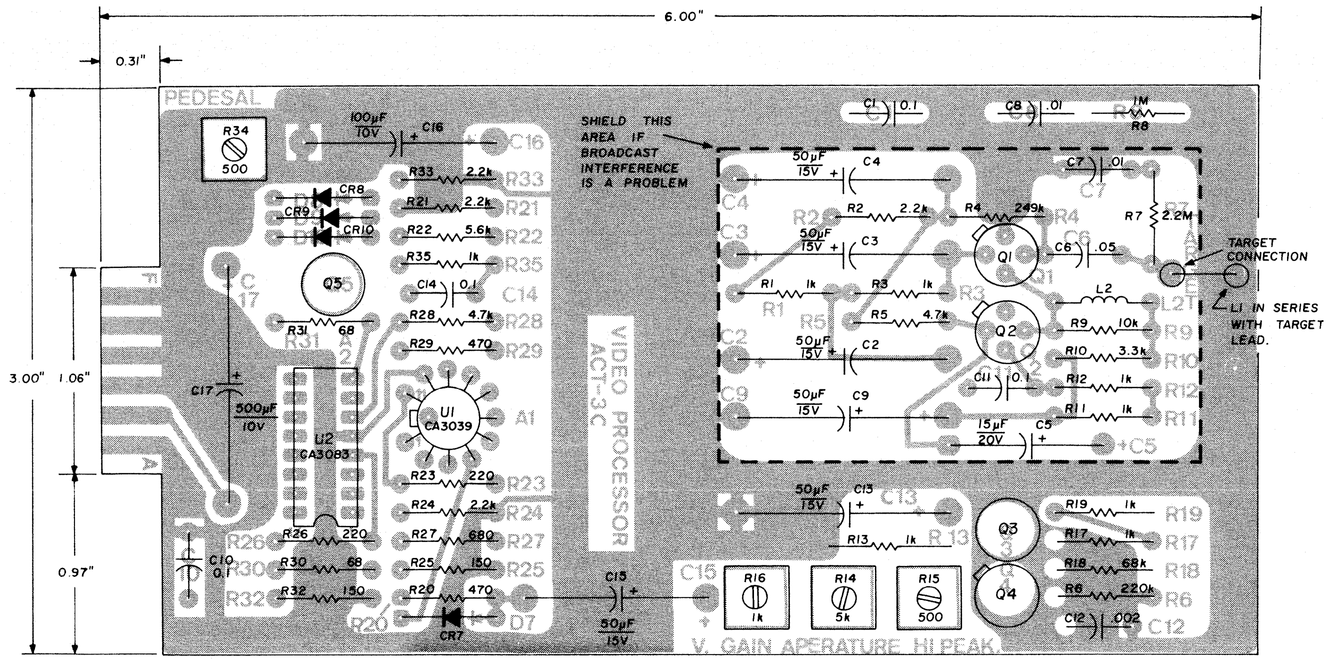


fig. 9. The area enclosed with a dotted line on the video processor board should be shielded if broadcast interference is a problem. L1 is not mounted on the board but runs directly between the board and the Vidicon. All other components are mounted as shown in this diagram.

were designed with surplus dealers in mind. Many substitutions of components are possible, but good judgement must be exercised. Hopefully, after reading the preceding technical descriptions of each board a knowledgeable substitution can be made knowing that it will work. If substitution of transistors is required for the 2N3053, 2N4036, and 2N3440, be sure to check package size, voltage rating, current gain, and power handling capacity.

The high-voltage transformer in the power supply must be wound with the Magnetics, Inc. cores detailed in **fig. 10.** The cores and bobbin are not generally available in small quantity, from Magnetics, Inc., but they can be obtained from Automation Engineering for $2.50, postpaid. Adding the required wire to these parts is quite simple. First, insert a 7/16-inch (11mm) bolt through the bobbin and secure with a nut. Now hold the bolt instead of the bobbin while winding. The primary must be wound on first, with the wire occupying the bobbin section closest to the pins. Wind with 20 turns of no. 26 AWG (0.4mm) enameled wire (approximately 3 feet or 1m) connecting the beginning of the winding to pin 1 and the finish to pin 3 (pin 2 is unused). Wind the first half of the secondary on the center section with 200 turns of no. 36 AWG wire (0.13mm). Connect the start to pin 4 and the finish to pin 5. Now, continue winding another 200 turns on the top section *in the same direction* as the first 200 turns, with the start at pin 5 and the finish at pin 6. After completion of the winding, slip spaghetti insulation over the leads and place the cores in position before soldering the leads to the pins. Careful positioning of the leads will prevent them from touching the cores. Finally, seal the wire in place on the bobbin by melting a small amount of beeswax or crayon on it.

Almost any 1-inch (2.5cm) Vidicon is usable in this camera, including the 8507 separate-mesh Vidicon. When this tube is used, however, tie G4 (pin 3) directly to power supply pin 10 through a 1-megohm resistor. G3 (pin 6) is then connected to the focus pot as shown. I recommend a yoke-focus coil assembly available from Denson Electronics (Part 2045) for this camera, but it must be altered and mounted as follows:

1. The wire on the focus coil must be unwound and rewound with 1 pound (0.45kg) of no. 28 AWG (0.3mm) or no. 29 AWG (0.27mm) wire (approximately 7000 turns).

2. A good target connection is not supplied with this assembly. However, if the front mount is removed by gently tapping it from inside and replaced with a Denson plastic front Vidicon mount and target connection (furnished with the coil as a loose item), this will make the assembly usable with a minimum of effort. The two plastic ears on the plastic front Vidicon mount, which extend behind the solid piece, must be sawed off flush so the yoke can be fully inserted.

3. A method of clamping the yoke and Vidicon to the focus coil case must be made by the builder.

4. The electrical alignment coils provided on this yoke must be powered to operate properly (many yokes have permanent magnet alignment magnets not needing electrical connections). No provision for alignment coil Dower is provided on the circuit boards, but the circuit shown in **fig. 11** can be used to power the coils.

Two other yokes that I've tested and found to be *satisfactory are* the Denson *2047* and 2013. They require approximately 300 mA p-p to operate the horizontal coils and 15 mA to operate the vertical coils. See fig. 12. If either yoke is used, you must furnish a satisfactory focus coil. The following tips will help if you feel confident enough to build a focus coil.

1. Wind the coil with 1 pound (0.45kg) of no. 28 (0.3mm) wire (approximately 7000 turns).

2. Use a plastic front Vidicon mount and target connection, available from Denson by description.

3. Use a permanent magnet alignment assembly, Denson 7138.

The power transformer is a special unit I wound for this application. Although standard transformers are available which could work, I found none with all of the required voltages. It is possible to hand wind a transformer using a Stancor P6466 as a core but this is somewhat laborious.

The 0.02 μF capacitors from Vidicon pins 2, 3, 5, and 6 to ground are 600 volt ceramic capacitors. They're installed on the Vidicon socket pins with as short leads as possible. The ground side of the capacitors should be connected together and then connected to the video processor socket pins C, D, E, and F at the ground lug with a short piece of shielding braid. This is for flexibility along with providing a low impedance to ground.

initial testing and setup

The testing of each circuit board should be done in a systematic manner to avoid the possibility of damaging good components on one circuit board due to a problem on another; the following steps will save a lot of time and headaches later.

First, after all frame wiring is completed (do not install the Vidicon or any circuit boards) plug the camera in and verity that, with respect to pin 2, approximately 20 Vac exists at the power supply socket pins A and C. Next, measure the voltage between pins E and F. It should be about 8 Vac. Last, check for about 7 Vac between Vidicon socket pins 1 and 8.

Now plug in the power supply circuit board and, with a clip lead, short pin 6 to ground. Turn the power on and check for ± 15 Vdc, $+5$ Vdc, ± 25 Vdc and -50 Vdc at the power supply socket and all other socket connections to which these voltages go. If an oscilloscope is available, make sure the ripple at pins 1 and 4 (± 15 volts) and pin 5 ($+5$ volts) is below 10 millivolts.

After all the tests are complete, remove the clip lead jumper, and plug in the sync generator. Again apply power and check for proper high voltages (approximately 500-600 volts at pin 10). I assume at this point that the sync generator has been previously checked and is operational. In normal opera-

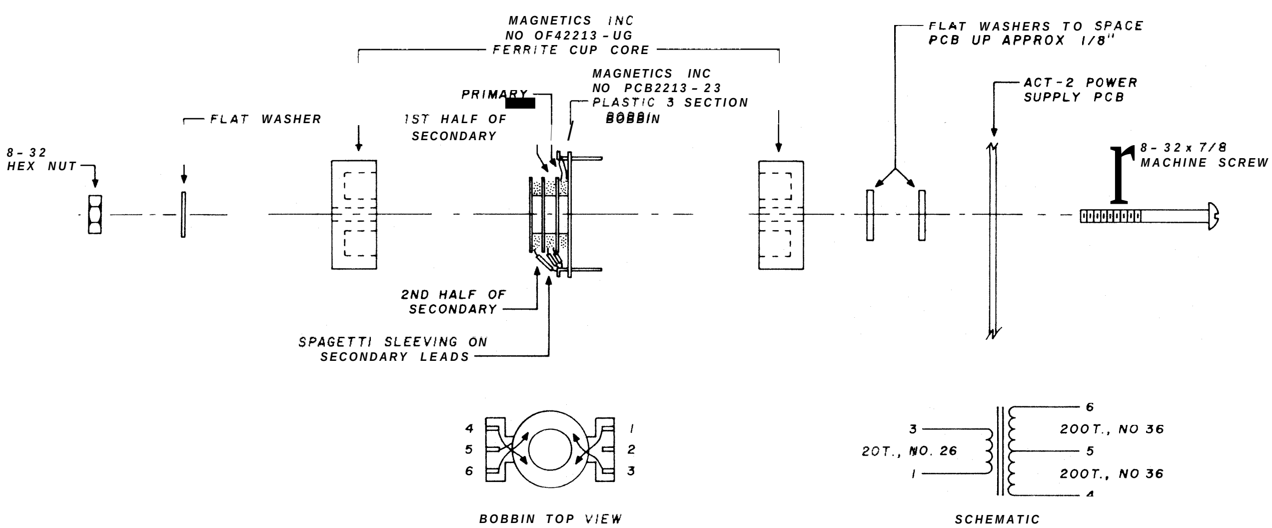


fig. 10. Construction details for the high-voltage transformer. The primary winding will require approximately 3 feet (1m) of no. 26 AWG (0.4mm) wire, while the secondary will require 50 feet (17m) of no. 36 AWG (0.13mm) wire. The core and bobbin are available from Automation Engineering for $2.50, postpaid.

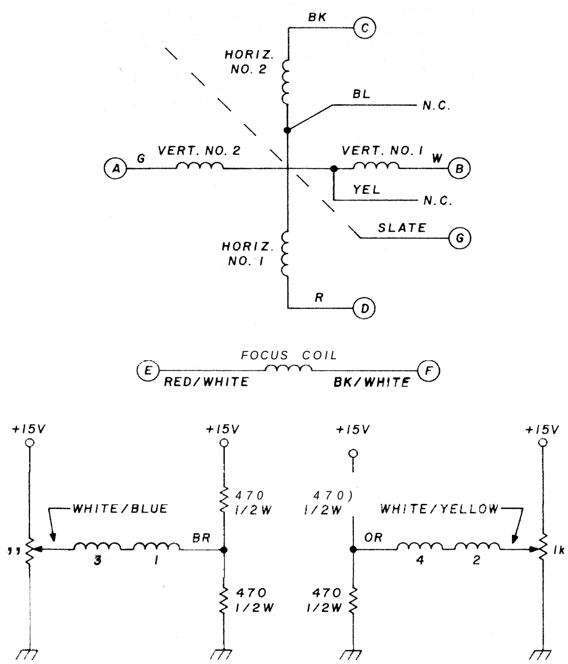


fig. 11. The upper diagrams show the lead identification colors for the Denson type 2045 yoke/focus coil assembly. After the focus coil has been rewound, the dc resistance should be approximately 250 ohms. R3, on the sweep generator board, can be reduced to about 1000 ohms when using this assembly to improve the shading. The circuits at the bottom should be used to power the electrical alignment coils. The components can be mounted on the circuit board connector socket in the frame of the camera. It is possible that in some extreme cases one of the 470-ohm resistors may have to be reduced to obtain the proper range.

tion, QI gets warm but never too hot to touch. If it does, either a secondary short exists or the transformer was improperly wound; too few turns on the primary will produce this problem. In a couple of cases I have found that the high voltage at pin 10 was low (about 400 volts) and Q1 ran very hot. I unwound the primary and, finding the proper number of turns, wound it back on with the same wire. When I checked, the voltage was up to 600 volts and Q1 was running cool. No explanations found — any suggestions? In any case, this could be a source of trouble. Also make sure the surfaces between the core halves are clean. My ears are sensitive to the 15 kHz whistle, so I found this early.

Next, plug in the sweep generator and connect all yoke/focus coil leads. Turn the power on and check for proper sawtooth waveforms at terminals A and C with an oscilloscope. Adjust R20 for approximately 1.4 volts p-p at terminal A (16 mA) and R43 for approximately 1.6 volts p-p at terminal C (162 mA) with the Denson 2045 assembly. Make sure the centering pots R10 and R32 are set to position the sawtooth waveform equally above and below zero. If either pot is adjusted to an extreme for an extended period of time, the output transistors will get very hot and may fail, so check this first! In normal operation Q8 will get very warm. A good check here is to place your finger directly on Q8. Count slowly to two. If you still have your finger on Q8, it is *not* too hot! This may be a crude test, but I like tests involving instruments that are handy (pun intended).

Now check the constant-current source supplying current to the focus coil. Set focus pot R60 to supply approximately 40 mA. This will produce a voltage drop of 0.4 volts across R59. Make sure that the focus coil's magnetic field polarity is correct by placing a compass at the outside of and at the image end of the focus coil. The north seeking pole must be *attracted* to the coil.

Finally, plug in the video processor and short the target lead to ground at C6 with a *very short* jumper. With a television monitor connected to the video output, vary R34. The monitor should have a blank raster that becomes lighter as the pot is rotated counter-clockwise. Remove the short at C6. By placing your hand near C6, herringbone patterns should occur, indicating that the processor is passing broadcast radio signals and is operating properly. If a vertical line is noticed near the center of the picture, it may be due either to improper grounding of the power supply or excessive pulse risetimes or the 15 kHz drive clock. If the latter is the case, a 0.01 μF capacitor *directly* across pins 7 and 8 of U19 on the sync generator will correct the problem. After all items have been checked, the camera is ready for the Vidicon and final testing.

final checks and calibration

Install the Vidicon in the camera. The short index pin should be oriented to position it at 9 o'clock when facing the camera. The Vidicon should be inserted into the yoke to a point where the tube face is approximately 112-inch (12.5mm) behind the lens. This dimension, however, is rough and final positioning will be necessary after the camera is operational. Apply power to the camera and perform the following steps:

1. Adjust the focus pot for +450 volts at the wiper.

2. When an image appears, adjust R60 on the sweep generator for proper magnetic focus. All electrical focusing should henceforth be done with the rear panel focus pot.

3. Rotate the yoke, if required, for proper picture orientation.

4. Adjust the centering pots. If either pot is rotated to its extreme, the round edge of the target will be seen. This can be used as a guide for proper centering along with proper positioning of the object being viewed (test pattern).

5. Adjust the horizontal and vertical size pots for correct image size. Note that *under* scanning the tube will produce a *larger* than normal picture. This is undesirable because resolution suffers, and if normal scanning is resumed a premanent burn line of the target will result.

6. With a picture in view, rotate R15 on the video processor clockwise until the picture "breaks up." Back off slightly and leave it there. In general, this pot increases the frequency response of the amplifier and a corresponding increase in resolution will occur until overcompensation is reached and oscillation takes place.

7. Next, rotate R14 (aperture) for no white tails following black-to-white transitions (pot fully counter-clockwise) or no black tails following black-to-white transitions (pot fully clockwise).

8. Finally, after all other items are satisfactory, the alignment magnets on the rear of the yoke should be adjusted to improve resolution and shading. These are not centering magnets and centering will possibly have to be touched up later. Rotate the focus pot on the rear panel. The image must not shift from the center position. Adjust these magnets until rotating the focus pot to each extreme produces a picture that rotates about an imaginary center axis as it goes through focus. No side-to-side or top-to-bottom shifts must take place.

conclusion

This television camera, although complex in some respects, is relatively easy to build and troubleshoot. A number of construction approaches were tried before settling on this one. Because many amateurs do not own a machine shop, as much of the design as possible takes advantage of standard workshop tools. This is not 100 per cent applicable, but I'm sure that alternate approaches, for the same result, with available tools will be devised. For this reason, I've gone into further detail than normally would be expected. The construction of this camera is not really as complex as it seems on the surface. Give it a try!

A number of extensions to this design are possible as discussed in the preceding text, and in most cases are limited only by your imagination. With a thorough understanding of this design, it is possible to produce a camera applicable to your situation. Because of the interlace quality, multi-camera con-

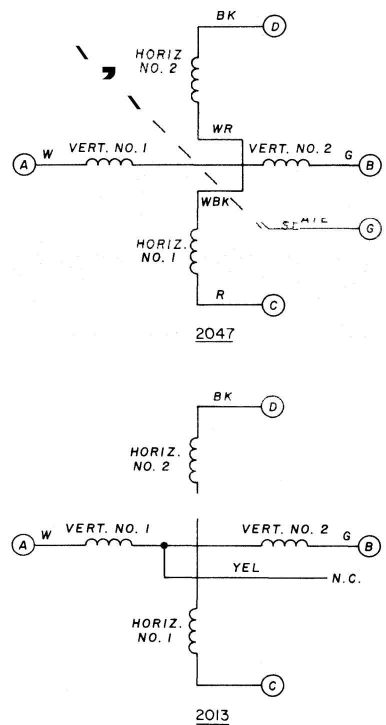


fig. 12. Connections for two additional yoke assemblies that have been tested in the camera. It is easier to mount a permanent magnet assembly on the 2047; it also has slightly better resolution capabilities than the 2013.

trol is possible (color anyone?). Portable operation, as mentioned earlier, is also possible because of the elimination of line-lock operation (how about mobile ATV?).

Before rushing to the workshop, or your local electronics store, re-read the article, understand the contents, and visualize your construction method. After all, you may already have most of the parts without realizing it. I will be happy to correspond with any individual about the existing design, construction difficulty, or improvements if a self-addressed, stamped envelope is enclosed with the correspondence. Good luck and happy construction!

references

1. William Neeland, "Function Generator has Variable Polarity Exponents," Electronic Engineer, January. 1971, page **44.**

2. Jeremy M. C. Tucker, "A Signal Processing Circuit for a Black and White TV Camera," Fairchild Semiconductor *Application* Brief 103, Fairchild Semiconductor Corporation, Mountain view, California 94040, May, 1969.

bibliography

1. Eastman, Gerald A., Measurement Techniques, Tektronix, Inc., Beaverton, Oregon 97077, 1969.

2. Smith, John A., Modern Operational Circuit *Design,* John Wiley and Sons, New York. 1971.

3. Reference Data for Radio Engineers, 5th Edition, Howard Sams and Company, Indianapolis, 1968.

4. Television Signal Analysis, 2nd edition, American Telephone and Telegraph Company. April, 1963.

ham radio

microstrip transmission line

A discussion of the operating parameters of microstrip transmission line and how to calculate its characteristic impedance and propagation velocity

Printed electronic circuits were originally developed for military equipment manufactured during World War II; in the years following the war printed-circuit manufacturing techniques were developed which both improved reliability and reduced cost. While much of this development work was accomplished under the watchful eye of the military services, the television set manufacturers also played an important part — they were exploring every avenue that offered the possibility of lowering the production costs of their sets.

In 1949 Robert Barrett of the Air Force Cambridge Research Center proposed that printed-circuit techniques be adapted to uhf and microwave circuits by using flat coaxial configurations with an air or solid dielectric.[1] The only known use of this technique at that time was in an antenna power divider.

In the early 1950s several manufacturers began developing Barrett's original suggestion. Airborne Instrument Laboratories (AIL) developed a system using air dielectric which they called *stripline*;[2] ITT introduced a single ground plane, solid-dielectric strip transmission line called *microstrip*;[3] and Sanders Associates began investigating a dual ground plane, solid-dielectric arrangement that subsequently became known as *Tri-Plate* transmission line.[4] Since all three forms of strip transmission line supported the TEM mode of propagation, they had similar electrical characteristics. The major differences were in size, shielding, insertion loss, and ease of construction.

Many companies and research groups contributed to the advancement of strip transmission line techniques, and by 1955 a wide range of uhf and microwave components were in limited production. Multifunction assemblies were developed by several manufacturers which were made available commer-

By James R. **Fisk, W1HR,** Communications Technology, Greenville, New Hampshire 03048

cially; strip transmission line couplers were developed,[5] and multiple section lines were developed to increase the bandwidth of individual components such as directional couplers, hybrid rings, and filters.[6]

In addition to strip transmission line layout techniques, improvements were also made in dielectric materials. Early efforts produced only a few suitable dielectrics such as fiberglass, Rexolite, and Teflon. By the 1960s irradiated polyolefin, Teflon-fiberglass, beryllia, and a variety of other laminates provided a wide range of dielectric constants and operating temperatures.[7] Dielectric substrates such as quartz, alumina, sapphire, and magnesium titanate were developed later and have found wide use in microwave integrated circuits.

The Tri-Plate strip transmission line developed by Sanders Associates has been used extensively in directional couplers and other uhf and microwave circuits where shielding is required, while ITT's air-dielectric stripline has been used primarily in high-power circuits such as amateur vhf/uhf power amplifiers and vhf fm broadcast transmitters. Microstrip, on the other hand, is used widely from vhf through microwave — it has been employed in such diverse applications as frequency counters, broadband solid-state amateur vhf power amplifiers, uhf rat-race mixers, matching networks, antenna baluns, and phased antenna arrays for radar and satellite communications.

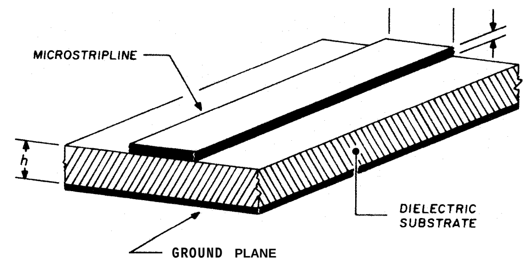


fig. 1. Microstrip transmission line consists of a thin conducting strip placed on one side of a dielectric substrate which has a ground plane on the opposite side. The characteristic impedance of microstrip is a function of the ratio of the strip width to dielectric thickness, w/h; t is the thickness of the conducting strip.

microstrip characteristics

A microstrip transmission line consists of a thin conducting strip placed on one side of a dielectric substrate which has a solid, conducting ground plane on the opposite side as shown in fig. 1. The substrate is usually a low-loss dielectric, but ferromagnetic and semiconductor materials have been used in some specialized applications.

The propagation characteristics of a strip transmission line are very similar to those of a coaxial transmission line, from which it evolved. The electric and magnetic field configuration of microstrip in fig. 2 is the final stage in a progressive modification of the conventional coaxial line. The solid lines are used to indicate the electric field; the dashed lines, the magnetic field. Both are entirely in the transverse plane (at right angles to each other and at 90° to the direction of propagation), so this is called the transverse electro-magnetic or TEM mode.

In the Tri-Plate line (fig. 2D) the electric field is bounded entirely by the flat outer conductors and there is essentially no electric field component to the sides of the center strip. If the ratio of the outer conductor width is more than about three times the strip width, *w*, sidewalls are not required.

The two properties of microstrip of most importance to rf circuit designers are velocity of propagation (phase velocity) and characteristic impedance. Whereas the propagation of rf energy in coaxial lines and Tri-Plate line is purely in the TEM mode, in microstrip the field lines are not entirely contained in the substrate (fig. 2E). For this reason, the propagation mode in microstrip is called quasi-

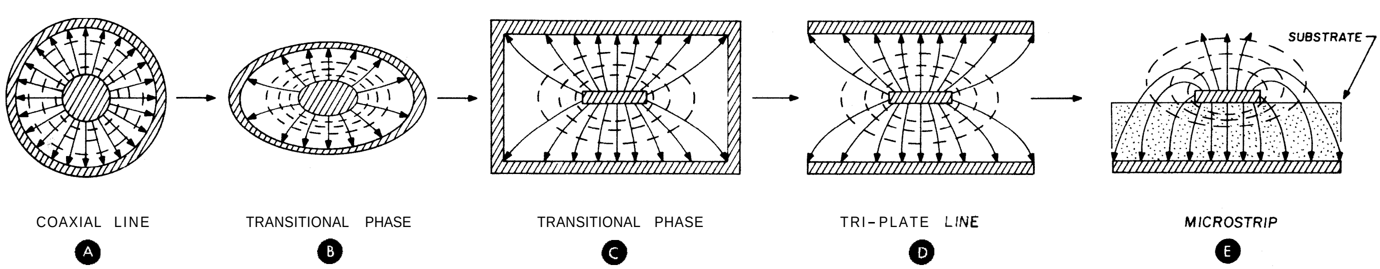


fig. 2. The propagation characteristics of microstrip (E) are very similar to coaxial line (A), from which it evolved. The solid lines indicate the electric field; dashed lines the magnetic field — both are in the transverse plane so this is the Transverse Electro-Magnetic or TEM mode. Since the microstrip field lines are not entirely within the substrate, propagation is not purely TEM but quasi-TEM (see text).

table 1. Characteristic impedance Z_o and propagation properties of microstrip etched on fiberglass-epoxy circuit board ($\epsilon_r = 4.8$) double clad with 1 ounce copper. The ratio w/h is microstrip width to dielectric height.

		microstrip width (1/32″ or 0.8 mm board)		microstrip width (1/16″ or 1.6 mm board)		
Z_o	w/h	mils	mm	mils	mm	v_p
10	14.93	422	10.7	887	22.5	0.481
15	9.35	264	6.7	556	14.1	0.490
20	6.59	186	4.7	392	10.0	0.498
25	4.96	139	3.5	295	7.5	0.505
30	3.89	109	2.8	230	5.8	0.510
35	3.13	87	2.2	185	4.7	0.516
40	2.56	71	1.8	152	3.9	0.520
45	2.13	59	1.5	126	3.2	0.524
50	1.79	49	1.2	105	2.7	0.528
55	1.52	41	1.04	89	2.3	0.532
60	1.30	35	0.88	75	1.9	0.535
65	1.11	30	0.76	64	1.6	0.538
70	0.955	25	0.64	54	1.4	0.541
75	0.823	21	0.53	46	1.2	0.544
80	0.711	18	0.46	40	1.02	0.546
85	0.614	15	0.38	34	0.86	0.548
90	0.532	13	0.33	29	0.74	0.550
95	0.460	11	0.28	25	0.61	0.552
100	0.399	9	0.23	21	0.54	0.553
105	0.346	7.6	0.20	18	0.46	0.555
110	0.299	6.4	0.16	15	0.39	0.556
115	0.260	5.2	0.13	13	0.33	0.557
120	0.225	4.3	0.11	11	0.28	0.559
125	0.195	3.4	0.09	9	0.23	0.560
130	0.169	2.8	0.069	7.6	0.20	0.561
135	0.147	—	—	6.5	0.17	0.562
140	0.127	—	—	5.4	0.14	0.563
145	0.111	—	—	4.5	0.11	0.564
150	0.096	—	—	3.7	0.09	0.565
155	0.0833	—	—	3.1	0.079	0.566
160	0.0723	—	—	2.5	0.064	0.567

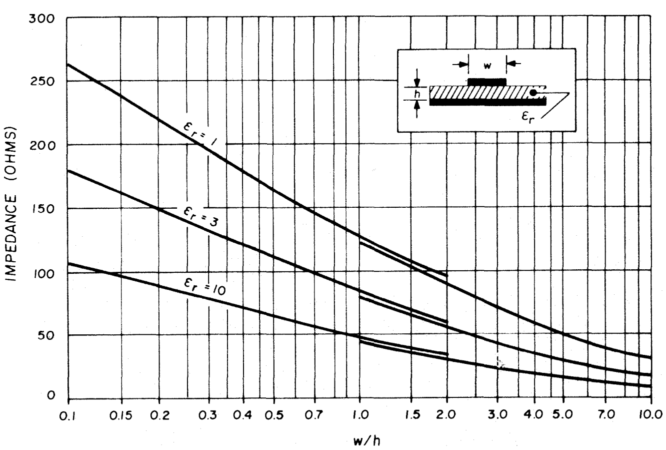


fig. 3. Characteristic impedance, Z_o, of microstrip vs the ratio w/h, as derived by Wheeler. Since the curves for wide strip don't intersect the curves for narrow strip, values which fall in the undefined region must be interpolated (w/h in the range between 1.0 and 2.0); many computer programs have been written to accomplish this task automatically, but Wheeler's rnicrostrip equations are difficult for the amateur to use. The impedance curves plotted from the more recent microstrip design equations of Schneider, Sobol, Hammerstad, and others don't have any undefined areas.

TEM. Assuming the quasi-TEM mode, the velocity of propagation in microstrip is given by

$$v_p = \frac{c}{\sqrt{\epsilon_{eff}}} \qquad (1)$$

where v_p is the velocity of propagation, c is the speed of light, and ϵ_{eff} is the effective dielectric constant of the substrate. The effective dielectric constant is always lower than the relative dielectric constant of the substrate, ϵ_r, because some of the field lines are outside the substrate.

The characteristic impedance of microstrip line, Z_o is given by the familiar transmission line equations

$$Z_o = \frac{1}{v_p C} = v_p L \qquad (2)$$

where v_p is the propagation velocity (**eq. 1**), C is the capacitance per unit length of line, and L is the inductance per unit length of line. Unfortunately, the calculation is not as simple as it may first appear because capacitance and inductance are both functions of microstrip geometry; and capacitance and table 2. Characteristic impedance Z_o and propagation properties of microstrip etched on Teflon-fibelglass circuit board ($\epsilon_r = 2.55$) double clad with 1 ounce copper. The ratio w/h is microstrip width to dielectric height (see text).

		microstrip width (1/32″ or 0.8 rnm board)		microstrip width (1/16″ or 1.6 rnrn board)		
Z_o	w/h	mils	mm	mils	mm	v_p
10	20.96	593	15.1	1246	31.6	0.646
15	13.27	375	9.5	789	20.0	0.654
20	9.47	267	6.8	563	14.3	0.661
25	7.21	203	5.2	429	10.9	0.667
30	5.72	161	4.1	340	8.6	0.672
35	4.66	131	3.3	277	7.0	0.676
40	3.88	109	2.8	231	5.9	0.681
45	3.28	92	2.3	195	5.0	0.685
50	2.80	78	2.0	166	4.2	0.688
55	2.42	67	1.7	143	3.6	0.691
60	2.10	58	1.5	124	3.1	0.694
65	1.84	51	1.3	107	2.7	0.697
70	1.61	44	1.1	95	2.4	0.700
75	1.42	39	1.0	83	2.1	0.702
80	1.26	34	0.86	73	1.8	0.705
85	1.12	30	0.79	64	1.6	0.707
90	0.991	26	0.66	57	1.4	0.709
95	0.882	23	0.60	51	1.3	0.711
100	0.785	20	0.51	45	1.1	0.713
105	0.700	18	0.45	39	1.00	0.714
110	0.625	16	0.40	35	0.89	0.716
115	0,558	14	0.35	31	0.78	0.717
120	0.498	12	0.31	27	0.69	0.718
125	0.445	11	0.27	24	0.61	0.720
130	0.398	9.2	0.23	21	0.54	0.721
135	0,356	8.0	0.20	19	0.48	0.722
140	0.318	7.0	0.18	17	0.42	0.723
145	0.285	6.0	0.15	15	0.37	0.724
150	0.254	5.1	0.13	13	0.32	0.725
155	0.228	4.4	0.11	11	0.28	0.726
160	0.204	3.7	0.094	10	0.25	0.727
165	0.182	3.1	0.078	8.5	0.21	0.727
170	0.163	2.6	0.066	7.3	0.19	0.728
175	0.146	—	—	6.5	0.17	0.729
180	0.131	—	—	5.6	0.14	0.730
185	0.117	—	—	4.8	0.12	0.730
190	0.105	—	—	4.2	0.11	0.731
195	0.094	—	—	3.6	0.09	0.732
200	0.937	—	—	3.1	0.08	0.732

propagation velocity are functions of the effective dielectric constant.

Early efforts to derive formulas for the characteristic impedance of microstrip were based on the quasi-TEM model, but there were serious difficulties. As pointed out in 1964 by Harold Wheeler, one of the first to derive practical microstrip design equations, "Because this was a problem in two-dimensional electric and magnetic fields, it was natural to apply the principles of . . . conformal mapping. The resulting formulas were usually so complicated that any practical utility resulted from simplified approximations for limited ranges of variables."[8] Later, Wheeler developed a set of approximate equations for microstrip and published design charts which were widely used by microwave designers.[9]

Although Wheeler published both analysis and synthesis equations,* the synthesis equations were apparently largely overlooked because most of the published articles which referred to Wheeler's work presented only the analysis equations. This meant that designers had to use lengthy, interactive trial-and-error solutions to determine the correct microstrip geometry for a required value of characteristic impedance.

One of the disadvantages of Wheeler's equations is that two different equations are required — one for

*Analysis equations give Z_o in terms of the ratio of strip width to substrate height, w/h; synthesis equations give w/h directly as a function of Z_o. Use of the analysis equation to find w/h requires ten or more iterations, a process that might require a half hour or more with a slide rule; it takes about a minute with the programmable HP-25 calculator — the synthesis equation provides an answer in less than 10 seconds.

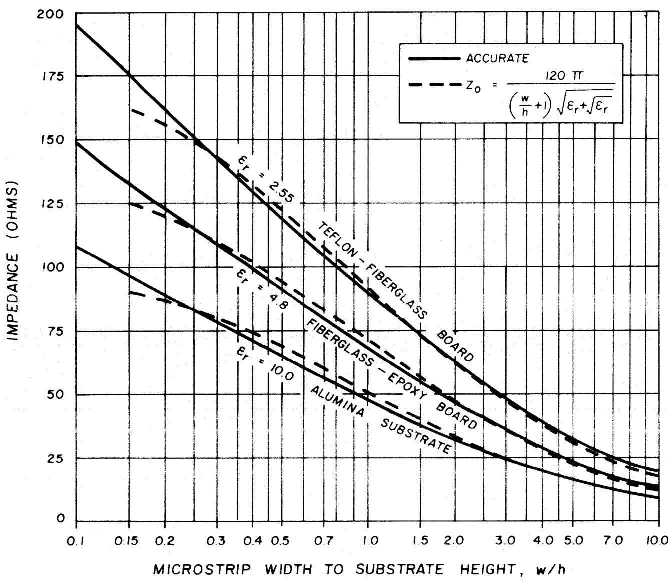


fig. 4. Plot of simplified rnicrostrip equation derived by Fisk (eqs. 3 and 5), as compared to the more accurate expression of Hammerstad. For $w/h > 0.2$ and $\epsilon_r > 4.0$, the Fisk formula is accurate to within a few per cent.

wide strip ($w/h > 1$), and another for narrow strip ($w/h < 2$). Furthermore, the boundary between the two cases is not clearly defined, as shown in fig. 3, because the two curves don't intersect. Therefore, if the required w/h ratio falls in this undefined region, it's necessary to interpolate the correct value.

A few years after Wheeler's formulas were published, Schneider developed more explicit equations for the free-space characteristic impedance of microstrip and effective dielectric constant.[10] Like the Wheeler equations, two formulas are required; one for narrow microstrip and another for wide. Unlike Wheeler's work, however, curves plotted from the Schneider equations intersect, so there is no "indefinite region." Unfortunately, Schneider published only analysis equations, so synthesis of w/h for a desired Z_o required the lengthy iteration process.

In 1967 Dr. Harold Sobol fitted curves to Wheeler's analysis and published an expression for rnicrostrip impedance which covered both narrow and wide lines.[11] This equation has been widely publicized in the literature, notably in application notes published by Motorola Semiconductor, and is the basis for many microstrip design charts. The Sobol formula is for analysis (Z_o from w/h), but many microwave designers have access to high-speed computers, and when synthesis is required, a computer can go through the necessary iterations very quickly; the single closed-form expression for both wide and narrow microstrip simplifies programming."

Although a computer can go through the iterations quickly, programmable hand-held calculators cannot. N6TX has written an HP-25 program which provides acceptable accuracy for most amateur work, but his program begins at $w/h = 1$ and iterates out to the required value. Therefore, for high and low values of Z_o the required calculation may require a minute or more of successive iterations. To reduce the calculation time, I set about to develop a simple equation for the approximate value of w/h which could then be refined with Sobol's formula. My first try greatly reduced the calculation time for values of Z_o less than about 75 ohms ($\epsilon_r \approx 5$), but offered no improvement for higher impedance microstrip.[12] Further refinement of the approximate formula resulted in the following simple expression, which gives quite good accuracy for microstrip substrates used in most amateur work ($\epsilon_r > 3.0$)

$$w/h \approx \frac{120\pi}{Z_o\sqrt{\epsilon_r + \sqrt{\epsilon_r}}} - 1 \tag{3}$$

For fiberglass-epoxy material ($\epsilon_r = 4.8$) this may be simplified to

$$w/h \approx \frac{142.6}{Z_o} - 1 \tag{4}$$

As can be seen in fig. 4, for $w/h > 0.2$ this simplified expression for microstrip impedance is within a few per cent of the impedance calculated with more accurate equations. This covers the microstrip impedance range most commonly used in radio communications work. With fiberglass-epoxy board ($\epsilon_r = 4.8$), **eq.** 3 is within about 1 ohm of the exact expression for all values of Z_o below 60 ohms. Accuracy falls off for $\epsilon_r < 3.0$, but is still acceptable for many applications.

When it's necessary to calculate Z_o from a given value of w/h, **eq.** 3 can be rearranged to

$$Z_o \approx \frac{120\pi}{(w/h + 1)\sqrt{\epsilon_r + \sqrt{\epsilon_r}}} \tag{5}$$

$$Z_o \approx \frac{142.6}{w/h + 1} \qquad (\text{for } \epsilon_r = 4.8) \tag{6}$$

The accuracy of this simplified equation is the same as that of **eq.** 3 (i.e., within a few per cent for $w/h > 0.2$ and $\epsilon_r > 3.0$).

accurate impedance caiculations

In 1975 E. O. Hammerstad reported to the European Microwave Conference that he had developed analysis *and synthesis* equations for microstrip which are more accurate than earlier work, and fall within 1 per cent of Wheeler's numerical results.[13] His formulas, which are based on the work of Wheeler and

'Other forms of microstrip design equations have been proposed by H. L. Clemm (*Frequenz* [Germany], July, 1968) and A. H. Kwon (*Microwave Journal,* January, 1976). Clemm's analysis equation is less accurate than Sobol's for high ϵ_r, and Kwon's equations, while purported to be *new,* are actually identical to those published by Wheeler in 1965.

Schneider, are considered to be the best available at the present time, and are shown here. First, the *analysis* equations:

For $w/h < 1$

$$Z_o = \frac{60}{\sqrt{\epsilon_{eff}}} \ln\,(8\,h/w + w/4h) \tag{7}$$

Where:

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[\left(\frac{1}{\sqrt{1 + 12h/w}} \right) + 0.04\,(1 - w/h)^2 \right] \tag{8}$$

For $w/h > 1$

$$Z_o = \frac{120\pi \,/\, \sqrt{\epsilon_{eff}}}{w/h + 1.393 + 2/3\,\ln\,(w/h + 1.444)} \tag{9}$$

Where:

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(\frac{1}{\sqrt{1 + 12h/w}} \right) \tag{10}$$

Hammerstad's microstrip *synthesis* equations for w/h in terms of Z_o and ϵ_r are given below:

For $w/h < 2$

$$w/h = \frac{8e^A}{e^{2A} - 2} \tag{11}$$

For $w/h > 2$

$$\frac{w}{h} = \frac{2}{\pi} \left\{ B - 1 - \ln\,(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left[\ln\,(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right] \right\} \tag{12}$$

Where:

$$A = \frac{Z_o}{60} \sqrt{\frac{\epsilon_r + 1}{2}} + \frac{\epsilon_r - 1}{\epsilon_r + 1} \left(0.23 + \frac{0.11}{\epsilon_r}\right)$$

$$B = \frac{377\pi}{2\,Z_o \sqrt{\epsilon_r}}$$

Hammerstad notes that for $\epsilon_r < 16$, the maximum relative error is less than 0.5 per cent for $w/h > 0.5$; for $w/h < 20$ the stated error is less than 0.8 per cent.

Although these formulas may look formidable, they can be solved easily with a hand-held scientific calculator. These equations can also be quickly solved with programmable calculators – solutions with my HP-25 require less than 10 seconds (plus programming time, of course).* If you don't have a calculator, the graph of **fig. 5** shows how the ratio of

*Copies of the HP-25 programs will be sent to interested readers upon receipt of a self-addressed, stamped envelope. Send requests to *ham radio*. Greenville, New Hampshire 03048.

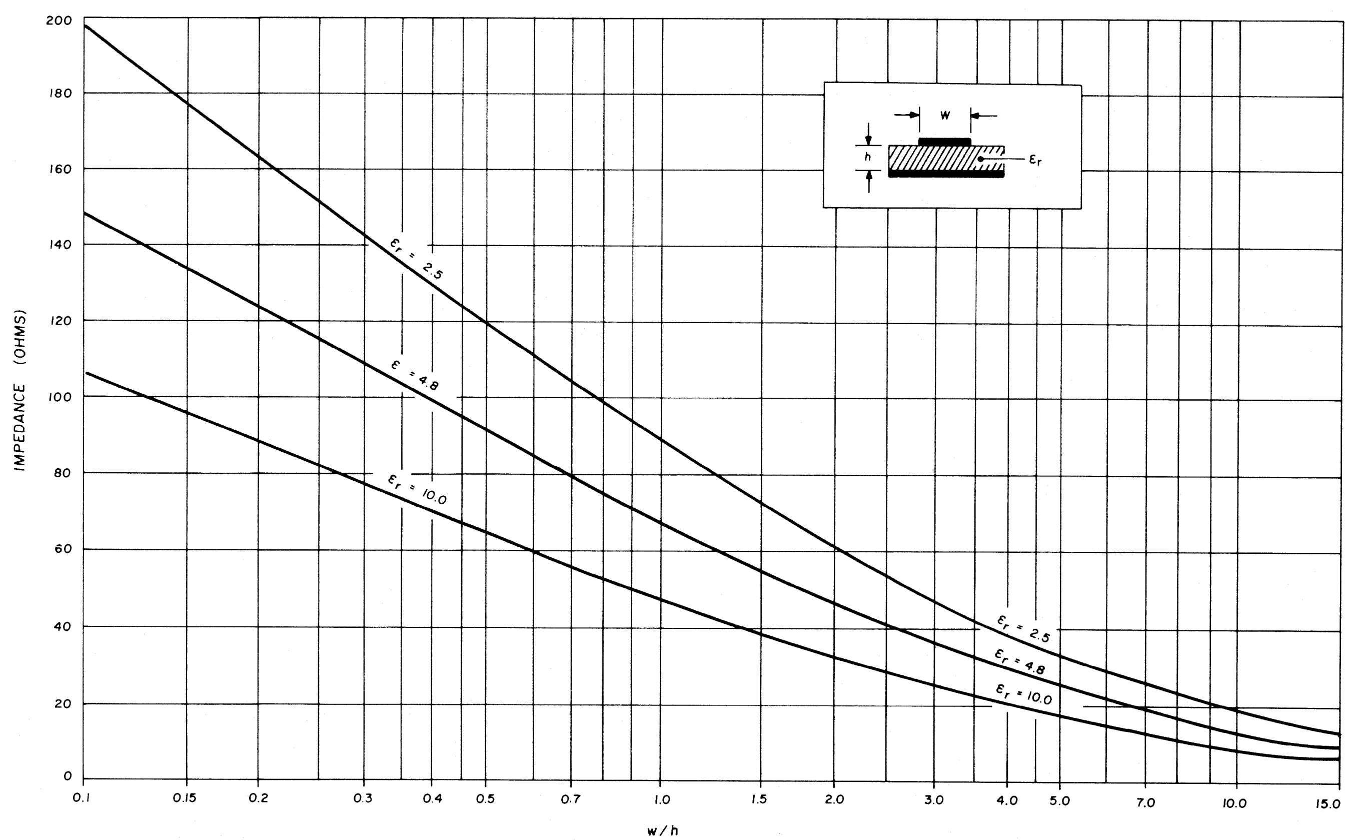


fig. 5. Characteristic impedance of microstrip as a function of w/h, plotted from the design equations derived by Hammerstad for glass-epoxy ($\epsilon_r = 4.8$), Teflon-epoxy ($\epsilon_r = 2.55$). and alumina ($\epsilon_r = 10.0$).

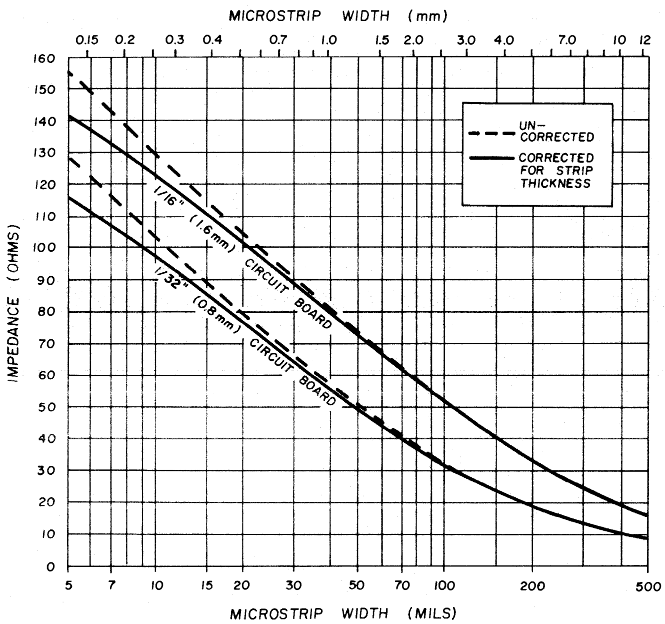


fig. 6. Effect of finite conductor thickness on microstrip impedance for 1/16" (1.6mm) and 1/32" (0.8mm) fiberglass-epoxy circuit board. In most applications the correction factor (eqs. 13 and 14) can be ignored for low impedance microstrip; at impedances above about 70 ohms the correction factor should be included in the impedance calculation.

strip width to dielectric thickness, w/h, affects the characteristic impedance of the line for fiberglass-epoxy circuit board ($\epsilon_r =$ 4.8) and Teflon-fiberglass board ($\epsilon_r = 2.55$). These are the materials most often available for amateur work. For more precise results, **table 1** lists the required microstrip geometry for Z_o in 5-ohm steps.

effects of strip thickness

The microstrip impedance equations assume a two-dimensional microstrip with zero thickness. For very thin strips ($t/h < 0.005$) the experimental and theoretical results have been shown to be in excellent agreement.[14] For thicker strips, the effect of finite strip thickness can be compensated by slightly reducing the strip width as suggested by Wheeler.[9] In other words, the *effective* microstrip width, w_{eff}, is somewhat wider than the strip's physical width. The Hammerstad impedance formulas can be modified to consider the thickness of the strip by replacing strip width, w, with effective strip width calculated from the following relationships

For $\frac{w}{h} > \frac{1}{2\pi} = 0.16$

$$\frac{w_{eff}}{h} = \frac{w}{h} + \frac{t}{\pi h}\left(1 + \ln \frac{2h}{t}\right) \qquad (13)$$

$$\frac{w_{eff}}{h} = \frac{w}{h} + \frac{t}{\pi h}\left(1 + \ln \frac{4\pi w}{t}\right) \qquad (14)$$

For microstrips with $w/h > 0.16$ etched on 1/16 inch (1.6mm) fiberglass-epoxy circuit board, double-clad with 1-ounce copper ($t = 35.6$ microns or 0.0014 inch), the correction factor Aw is 0.041 (i.e., $w_{eff} = w + 0.041$); for 1/32 inch (0.8mm) circuit board the correction factor Aw is 0.074. For very thin microstrip ($w/h > 0.16$), the width correction is somewhat grearer; it also varies with strip width so must be calculated separately for each different width.

Except for very precise work where photo-etching techniques are used and etching is carefully controlled, the width correction factor may be ignored for low-impedance microstrip etched on common copper-clad board. For impedances above about 70 ohms, however, the effect of strip thickness should be considered (see **fig. 6**); at impedances greater than 100 ohms the 1.4 mil (35 micron) conductor thickness reduces the characteristic impedance by as much as 10 ohms. (Note that the strip widths given in **tables 1** and **2** have been compensated for strip thickness.)

microstrip materials

Of the many materials used in commercial microstrip circuits, fiberglass-epoxy (G-10) circuit board is the one available to most amateurs. Professional designers shy away from G-10 board at frequencies above 200 or 300 MHz, but amateurs have used it quite successfully in low-power circuits up to 1300 MHz. The losses of G-10 increase substantially above 1300 MHz, however, so more expensive Teflon-fiberglass board should be used for microstrip circuits designed for 2300-MHz and the higher amateur bands.

In addition to its loss, microscopic air pockets in the fiberglass-epoxy cause small changes in dielectric constant which can be troublesome in precision circuits. Some users have reported that the characteristics of G-10 tend to vary widely from one manufacturer to another, but this problem is not confined to G-10 — similar problems have been reported for Teflon-fiberglass and alumina.[15] One commercial user I have talked to buys his substrate materials from one supplier, and then purchases material only in batch lots to ensure consistency.

Fiberglass-epoxy circuit board, copper-clad on both sides with 1 or 2 ounce copper, can be obtained from many supply houses in thicknesses of 1/32 inch (0.8mm) and 1/16 inch (1.6mm). This material is also manufactured in 1/8 inch (3.2mm) and 1/64 inch

(0.4mm) thicknesses, but it's usually available only on special order. Note that the specified thickness is the overall dimension and includes the copper foil. One-ounce copper* is 1.4 mils 10.0014 inch or 36 microns) thick; therefore, the dielectric thickness of a double-clad 1/16 inch (62.5 mils or 1.6mm) board is 59.7 mils (1.5mm) [$62.5 - 2(1.4)$]. This is the h dimension shown in **fig. 1.** To find the required microstrip width, simply multiply w/h times the dielectric height h.

For example, assume you need a 75-ohm microstrip. From **table 1,** for 1/16 inch (1.6mm) fiberglass-epoxy board, $w/h = 0.823$ for $Z_o = 75$ *ohms*. Therefore, the required microstrip width (not including the correction factor for finite strip thickness) is

$$w = 0.823 \cdot 59.7 = 49.1 \text{ mils}$$

(about 0.049 inch or 1.3mm)

The dielectric thickness of 1/32 inch 131.25 mils or 0.8mm) board, double clad with 1 ounce copper, is 28.45 mils (0.7mm). Therefore, the microstrip width will be a little less than half that of the same impedance line on 1/16 inch (1.6mm) board — 23.4 mils or 6mm for a 75-ohm microstrip.

*Called 1-ounce copper because one square foot of foil weighs 1 ounce.

Most samples of fiberglass-epoxy circuit board which I've measured with a micrometer have been very close to the specified thickness; I've seldom found circuit-board material so far off that it would seriously affect microstrip width. If you want to make this measurement yourself, carefully strip back the copper foil from a corner of a sample and measure the thickness of the dielectric — this is the critical dimension.

You may also want to check the relative dielectric constant of the board you're going to use. This can be easily done by measuring the capacitance of a carefully cut section of circuit board. The relative dielectric constant is calculated by rewriting the well known formula for parallel plate capacitors with ϵ_r as the unknown

$$\epsilon_r = \frac{C\,h}{0.8842\,A}$$

where C is the capacitance in pF, h is the dielectric thickness in mm, and A is the area in square centimeters. For dimensions in inches the formula is

$$\epsilon_r = \frac{C\,h}{224.4\,A}$$

where C is the capacitance in pF, h is dielectric thickness in mils, and A is the area in square inches. For example, the capacitance of a 1 inch square

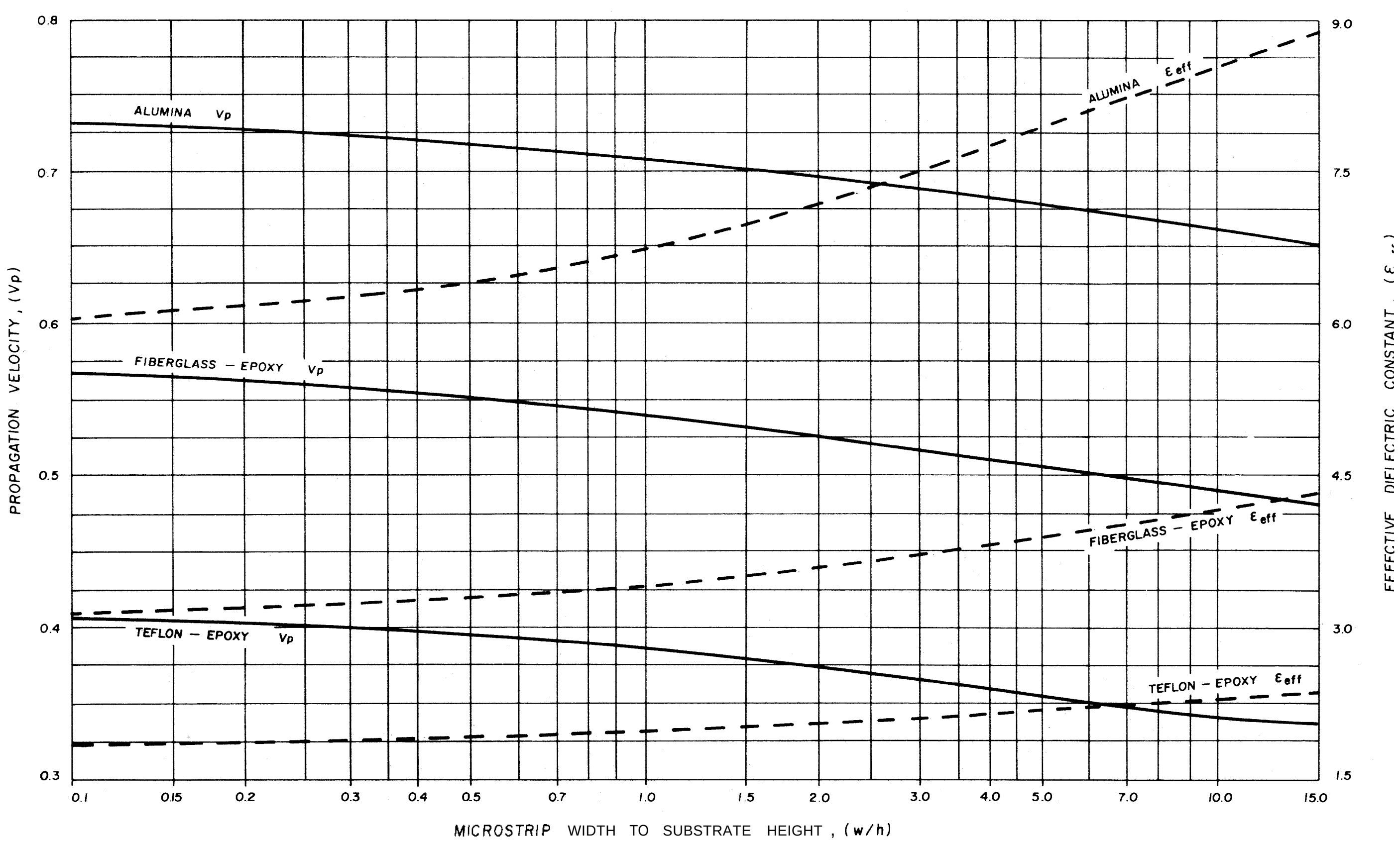


fig. 7. Propagation velocity, v_p (solid lines), and effective dielectric constant, ϵ_{eff} (dashed lines), as a function of the ratio of microstrip width to substrate thickness, w/h, for fiberglass-epoxy ($\epsilon_r = 4.8$), Teflon-epoxy ($\epsilon_r = 2.55$), and alumina ($\epsilon_r = 10.0$).

(6.5cm²) of 1/16 inch (1.6mm) board, double clad with 1 ounce copper, should be about 18 pF for $\epsilon_r = 4.8$. Note, however, that the relative dielectric constant of most materials decreases with increasing frequency. This means your low-frequency capacitance measurement may indicate a dielectric constant between 5.0 and 5.5. If the capacitance of the l-inch square (6.5cm²) sample is between 19 and 21 pF, ϵ_r will be approximately 4.8 at the vhf and uhf frequencies.

One of the advantages of Teflon-fiberglass circuit board, other than its lower loss, is that physical microstrip width and electrical length are larger than substrates with higher dielectric constants. This is very useful at microwave frequencies where high impedance (narrow microstrip) and fractional wavelength lines are required. On the other hand, substrates with higher dielectric constants such as alumina ($\epsilon_r = 10$) are used extensively in microwave ICs where miniaturization is important."

microstrip phase velocity

In addition to the characteristic impedance of microstrip, phase velocity is the one other property of major importance in vhf and uhf designs where precise electrical lengths are needed for impedance matching or other circuit requirements. As was shown in eq. **1,** the phase velocity, v_p, is given by

$$v_p = \frac{c}{\sqrt{\epsilon_{eff}}}$$

where c is the speed of light and ϵ_{eff} is the effective dielectric constant. This may be rewritten to give microstrip wavelength, λ_g, as a function of free-space wavelength, λ_o

$$\lambda_g = v_p\ \lambda_o = \frac{\lambda_o}{\sqrt{\epsilon_{eff}}} \qquad \textbf{(15)}$$

where the free-space wavelength, λ_o, can be found from

$$\lambda_o = \frac{29980}{f_{MHz}}\ (cm) \qquad \textbf{(16)}$$

$$\lambda_o = \frac{11803}{f_{MHz}}\ (inches) \qquad \textbf{(17)}$$

The velocity factor, v_p, which is a function of w/h, is given in tables **1** and **2.**

Example. Assume you need a 75-ohm quarter-wavelength matching transformer at 432.1 MHz; your circuit is to be built on fiberglass-epoxy circuit board ($\epsilon_r = 4.8$). From table **1,** w/h for 75 ohms is 0.823; $v_p = 0.544$. A free-space quarter-wavelength at 432.1 MHz is

$$\frac{\lambda_o}{4} = \frac{29980}{4 \cdot 432.1} = 17.35\ cm\ (6.83\ inches)$$

Therefore $$\frac{\lambda_g}{4} = 0.544 \cdot 17.35 = 9.44\ cm$$

If you use the Hammerstad equations to calculate the required microstrip geometry for a given value of characteristic impedance, ϵ_{eff} can be calculated accurately with either eq. **8** (narrow microstrip) or eq. **10** (wide microstrip).' These same expressions can also be employed if you use the simplified formulas (**eqs.** 3 and **5**), but there's a simpler way if you're not doing precision microstrip work.

The following formula, which was derived from the work of Wheeler and Schneider, is somewhat easier to use than the Hammerstad expressions, covers the entire range of microstrip widths, and gives good accuracy for practical work.

$$\epsilon_{eff} = 1 + (\epsilon - 1)\left[\tfrac{1}{2}\left(1 + \frac{1}{\sqrt{1 + \frac{10}{w/h}}}\right)\right] \qquad \textbf{(18)}$$

Consider the previous example where a 75-ohm quarter-wavelength matching transformer was required at 432.1 MHz. Using the simplified formula (**eq. 3**) to calculate w/h

$$w/h = \frac{142.6}{75} - 1 = 0.90$$

From eq. **18**

$$\epsilon_{eff} = 1 + (3.8)\left[\tfrac{1}{2}\left(1 + \frac{1}{\sqrt{1 + \frac{10}{0.9}}}\right)\right] = 3.446$$

$$\sqrt{\epsilon_{eff}} = \sqrt{3.446} = 1.856$$

$$v_p = \frac{1}{1.856} = 0.54$$

$$\frac{\lambda_g}{4} = 0.54 \cdot 17.35 = 9.37\ cm$$

The small length difference of 0.7mm (0.028 inch) will make no practical difference in circuit operation; and don't be concerned with what appears to be a large discrepancy between w/h calculated with the simple formula, and the accurate value of w/h from table **1.** The actual difference in strip width on 1/16 inch (1.6mm) circuit board is only about 0.004 inch

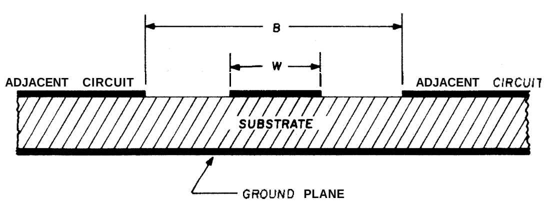
"A new microstrip substrate has been announced by the 3M Company. Called *Epsilam-10,* it's based on a ceramic-filled Teflon material and has the electrical properties of alumina, but can be cut with a razor blade or shear and may be etched like any other printed circuit.

(115 microns). Unless you are using photo-reduction techniques and carefully control both etchant temperature and etching time, you can't maintain this accuracy in your home workshop.

practical considerations

Theoretically, you can design a microstrip line for any desired characteristic impedance, but at very high values of Z_o the conducting strip becomes so narrow that it's impossible to compensate for the effect of finite strip thickness. With a strip width of 1.4 mils (36 microns) on 1 ounce copper, for example, the microstrip is actually square; the impedance of a zero-thickness conductor of this width is about 200 ohms – the impedance of the rectangular conductor is 173 ohms or less; it can't be calculated directly (the width correction factor, eq. **14,** is not accurate for strip widths less than 2.8 mils or 71 microns etched on 1 ounce copper).

Conversely, at very low values of Z_o the microstrip becomes so wide the rf current isn't distributed evenly across the conductor, so it's impossible to accurately predict performance. In addition, it's difficult to prevent coupling to nearby circuitry. For microstrip impedances up to about 50 ohms, other nearby circuit traces should be spaced a minimum of one microstrip width away (i.e., in the illustration below dimension B should be a minimum of three times the strip width). Depending on the application and the required circuit density, this places a lower limit on Z_o at about 10 ohms. Note that dimension B should be about 10 times the microstrip width for impedances above 50 ohms.



Also to be considered are the effects of placing the microstrip circuit in a metal enclosure. Experimental work has shown that a conducting enclosure tends to lower both the impedance and effective dielectric constant because the field lines are prematurely terminated, thereby increasing the density of the field lines in air. When the distance between the upper and lower walls is greater than five times the substrate thickness, and when side-wall spacing is five times the strip width or more, effects on microstrip characteristics are negligible.

It's not important for most amateur applications, but it should be mentioned that the formulas and tables for Z_o and ϵ_r presented in this article are valid only to about 4000 MHz. Above 4000 MHz both Z_o and ϵ_r begin to change with frequency due to the propagation of hybrid modes. This has been discussed in the engineering literature by a number of researchers; interested readers are referred to references 16 and 17.

conclusion

Microstrip transmission line is being used in many applications at vhf and uhf. Although microstrip is a relatively low Q transmission line and has greater losses than air-spaced coaxial or troughline structures, it provides excellent performance in many low-power circuits – indeed, there are some circuits which would be impractical to duplicate in the home workshop without the use of microstrip.

The design of microstrip circuits is the same as that for more conventional transmission-line circuits; if there is sufficient reader interest this will be the subject of a future article.

references

1. R. M. Barrett, "Microwave Printed Circuits – An Historical Survey," IRE Transactions on Microwave Theory and Techniques, MTT-3, March, 1955, page 1
2. H. S. Keen, "Scientific Report on the Study of Strip Transmission Line," AIL Report 2830-2, December, 1955.
3. D. D. Grieg and H. F. Engelmann, "Microstrip – A New Transmission Technique for the Kilomegacycle Range," Proceedings of the IRE, December, 1952, page 1644
4. N. R. Wild et al, Handbook of Tri-Plate Microwave Components, Sanders Associates, Inc., Nashua, New Hampshire, 1956, ASTIA no. AD110157.
5. S. B. Cohn, "Shielded Coupled-Strip Transmission Line," IRE Transactions on Microwave Theory and Techniques, MTT-3, October, 1955, page 29.
6. G. Matthaei et al, Microwave Filters, Impedance-Matching Networks, and Filters, McGraw-Hill, New York, 1964.
7. D. R. Ayer and C. A. Wheeler, "The Evolution of Strip Transmission Line," The Microwave Journal, May, 1969, page 31.
8. H. A. Wheeler, "Transmission-Line Properties of Parallel-Wide Strips by a Conformal-Mapping Approximation." IEEE Transactions on Microwave Theoryand Techniques, May, 1964, page 280.
9. H. A. Wheeler, "Transmission-Line Properties of Parallel Strips Separated by a Dielectric Sheet," IEEE Transactions on Microwave Theory and Techniques, March, 1965, page 172.
10. M. V. Schneider, "Microstrip Lines for Microwave Integrated Circuits," *Bell System Technical Journal,* May-June, 1969, page 1421.
11. H. Sobol, "Extending IC Technology to Microwave Equipment," Electronics, March 20, 1967, page 112.
12. J. R. Fisk, WIHR, "Simple Formula for Microstrip Impedance," ham radio, December, 1977, page 72.
13. E. O. Hammerstad, "Equations for Microstrip Circuit Design," Proceedings of the European Microwave Conference. Hamburg (Germany), September, 1975, page 268.
14. M. A. R. Gunston and J. R. Weale, "Variation of Microstrip Impedance with Strip Thickness," Electronic Letters, December 27. 1969, page 697.
15. S. Hagelin and L. Wernlund, "Properties of Microstrip Transmission Lines," Swedish Research Institute of National Defense Report C3769-E2, June, 1974, page 60.
16. W. J. Getsinger, "Microstrip Dispersion Model," IEEE Transactions on Microwave Theoryand Techniques, January, 1973, page 34.
17. R. P. Owens, "Predicted Frequency Dependence of Microstrip Characteristic Impedance Using the Planar-Waveguide Model," Electronic Letters, May 27, 1976, page 269.

ham radio

microwave path evaluation

How to plot line-of-sight microwave paths under varying propagation conditions to determine if the path is clear

Interest in **amateur microwave** radio seems to be at an all time high. Amateurs in the United Kingdom recently set a new **10-GHz** DX record of 324 miles (521km) and with the recent introduction of inexpensive Gunn diode transceivers, equipment availability has never been better.[1]

The communications capabilities of microwave equipment now within easy reach of the average amateur are initially rather astounding to the commercial microwave user. The difference is, of course, the luxury of working with 0 dB fade margins (on DX paths) and substantially narrower bandwidths. Those two factors allow far greater working range with a given power output.

One area of concern to both amateur and professional is proper path clearance. Since obstacles along the path introduce loss, a signal should clear path obstacles, either optically or effectively through beam bending.

This article reviews the essential information an amateur will need to evaluate a potential microwave path before trying it. The information will be useful to both microwave DX enthusiasts and to those who wish to install permanent microwave links for repeater control or other purposes.

distance and azimuth

For the short distances involved in nearly all microwave paths, great-circle formulas are normally not used to calculate distance and azimuth. This is because these formulas assume a perfect spherical model of the earth when, in fact, the earth's surface is oblate. In addition, it is difficult on very short paths with small differences in latitude and longitude to achieve the necessary degree of trigonometric function accuracy. The usual method used for short microwave paths up to about 300 miles (482km) is to first determine the number of miles per degree of latitude and longitude in the vicinity of the path, and then use a plane right triangle to determine the distance and azimuth.

By incorporating a table of these distances into linear regression formulas, it is possible to devise equations which allow you to dispense with the tables, within certain limits. The regression-derived

By Dennis L. Haarsager, N7DH, 483 South Walnut, Boise, Idaho 83706

formula for the length of the horizontal portion of the triangle is*

$$H = (0.004009m + 69.108348)(\cos m)(h) \qquad (1)$$

and for the vertical part of the triangle

$$V = (0.011993m + 68.513612)(v) \qquad (2)$$

The distance along the path is given by

$$D = \sqrt{H^2 + V^2} \qquad (3)$$

where m = the arithmetic average of the two latitudes
v = the difference between the two latitudes
h = the difference between the two longitudes

It is easiest to perform the arithmetic if minutes and seconds are converted to decimal degrees before performing the mathematical operations. These regression equations are normalized for latitudes between 25°N and 60°N.

To find the azimuth, use **equation 4**

$$A = \arctan V/H \qquad (4)$$

Note that A will be positive in the second and fourth quadrants, and negative in the first and third quadrants. Add this angle algebraically to 90° if the destination is east of the starting point, and 270° if the destination is west of the starting point. The result will be the azimuth of the destination in degrees east of north from the starting point. A bit of ingenuity on your part will enable you to incorporate angle A (not the azimuth) into a procedure for finding intersecting points on map edges where you need to use two or more maps to draft your path profile.

map usage

An essential step in path evaluation is drafting a vertical profile of the path. Use standard graph paper that has at least 10 squares to the inch or, better yet, millimeter divisions. Choose vertical and horizontal scales which will enable you to conveniently plot the range of your path elevations and distance. The smaller the scale you choose, the better the resolution will be.

For rough initial path evaluation, especially in hilly or mountainous areas, the 1:250,000 scale topographic maps published by the United States Geological Survey (USGS) are adequate. Final evaluation of potential obstruction points requires use of the USGS's 7.5- or 15-minute series quadrangle maps. The latter are available for most areas in the United States. Maps and map indexes are available at nominal prices from private map dealers, from USGS sales counters in major cities, or by mail order from certain USGS offices."

Another source of topographical maps, although they are less useful for plotting microwave paths, is your local airport's private aviation flight shop; they sell Sectional and World Aeronautical Charts (1:500,000 and 1:1,000,000 scales, respectively) published by the National Oceanic and Atmospheric Administration. The scale of these maps is too large for accurate microwave path profiles, but significant changes in elevation are shaded on these maps, making them useful for identification of possible DX paths.

In making a path profile, assume the earth is flat and choose some arbitrary elevation — normally the lowest one along your path — as the base elevation for your profile. Enter it on the bottom of your graph paper and add elevation indicators from there according to your vertical scale as illustrated in **fig. 1.**

The next step is to draw a line between the two points which define your path on the map or maps you are using. In most cases you will need to use more than one map to cover your path. You may either tape adjoining map sections together or determine where your line intersects map edges by application of the angle A and **eqs. 1** through **4.** It will be very important to the mathematical procedure to determine the latitude and longitude of the two end points as close as you can. Use the smallest scale maps available for this determination. Carefully

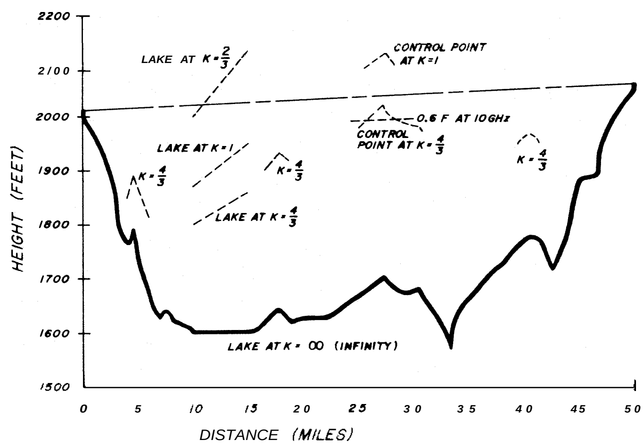


fig. 1. Simple microwave path evaluation between two sites 50 miles (80km) apart. The dashed lines show the effects of different values of K, a factor which indicates the earth's effective radius and results from varying degrees of refraction (see text).

*To convert the results of **eqs. 1.** 2, and 3 to kilometers, multiply **by** 1.609344.

"For maps of areas east of the Mississippi write to the USGS at 1200 South Eads Street, Arlington, Virginia 22202; for areas west of the Mississippi, including all of Louisiana and Minnesota, write to USGS c/o Federal Center, Denver, Colorado 80225.

transfer elevations shown on the maps to the appropriate distances on your profile.

Once the path profile is constructed, it will be necessary to add a correction factor on top of the plotted terrain to allow for the curvature of the earth. This is a bit more complicated than it sounds because the *effective* curvature of the earth changes with varying atmospheric conditions.

Most amateurs are familiar with something called the "4/3 radio horizon," but it is less well known that this horizon is different in different geographical areas, at different times of the day and year, and at different elevations.

The amount and direction of bending or refraction of a radio beam is dependent upon several atmospheric variables. A factor called K is used to indicate the effective earth's radius which results from these varying degrees of refraction. On a *grazing* path, that is, on a path where the microwave beam just barely clears an obstruction, small changes in K can make or break the path for communications by "moving" that obstruction in and out of the beam path.

The K factor is determined by how much another variable, the radio refractivity index, changes over a given change in altitude. The radio refractivity index is normally designated as N and is calculated from pressure, temperature, and water vapor data. When N decreases by 40 units with a one kilometer increase in altitude (the so-called *standard atmosphere)* K equals 4/3. When N decreases by more than 40 units per kilometer, the earth effectively flattens. At – 157 N units/km, K equals infinity and the earth is effectively flat. When N decreases by more than – 157 N units/km, K adopts a negative value and the earth becomes effectively concave. Such extreme conditions are called *superrefractive.*

Under certain atmospheric conditions the index decreases by less than 40 N units/km or even increases with increasing altitude. Under these conditions the earth's effective radius decreases (as measured by K) and path obstructions are effectively higher. When the atmosphere is homogenous, N remains constant over the 1-km range and K equals unity. In this case, there is no refraction and the effective earth's radius is the same as the true earth's radius. During a humidity inversion, N can increase by over 200 N units/km and K is reduced to less than 112. Such extreme conditions are called *subrefractive*.[2]

table 1. Average K values for different localities and atmospheric conditions (adapted from reference 3).

K	description
1.17	Normal K for light refraction; radio refraction in dry, mountainous areas above 7500 feet (2290 meters) elevation
1.20	Dry mountainous areas between 5000 and 7500 feet (1525 to 2290 meters)
1.25	Dry mountainous areas up to 5000 feet (1525 meters)
1.30	Inland plains during winter
1.34	Standard atmosphere K
1.50	Inland plains and northern coastal areas during summer; southern coastal areas during winter
1.60	Southern coastal areas during summer
1.75	Extreme southern coastal areas during summer

The amateur microwave DX enthusiast will be most interested in typical values of K for the path in use. The values for K given in **table 1** will suffice for this purpose.[3] These values of K should be used in **eq. 5** to determine the earth's effective bulge.*

$$B = d_1 d_2 / 1.5K \tag{5}$$

where: B = earth's effective bulge in feet
d_1 = distance from starting point to test point in miles
d_2 = distance from test point to destination in miles

This earth bulge should be added on top of the point you are testing for an obstruction as shown in **fig. 1**. Amateurs interested in establishing permanent microwave links should add an amount for earth bulge at $K = 1$ to allow for adverse propagation conditions at certain times.[4]

other clearance requirements

In evaluating the path profile, an allowance should be made for any trees or buildings known to be at the *controlpoint,* That's the point along the path which gives the least vertical clearance for the microwave beam. Consideration should also be given to these non-terrain obstructions for points close in elevation to the control point.

Finally, an allowance should be made for the space taken up by the microwave beam itself. Without going too deeply into wave theory, every radio beam is composed of concentric Fresnel (pronounced *fray-NEL*) zones. The first Fresnel zone is defined by a grouping of points representing all possible paths which are one-half wavelength longer than a straight line between transmitter and receiver.[5] If it were possible to view these points from the side of the path they would represent a cigar-shaped ellipse. The width of a Fresnel zone is therefore greater in the middle of a path than at the ends. To avoid obstruction loss, it is necessary to have a beam clearance equal to at least 0.6 first Fresnel zone radius over the

*In metric form, **eq. 5** is

$$B = d_1 d_2 / 12.75\, K$$

where B is in meters, and d_1 and d_2 are in kilometers.

potential control point (terrain + earth's bulge + trees or buildings). The formula for this Fresnel zone clearance is given as*

$$0.6F_1 = 43.25\sqrt{d_1 d_2/Df} \qquad (6)$$

where $0.6F_1$ = 0.6 first Fresnel zone radius in feet
D = path length in miles
f = frequency in GHz
d_1, d_2 are defined in **eq. 1**

To plot this additional clearance draw a straight line from the antenna elevations at both the starting (transmitting) and destination (receiving) points. The Fresnel zone clearance is then added below this line as shown in fig. **1.** If there is 0.6 first Fresnel zone radius clearance over your obstruction at the normal K for your area, you will have near free space conditions over the path most of the time. If there is 0.6 first Fresnel zone radius over $K = 1$ (true earth curvature), near free space conditions will exist over the path for all but the most extreme atmospheric conditions.

reflection analysis

Microwave signals may be reflected off certain surfaces such as open water or dew-laden fields of grain — surfaces that appear smooth relative to the wavelength of the microwave beam. If this reflected energy arrives at the receiving antenna at or near 180° out of phase with the direct beam (when clearance over the point is an even-numbered Fresnel zone radius), severe fading can take place. Because of this, it is recommended that amateurs interested in installing permanent, reliable microwave links also evaluate their potential paths for such reflection points.

Since the position of a reflection point varies with any change in the earth's effective radius (K), reflection analysis begins with sketching in the potentially reflective surface and other terrain features at differing values of K. This procedure makes use of **eq. 5** and is illustrated in fig. **1.** If the reflective surface is not clearly blocked by intervening terrain at the same K factor, it is advisable to make some computations to determine if a reflected wave will reach your desired receiving antenna.

The following formula may be used for such computations.†

$$h_2 = d_2 \left[\frac{h_1}{d_1} + \frac{(d_2 - d_1)}{1.5k}\right]$$

where h_1, h_2 = higher antenna and reflection heights, respectively, in feet, above the reflective surface plotted at $K = \infty$.
d_1, d_2 = distance from each antenna to the potential reflection point, in miles

In fig. **1,** the potential reflecting surface is the lake which runs from mile 10 to mile 15 along the path. Assume that the obstructing point for $K = 4/3$ at about 27.5 miles is not there. To calculate where the reflection crosses mile 50, use **eq. 7.** Substituting, $h_1 = 410$ feet, $d_1 = 10$ miles, $d_2 = 40$ miles, and $K = 4/3$. This makes $h_2 = 2240$ feet. Adding this to the height of the reflecting surface at $K = \infty$ (1600 feet) gives 3840 feet above sea level. Since this is well above the planned receiving antenna height, no reflection problems will be experienced because of this reflection point. Similar calculations should be made for other unobstructed potential reflection points and at other projected values of K.

summary

In fig. **1**, the example given assumes an operating frequency of 10 GHz, and a *normal K* of 4/3. The path profiled is 50 miles (80km) long.

At this K value, the user will have little more than grazing clearance over the control point, which is located 27.3 miles (44km) from the transmitting antenna. There is not 0.6 first Fresnel zone clearance over the control point, however, indicating that a small amount of obstruction loss will be experienced. For casual use, this loss will pose no major problem, but for permanent installations the small clearance will give fading problems whenever K goes below 4/3. Note that the receiving point is *shadowed* when K decreases to about 1.22 or less.

The techniques outlined here will give the amateur microwave user the basic information needed to evaluate any potential path for normal communications possibilities. It does not preclude use in path analysis for tropospheric scatter or refractive duct communications, but such techniques require additional analysis beyond the scope of this article.

*In metric form, **eq. 6** is

$$0.6F_1 = 10.39\sqrt{d_1 d_2/Df}$$

where d_1, d_2, and D are in kilometers.

†Adapted from reference **4,** page 17.

references

1. J. Fisk, W1HR, "Solid-State Microwave RF Generators," *ham radio,* April, 1977, page 10.
2. "Anomalous Propagation — Part I," *GTE-Lenkurt Demodulator,* July/August, 1975, page 280.
3. B. R. Bean, J. D. Horn, and A. M. Ozanich, Jr., *Climatic Charts and Data of the Radio Refractive Index for the United States and the World,* National Bureau of Standards Monograph 22, 1960, page 80.
4. R. F. White, *Engineering Considerations for Microwave Communications, GTE* Lenkurt, San Carlos, California, 1970, page 51.
5. I. T. Corbell, "Microwave Engineering for the Broadcaster," NAB *Engineering Handbook,* 6th edition, page 615.

ham radio

solution to the
low-band antenna problem

The Marconi antenna works great on the lower amateur frequencies — here's how to make it play using simple methods and inexpensive parts

Since many transceivers include 160 meters, many amateurs need a low-band antenna. However, this can raise a number of questions or comments, such as:

1. Most trap or shortened antennas are very narrow band. Is there something simple that will cover the entire low bands?

2. The cost of an antenna and transmatch may be wasted if I find I don't like 160 meters.

3. My lot is too small for even a quarter wavelength on 160, so how can I put up something effective?

4. I tried to load a "compromise" antenna but burned up my final amplifier before I got it to load. I don't intend to buy an antenna lab just to check out an antenna!

background

Radio Handbook has, for years, included one antenna — a 180-foot (55m) Marconi. This length works as a 314-wavelength antenna on 80 meters and as a 114-wavelength on 160 meters by shortening its electrical length with a variable capacitor. This antenna works nicely, but may still be too long. The Handbook suggests a 90-foot (27.4m) length for 80 and 40, and may make you think about loading it for 160.

The latest edition of *Radio Handbook* suggests the antenna be pruned to favor desired band segments. This seemed to fit the needs of a friend who needed a modest 160-meter antenna, so I agreed to "wring it out." The antenna started as a 94-foot-long (28.7m) antenna and was a real bear on 80, since the 1/2-wavelength point was at 4 MHz. Since it was end-to-end with my Bobtail Curtain, I assumed it was coupling to that antenna, so I re-erected it in the opposite direction completely clear of all antennas and utility lines. This was no improvement, so the antenna was progressively shortened until it behaved.

taming the antenna

Table **1** shows the results. The column labeled compensation indicates the component in series with the antenna to make it purely resistive at that frequency. At that point the *R* value was determined. Notice that the 82-foot (25m) version will provide operation on the high end of 160, all of 40, and all of 15 with low swr, and no matching is needed other than the simple compensating element. You might feel that the 94 footer (28.7m) should behave exactly as the 180 footer (55m) does, with respect to impedances. It won't for this reason: 160 and 80 are very nearly harmonically related (1.8 to 3.6 MHz and 2 to 4 MHz). 80 and 40 are not: (3.5 to 7 MHz and 4 to 8 MHz). If you set up the 94 footer (28.7m) for resonance at 8 MHz, it will perform nicely at 80 meters. The 82 foot (25m) version in table **1** is resonant at 7450 kHz.

It might be a temptation to let the antenna behave as a 1/2 wavelength at 75 meters. It could be loaded with a parallel-resonant tank circuit link coupled to the rig. If your version had a flat top at a height of 1/8 wavelength or less, if it was built of no. 12 (2.1mm) or smaller, and if it were located over soil having excellent conductivity (or a radial ground system), its

By Bill Wildenhein, W8YFB, 41230 Butternut Ridge, Elyria, Ohio 44035

table 1. Values of compensating components for use with the low-band Marconi antenna (160 through 10 meters).

frequency kHz	R (ohms)	94 ft (28.7m) to north compensation	R (ohms)	94 ft (28.7m) to south compensation	R (ohms)	82 ft (25m) to north compensation	R (ohms)	82 ft (25m) to sou compensation
1803	22	15.5μH	25	19 5μH	25	22 0μH	20	20 5μH
1890	30	11.5μH	30	18.0μH	30	16.0μH	40	17 5μH
3557	300	100pF	250	100pF	160	120pF	145	120pF
3646	500	100pF	350	80pF	200	90pF	160	100pF
3180	900	80pF	500	70pF	250	75pF	240	80pF
3980	5k	30pF	1.5k	36pF	1k	35pF	500	50pF
7005	100	240pF	48	3 35μH	75	1.2μH	58	2.3μH
7150			50	2.7μH			60	1 8μH
7200	115	150pF			63	RESONANT		
7292	125	125pF	55	1 9μH	65	520pF	63	1.2μH
14,010			750	RESONANT	750	SEE TEXT	350	75pF
14,240			900	0.54μH	800	SEE TEXT	1000	RESONANT
14,300			900	1.2μH	600	SEE TEXT	800	0.26μH
21,015	90		100	1.2μH	65	0.7μH	75	0 85μH
21,360	125		95	1 0μH	58	0.25μH	60	0.72μH
21,450	130		90	0.85μH	60	0.25μH	58	0.68μH
28,020	250	1.4μH	90	325pF	175	0.25μH	110	190pF
28.600	130	1.0μH	120	170pF	250	0.4μH	80	100pF
29,560	100	0.54μH	150	RESONANT	100	0.55μH	110	75pF

end impedance will be well over 10,000 ohms. That means very high rf voltage in the station, BCI and TVI become more severe, lead-in losses rise, capacitor spacing would have to be much greater (and cost much more). There would be a possibility of very serious rf burns if a child happened to touch the lead-in. It would be beyond the capabilities of commercial transmatches.

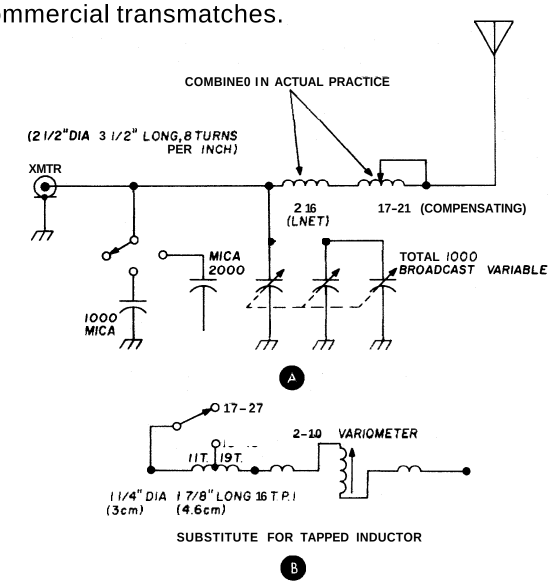


fig. 1. Loading circuit for the low-band Marconi. The L network and compensating coil are one unit in the practical circuit. Voltage is low in this system, so an ordinary capacitor from a broadcast receiver can be used for tuning. Sketch B shows an alternative method for a tapped inductance.

Instead, deliberately detune it by pruning it well above 40 meters. The 82-foot (25m) version is resonant at 7450 kHz in my backyard. Notice that the 80-meter numbers become quite reasonable. Also notice that the 40- and 15-meter numbers do not change appreciably with changes in length. You might be concerned that the antenna will lose efficiency if not cut to exact resonance. By adding the small compensating coil, you are resonating the entire system.

the ground system

A good ground is essential for a Marconi antenna. I use three radials for each band, 1/4-wavelength long, and fanned out under the antenna. These radials are tied to a ground stake just outside the basement window. The lead-in is a scrap of no. 4 AWG (5mm) wire but could just as well be a scrap of RG-8/U coax cable with shield and center conductor tied together. A second ground stake about 8 feet (2.4m) away is tied to the first with another scrap of no. 4 AWG (5mm) wire. Inside the window, my copper plumbing is also tied in with a piece of no. 4 AWG (5mm) wire. If you have hot water heating, the plumbing in the basement may be copper pipe buried in the concrete. This results in a large area of copper capacitively coupled to ground. The ground wire may be fastened to the copper plumbing with hose clamps. A ground system such as this is very effective and results in an efficient antenna.

An easy way to enter a basement window is to replace a pane with plexiglass. This material may be drilled (cautiously) for 1/4-20 (M7) brass bolts for antenna and ground lead. Keep the antenna lead well separated from the sash and ground lead. The most desirable location for the tuner is on a shelf at the basement window, with coax running to the rig. A long ground lead is undesirable. In short, when using a Marconi, don't cheat on the ground! I have used Marconis with only one radial cut for the lowest band; another had no radials and only a water-pipe ground. With the latter antenna I worked over 20 states on 160 with 12 watts on a-m. The present antenna has convinced me of the value of a good ground. I get good, solid contacts over at least a 500-mile (800km) radius running 35 watts PEP on ssb. This, again, on 160 meters.

leading circuit

Now let's see how cheaply we can load this thing. If you want the entire 160-meter band with unity swr, the arithmetic goes like this:

$$X_L = \sqrt{R\,R_{IN} - R^2} = \sqrt{20 \times 50 - 20^2}$$

$$= \sqrt{1000 - 400} = 24.5\ ohms$$

$$X_C = \frac{R\,R_{IN}}{X_L} = \frac{20 \times 50}{24.5} = 40.8\ ohms$$

$$L_{\mu H} = \frac{0.159160\ X_L}{f_{MHz}} = \frac{0.159160 \times 24.5}{1.8} = 2.17 \mu H$$

$$C_{pF} = \frac{159160}{f_{MHz} X_C} = \frac{159160}{1.8 \times 40.8} = 2167\ pF$$

This is for the case where the antenna presents 10 ohms. **Fig. 1** shows the circuit. The voltage across the capacitor is low so an ordinary 3-gang broadcast capacitor will do nicely. The inductance above can be added to the compensating inductance in the form of one single coil.

Next, there are two ways to match the antenna for 80 meters. If you have an Amidon balun core it can be rewound with the same number of turns, trifilar, and connected as an autotransformer. Connect the *finish* end of one winding to the *start* of the next. **Fig. 2** shows the hookup. If you jockey the impedance values to duplicate mine (by varying antenna length), this system will result in a maximum swr of 1.5 at any point in the 80-meter band.

The total expense of matching components for 160 and 80 is low enough that you can have separate, pretuned networks for each band. On 40 meters the antenna is so close to the proper impedance that you need only a 2.5 μH coil in series with the antenna lead, and with tap set for minimum swr. This will allow you to assess the capabilities of the setup. If you like it, you can buy a *Transmatch* later. You may have to add series inductance to get a commercial *Transmatch* to tune 160.

Another solution for 80-meter matching is shown in **fig.** 3. This is the reverse of the network shown in **fig. 1,** since we want to step up the impedance. This network will allow unity swr at any point in the 80-meter band. Component values allow reasonable cost if you are operating a transceiver barefoot. Values are computed in this manner:

$$X_C = R\sqrt{\frac{R_{IN}}{R - R_{IN}}} = 500\sqrt{\frac{50}{500 - 50}}$$

$$= 500\sqrt{0.1111} = 166\ ohms$$

$$X_L = \frac{R\ R_{IN}}{X_C} = \frac{500 \times 50}{166} = 150\ ohms$$

$$C_{pF} = \frac{159160}{f_{MHz} X_C} = \frac{159160}{4 \times 166} = 240\ pF$$

$$L_{\mu H} = \frac{0.159160\ X_L}{f_{MHz}} = \frac{0.159160 \times 150}{4} = 6 \mu H$$

This gives us the values for the 500-ohm point. In the same way you can calculate the values for the 3.5-MHz point.

Before you run away from the simple math shown here, let me say that the amateur who usually uses guesswork would look at that choice of coil and say, "You're nuts! That coil might work on 20 but will never tune on 80!" Substituting a coil that "looked" like an 80-meter coil, he'd hunt at length for a tap position that would result in a match. Finding it with the coil nearly shorted out, he'd say, "Anyone can see this isn't right! It must be loading on a harmonic or something! Better junk that antenna before you get in trouble with the FCC!"

Component values for an L network are, to a large extent, controlled by the ratio of impedances to be matched. Another reason for calculating the values is that it often enables you to see trouble ahead. If you didn't take the time to calculate the values, you might say, "Most commercial transmatches seem to use big capacitors. I have a pair of 450 pF variables here. With these I should be able to match ANYTHING." A good example of this is the "Moose" shown in the photo. Since I do a lot of antenna experimenting I built this thing to give me a wide range of capabilities as a pi network, or either version of an L network. I calculated the values needed for my antenna on 20 meters. The minimum capacitance required made me suspicious. I wasn't disappointed! The network in the photo has a minimum capacitance too high to permit using it on

20! The minimum plus distributed capacitance of the 450-pF capacitors would certainly also be too high,

While this antenna is primarily intended for 160-80-40-meter operation, it will work the higher-frequency bands. On the higher bands it begins to take on the characteristics of a long-wire antenna. For example, as can be seen from **table 1,** variations in impedance between voltage-feed and current-feed points become progressively less as frequency increases. The antenna will begin to exhibit directional tendencies and can supply occasional surprises when you hit the correct conditions. You'll probably want to use a more effective antenna on the higher bands, but this one is quite useful as a standby antenna. By the same token, it will, obviously, do a good job of radiating harmonics.

If TVI is a problem, a lowpass filter in the lead from the transmitter to the matching network should do the job. In my case, the TV set is located directly above the rig. Total separation is less than 10 feet (3m). The rig can be operated with shield covers removed, but no TVI occurs on 160 through 40. Of course, I use low power — 35 watts PEP, while the big amplifier is down for modification. If you do experience trouble, the lowpass filter should be the answer unless you have excessive ground-lead length. This will tend to couple the rf to the ac line. In that case a lowpass filter may still help but may not be quite as effective.

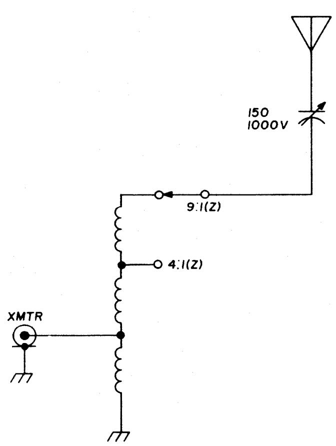


fig. 2. Schematic showing a method for matching the Marconi to your rig for 80-meter operation. By varying impedance values, this circuit will provide an swr of 1.5 across the entire band.

While on the subject of lowpass filters, it might be a temptation to build one of the simple lowpass filters described in the ARRL *Handbook* (late editions) for use in cleaning up vfo outputs. This little filter is very useful for its intended purpose but will behave as a halfwave filter when used in the output of a transmitter. For example, if the cutoff frequency were set a little beyond 20 meters (for instance, if you used a combination of 620 and 820 pF capacitors to yield a 14.5 MHz cutoff), it will work beautifully on 20 meters. If you check it with the signal going into a good 50-ohm load you'll find that each lower band will show greater reflected power, making it useless as a wide-range unit. The halfwave filter is a very effective single-band device. If you have TVI problems on a single band, the filter can be an inexpensive and effective cure. Just don't be misled into thinking it will perform well on all bands below the cutoff frequency!

instrumentation and design aids

If you aren't the experimenter type, you're probably worrying about all the L and C calculations. Be smart! Send $2.00 to the ARRL for their Type A *Calculator.* With this thing you can solve your L, C, and F calculations in seconds. I've worn out three of them over the years, and keep one at work and one at home. Once you get used to the thing, you're hooked. Instructions are printed right on the calculator, so it's impossible to mislay or forget this information.

A good practical example of the savings in time and money is this: You might not have a grid dipper but may want a cheap and dirty model just for this antenna work. Perhaps you have a little two-gang broadcast capacitor but don't know its capacitance range. Be assured it will work very well in a conventional Colpitts oscillator. Perhaps you leave it on the original chassis, use an existing socket, and haywire the circuit. You may have some defunct octal tubes. The bases can be the plugin coil forms. Take a guess that in the Colpitts circuit the little capacitor will hit 100 pF somewhere in the vicinity of a capacitor setting that puts it two-thirds meshed.

The ARRL calculator gives you the correct number of turns on the tube base to hit some desired frequency. Once you have the instrument running, you can determine the total range of frequencies the coil will cover.

Now you know the coil inductance and the end frequencies. With the calculator you can determine accurately the exact number of turns for any range. The calculator will also tell you the true maximum and minimum capacitance you have, which will include stray capacitance. You can use the calculator to juggle turns to give a desired amount of overlap in coil ranges. In other words, it takes out the guesswork, and the time-consuming work of removing turns or rewinding coils repeatedly to achieve a specific range.

The calculator also gives positive verification of a hoped-for cost saving. For instance, in making the dipper, you might look at the tube base and say, ''I'll never get a 160-meter coil on that form!" But the calculator says you certainly will. For example, very often an article might specify a certain length of husky coil stock. Your calculator tells you what inductance this represents. It enables you to say confidently, "That coil from a surplus tuning unit will be correct, too." Think about it. You saved the price of the calculator with that one decision!

measuring antenna impedance

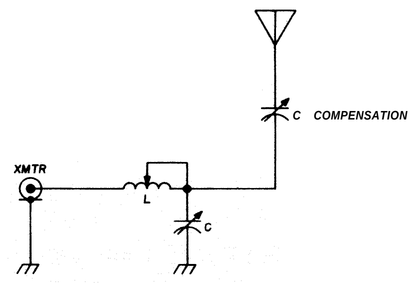
Possibly in your location the antenna impedances will be somewhat different from mine. Or you might have heard that two or three spaced wires, instead of a single wire, will decrease the impedance at the halfwave points. How much lower? You need to measure the impedance but have no tools? The *Radio Handbook* has carried a description of a simple little bridge for a number of years. It is the *Antennascope.* The latest edition has a version that only goes to 100 ohms. Earlier editions had a model that went to 1000 ohms. For hf antenna work, that model is entirely adequate. It is an inexpensive, two-evening project and can be made from standard parts.

The meter need not be an expensive one since we're concerned only with a null. You can substitute one of the $1.50 surplus tuning meters used in stereo amplifiers. It should have a 100 or 200 microampere movement. Calibration requires nothing more than a handful of resistors between 10-1000 ohms.

If you really want to go all out on antenna measurements, Hank Keen described a bridge that independently measures the R and X components.[1] I described a similar one in reference 2. While only a little more costly and complex, both instruments will do an excellent job. However, if properly used, the little *Antennascope* is entirely adequate.

In addition to the *Antennascope,* you'll need a source of low-power rf. This source can be a vacuum-tube-type grid dipper (the solid-state models don't have the power output necessary to serve as bridge drivers). You could also use a simple crystal oscillator. This oscillator should be a vacuum-tube type with a tuned-plate circuit link coupled to the *Antennascope.* No tuning meter is required since the *Antennascope* meter can be used to indicate maximum output from the oscillator. Again, this could be built out of a scrap ac-dc set. The tuning capacitor in the ac-dc set will be adequate. Plug-in coils wound on tube bases will be satisfactory.

With the original tuning capacitor you can hit two adjacent bands with a single coil. For instance, one coil will hit both 160 and 80 meters. A second coil will hit 40 and 20 meters. If you don't have an rf choke handy, use one pie from an old 455-kHz i-f transformer.



band	L	C	C compensation
80	3 6μH 12 turns 2" dia (51mm) 1 1 2" (38mm) long	240 440pF 1kV	50 120pF 1kV
20	1 4-2 5μH 12 turns 2" dia (52mm) 1 1/2" (38mm) long	40-80pF, 1kV	shorted out
10	0 25 0 35μH 12 turns 2" dia (51mm) 1-1/2" (38mm) long	50 60pF, 500V	75-200pF, 500V

fig. 3. Alternative matching circuit for 80 meters.

Another solution is possible if you have a service-type signal generator. Many of these units have an open circuit output of about 0.3 volts. Make a "mini linear" using a hot pentode such as a 6AG7, 6CL6, or 6GK6. Don't tune the input — just use a 10k resistor to ground. Tune the output and link couple it to the *Antennascope.* Set it up according to the tube manual values for class A operation. If output is marginal, tune the link. If you're using this technique or a grid dipper, don't rely on dial calibration; verify the frequency with your communications receiver.

measurement technique

The proper technique to ensure accurate measurements is this: In each case, the antenna must first be made resonant. **Table 1** will give you a rough idea of the compensating"element needed to ensure exact resonance. Example:

1. On 160 meters, connect about 30 μH of inductance between antenna and ground.

2. Couple a dipper to the coil and carefully short turns until a dip occurs at the desired frequency.

3. Disconnect the grounded end of the coil from ground and insert the *Antennascope* between ground and the end of the coil.

4. Feed power to the *Antennascope* at the same frequency.

5. Adjust the *Antennascope* for a null on its meter and read the resistance indicated. Also note the inductance needed to obtain resonance.

On 40, if your antenna is similar to mine, you could use the same technique but with a smaller coil. On 80 you'd use a section or two of a broadcast capacitor in series with the antenna. The free end of the capacitor goes to a 2-turn link coil to ground. The dipper is coupled to the link, and the series capacitor is adjusted for a dip. Then the *Antennascope* is connected in place of the link. At halfwave resonance, the impedance is very high. If you connect a link to ground, you may not be able to find a dip — or the dip may be very shallow. Thus, on 80 meters, it would be best to start at 3.5 MHz. The antenna will be far enough from the half-wavelength resonant point to enable you to get a dip. The dip may still be shallow, but it can be found. A similar situation exists on 20 meters.

Another technique is to judge resonance by deepness of dip on the *Antennascope.* For instance, the *Antennascope* could be connected directly from antenna to ground on 80 meters. You may find that its null is shallow. This will make it difficult to determine the exact R value. You can connect the series capacitor between the *Antennascope* and antenna. Adjust the capacitor carefully to find the point where the null on the *Antennascope* is deepest. If this is at the point of minimum capacitance on the series capacitor, you may want to use a smaller capacitor. Finally, find a point with the series capacitor where the null is deep and sharp. This makes it easy to obtain the exact R value.

On 20 meters you may have to use either a small inductance or fairly large capacitance to find resonance (depending upon the frequency in the 20-meter band). At some point the antenna may be resonant and will require no compensation. (This is the advantage of using the more complex bridge mentioned earlier.) Incidentally, in a situation such as the 82-foot (25m) antenna at 20 meters, the reactive component is small enough so that no compensating element must be used with an L network. A standard L network will accommodate this reactive component, which simplifies matching.

This antenna does a good job on the low-frequency bands. I hope this description makes it easy for you to obtain top performance. With the simple equipment and techniques described, you can be assured of an exact match without even turning on your rig. One last warning: The antenna length is measured from the far end to the point where it connects to the tuner. Have fun!

references

1. Henry S. Keen, W2CTK, "A Simple Bridge for Antenna Measurements," ham radio. September. 1970. pages 34-38.
2. Bill Wildenheim, W8YFB, "Low-Cost RX Impedance Bridge." ham radio, May. 1973, pages 6-15.

ham radio

a review of
ssb phasing techniques

Phasing methods for ssb signal generation have provoked much controversy — here's an article that provides some interesting information to the contrary

A revival **of** interest in ssb phasing systems seems to have occurred in recent years, a trend I wholeheartedly endorse. Thus it's time to discuss the various phasing techniques used and to introduce some new or little-known techniques, which in my opinion, may revolutionize traditional approaches for obtaining ssb signals by this method.

Using the rf-phasing methods discussed near the end of the article, it is my opinion that direct-conversion ssb generators and receivers can be made that cover an octave or more in bandwidth. This is the so-called "third-method" of ssb-signal generation in which the desired output signals can be obtained without using cumbersome heterodyne methods with filters and their problems of frequency drift, which require periodic realignment.

economic considerations

A block diagram of the classic phasing method of ssb generation is shown in **fig. 1(A)**. More usually for practical reasons it's implemented as shown in **fig. 1(B)**. The key to the whole technique is the phasing networks – this may be stating the obvious, but their design, construction, and adjustment, and the difficulties encountered therein account for much constructor resistance to phasing ssb. Phasing ssb is mostly looked on as not a "proper" method of generating ssb, an opinion I entirely reject. I've always thought that the filter method is really a brute-force technique.

It's possible to obtain opposite-sideband suppression of more than 50 dB with narrow bandwidths, but the subsequent amplifier stages degrade this

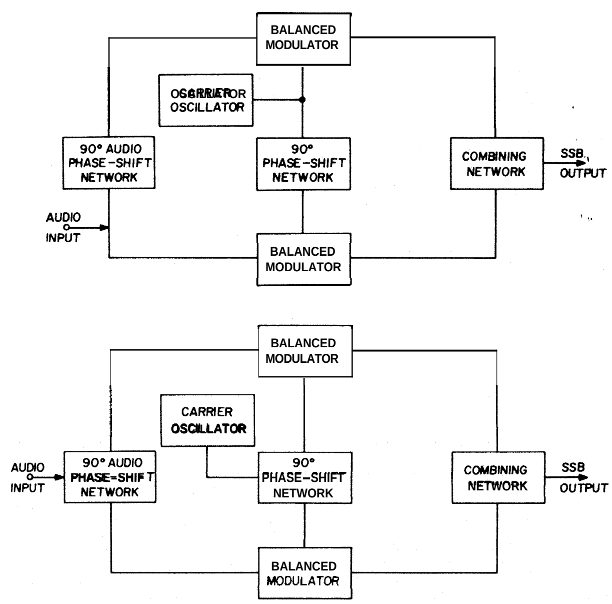


fig. 1. Classic phasing method for generating ssb signals (upper). For practical reasons, the method is implemented as shown in the lower drawing.

suppression and, as they generate intermodulation products at high levels, some of the advantage is lost.

As the state of the art existed some 20 years ago, when most ssb equipment was homebrewed, the

By Roger Harrison, VK2ZTB, 14 Rosebery Street, Balmain 2041, Australia

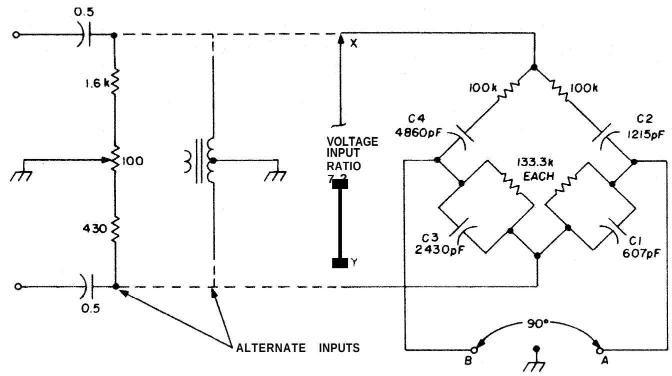
filter method was somewhat easier to implement. However, cost is almost always a consideration in homebrew projects, and the phasing method has a great advantage here. Furthermore, using some circuit techniques described later, my opinion is that the phasing method is easier to implement than the filter method. Although you must spend $30 - $50 on a filter and matched upper and lower sideband crystals and a further $5 - $10 on components, a phasing generator can be built for around $10 or less.

Even if you buy a batch of surplus crystals and make your own filter the cost will be considerable. Furthermore, the filter will require a great deal of time and effort to align — without guaranteed results. In my opinion, simplification of both circuit and alignment, as in the phasing system, is a step forward in which you can do the job required and still achieve adequate specifications.

ssb requirements

Let's look at what specifications are considered "adequate" for ssb. Opposite sideband suppression is important; after all, you have to live with your neighbors. Opposite sideband suppression of −40 dB is quoted in many texts as reasonable. However, with moderate output power −35 dB can be tolerated. Such suppression can be obtained with phasing techniques, but with some circuits it's difficult to maintain this number; in other circuits it can be exceeded.

Carrier suppression with phasing ssb depends on balanced-modulator performance, as with filter systems. Suppression of −50 dB may be obtained with filter systems, depending on the type of mixer. The bandwidth of filter-type ssb systems is mostly determined by the filter. These bandwidths range between 2.1 and 3.2 kHz for most commercially available equipment, which usually has a 6-60 dB shape factor of more than 2.



C1	607 pF (560 pF and 47 pF, 5% capacitors in parallel)
C2	1215 pF (390 pF and 820 pF, 5% capacitors in parallel)
C3	2430 pF (2200 pF and 220 pF, 5% capacitors in parallel)
C4	4860 pF (4700 pF and 150 pF, 5% capacitors in parallel)
R1,R2	133.3k (E96 series) or 100k and 33k, 1% resistors in series, or 1.2 meg. 5%, and 150k, 1% resistors in parallel

fig. 2. An audio phase-shift network (psn) for homebrew ssb projects popular from the early 1950s to the present. It was marketed by the Millen Company and by Central Electronics. Preferred capacitors are 1% or 2% silver mica, NPO ceramic, or polystyrene.

With phasing systems bandwidth depends on audiostage bandwidth and the audio phase-shift network. Unless elaborate sharp cutoff audio filters are used, the shape factor is not as good as in the filter system. However, this isn't a major problem, and many operators find phasing-type ssb easier to tune (on receive) and often describe it as having a more natural quality. It's easy to weight the audio response of a phasing generator to provide improved intelligibility.

Maintaining the specifications, particularly opposite sideband suppression, has always been a problem with phasing-type ssb systems, requiring periodic realignment of the phase-shift circuits. Opposite-sideband suppression was largely a function of component stability (that is, frequency drift).

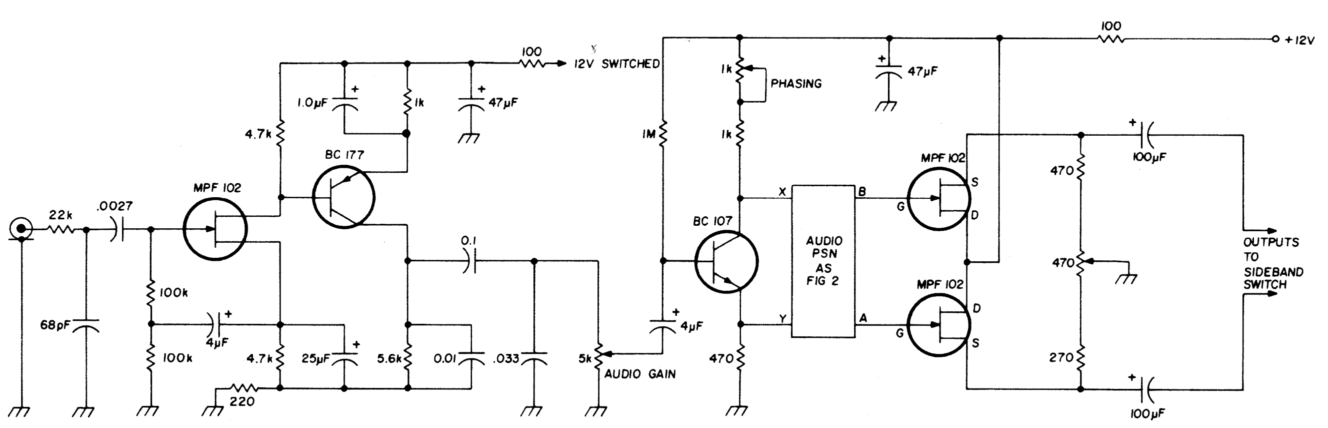
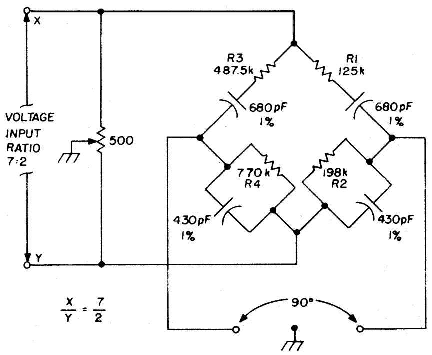


fig. 3. Audio stages of the "Tucker Tin Mark II" (reference 1), a phasing ssb transmitter using the phase-shift network shown in fig. 2.



C1	24.2 nF or 24200 pF (22 nF polycarbonate or polystyrene and a 470 pF silver mica. NPO ceramic. or polystyrene, connected in parallel)
C2	8.06 nF or 8060 pF 14.7 nF and 3.3 nF in parallel, or 6.9 nF and 1.2 nF in parallel, or 8.2 nF and 0.47 μF in series, polycarbonate or polystyrene capacitors)
C3	5.35 nF or 5350 pF (5 6 nF and 0.12 μF in series. or 4.7 nF and 680 pF in parallel, polycarbonate or polystyrene capacitors)
C4	4.03 nF or 4030 pF (3.9 nF and 120 pF in parallel, polycarbonate or polystyrene capacitors)
C5	1.78 nF or 1780 pF (1.8 nF and 0.18 μF in series, or 1.5 nF and 270 pF in parallel)
C6	892 pF (560 pF and 330 pF in parallel: use silver mica. NPO ceramic, or polystyrene capacitors)
R1	20k E96 series or two 10k, 1% resistors in sertes
R2.R3	60.4k E96 series, or 27k, and 33k, 1% in series, or two 120k in parallel

fig. 4. Phase-shift network first popularized by WZKUJ (reference 1).

With modern components and circuit techniques, these problems can be overcome, as we shall see.

audio phase-shift networks

Two types of phase-shift networks are used — active and passive. The latter are most widely used, but we'll examine both.

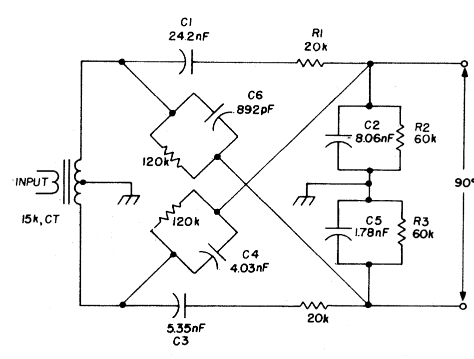
In the heyday of homebrew ssb, several commercially made audio phase-shift networks were available; the Millen and B&W 2Q4 being perhaps most widely used. The Millen network is shown in **fig. 2.** Its popularity over the past 20 years is probably due to its relative simplicity. When properly adjusted the differential phase shift between outputs can be maintained within ± 1.3° of 90° over the audio range from 225 to 2750 Hz. This results in an average opposite-sideband suppression of 45 dB. Using off-the-shelf 1% or 5% resistors and 2% or 5% capacitors, an opposite-sideband suppression of 40 dB can be achieved. The circuit has two other distinct advantages: it requires a minimum of 12-14 components, and the overall loss is about 10 dB, which is the lowest of all the RC networks to be described.

The Millen circuit requires unequal drive voltages at inputs X and Y **(fig. 2)** in the ratio of 7:2. The source impedance is about 2k. The circuit may be driven by a specially wound transformer (well-nigh impossible to obtain today), which would have to be built. Another alternative is to drive the circuit from a phase splitter (that is, 180° out of phase) through the RC network shown. The 100-ohm trimpot is then adjusted so that the audio voltage on input Y is only 28.5% of that on X. Alternatively, the pot may be adjusted to provide equal-amplitude signals at the outputs, A and B.

The audio bandwidth must be restricted as the differential phase shift between A and B departs further from the required 90° outside the bandwidth mentioned, thus markedly degrading the opposite sideband suppression. A rolloff of at least 12 dB per octave above about 2.5-3.0 kHz is recommended – preferably 16 dB per octave. The low frequency should be rolled off at about 10-12 dB per octave below 300 Hz.

The outputs must drive a very high impedance, preferably an fet source follower. The "Tucker Tin Mark II" ssb transmitter' used this phase-shift network, which was driven by a phase splitter. The outputs drove two fet source followers as shown in **fig.** 3. The audio amplifier, which has a high-impedance input, is arranged to provide the appropriate frequency response and gain, making it suitable for use with either crystal, ceramic, or dynamic microphones.

The phasing and audio-frequency balance pots are adjusted to provide minimum opposite-sideband suppression during alignment and rarely need readjustment. The phase-pot is used to adjust the input drive voltages to the phase-shift network in the correct ratio. The Tucker Tin was a highly successful kit. It was made available by the Upper Hutt branch of the NZART. Phase-shift network components were standard off-the-shelf 1% resistors and 5% capacitors.



R1	125k (124k or E12 series, 120k and 4.7k, 1% resistors in series)
R2	198k (196k E96 or E12 series, 180k and 18k, 1% resistors in series)
R3	487.5k (487k E96 or E12 series, 470k and 18k, 1% resistors in series, or 3.9 meg, 5%, and 560k, 1% resistors in parallel)
R4	770k (768k E96 or E12 series, 12 meg. 5%. and 820k, 1% resistors in parallel, or 680k and 82k, 1% resistors in series)

fig. 5. Network designed by Dome (reference 4). Circuit must be driven by equal-amplitude, opposite-phase signals as shown or from a phase splitter. Output must be a very high impedance.

C1	10.5 nF or 10500 pF (10 nF and 470 nF in parallel)
C2	32.3 nF or 32300 pF (27 nF and 560 pF in parallel)
C3	7 14 nF or7140 pF (6.8nF and 330 pF in parallel)
C4	5.38nF or 5380 pF 14.7 nF and 680 pF in parallel)
C5	2.38 nF or2380 pF (2.2nF and 180 pF in parallel)
C6	1.2 nF or 1200 pF (1 nF and 200 pF in parallel)
R1,R2	90k (22k and 68k, 1% in series)

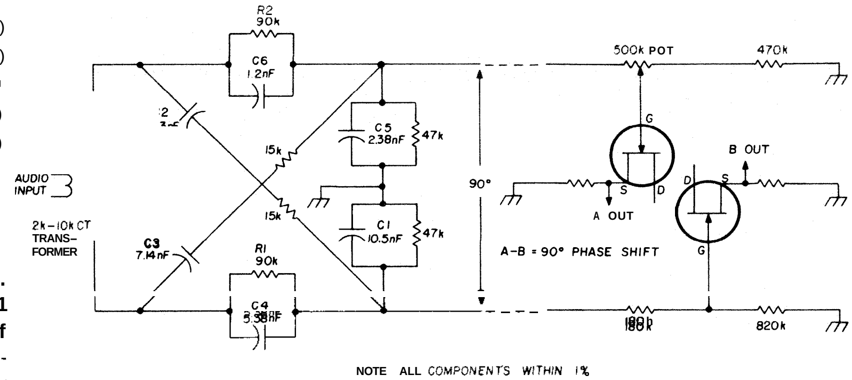


fig. 6. Modified Dome phase-shift network. Outputs may drive an impedance as low as 1 meg, which swamps the input impedance of the following stages (usually a source follower or similar circuit). Polycarbonate or polystyrene capacitors should be used for values above 1 nF (1000 pF); polystyrene, silver mica, or NPO ceramics below 1000 pF.

A similar circuit (fig. 4), having different values to accommodate a lower input impedance, was first popularized in the 'SSB Jr.," a phasing-type ssb transmitter designed and described by Don Norgaard, W2KUJ.[2] Bandwidth and differential phase shift characteristics are much the same as in the circuit of fig. 2, which has the distinct advantage of a minimum component count of 14. Standard E24-series 1% or 5%, silver mica, or NPO ceramic capacitors are preferred and are readily available.

The resistors may be 1% or 5% with values from the E96 series as indicated. Alternatively, they may be derived from 1% or 2% tolerance types from the E12 series. If E96-series resistors are used, the component count is 9, whereas if E12 series are used with values in series or parallel combination, the component count is 13 (including the 100-ohm pot). The pot can be a carbon or wirewound type. A carbon type with a Cermet element is preferred.

As for the circuit in fig. 2, the phase-shift network requires the X and Y input voltages to be in the ratio of 7:2. The pot is used to set this ratio. The circuit can be driven from a simple fet phase splitter with equal-amplitude outputs or by a transformer. A very high impedance must be presented to the outputs, and fet source followers are again recommended.

For maximum results the two networks just described (figs. 3 and 4) can be aligned by making a portion of each capacitance a trimmer, then the entire phase-shift network can be adjusted for minimum phase deviation (from 90°) over the frequency range. An audio oscillator and scope are necesssary. The procedure is described in reference 3. Details for aligning the Tucker Tin circuit are given in reference 1.

A detailed and very useful discussion on the design, construction, and alignment of audio phase-shift networks is given in reference 3. An alternative procedure is to measure a group of components on a precision bridge, selecting those within 1% of the values given in the circuit. The only adjustment then is to get the input-voltage ratios correct.

The Dome network4 shown in fig. 5 has the advantage that it can be driven directly from a balanced input such as a center-tapped transformer or phase splitter. However, as with the two previous circuits, the outputs must be presented to a very high impedance. If components are close to the values specified, deviation from 90° phase shift will be about ±1.5° between about 270 Hz and 2.9 kHz. Restricted audio bandwidth must be used, as discussed previously.

Another disadvantage of the circuit in fig. 5 is its high loss, which is about 15 dB compared with 10 dB

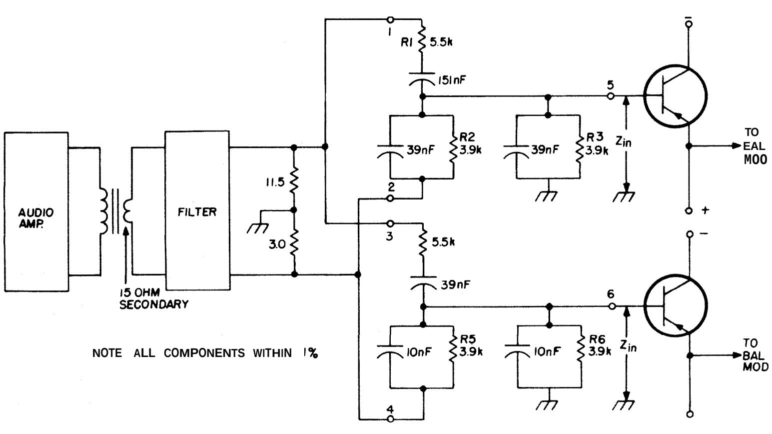


fig. 7. Network designed by Van Heddgem (reference 5) for solid-state applications. Values of R3 and R6 include Z_{in}. All values shown in the schematic should be within 1% or better. The ratio of R7/R8 = 3.83 (within1% or better).

for the previous two circuits. However, it's usually not too difficult to provide sufficient gain margin in the audio stages to compensate.

One of the disadvantages with all the circuits described thus far is the necessity for the outputs to be presented with a very high impedance. The phase shift is affected by load-impedance variations, with a consequent degradation in opposite-sideband suppression. If the output impedance can be defined, and a practical value lower than the input impedance of the following stage selected, then the effect of any variations in load impedance can be swamped.

Southwell[3] discusses a variation of the Dome network[4] in which the network outputs drive 1-megohm loads placed across the inputs of the following stages. The circuit is shown in **fig. 6.** Amplitude balance at the outputs may be obtained by a fixed voltage divider and by a pot in the other load resistance. The original circuit was designed to drive a cathode follower, whereas I've shown a source-follower stage.

Van Heddegem[5] discussed modifications to the basic Dome network in which a suitable network could be built for solid-state applications. See **fig. 7.** The input impedance of the following emitter-follower stages is taken into account when calculating R3 and R6. Note that input and output impedances are quite low compared with those in the previous circuits. Many standard component values may be used.

In contrast with the original Dome network, the circuit of **fig.** 7 requires that the input drive voltage be in the ratio of 3.83:1. The circuit was later used by Doyle[6] in his passive ssb generator. The ratio of the two input resistors, which determine the input-voltage ratios, must be accurate to within ±1% or better.

Input impedance is noncritical but should be low. Doyle[6] uses two 5k pots to adjust output levels and impedance for best opposite-sideband suppression. His circuit is shown in **fig. 8.** According to Van Heddegem[5] the network should work well between 280 Hz and 2.8 kHz. Audio must be restricted in bandwidth to maintain opposite sideband suppression. No numbers are given as to how close the phase shift remains at 90°. Component count is

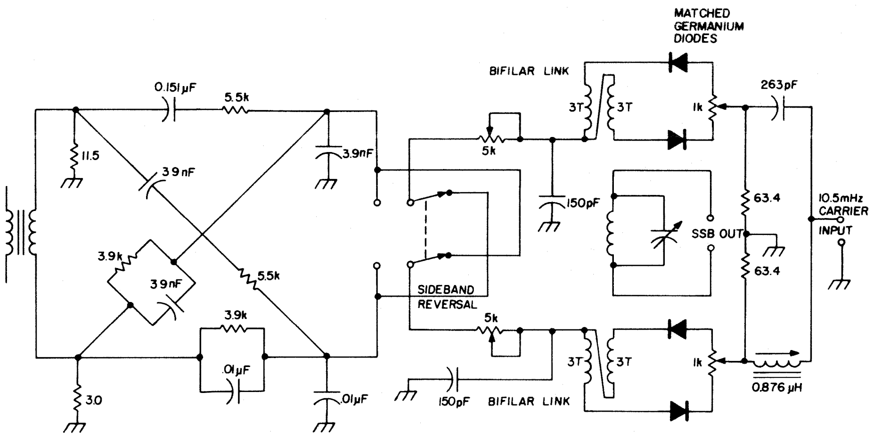
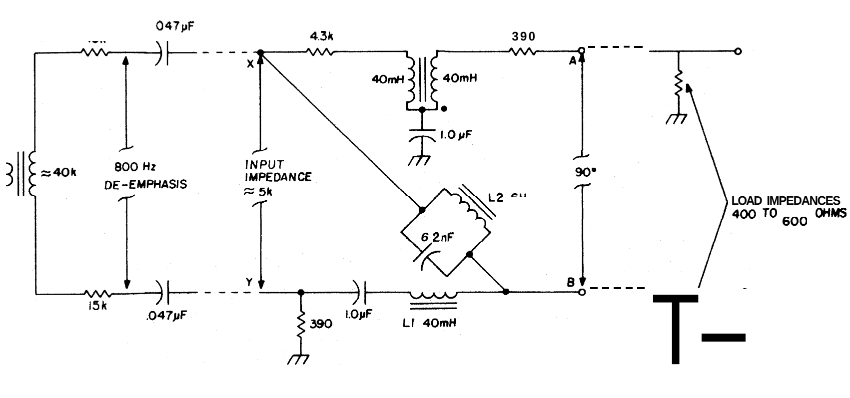


fig. 8. Passive phasing generator by W. Doyle, W7CMJ (reference 6).

fig. 9. Phase-shift network using RLC components developed by Westinghouse in 1944 and described by Cheek (reference 7). Circuit has a minimum of 9 components. The 40-mH inductors may be made from 88-mH toroids. The 6-henry inductor may be an ordinary iron-core choke. Alternatively, all inductors may be wound on pot cores or low-frequency toroid cores.



1 176 turns no. 26 (0.3mm) enameled wire on a single bobbin in a Vinkor LA2330 pot core

L2 2090 turns no. 42 (0.06mm) on a single bobbin in a Vinkor LA2330 pot core

T1 2 windings, 176 turns each, no. 34 (0.16mm) enameled wire, in each half of a double bobbin in a Vinkor LA2330 pot core

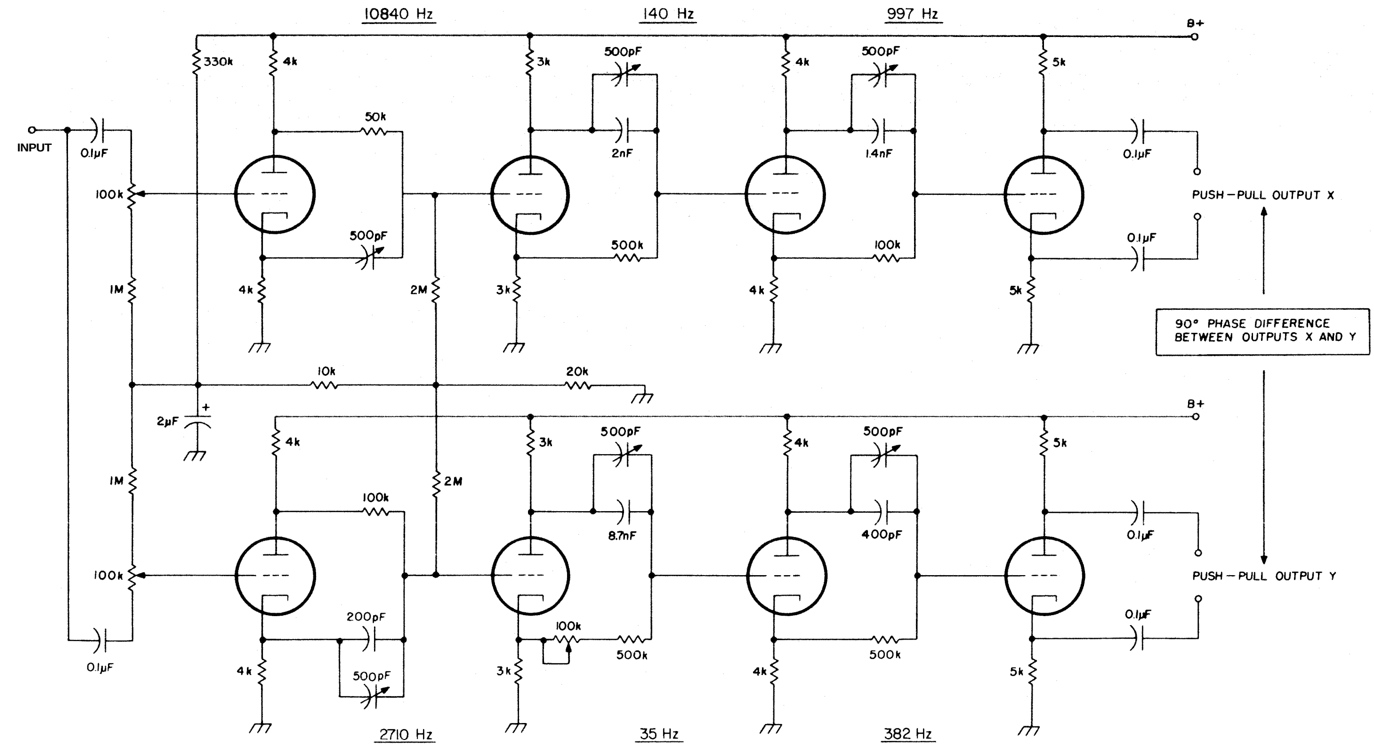


fig. 10. **Wideband** active audio phase-shift network described by Norgaard (reference 2) and Southwell (reference 3). Each RC network is adjusted **for 45**° phase shift, grid-to-grid, at the frequencies indicated. Circuit loss is about 8-10 dB.

only 12 if E96-series or selected E12-series components are used.

If you wish to build this ssb generator, I recommend that you read reference 3 for adjustment and alignment. (The rf phase-shift circuit is discussed later.) No numbers are given for opposite-sideband suppression, but it appears that at least 30 dB is obtainable.

So far all phase-shift networks considered have been made of RC combinations. Networks using R, L, and C combinations are quite rare in the literature, probably because the inductances in a phase-shift network of this type are not off-the-shelf items. Nevertheless, such a circuit has certain merits; one in particular is shown in **fig. 9.** This circuit was developed by Westinghouse in 1944 and subsequent-

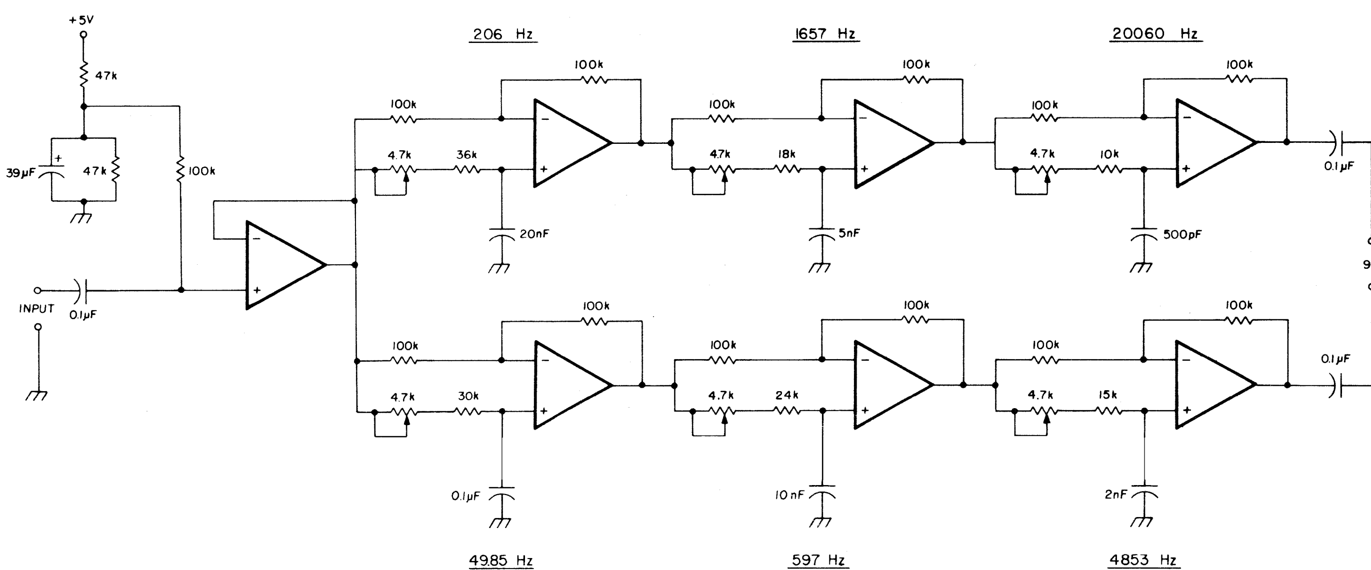


fig. 11. Modern **wideband** active phase-shift network described by Dickey (reference 8) uses two **LM324** quad op amps. Input circuit provides operation from a single 5-volt supply. Each stage is adjusted for 90° phase shift, input-to-output, at the frequencies indicated. Circuit has unity gain (no loss).

ly described by Cheek.[7] Components are noncritical. Resistors and capacitors can be standard 5% or 10% components. Composition resistors and paper capacitors were used in the original circuit. The main requirement is that each 40-mH inductor resonate with the 1-μF capacitor at 800 Hz. Exact values are not critical as long as components of the specified nominal values are used.

The 6-henry inductor and the 6200-pF capacitor must resonate at 800 Hz. The 40-mH inductors may be made from 88-mH toroids, which are readily available in the surplus outlets. These inductors consist of two 44-mH coils wound on a toroid core and connected in series.

Using a scope or vtvm and an audio oscillator, it's easy to resonate a 44-mH inductor and a 1-μF capacitor to 800 Hz. Just remove turns from the 44-mH winding until resonance is obtained. The exact frequency has no magic about it; 800 Hz is the geometric mean of 160 and 4000 Hz, which adequately covers the speech band. Ensure that each LC circuit resonates to the same frequency. This frequency could just as easily be 750 Hz (geometric mean of 200 and 2800 Hz) or 900 Hz (geometric mean of 270 and 3000 Hz).

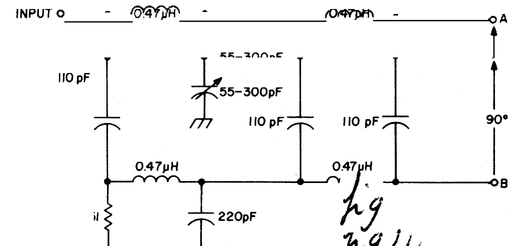


fig 13. Coaxial-cable quadrature phase shift network

Transformer T1 consists of two windings having equal numbers of turns wound on the same core resonated at 800 Hz with the 1-μF capacitor. The two windings are connected in series: dots in the circuit in **fig. 9** for T1 indicate the start (or finish) of each winding. As an alternative, each inductor could be wound on a standard pot-core assembly or a low-frequency toroid.

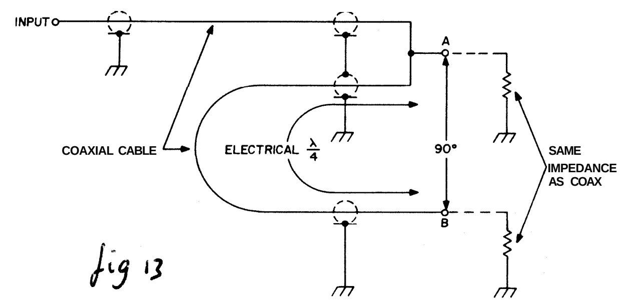


fig. 14. Quadrature phase-shift network after Taylor (reference 13) centered on 14.25 MHz. Output amplitudes are within 0.8 dB between 13.8-14.6 MHz.

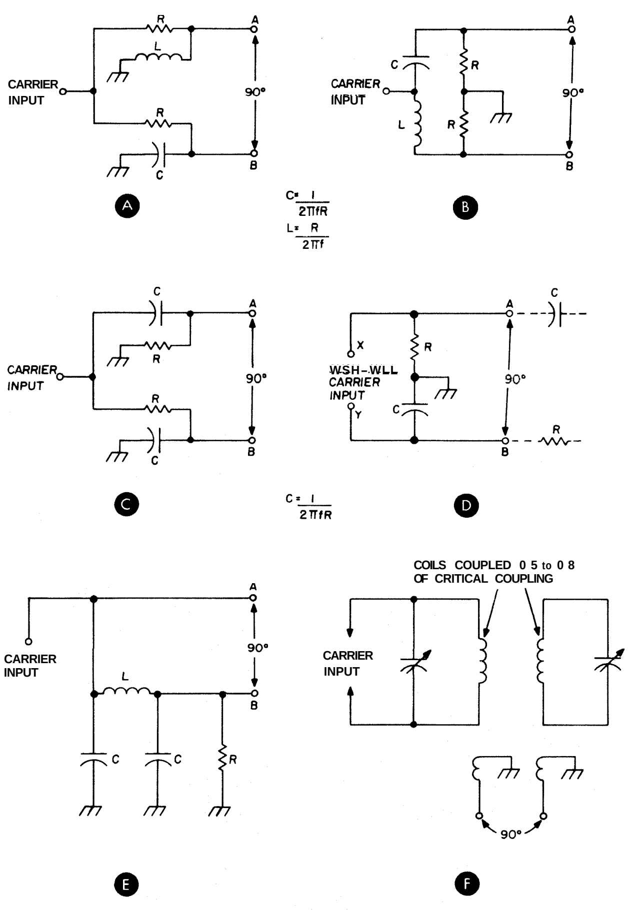
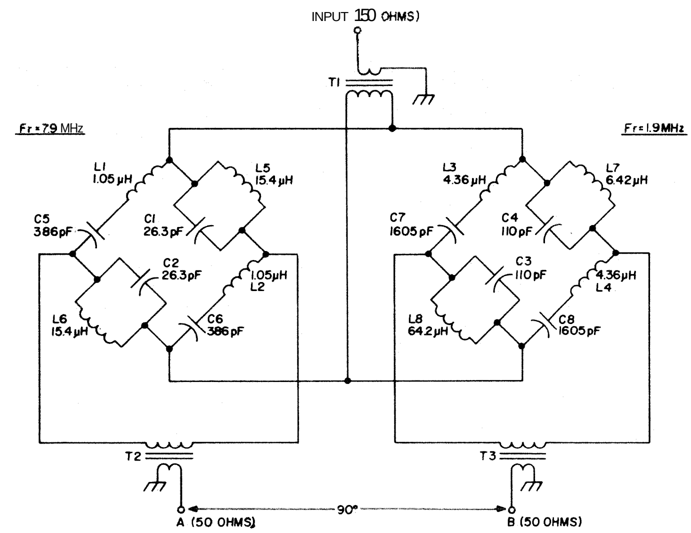


fig. 12. Passive rf phase-shift networks commonly used in phasing ssb designs over the past 30 years, which are discussed in the text. All are suitable only for single-frequency use or for a very small frequency range.

The two quadrature outputs, A and B, can drive low-impedance loads. The characteristics of the phase-shift network are unaffected by the load impedance, which may be between 400 and 600 ohms. Input impedance is about 5k and should be floating with respect to ground. The input should be driven by a transformer or a differential amplifier.

The speech amplifier preceding the phase-shift network should include deemphasis for frequencies below 800 Hz. If the network is transformer driven, a deemphasis network consisting of two 15k resistors and two 47 nF (= .047 μF) capacitors, connected in series with each input terminal, serves this purpose. The input impedance then will increase to about 40k and the input transformer should be selected to drive such an impedance. This is suggested by Cheek.[7]

The network will maintain the phase shift within 1° or better between 300 Hz and 3.5 kHz. The amplitude balance between the quadrature outputs is within 2% or better between 200 Hz and 4 kHz. Thus, it's easy to achieve an opposite-sideband suppression of about 40 dB, which is certainly one of the advantages of this particular circuit. That and the quite low



C1,C2	26.3 pF (27 pF, 5% NPO ceramic or silver mica)
C3,C4	110 pF, 5% NPO ceramic or silver mica
C5,C6	390 pF, 5% NPO ceramic or silver mica
C7,C8	1605 pF (2700 pF and 3900 pF, 5% polyfilm capacitors in series)
L1,L2	1.05 μH. 5 turns no. 26 (0.4mm) enameled, closewound on Philips 020-91010 toroid core
L3,L4	4.36 μH. 12 turns no. 26 (0.4mm) enameled wound on Philips 020-91010 toroid core, turns spread evenly around circumference of core
L5,L6	15.4 μH. 24 turns no. 26 (0.4mm) enameled, wound on Philips 020-91010 toroid core
L7,L8	64.2 μH. 48 turns no. 30 (0.25mm) enameled, wound on Philips 020-91010 toroid core
T1,T2 T3	See text

fig. 15. Wideband 90° phase-shift network using two 45° bridge circuits (courtesy of Jim Koehler, VE5FP/VK2BOV).

input and output impedances together with the relatively noncritical nature of the components, gives this circuit quite an edge on the RC circuits discussed.

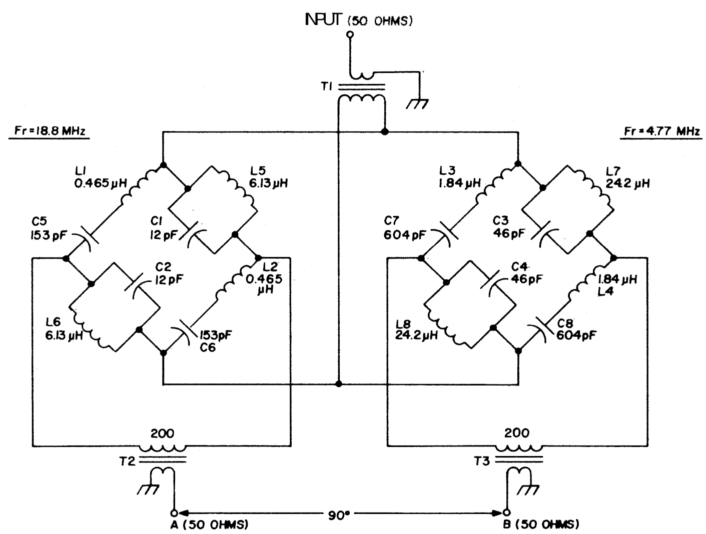
The overall loss is 12-14 dB (excluding the de-emphasis circuit), which is comparable to the RC circuits. The audio stages preceding the network must have a sharp cutoff above 3 kHz, a common feature of all networks discussed. The minimum component count of 9 is also an attractive feature and is competitive with the circuit in **fig. 4.**

A point worth noting is that the low output impedance of the network in **fig. 9** makes it suitable for driving low-impedance diode bridge balanced modulators.

Active audio phase-shift networks were used in many early designs for ssb exciters. These circuits generally consisted of a cascaded series of triode phase splitters with RC networks coupling each stage as shown in **fig. 10.** This circuit is discussed by both Sothwell[3] and Norgaard,[2] amongst others. Each stage produces a 45° phase shift at a particular frequency. The frequency of each RC network is chosen so that the entire network produces a differential phase shift within ±1° between 70 Hz and 5.5 kHz.

This type of network needs alignment, but the procedure is more complicated to explain than to accomplish and is not outside the expertise of most amateurs. All you require is a passing acquaintance with a scope and an audio oscillator. Once adjusted, the network will maintain its alignment for considerable periods. The wide bandwidth allows good opposite-sideband suppression over the speech bandwidth of 300 Hz to 3 kHz. The circuit has excellent phase and amplitude stability. Overall loss is about 8-10 dB. The circuit can obviously be adapted to use modern fets.

A more recent circuit, using two quad op-amps, was described by Dickey.[8] He claims this circuit will provide two equal-amplitude outputs that differ in phase by 90° within ±2° over the frequency range



C1,C2	12 pF, 5% NPO ceramic or silver mica
C3,C4	46 pF (47 pF, 5% NPO ceramic or silver mica)
C5,C6	153 pF (150 pF, 5% NPO ceramic or silver mica)
C7,C8	604 pF (680 pF, 5% NPO ceramic or silver mica in series with 5600 pF, 5% polyfilm capacitor)
L1,L2	0.465 μH. 5 to 6 turns no. 22 (0.6mm) enameled on 579x250x312/900 Neosid toroid, turns spread evenly around circumference
L3,L4	1.84 pH. 6 turns no. 30 (0.25mm) enameled wire, closewound on Philips 020-91010 toroid core
L5,L6	6.13 pH. 12 turns no. 26 (0.4mm) enameled, wound on Philips 020-91010 toroid core, turns spread around 213 the circumference
L7,L8	24.2 μH. 27 turns no. 30 (0.25mm) enameled, wound on Philips 020-91010 toroid core
T1,T2, T3	Wound on Neosid 1050-1-F14 of Indiana General F684-1 balun core. Twist together three 7″ (180mm) lengths of no. 26 (0.4mm) enameled wire and wind 3 turns through 2 holes; connect two wires in series for the 200-ohm winding

fig. 16. 3-30 MHz quadrature phase-shift network. Maximum phase error is about 1°. Overall loss of this network and that of fig. 15 is about 6 dB.

100 Hz to 10 kHz. High-fidelity ssb! The circuit is shown in **fig. 11.** Two LM324 quad op-amps are used. Each stage is adjusted, using the 47k trimpot, to produce a 90° phase shift at the frequencies shown. The design values were calculated from data published by S. D. Bedrosion,[9] if you're interested in getting into heavy phase-shift network design. The circuit in **fig. 11** has the advantage of having minimal loss.

Alignment techniques for both circuits in **figs. 10** and **11** involve the use of an audio oscillator and a scope, as mentioned above, a phase meter, or a network analyzer. A technique using the oscillator and scope is discussed in detail in the references.

For further reading the article by Wade[10] is recommended. An excellent description of simple methods for aligning phasing-type ssb exciters is given by Fred Johnson.'

rf phase-shift networks

Again, both active and passive phase-shift network designs are available. The networks commonly used in phasing ssb designs over the past 30 years are illustrated in **fig. 12.** Popularity seems evenly divided among the various circuits with the exception of E, the pi network. The RLG circuit in B is simply a variation of that in A. Input or output impedances are a consideration in all cases; the component values are dimensioned to accommodate external circuit conditions.

The networks in **figs. 12 A, B, C** and **D** are quite simple to set up. Usually, one of the components is made variable to provide phase adjustment for final

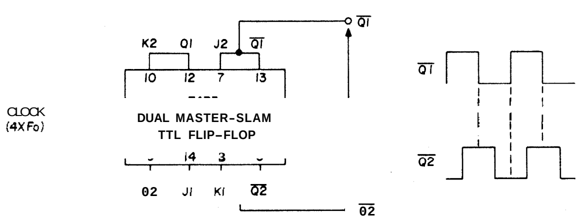


fig. 17. The 7473 IC connected to produce quadrature square waves.

circuit trimming. Extra components may be added to account for circuit strays as necessary.

The first two circuits exhibit an output phase characteristic that does not vary with frequency, but the relative amplitudes vary markedly either side of the design frequency. The circuits in C and D also exhibit the same characteristics, but the impedance varies also. The network in C exhibits least variation in this respect. Alternatively, another capacitor and resistor may be inserted in series with output terminals A and B respectively in the circuit of **fig. 12D** to reduce output impedance variation.

The technique of using two under-coupled tuned circuits, as in **fig. 12F**, was first popularized in reference 11 in 1950 and has been used in several phasing ssb transmitter designs since then. In practice, the two tuned circuits are coupled by 50% and 80% of critical coupling, and the links are adjusted to achieve equal-amplitude output. Secondary tuning is adjusted to provide the correct phase shift between the outputs. A very complete discussion on theory and practical considerations of this technique is given by R. W. Martin, VK2AHI.[12]

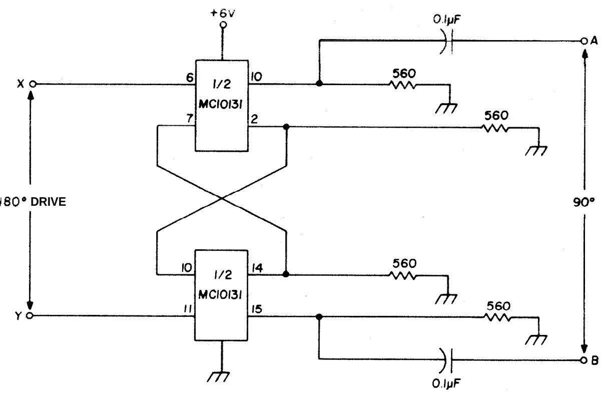


fig. 18. Digital quadrature-phase rf circuit given by Shubert, WAØJYK (reference15).

All circuits shown in **fig. 12** suffer from three disadvantages:

A. They work only over a narrow range of frequencies and are thus limited to fixed-frequency applications or operation over bandwidths of 100 to 200 kHz at best.

B. All need periodic realignment as they are subject to drift due to environment (temperature, etc.) and component aging.

C. The practical upper-frequency limit is about 15 MHz at best, depending mostly on strays, external-circuit conditions, and component performance. In any case, using these circuits above 10 MHz is not recommended.

A technique that has occasional mention in the literature, but which I've not yet seen applied, is the use of coax cable as a phase-shift element in an ssb exciter. An electrical quarter wavelength of coax exhibits a phase shift of 90° at the design frequency and will remain within ±1° of this amount over a small bandwidth. Coax cables are relatively unaffected by temperature changes that would cause marked changes in the circuits discussed so far. The low impedance is an advantage in some instances (e.g., where diode-ring balanced modulators are used). Amplitude differences between the two outputs are not a consideration. This technique is extensively used in antenna phasing applications.

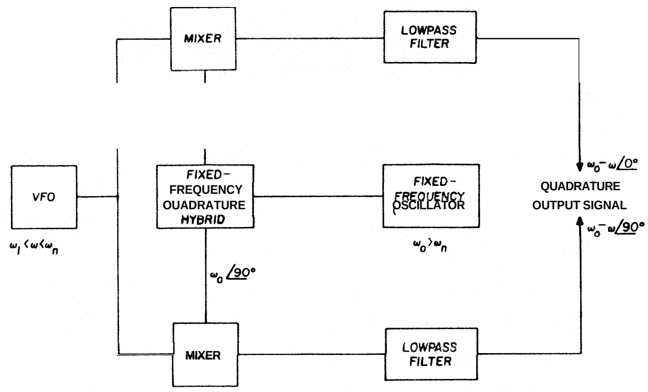


fig. 19. Quadrature signals can be generated over a wide band by adapting the third method of ssb-signal generation.

A coax-cable quadrature phase shifter is shown in **fig. 13.** No real theoretical upper frequency limit exists for this method, but practical limitations put it at about 150 MHz. A coaxial line-stretcher would be useful here. A coax-cable quadrature phase-shift network has the same disadvantages as those of the circuits in **fig. 12,** only slightly less so. An electrical quarter wavelength of coax on 10 MHz is nearly 16.5 feet (5m) long. This technique is probably best for fixed-frequency or narrowband use above about 15 MHz. Small-diameter cables, such as Microdot, of a suitable impedance are best as they are less bulky than standard cables such as RG-58/U.

Passive wideband rf phase-shift networks for application in phasing ssb exciters are rare in the literature. Richard Taylor, W1DAX, described a circuit in the Septmember, 1969, issue of *QST*[13] used in a direct-conversion ssb receiver for 14 MHz. The article was reprinted in the ARRL's Single Sideband for the Radio *Amateur,* fifth edition, 1970.

This circuit, **fig. 14,** will maintain 90° phase-shift and output amplitudes within 0.8 dB between 13.8 and 14.6 MHz. A bandwidth of 800 KHz isn't exactly wideband, but is certainly much better than the 100-200 kHz bandwidth of the circuits in **fig. 12.** The trimmer provides phase adjustment. As mentioned, Taylor[13] used this circuit in a direct conversion receiver, but it could be used in a transmitter as well. You could generate ssb signals directly on the desired output frequency rather than on a fixed frequency, which requires heterodyning to the desired output frequency, as in common practice.

Real advantages exist when generating ssb signals on the desired output frequency. The only spurs to contend with are those associated with opposite-sideband suppression and with intermodulation distortion, both of which must be considered in any heterodyning system.

Then there's the simplicity of the circuitry. A major push behind the development of modern IC circuits is the simplicity of the following circuitry; therefore, circuit simplicity is certainly an advantage. Circuit complexity isn't necessarily synonymous with sophistication or "the state of the art."

Quadrature rf phase-shift networks that operate over an octave or more in frequency were described many years ago. However, you must search the literature on antennas and circuit theory to find them.

The network in **fig. 15** is through the courtesy of Jim Koehlor, VE5FP/VK2BOX, who designed it for a circularly polarized antenna system. Two bridge networks each provide 45° phase-shift between 1 and 15 MHz, resulting in a differential phase shift of 90° over that range. Phase error is less than 1°, and the amplitude differences between outputs is less than 0.5 dB over the range. This rf phase-shift network

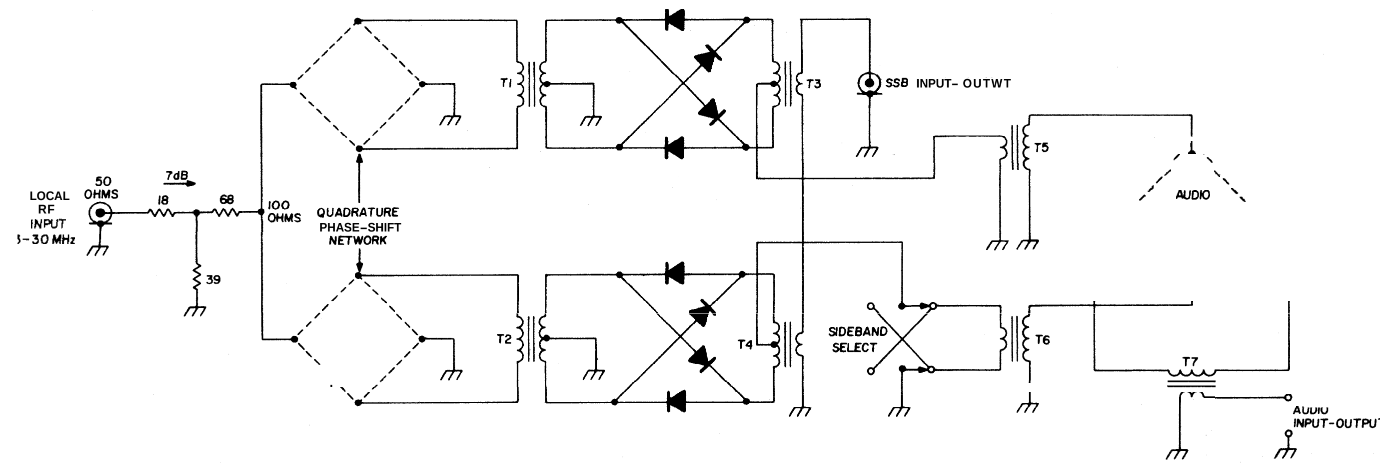


fig. 20. Suggested bilaterial direct-conversion phasing ssb generator-detector. T1-T4 are wideband rf transformers. T5-T7 are audio transformers to suit the audio phase-shift network used. A wideband rf transformer may be used instead of the 7-dB pad.

makes direct-conversion phasing ssb generation possible and has application in direct-conversion receivers. Third-method ssb generation, with output directly on any desired frequency between 1 and 15 MHz, is also a possiblity.

A network designed to cover 3 to 30 MHz is shown in **fig. 16.** It has characteristics similar to those of **fig. 15.** Input and output impedances of each bridge in both networks is 200 ohms. Transformers T2 and T3 transform the impedance to 50 ohms, which is convenient.

Although the inputs of each bridge are in parallel, making the input impedance 100 ohms, T1 may be the same as T2 and T3, as the mismatch has no serious effect on network performance. The three transformers are constructed as wideband baluns having a turns ratio of 2:1. Small toroids or dual-hole balun cores, such as the Neosid 1050/1/F14 or Indiana General F684-1, are suitable. The input and output windings must be isolated. To use dual-hole balun core, twist together three 7-inch (180mm) lengths of 26 or 30 B&S or AWG (0.3 or 0.25mm) enameled copper wire at about two twists per 3/8 inch (10m). Wind three turns through the two holes and connect two of the wires in series to make the 200-ohm winding. If desired, the secondaries of T2 and T3 may be arranged to drive diode-ring balanced modulators directly.

It's important that coupling between the tuned circuits in each arm of the bridge, and between each bridge, be kept to a minimum. Also, the *Q* of each coil must be at least above 50 or 60. Consequently, toroids have been suggested, although standard coil-former and screened-can assemblies (with ferrite cup cores) have been used successfully. Each arm is constructed individually and the inductor adjusted to resonate with the capacitor at the frequency indicated. Each series arm is temporarily connected as a parallel-tuned circuit to enable adjustment. This is very simply done with grid-dipper and a monitoring receiver. Sufficient accuracy is easily obtained. Of course, if you have a network analyzer or phase meter, the job is a little simpler.

Wideband active rf phase-shift networks involve digital techniques. This technique involves cross-coupled JK flip-flops and was described by A. J. Turner.[14] The circuit is shown in **fig. 17.** The upper frequency of such circuits is limited by the phase jitter between the two outputs and is somewhat below the upper clock speed limit of the device used. The clock frequency of the circuit shown in **fig. 17** must be four times the desired output frequency.

A circuit that requires a clock frequency only twice the desired output frequency is presented in **fig. 18.** This circuit is by G. K. Shubert.[15]

The disadvantage of the digital technique is the nonsinusoidal output waveform and the attendant harmonics that must be removed. Although these may be reduced with simple low-pass filters, extra spurs are undesirable. The digital technique has the big advantage of requiring no adjustment.

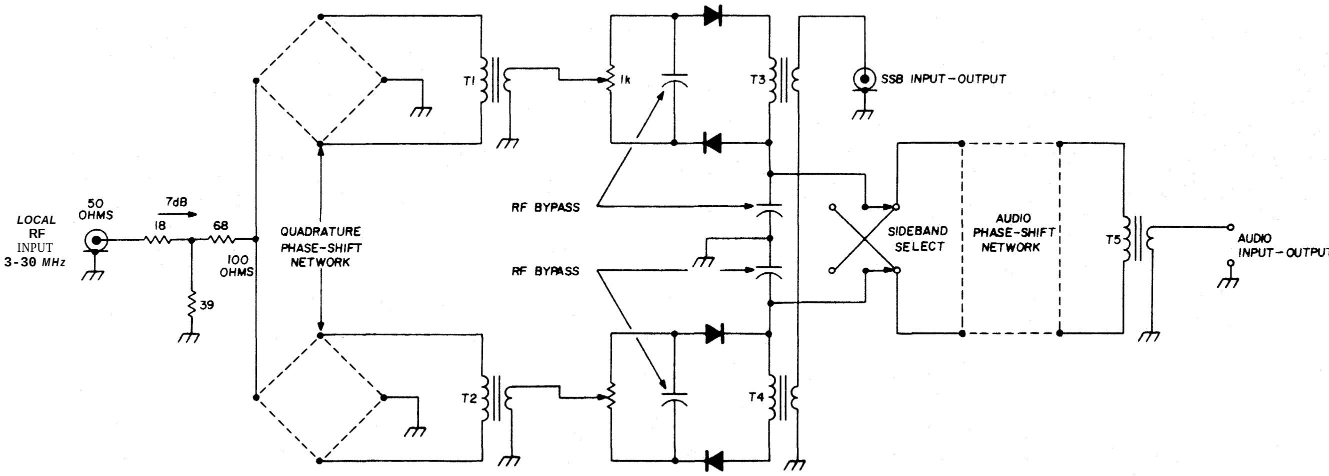


fig. 21. Another suggested bilateral direct-conversion phasing ssb generator. T1-T4 are wideband rf transformers. T5 is an audio transformer chosen for the phase-shift network used. As in the circuit of fig. 20, the 7-dB pad may be replaced by a wideband rf transformer.

Another technique for producing broadband quadrature rf signals, adapted from third-method ssb generation, is suggested by Taylor.[13] A block diagram, **fig. 19,** illustrates this. However, its relative complexity puts this technique at a disadvantage.

The networks in **figs. 15** and **16** and the circuits in **figs. 17** and **18** may be used for direct-conversion generation or reception of ssb signals using either the phasing method or the third-method as already mentioned. Indeed, it should be possible to build a passive phasing exciter using a combination of the techniques discussed. The third method produces superior performance with regard to opposite side-band and carrier suppression than either the phasing or filter techniques.

Fig. 20 shows a suggested bilateral direct-conversion phasing-type ssb generator/detector. It may be possible to use all-passive techniques. The audio phase-shift network may exhibit too much loss for successful operation and the bilateral feature of the circuit may be impossible to realize. T1, T2, T3, and T4 are wideband rf transformers as suggested previously. T5, T6, and T7 are audio transformers to suit the audio phase-shift network. A 7-dB resistive pad may be used to isolate the local rf input. Alternatively, a wideband transformer may be substituted.

Fig. 21 is a somewhat simpler circuit using series-bridge-diode balanced modulators instead of the ring-diode balanced modulators. Comments similar to those for **fig. 20** apply. Performance may not be quite as good as the previous circuit, but the simplicity may be an advantage. The phasing of the secondaries of T3 and T4 in both circuits is important.

references

1. Fred Johnson, ZL2AMJ, "The Tucker Tin Mark II, An SSB Transmitter of Simple Design," *Break-in* (New Zealand Association of Radio Transmitters Journal), August, 1971, pages 251-272.
2. Don Norgaard, W2KUJ, "Versatile Single Sideband Exciter," *CQ,* March, 1949, page 34.
3. N. Southweil, VK2ZF, "Wideband Audio Phase-Shift Networks." *Amateur Radio* (Wireless Institute of Australia Journal), June, 1955, pages 2-5 and July, 1955, pages 2, 3, and 12.
4. R. B. Dome, W2WAM, "Wideband Phase-Shift Networks," *Electronics,* December, 1946, page 112.
5. W. van Heddgem, "Audio Phase-Shift Network for Transistorized SSB Transmitters and Receivers." *QST,* December, 1964, page 27.
6. W. Doyle, WCMJ, "Phasing-Type SSB Generator," *ham radio,* April, 1973, pages 22-25.
7. Robert C. Cheek, W3LOE, "Single Sideband for Everyone," *CQ,* November, 1948, page 17.
8. Richard K. Dickey, "Outputs of Op-Amp Networks Have Fixed Phase Difference," *Electronics,* August 21, 1975.
9. S. D. Bedrosion, "Normalized Design of 90° Phase Difference Networks," *IRE Transactions (Circuit Theory),* June, 1960, pages 128-136.
10. A.I.H. Wade, G3RNW, "Design Criteria for SSB Phase-Shift Networks," *ham radio,* June, 1970, page 34.
11. "SSB Junior," *GE Ham News,* November - December, 1950.
12. R. W. Martin, VK2AHI, "The Coupled Tuned Circuit RF Phase-Shift Network," *Amateur Radio* (Wireless Institute of Australia Journal), September, 1967, pages 7-10.
13. Richard S. Taylor, "A Direct-Conversion SSB Receiver," *QST,* September, 1969. Reprinted in *Single Sideband for the Radio Amateur,* ARRL, fifth edition. 1970.
14. A. J. Turner, G3UFP, "Single-Sideband Suppressed Carrier Generation," *Wireless World* (England), September, 1973, pages 453-455.
15. G. K. Shubert, WA0JYK, "Phasing-Type Single Sideband Transmitter," *ham radio,* June, 1975, pages 8-22.

ham radio

RTTY test generator

Construction details for the *Digiratt* RIY generator that can be used with the *Digiratt* afsk generator or as a stand-alone accessory

The Digiratt R/Y test generator is a companion unit to the original Digiratt precision AFSK generator and phase-locked-loop terminal unit featured in an earlier issue of *ham radio* magazine (September, 1977); it can be used with that unit or as a stand-alone accessory for RTTY enthusiasts.

While it would be simpler, from a design standpoint, to use a PROM (Programmable Read Only Memory) as the heart of the unit, I decided not to go this route for two reasons. First, few hams possess the required equipment necessary to program the PROM. Second, few people have the patience to do the programming. I also thought that more people would be interested in building a unit which used readily obtainable ICs. Those who do build the complete unit will have a non-mechanical device which generates 64 RYs, a carriage return, and a line feed.

The Digiratt R/Y generator is an eleven IC, TTL-based device for generating the 5-level Baudot code. It has on-board encoding for automatic sequential generating of the Baudot code necessary to print the letters R and Y. Additionally, encoding is provided for carriage return and line feed code generation. Logic is provided which keeps track of the number of characters printed and steers the output port to select either the RY message, carriage return (CR) or line feed (LF) code.

The unit was designed in such a way that by constructing only that portion of the schematic **(fig. 1)** enclosed by the dashed line, an RY generator only, can be built which deletes the CR and LF provisions. If this is done, the unit will print RYs continuously without regard to line length. Additionally, provision is made for "normal" and "inverted" output data to key transmitters with either mark high or space high signals. Finally, the unit is designed to operate at slightly less than the full 60 wpm. Older ***Teletype*** machines and those slightly out of adjustment should be able to copy the test message with little difficulty.

shift registers

Before examining the details of the schematic diagram **(fig. 1)**, the basic operation of a shift register should be understood. The SN74165 registers used in this design are capable of changing an 8-bit parallel data bus into a serial stream of pulses. The parallel information is first loaded into the registers by the application of the load data pulse. Next, for each clock pulse that is received the bit pattern is serially shifted one register to the right. In this way, at the end of eight clock pulses the entire 8-bit data pattern is now in a serial form. By hardwiring the parallel input ports to a known pattern, a specific character, in this case CR and LF, can be generated. The hardwired pins, from right to left, or the first to last bit out are, 6, 5, 4, 3, 14, 13, 12, and 11.

The Baudot code used for ***Teletype*** is composed of 5 either mark or space conditions. Different combinations of the marks and spaces represent the actual letters, symbols, and functions. In addition to the first 5 bits, a start mark precedes the actual information. Finally, a stop pulse is used to indicate the end of the character. For the RY test generator I've combined the last two bits from the shift register into a

By John Loughmiller, WB9ATW, Route 1, Box 480C, Borden, Indiana 47106

slightly longer than normal stop pulse, 44 vs 31 ms. While this will slightly reduce the speed of the machine, it insures that older machines will print correctly.

circuit description

U1 is the clock pulse generator and, as such, is the heart of the system. The output from pin 3 provides a pulse train with a 45.45 hertz rate which is applied to the clock inputs of shift registers U4, U9, and U10. These pulses are also applied to pin 4 of U2 which divides the rate by a factor of eight (5.68). The divided signal then becomes the load data pulse for the shift registers.

To generate the required pulse configuration for the CR and LF functions, the eight parallel inputs of the shift register are hardwired to either 1 or 0. Initially, the load data pulse loads this hardwired information into the shift register. Each clock pulse then shifts the information one position. On the eighth pulse, new information is again entered into the shift register. Every eight pulses you will have a complete bit pattern available for use.

In addition to functioning as the load data pulse, the 5.68 hertz pulse rate is inverted and then used to clock U5, U6, and U7. U5 is a J-K flip-flop that will alternately change the hardwired pattern of the RY shift register. In this way, the shift register will produce an R and then a Y as the IC toggles. US and U7 are wired to divide by the fixed rate of 66. And, with the addition of U8, are the basis for producing the 32 RYs and the CR, LF on each line.

With the count initially at 66, a CR is generated, count 65 produces a LF with RYs being produced on the rest of the counts. At the end of the count cycle, the counters are preset to 66 and then start decrementing again. The actual selection of the RY, CR and LF bit pattern is done by a 4:1 multiplexer, U11. The signals from U8 determine which pattern is selected as the ultimate output.

Switch S1 is used as a manual reset to ensure that each line starts with an R. Normally, the reset is performed after the counter decrements down from 66. Holding the switch closed will result in continuous CRs being sent. This technique is not recommended if 74LS series ICs are used.

construction

The construction of the RY test generator is non-critical. The only special precaution that should be

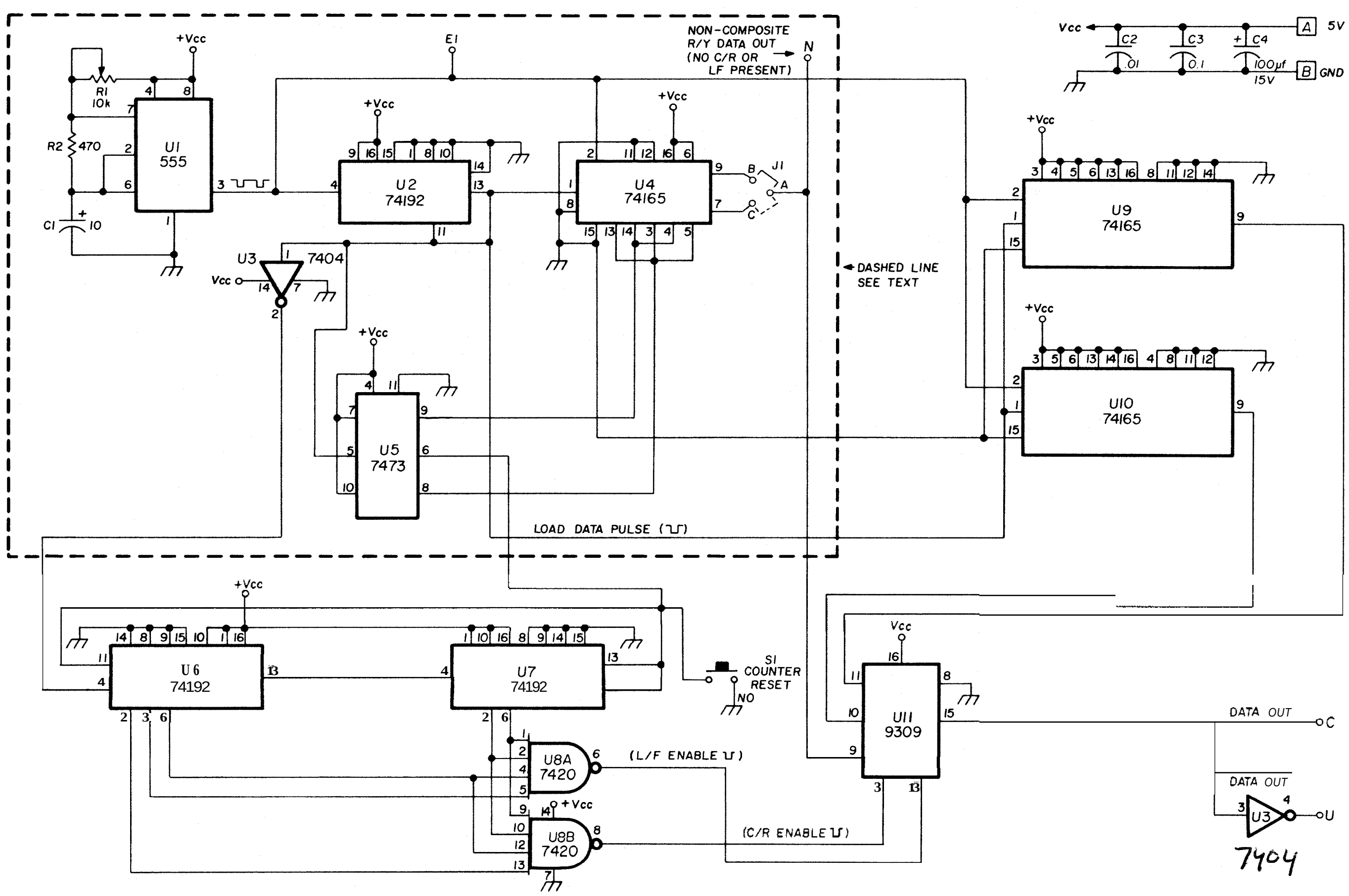


fig. 1. Schematic diagram of the RY test generator. The position of the jumper will permit you to select either inverted or normal RY information. A suitable loop keyer is shown in fig. 1, ham radio, September, 1977, page 27.

9309 -> 74153
pin 3 -> 2 9 -> 3
13 -> 14 10 -> 4
11 -> 5

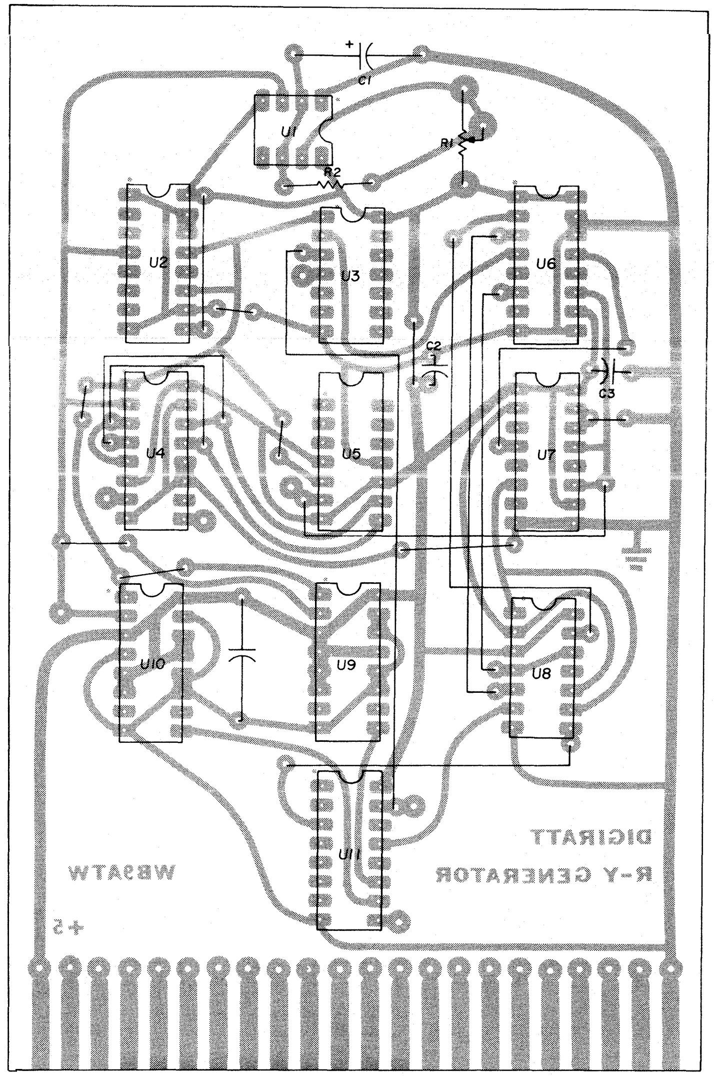


fig. 2. Component placement diagram for the board supplied by Circuit Specialists.

observed is to install the V_{CC} bypass capacitors on the board with the ICs **(fig. 2**). In extreme cases of rf interference, you may have to install additional bypass capacitors on the ICs. The power supply can be based on the popular LM309 with adequate heat sinking.

The first portion that should be assembled is indicated by the dotted line in **fig. 1.*** This portion will send a continuous stream of RYs. R1 should be adjusted to have the oscillator running at 45.45 Hz. If a counter is not available, R1 can be adjusted until the machine starts to print correctly.

*A drilled and plated circuit board is available from Circuit Board Specialists, Box 969, Pueblo, Colorado 81002 for $6.50. A complete kit of parts is also available from Circuit Specialists for $22.00.

ham radio

microwave bibliography

If you're interested in trying microwaves, but don't know where to start, this bibliography of amateur microwave articles will get you off on the right foot

Many amateurs have expressed an interest in microwave communications, but the techniques at microwave frequencies are considerably different than those used at vhf and uhf, so many amateurs don't know where to start. The following bibliography was prepared with this in mind. Far more good microwave information is documented in the amateur radio publications than is generally realized. Also included is a number of excellent books on the subject, as well as a short list of articles in other publications which are especially useful to the amateur microwave enthusiast; these publications can often be found in a local library.

If you're looking for the maximum amount of microwave information in the least amount of spac obtain a copy of the 3rd edition of the RSGB': *VHF/UHF Manual* and read Chapter 3 — it contains more amateur microwave data per page than any other single publication.

In the field of microwave textbooks, there are a great many which are of limited use to amateurs; those books are not listed in the following bibliography. Microwave books which are listed in the bibliography were chosen because they had something to offer to the amateur microwave enthusiast. Lance's *Microwave Measurements,* for example, is an excellent introduction to microwave techniques for those readers who are looking for the non-mathematical approach. *Microwave Transmission Design Data* is a paperbound reference which covers many microwave subjects but is especially valuable for its explanation of circular waveguides. *Very High Frequency Techniques* is a compilation of a number of experiments from which the amateur can obtain cavity design information and practical transmission line information for new designs. The book, *Principles and Applications of Waveguide Transmissions,* provides excellent coverage of transmission lines and conical antenna design.

Microwaves are really simple, when you get to know them, and microwaves are far superior to the lower frequencies for line-of-sight point-to-point communications. Microwaves are also an experimenter's paradise, and far less expensive than 432 MHz was fifteen years ago — and much more satisfying. Microwave is one area where amateurs can still contribute to the art of radio communications — the following bibliography will head you in the right direction.

By Cliff Buttschardt, W6HDO, 1488 Warner Street, Chico, California 95926

ham radio magazine

Adams, F. E., W6BPK, "Pogo Stick for Reflex Klystrons," *ham radio,* July, 1973, page 71.

Fisk, J. R., W1DTY, "Simple Untuned Crystal Mount," *ham radio,* June, 1968, page 68.

Fisk, J. R., W1HR, "Solid-State Microwave RF Power Generators," *ham radio,* April, 1977, page 10.

Foot, N. J., WA9HUV, "Narrow-band Solid-State 2304-MHz Preamplifiers," *ham radio,* July, 1974, page 6.

Foot, N. J., WA9HUV, "2304-MHz Power Amplifier," *ham radio,* February, 1975, page 8.

Franke, J. M., WA4WDL, "Marker Generator for 3cm Amateur Band," *ham radio,* June 1969, page 69.

Franke, J. M., WA4WDL, "Finding the Focal Length of Surplus Microwave Dish Antennas," *ham radio,* March, 1974, page 57.

Franke, J. M., WA4WDL, "Waveguide Klystron Cooler," *ham radio,* October, 1974, page 64.

Franke, J. M., WA4WDL, "Dielectric Antenna for 10 GHz," *ham radio,* May, 1975, page 79.

Franke, J. M., WA4WDL, "Microwave Frequency Doubler," *ham radio,* March, 1976, page 69.

Franke, J. M., WA4WDL, and Cohen, N. V., WB4LJM, "Simple 10-GHz Antenna," *ham radio,* May, 1977, page 40.

Jackson, H., K7PMY, "New DX Record for 40,000 MHz," *ham radio,* December, 1968, page 70.

Kolbly, R. B., K6HIJ, "Standards for Amateur Microwave Communication," *ham radio,* January, 1969, page 54.

Kolbly, R. B., K6HIJ, "Low-Cost Microwave Antenna," *ham radio,* November, 1969, page 52.

Kowhler, J. A., VE5FP, "Proportional Temperature Control for Crystal Ovens," *ham radio,* March, 1972, page 16.

Laakman, P., WB6IOM, "A Large Homebrew Parabolic Reflector," *ham radio,* August, 1969, page 8.

Roubal, W. T., "Getting Started in Microwaves," *ham radio,* June, 1972, page 53.

Stanton, W., K2JNG, and Moser, D., WA2LTM, "Solid-State 2304-MHz Converter," *ham radio,* March, 1972, page 16.

Tillitson, G., K6MBL, "Microwave Polaplexer Design," *ham radio,* March, 1977, page 40.

Vilardi, D., WA2VTR, "Solid-state 2304-MHz Preamplifier," *ham radio,* August, 1972, page 20.

Wade, P. C., WA2ZZF, "High Performance Balanced Mixer for 2304 MHz," *ham radio,* October, 1975, page 58.

QST

Badger, G., W6RXW, "An Introduction to the Klystron," *QST,* August, 1961, page 11.

Baird, W. H., W6VIX, "A Radio Club for Microwave Enthusiasts," *QST,* December, 1957, page 45.

Biancomano, V., WB2EZG, "A Prototype Pulse-Code Modulation System," *QST,* January, 1977, page 24.

Bray, D. W., K2LMG, and Kirchuwer, P. H., W2YBP, "Antenna Patterns from the Sun," *QST,* July, 1960, page 11.

Bredon, A. D., W6BGK, "Let's go Microwave," *QST,* June, 1958, page 11.

Fisher, R. E., W2CQH, "Interdigital Converters for 1296 and 2304 MHz," *QST,* January, 1974, page 11.

Goodman, B., W1DX, "A 2400-Mc Oscillator Cavity," *QST,* October, 1947, page 65.

Guba, R. F., WIQMN, "Pulse: A Practical Technique for Amateur Microwave Work, Part 1," *QST,* February, 1963, page 11.

Guba, R. F., WIQMN, "Pulse: A Practical Technique for Amateur Microwave Work, Part 2," *QST,* March, 1963, page 26.

Guba, R. F., WIQMN, "Pulse: A Practical Technique for Amateur Microwave Work, Part 3," *QST,* April, 1963, page 31.

Guba, R. F., WIQMN, "Pulse: A Practical Technique for Amateur Microwave Work, Part 4," *QST,* May, 1963, page 58.

Hatherell, G. A., K6LK, "Putting the G-Line to Work," *QST,* June, 1974, page 11.

Jensby, W., WA6BQO, "Stable Microwave Oscillators," *QST,* July, 1966, page 33.

Knadle, R. T. Jr., K2RIW, "A Twelve-Foot Stressed Parabolic Dish," *QST,* August, 1972, page 16.

Kolby, D., K6HIT and Munn, E., W6OYJ, "Microwave DX — California Style," *QST,* September, 1970, page 88.

May, L., W5AJG/W5HN and Lowe, B., K4VOW/WA5UVM, "A Simple and Efficient Mixer for 2304 MHz," *QST,* April, 1974, page 15.

McGregor, J. A., W2RIM, "Dishing out the Milliwatts on 10 KMc," *QST*, February, 1947, page 19.

Merchant, R., W2LFG, and Harrison, A. E., W6BMS, "Duplex phone on 5300 Megacycles," *QST*, January, 1946, page 19.

Peterson, K. E., K3KRU, "Practical Gear for Amateur Microwave Communication," *QST*, June, 1963, page 17.

Petit, R. C., W7GHM, "Coherent CW — Amateur Radio's New State of Art," *QST*, September, 1975, page 26.

Prechtel, C. J., W8DRR, "Experimental Transceivers for 5600 Mc," *QST*, August, 1960, page 11.

Sharbaugh, A. H., W1NUL/2 and Watters, R. L., W9SAD/2, "Our Best DX — 800 Feet," *QST*, August, 1946, page 19.

Sharbaugh, A. H., W1NUL/2 and Watters, R. L., W9SAD/2, "The World above 20,000 Megacycles," *QST*, May, 1959, page 11.

Smith, B., W5TUB, "Simple Approach to 2300-MHz Space Reception," (The World Above 50 Mc), *QST*, March, 1974, page 88.

Sumner, D., K1ZND, "How to Promote Microwaves, British Style," *QST*, November, 1976, page 71.

Tilton, E. P., W1HDQ, "To the Moon and Back on 2300 MHz," *QST*, July, 1969, page 54.

Vilardi, D., WA2VTR, "Simple and Efficient Feed for Parabolic Antennas," *QST*, March, 1973, page 42.

Wilson, P. M., W4HHK and Knadle, R. T. Jr., K2RIW, "Houston, This is Appollo," *QST*, April, 1971, page 34.

Zimmer, J., W2BVU, "225 Miles on 2300 MHz," *QST*, February, 1969, page 53.

"Preview — DX Contest high 'Phone Scores," *QST*, June, 1954, page 54.

CQ Magazine

Harris, F. S., W1FZJ, "A 3800 Mc Crystal Oscillator," (VHF Column), *CQ*, September, 1958, page 102.

Harris, F. S., W1FZJ, "The Parametric Amplifier," *CQ*, November, 1958, page 74.

Katz, A., K2UYH, "Simple Parabolic Antenna Design," *CQ*, August, 1966, page 10.

Kretzman, B. H., W2JTP, "Microwave and the Amateur," *CQ*, August, 1967, page 53.

Massey, W., W4EXL, "A Coax Wavemeter for 300-2000 Mc," *CQ*, February, 1960, page 40.

Schauers, C. J., W4VZO, "Klystron Discussion," (Ham Clinic), *CQ*, September, 1963, page 66.

73 Magazine

Hoisington, B., K1CLL, "The World of X-Band," *73 Magazine*, March, 1973, page 33.

Olbert, S. M., W1SNN, "Mobile Smokey Detector," *73 Magazine*, Holiday issue, 1976, page 32.

"The Reach for Higher Frequencies," *(Lenkurt Demodulator*, October, 1969), Reprinted in April, 1971 issue, *73 Magazine*, page 34.

Radio Communications

Burt, G., GM3OXX, "The GM3OXX portable 3cm Transceiver," *Radio Communications*, June, 1975, page 450.

Evans, D. S., G3RPE, "Gunn diodes at 10,000 MHz," (Microwaves column), *Radio Communications*, April, 1971, page 271.

Evans, D. S., G3RPE, "A Gunn-diode power supply unit and modulator," *Radio Communications*, March, 1972, page 158.

Evans, D. S., G3RPE, "A Wideband 30 MHz i-f amplifier stroke detector," *Radio Communications*, July, 1972, page 430.

Evans, D. S., G3RPE, "A Self-calibrating wavemeter for 3cm," (Microwaves column), *Radio Communications*, February, 1973, page 106.

Evans, D. S., G3RPE, "Operating in the new 24 GHz (12mm) band," (Microwaves column). *Radio Communications*, March, 1973, page 192.

Evans, D. S., G3RPE, "Putting together waveguide equipment," (Microwaves column), *Radio Communications*, October, 1973, page 692.

Evans, D. S., G3RPE, "Practical 10-GHz Gunn Oscillators," *Radio Communications*, May, 1974, page 288.

Evans, D. S., G3RPE, "The 1-3 GHz Band," (Microwaves column), *Radio Communications*, November, 1975, page 852.

Evans, D. S., G3RPE, "Gunn diode Power Supply Units," (Microwaves column), *Radio Communications*, December, 1976, page 919.

Wakeman, A., G3EEZ, "A Simple 3cm Polaplexer," *Radio Communications*, November, 1970, page 752.

other publications

Myers, F. A., "Gunn Effect Technology — Diodes," *Microwave Systems News*, April/May, 1975, page 67.

Myers, F. A., and Foster, W., "Gunn Effect Technology — Oscillators," *Microwave Systems News,* August/September, 1975, page 83.

Myers, F. A., and Foster, W., "Gunn Effect Technology — Tuning and Temperature Effects," *Microwave Systems News,* October/November, 1975, page 57.

"The Reach for Higher Frequencies," *Lenkurt Demodulator,* October, 1969.

"Anomalous Propagation," *Lenkurt Demodulator,* July/August, 1975.

recommended microwave books

Evans, D. S., G3RPE, and Jessop, G. R., G6JP *VHF-UHF Manual,* 3rd edition, Radio Society of Great Britain, London, 1976.

Ginzton, E. L., *Microwave Measurements,* McGraw-Hill, New York, 1957.

Lance, A. L., *Introduction to Microwave Theory and Measurements,* McGraw-Hill, New York, 1964.

Mattaei, G., Young, L., Jones, E., *Microwave Filters, Impedance-Matching Networks and Coupling Structures,* McGraw-Hill, New York, 1964.

Montgomery, C. G., *Techniques of Microwave Measurements,* (MIT Radiation Lab Series), McGraw-Hill, New York, 1947.

Montgomery, C. G., Dicke, R. H., and Purcell, E. M., *Principles of Microwave Circuits,* (MIT Radiation Lab Series), McGraw-Hill, New York, 1947.

Moreno, T., *Microwave Transmission Design Data,* Dover Publications, New York, 1958, especially pages 120-122 and 126-129.

Reich, H. J., *Very High-Frequency Techniques,* Harvard Radio Research Laboratory, Boston Technical Publishers, Cambridge, Massachusetts, 1965.

Shrader, R. L., *Electronic Communications,* 3rd edition, McGraw-Hill, New York, 1975, Chapter 26, "Microwaves."

Southworth, G. C., *Principles and Applications of Waveguide Transmission,* Bell Laboratories, New Jersey, 1966, especially pages 415-419.

Thomas, H. E., and Clark, C. A., *Handbook of Electronic Instruments and Measurement Techniques,* Prentice-Hall, Englewood, New Jersey, 1967.

Wheeler, G. J., *Introduction to Microwaves,* Prentice-Hall, Englewood, New Jersey, 1963.

Reference Data for Radio Engineers, 5th edition, International Telephone and Telegraph, Howard A. Sams, Indianapolis, 1969.

ham radio

introduction to
GaAs field-effect transistors

What's a GaAs fet? Certainly not a gaseous field-effect transistor, as a misguided soul recently asked. *GaAs fet* is short for Gallium-Arsenide field-effect transistor; it is the hottest new uhf and microwave component, and is being widely used in industrial and military applications.

Gallium Arsenide is one of the newer semiconductor compounds, which until recently was used primarily for LEDs and microwave diodes. If you remember the *Periodic Chart of the Elements* from basic chemistry, you'll find the common semiconductors, silicon (Si) and germanium (Ge) in column IV of the chart, indicating they have four free electrons. It has also been found that semiconductors can be made by combining an element from column III, such as gallium (Ga) or indium (In), with an element from column V, such as phosphorous (P) or arsenic (As). The combined compound apparently has an average of four free electrons and acts as a semiconductor. The more successful combinations are gallium arsenide (GaAs), indium phosphide (InP), and gallium phosphide (GaP).

What is the advantage of using these exotic semiconductors? It arises from the higher carrier mobility of these materials — the electrons move faster than they do in silicon or germanium. This is the key to high-frequency performance; the maximum operating frequency of any amplifying device is limited by the time it takes a signal to pass through it (transmit time in a vacuum tube for example).

GaAs fet construction

Gallium arsenide is the most commonly used of these III-V semiconductors; with appropriate doping, it is used for field-effect transistors, infrared LEDs, and many types of microwave diodes including Gunn oscillators. The fets are fabricated on an epitaxial layer of the proper doping, with the channel defined between the high-conductivity source and drain areas **(fig. 1)**. To take full advantage of the high carrier mobility for high-frequency performance, very short channels or gate lengths, are used. Typical gate length is one micron (10^{-6} meter) for a microwave GaAs fet; some devices are available with a half-micron gate length. The fundamental limitation is the wavelength of the ultraviolet light used to expose the photoresist, which is approximately 1/3 micron.

A one-micron gate seems very small, and it is, even though it is much wider than it is long — typically 100 to 150 microns wide. The current path is therefore relatively short and wide, a good high-frequency configuration. To increase power capability, several gates are paralleled for higher current. Fig. 2 shows a typical GaAs fet structure with four gates in parallel (hidden by metallization). The gate itself is a Schottky junction, as opposed to the common P-N junction found in low-frequency fets.

How high in frequency do GaAs fets work? In the laboratory they have been operated to 22 GHz and higher. Commercially available devices work well up to about 12 GHz as low-noise amplifiers and power

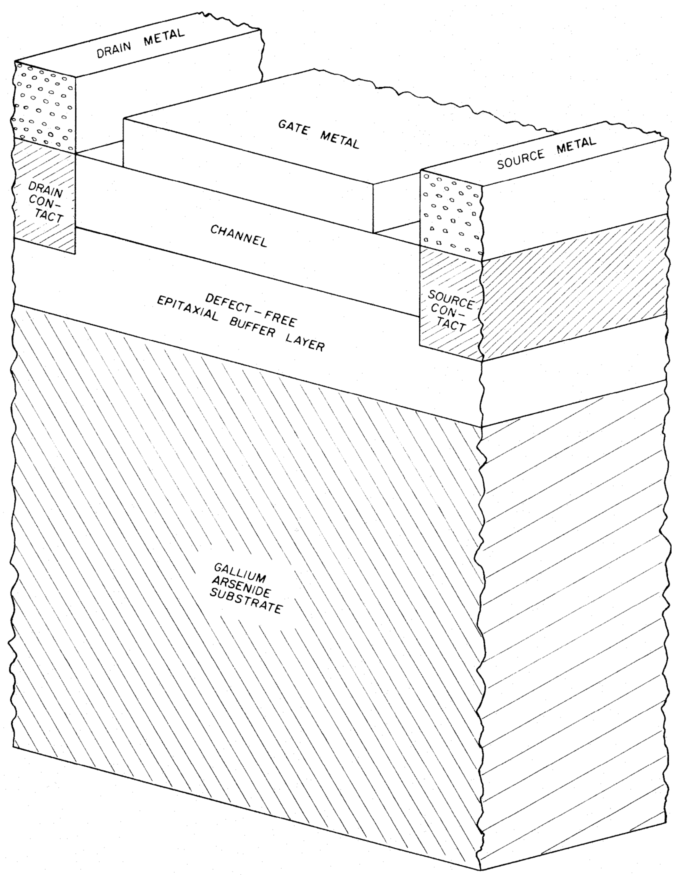


fig. 1. Construction of the GaAs fet, showing the channel between the high-conductivity drain and source areas.

By Paul C. Wade, WA2ZZF, GaAs FET Applications Leader, Microwave Semiconductor Corporation, 100 School House Road, Somerset, New Jersey 08873

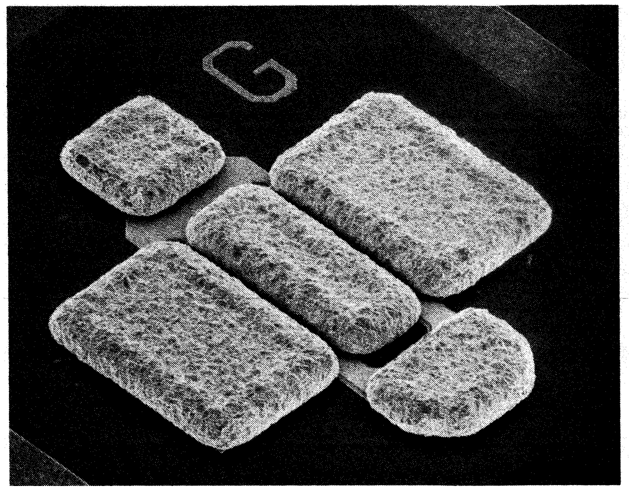


fig. 2. GaAs fet chip structure as seen by a scanning electron microscope.

amplifiers, and at even higher frequencies as oscillators. And this performance is obtained at low supply voltages, 3 to 12 volts.

At present, it is as low-noise amplifiers that GaAs fets really shine. Available noise figures were previously only obtainable with the best parametric amplifiers and MASERs. For instance, some GaAs fets offer noise figures under 2 dB at 4 GHz, and under 4 dB at 12 GHz. A few of these devices are finding their way into amateur hands; at the recent Eastern VHF/UHF Conference, K2UYH's 432-MHz preamp, using a NEC V244 GaAs fet, had a measured 0.8 dB noise figure! However, the prices for these devices, while dropping, are still rather steep.

Power GaAs fets are newer, but are also showing respectable performance. Devices are commercially available with 1 watt output up to 8 GHz, and more than 6 dB power gain. Since these are linear devices, unlike bipolar microwave transistors, the power and gain are specified at the standard 1 dB compression point.

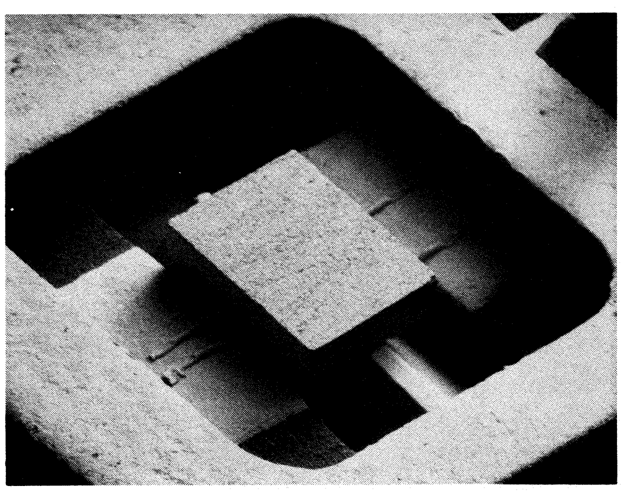


fig. 3. GaAs fet chip flip-chip mounted in package.

One problem with power GaAs fets is adequate heat sinking, since gallium arsenide has a thermal conductivity much lower than silicon. Normally, planar transistors are fabricated with the active area up and the heat is conducted through the bulk semiconductor material underneath. With power GaAs fets, however, some manufacturers are using an inverted mounting technique, with the source metallization attached directly to ground, as shown in **fig.** 3. This technique not only halves the thermal resistance, but also reduces the source inductance, which improves stability.

Investigation of GaAs fets as oscillators has only begun recently. To date, we have obtained as much as 0.6 watt output at 9 GHz from an oscillator. The highest frequency oscillator we have made so far was at 17.1 GHz, where we obtained 100 milliwatts from the waveguide oscillator shown in **fig. 4.**

precautions

GaAs fets also have a reputation for being fragile, but all new semiconductor devices pass through this stage; they inevitably become more rugged as better manufacturing techniques are developed. The latest

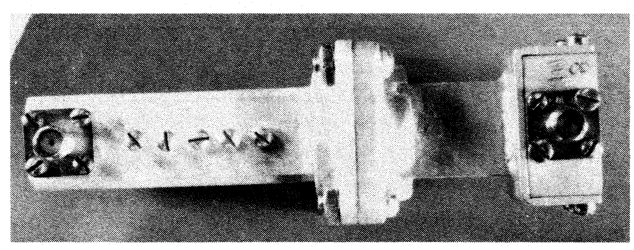


fig. 4. A GaAs fet 17-GHz waveguide oscillator.

GaAs fets are damaged only by excess voltage or extremely high temperature (>300°C). An excessive voltage applied between the source and drain causes a bulk breakdown, which unlike avalanche breakdown in a transistor, is irreversible. With the addition of protective zener diodes, however, GaAs fets are as rugged as most microwave semiconductor devices.

It may seem that GaAs fets are too rare, exotic, and expensive for amateur use, but many new devices started out this way. Then, radio amateurs like K2UYH, who needs the improved low-noise performance for his moonbounce work, began using them. Finally, after a few years, price and availability become more reasonable and formerly exotic devices come into general usage. This may or may not happen with GaAs fets, but amateurs should keep an eye on new technologies for the future.

*All photographs courtesy Microwave Semiconductor Corporation.

ham radio

new op amp challenges the 741

The new CA3140 IC op amp from RCA features a high-impedance mosfet input stage, improved slew rate, and wider frequency response at comparable cost to the popular 741

For several years the 741 op amp IC has been the popular workhorse for both industrial and hobbyist circuit designers. Why? Because it's inexpensive and simple to use. Of course, its slew rate isn't too great, and its input bias current isn't anything to write home about, but what do you expect for twenty-five cents?

Now there's another op amp IC on the market which I think deserves as much attention as the 741; it's the new RCA CA3140. Like the 741, it requires no external frequency compensation components, and its output is short-circuit proof. Its pin configuration is the same as the 741, and RCA claims the CA3140 is a direct plug-in replacement for the 741 in most applications.

So, what's so special about the CA3140? For openers, it has mosfet input transistors (diode protected) which means you can use much higher value resistors in the input circuit without worrying about their effect on output offset voltage. Another big advantage is that the slew rate of the CA3140 is an order of magnitude faster than that of the 741. Supply voltage range for both op amps is the same; ±2 to ±18 volts.

Now for the price. At this writing, it's available in an 8-pin TO-5 can for 80 cents in small quantities. It is reported that it will soon be available in the popular 8-pin minidip plastic package for 72 cents. The slight difference in cost as compared to the 741 seems very, reasonable for the higher performance of the CA3140.

Table 1 lists some important parameters for both the 741 and the CA3140, so you can quickly see what you're getting for your money. Specs given are for the commercial versions.

The big differences between the two op amps are clearly input resistance and bias current, and slew rate. Typical curves for the devices show that maximum output voltage swing for the 741 is flat out to 10 kHz before it starts falling off at higher frequencies, while the CA3140 is flat out to 100 kHz. **Fig. 1** shows a block diagram of the CA3140, and **fig.** 2 shows the schematic diagram.

a disadvantage

There is one point on which the 741 is superior to the CA3140: the 741 will drive a lower resistance load than the CA3140 will. My experience shows that severe clipping of the output occurs on negative peaks when the load resistance on the CA3140 is 1200 ohms. If the load resistance is increased to 2000 ohms, this problem disappears. The 741 output circuit is a complimentary npn-pnp emitter follower, while the CA3140 has an npn emitter follower with a current source in the emitter circuit.

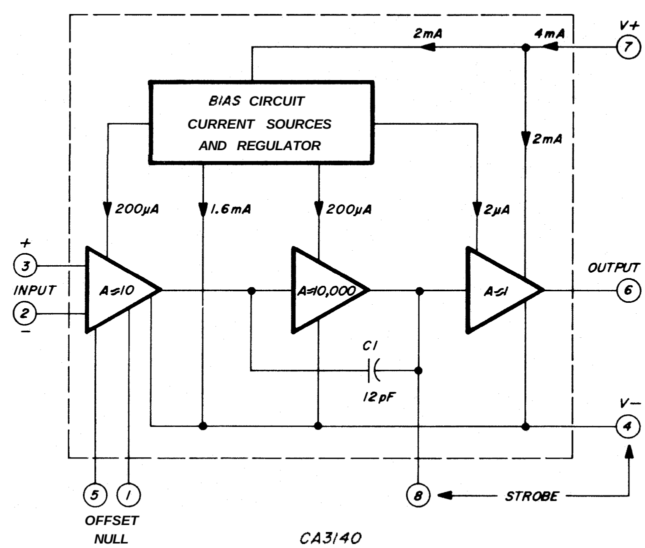


fig. 1. Block diagram of the new RCA CA3140 op amp IC which is a direct plug-in replacement for the popular 741 in most applications, but features a mosfet input stage for high input impedance.

application

Since the input current to the CA3140 is so low, many megohms of unbalanced resistance may be

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248

table 1. Specifications of the 741 and RCA CA3140 op amp ICs, compared at +25°C.

parameter	741	CA3140	units
Input resistance (typical)	2.0	1.5×10^6	megohms
Input bias current (max)	500,000	50	picoarnps
Input offset voltage (max)	6	15	millivolts
Slew rate (typical)	0.5	9	volts/μs
Large signal voltage gain (min)	20.000	20,000	volts/volt
Output resistance (typical)	75	60	ohms
Power supply rejection (max)	150	150	μ volt/volt
Common mode rejection ratio (min)	70	70	dB

used in the input circuit with no appreciable dc output offset due to bias current. Consider the circuit of **fig.** 3. The inverting input terminal sees a parallel equivalent resistance of 10 megohms. Since the maximum input bias current is 50 pico-amperes (0.00005 microamp), the offset voltage due to bias current will be no more than 0.5 millivolt. Therefore, for most applications, you can use just about any resistor network you choose on the input and forget about its effect on the offset voltage.

I took advantage of this feature of the CA3140, plus its high slew rate, to build the simple Wien bridge sine wave generator shown in **fig. 4.** Both ICs are CA3140s. U1 is the oscillator, and U2 provides a constant 600-ohm output impedance, regardless of the amplitude setting.

The resistor network at the output lets U2 see a load resistance of 2000 ohms when the output terminals are connected to a 600-ohm load; it also causes the output terminals to look like a 600-ohm source. Maximum output amplitude into a 600-ohm load is about one volt rms.

Frequency range of the Wien bridge oscillator is 30

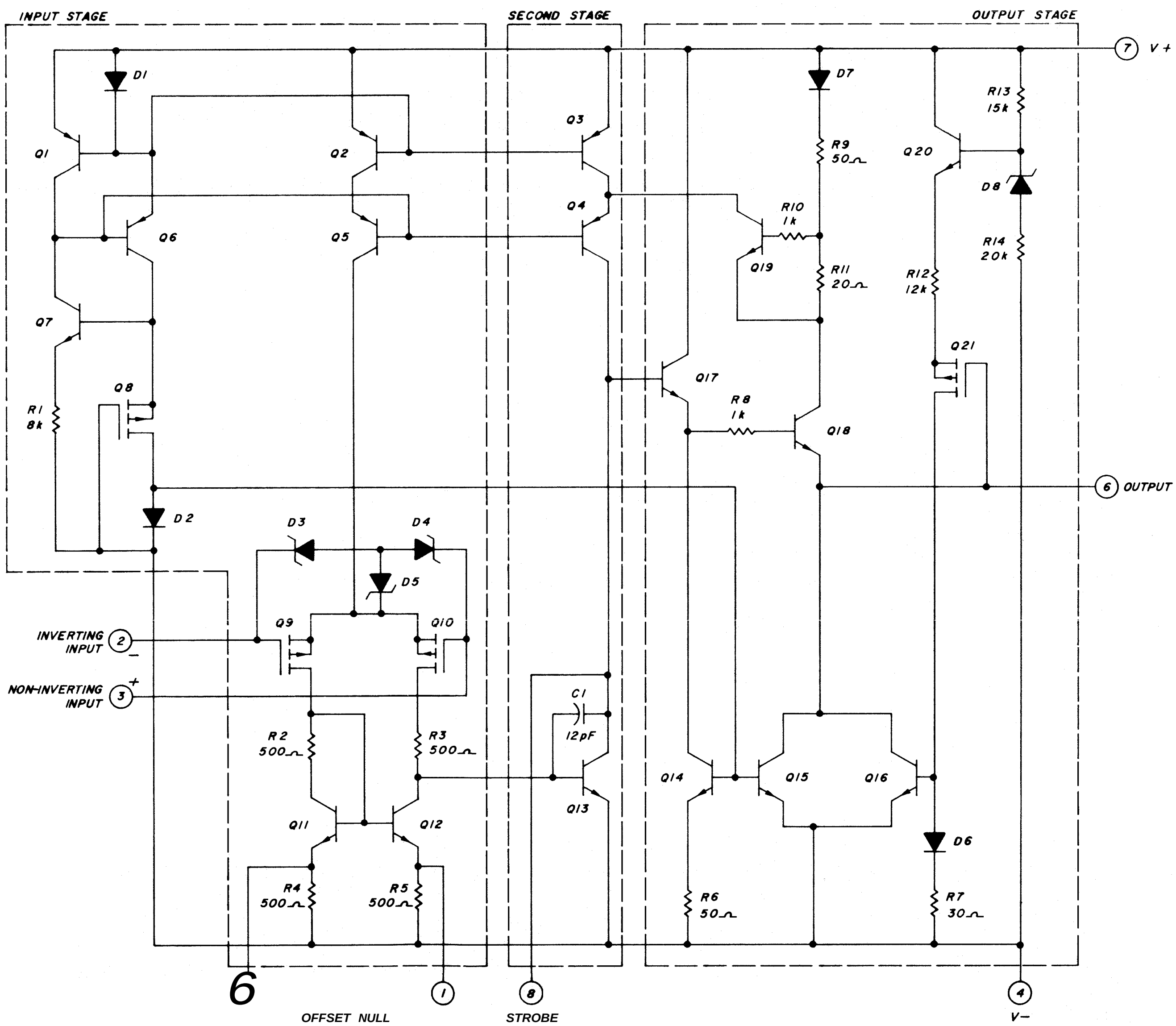


fig. 2. Schematic of the RCA CA3140 op amp IC with diode-protected mosfet input stage. As compared to the 741 op amp, the CA3140 offers higher input impedance and improved slew rate. Typical specifications for the two devices at room temperature are listed in *table* 1.

Hz to 100 kHz and is flat within 0.5 dB, due to the excellent gain control characteristic of the thermistor in the feedback circuit of U1. The 10k pot is adjusted for best waveform. Total harmonic distortion is less than 0.5 per cent at all frequencies. C1 and C2 is a two-gang 450-pF air variable. Its frame must be insulated from ground; I mounted it on a piece of plexiglass and used an insulated shaft coupling to connect it to the dial. A trimmer capacitor is needed to balance out stray capacitance from the capacitor frame to ground.

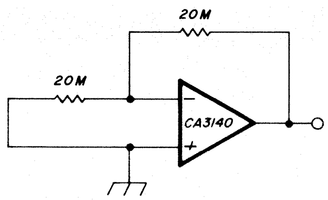


fig. 3. Maximum **dc offset** of this circuit due to bias current is 0.5 **millivolt**.

a word of caution

The input bias currents given above are for +25°C ambient temperature (room temperature). As temperature increases, the input bias current of the CA3140 will approximately double for each 10°C rise. At +125°C, its value will be roughly 1000 times greater than at room temperature. Input bias current for the 741, however, actually decreases as temperature rises.

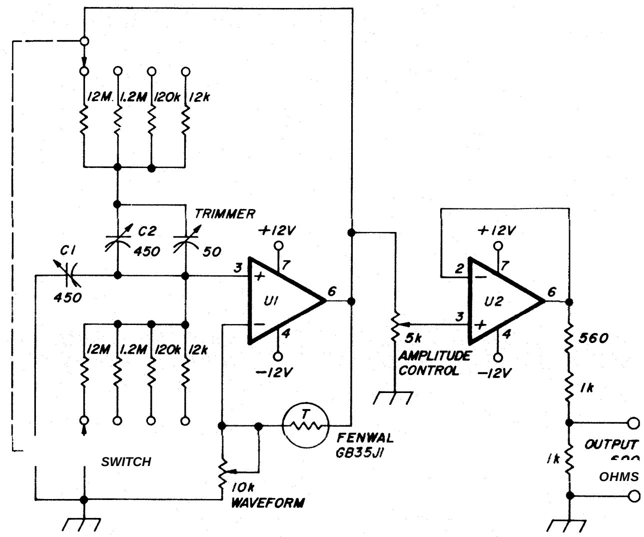
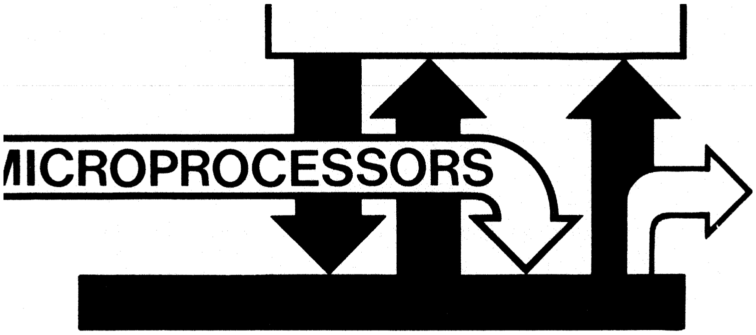


fig. 4. Wien bridge sine-wave oscillator using two RCA **CA3140** op amps covers 30 Hz to 100 kHz with less than 0.5 per cent total harmonic distortion. The **10k** pot is adjusted for best waveform. Capacitor **C1** and C2 is a two-gang 450-**pF** variable with its frame isolated from ground. Maximum output into a 600-ohm load is about 1 volt **rms.**

ham radio



microprocessors: a microprocessor controlled CW keyboard

Now that the microprocessor has made the homebrew computer possible, its popularity should tend toward dedicated applications in amateur radio. In this article, a preprogrammed microcomputer is designed to function as a Morse Code keyboard with extra features providing the utmost in flexibility. This project is an attractive alternative to its discrete equivalent with numerous gates, flip-flops, binary counters, and diode matrices.

The code computer was designed around the MCS-6504 microprocessor by MOS Technology. The other devices connected to the processor chip comprise a software simulation of a discrete logic system. The software for this project was developed and debugged with the aid of a KIM-1 microcomputer. After the program was working to my satisfaction, the source listing for the software package was transferred to the 1702A EPROMs for permanent storage. Thus, the system is running upon application of power.

However, this system does more than synthesize Morse code from an ASCII keyboard. It also provides control functions, for operator convenience, which are unheard of in similar units of discrete design. The features of this system are:

1. Variable code speed, 5 to 99 wpm range. Code speed is entered digitally from the numeric keys on the keyboard.

2. 256-character first-in-first-out (FIFO) buffer memory. This allows the operator to type faster than the machine is sending. At 10 wpm, it's possible to get five minutes ahead of the machine.

3. Automatic character spacing. Word spacing is provided by the operator's depressing the SPACE bar on the keyboard.

4. 64-character auxiliary buffer for storing repeated messages like CQ or call-up sequences. Data can be entered into the auxiliary buffer without causing interference to the FIFO.

5. BACKSPACE command. A backspace routine is included in the software for FIFO error correction, and operates like the BACKSPACE key on a typewriter. For ASCII keyboards without a BACKSPACE key, CONTROL H can be used.

6. Automatic default. Illegitimate control characters are ignored by the program to prevent a software lock-up condition. Control characters are used for code speed entry, auxiliary buffer data entry, auxiliary buffer data transfer to FIFO, and backspacing.

7. TRANSMIT/RECEIVE output. This reed output automatically switches your transceiver from receive to transmit when you begin typing. The rig will stay on the air until the FIFO is empty.

8. Warning lights. Two indicator lamps are provided to prevent the operator from filling the FIFO to the OVERFLOW point. Lamp 1 lights when there are less than 64 character spaces left in the FIFO. Lamp 2 (condition red) turns on when the operator is within 16 characters of OVERFLOW.

9. FIFO full output. This signal from U13 goes to a logic 1 when the condition red lamp goes on. It stays high until the condition yellow lamp turns off. This output can be used to control a paper-tape reader.

By James W. Pollock, WB2DFA, 6 Terrace Avenue, New Egypt, New Jersey 08533

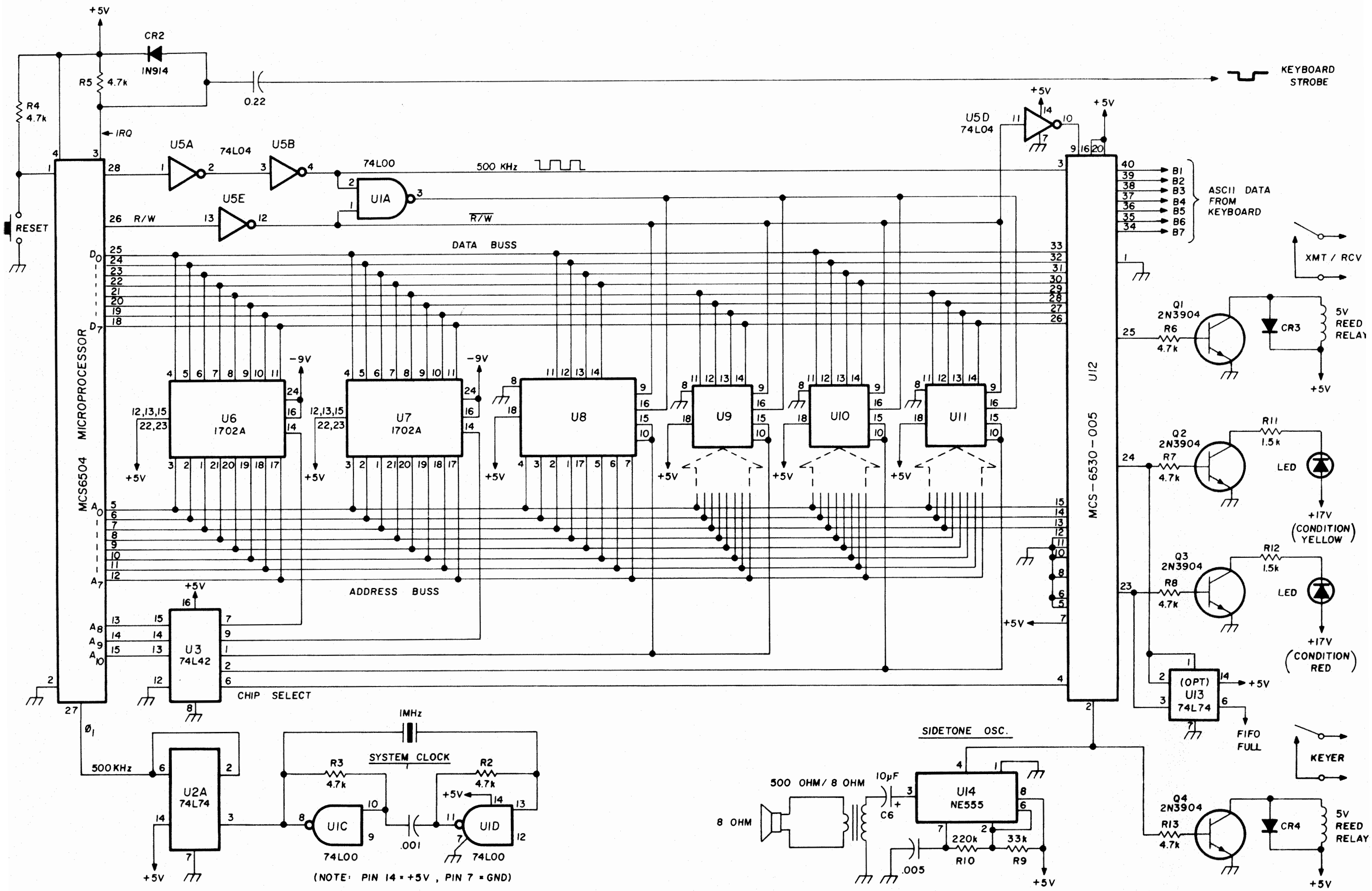


fig. 1. Schematic diagram of the microprocessor controlled keyboard. The MCS-6504 and MCS-6539-005 are available from Falk-Baker Associates, 382 Franklin Avenue, Nutley, New Jersey 07110, for $20 and $14, respectively. The 2111 RAMs can also be obtained from the same source.

10. By adding a serial-to-parallel converter, like a UART, the system can be used with an ASR-33 or similar Teletype, CRT serial data terminal, etc.

Although construction is not especially critical, this project is not recommended for the beginner; an experienced hardware hacker should encounter no difficulty. The cost of this project including power supply, ASCII keyboard, enclosure, and all parts is less than $200. About half of that figure will be invested in the ICs alone. Knowledge of basic microprocessor operation and programming is recommended, but is not a requirement to build the system since the EPROMs store the system's operation program.

It's important that the builder be very meticulous while assembling the system. Since computers do only what they are told to do, a miswired address or data line will wreak havoc. Interchanging EPROMs U6 and U7 will have the same effect. The instructions for the program have been listed in a specific order to define the behavior of the Morse keyboard. Interchanging the EPROMs, in effect, scrambles the order in which the instructions are to be executed.

system operation

The heart of this code computer is a 6504 microprocessor which is a software compatible cousin to the 6502. The 6504 was chosen for its lower cost and compact 28-pin package design.

The crystal oscillator (U1C and U1D) functions as the system clock (see **fig. 1**). U2A divides the 1-MHz clock down to 500 kHz to compensate for the slow speed of the 1702A EPROMs. Since the access time for the 1702A is usually specified at 1 μsec, the operation of surplus units may be marginal at the full 1-MHz clock rate. Thus, a system clock of 500 kHz was chosen to prevent EPROM access timing problems without resorting to buying factory prime units.

Pin 28 ($\phi 2$ out) on the 6504 is used to coordinate the readlwrite timing of the RAMs (U8-U11) and the peripheral interface adapter (PIA), U12. U5A, U5B, and U5C buffer the readlwrite and $\phi 2$ signals to prevent loading effects on the microprocessor.

The PIA (U12) is used to interface the data bus of the microprocessor to the ASCII keyboard, keying relay, xmt/rcv relay, side tone oscillator, and the FIFO warning lamps. Thus, the PIA chip is used as an I/O port for the system. Port A is used to read the seven-bit input from the ASCII keyboard. Pin 2 of the PIA is bit 0 of Port A and is used as the serial output for the Morse code information that switches the side tone oscillator (U13) and the keying relay driver transistor (Q4). Port B is used as an output latch for the FIFO status flags.

1. Pin 25 — TRANSMIT/RECEIVE output
2. Pin 24 — Condition yellow output
3. Pin 23 — Condition red output

In addition to two I/O ports, the PIA is equipped with a readlwrite interval timer that is used extensively for the timing of dots, dashes, and spaces. The timer is programmable in discrete steps of two milliseconds with a 500-kHz time base.

The PIA also has a 64 by 8 bit RAM that can be used for scratch pad, or temporary program storage. The RAM, however, was not used on this system.

The RAMs (random access memories) chosen for this project are 2111's (U8, 9, 10, 11). These chips are crganized as 256 by 4 bit devices, and are used in pairs to accommodate the 8-bit data bus of the microprocessor. Thus each pair (U8, U9, and U10, U11) makes up a 256 by 8 bit memory page for a total RAM storage of 512 bytes. As seen in **fig. 1**, the RAMs are used as temporary storage for the system scratch pad and messages entered via the keyboard.

These particular RAMs, like the microprocessor, have a bi-directional data bus that permits OR tying to the CPU for ease of construction. The $\overline{R/W}$ signal from U5E, when a logic 0, allows the RAMs to send data to the CPU; when a logic 1, they will accept data.

The bi-directional data bus of the CPU, pins 18-25, is OR tied with the data bus pins of the EPROMs, RAMs, and the PIA. Since these lines are tri-state, ORing them in this fashion greatly simplifies construction. These I/O pins are in a high impedance state when the chip select $(\overline{CS})$ pin of the IC is a logic 1. Since the address bus is a "one-way street," these lines are also tied together.

Selection of the support devices is accomplished by U3. The high order address lines of the CPU, are decoded by U3 which in turn presents a logic 0 at the $\overline{CS}$ pin of the appropriate device. Address lines A8-A10 select the support device (RAM, EPROM, PIA) while the low order address lines A_0-A_7 select a memory cell within that device. **Table 1** shows the selection scheme with regard to the address lines.

table 1. Device selection by the microprocessor.

device selected	A10	A9	A8
U8 and U9	0	0	0
U10 and U11	0	0	1
u12	1	0	1
U6	1	1	0
U7	1	1	1

'A PROM programming service is available from Keith Petersen. 1418 Genesee Street, Royal Oak, Michigan 48073. For this project only, the cost is $6 for programming plus $1 for shipping and handling, per pair. Send the PROMs in a conductive carrier to prevent static discharge damage.

In order to have data from the keyboard processed by the CPU, the strobe output from the keyboard drives the Interrupt Request pin (IRQ) with a negative-going pulse. The negative-going pulse at the IRQ input is about 10 μsec long. Using the IRQ input in this manner alerts the CPU to the fact that keyboard data has been entered for processing. A

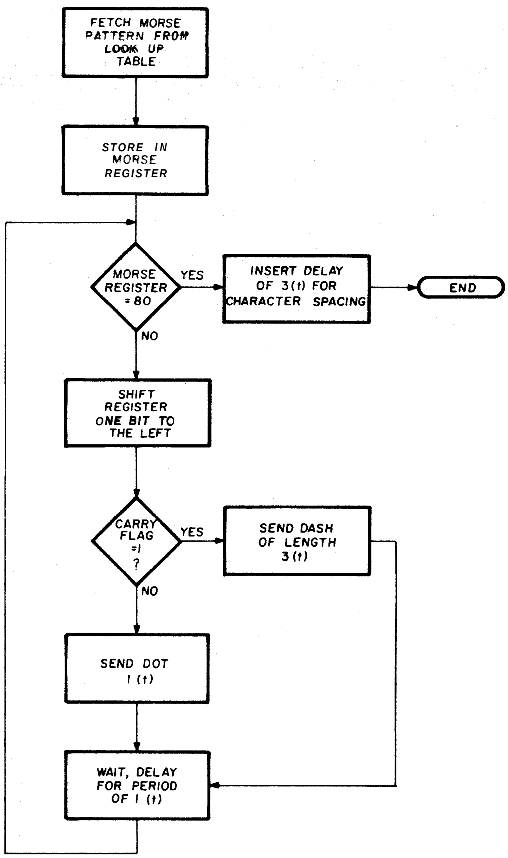


fig. 2. This flow chart shows the ASCII to Morse conversion used in the keyboard.

subroutine then fetches the keyboard data from the PIA and stores it in memory each time a key is pressed.

Any ASCII encoded keyboard with TTL compatible output levels will easily interface with the computer. Pins 34-40 are the ASCII inputs to the PIA, with pin 40 being the least significant bit, B1. The KBD-5 ASCII keyboard kit by South West Technical Products is a good choice. Regardless of the keyboard you select, bear in mind that the keyboard must be programmed for upper case characters; lower case characters will not work in this system.

On the KBD-5, the BACKSPACE key is uncommitted and must be connected to the on-board encoder. In my case, I connected the leads for the BACKSPACE key to pins 26 and 36 of the AY-5-2376 encoder. The ASCII output will then be 0001000 when the key is pressed. One additional point should be remembered: the ASCII outputs are not latched, and will only be present when the key is pressed.

ASCII to Morse conversion

The generation of the Morse character for its ASCII equivalent begins by using **table** 2 for U6. As an example, the ASCII code for the letter F is 1000110 or 46_{hex}. The CPU looks at the 46th position in **table 2.** At position 46_{hex} the number is 00101000 or 28_{hex}. The CPU stores this value in a memory location for shift operations during the code synthesis process and for future reference.

table 2. Look-up table listing for U6

	ASCII	hex	Morse code group	hex equivalent	
A	1000001	41	01100000	60	
B	1000010	42	10001000	88	
C	1000011	43	10101000	A8	
D	1000100	44	10010000	90	
E	1000101	45	01000000	40	
F	1000110	46	00101000	28	
G	1000111	47	11010000	D0	
H	1001000	48	00001000	08	
I	1001001	49	00100000	20	
J	1001010	4A	01111000	78	
K	1001011	4B	10110000	B0	
L	1001100	4C	01001000	48	
M	1001101	4D	11100000	E0	
N	1001110	4E	10100000	A0	
O	1001111	4F	11110000	F0	
P	1010000	50	01101000	68	
Q	1010001	51	11011000	D8	
R	1010010	52	01010000	50	
S	1010011	53	00010000	10	
T	1010100	54	11000000	C0	
U	1010101	55	00110000	30	
V	1010110	56	00011000	18	
W	1010111	57	01110000	70	
X	1011000	58	10011000	98	
Y	1011001	59	10111000	B8	
Z	1011010	5A	11001000	C8	
Ø	0110000	30	11111100	FC	
1	0110001	31	01111100	7C	
2	0110010	32	00111100	3C	
3	0110011	33	00011100	1C	
4	0110100	34	00001100	0C	
5	0110101	35	00000100	04	
6	0110110	36	10000100	84	
7	0110111	37	11000100	C4	
8	0111000	38	11100100	E4	
9	0111001	39	11110100	F4	
(,)	0101100	2C	11001110	CE	
(=)	0101101	2D	00010110	16	$\overline{SK}$
(.)	0101110	2E	01010110	56	
(?)	0101111	2F	00110010	32	
([ )	1011011	5B	01010100	54	$\overline{AR}$
( ])	1011101	5D	10010100	94	$\overline{DN}$
(↑)	1011110	5E	10001100	8C	$\overline{BT}$
(:)	0111010	3A	11100010	E2	
(;)	0111011	3B	10101010	AA	

The Morse buffer register will contain the information from **table** 2 as:

C	B7	B6	B5	B4	B3	B2	B1	B0	
X	0	0	1	0	1	0	0	0	$=28_{hex}$

The 0s represent dots and the 1s represent dashes. The actual code pattern is determined by inspecting the bits, from most to least significant. The remaining bits are used to denote completion of the character by putting a 1 after the character and 0s after the one. The software will then check for 10000000 (80_{hex}).

The ***arithmetic shift left*** (ASL) instruction is used to shift the code group into the CARRY flag one bit at a time.

	C	B7	B6	B5	B4	B3	B2	B1	B0
START	X	0	0	1	0	1	0	0	0
1st ASL	0	0	1	0	1	0	0	0	0
2nd ASL	0	1	0	1	0	0	0	0	0
3rd ASL	1	0	1	0	0	0	0	0	0
4th ASL	0	1	1	0	0	0	0	0	0

Upon execution of ASL, B7 is shifted into the CARRY flag and the other bits in the Morse register are subsequently shifted to the left. The shifts are performed until the content of the register is 1000 0000 (80_{hex}). In the case of the letter F, four shifts are used.

Morse characters are synthesized in dot-space and dash-space pairs. The state of the CARRY flag determines whether a dot-space or dash-space pair will be sent by the timing loop software to the keying relay. When the final shift occurs, the program inserts a time delay equivalent to three dots to provide proper spacing before the next character.

The flow chart illustrating the entire ASCII to Morse conversion technique is shown in **fig. 2.**

control characters

Control characters lend a greater flexibility to the system by allowing the operator to change the

The CW keyboard is housed in a 14 x 11 x 3 inch (36x28x8 cm) cabinet available from Nu Data Electronics, 104 North Emerson Street, Mount Prospect, *Illinois* 60056.

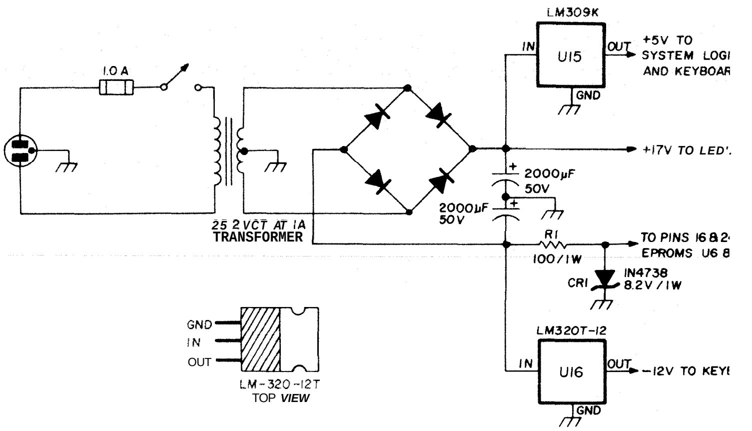


fig. 3. Schematic diagram of the power supply. The LM309K should be mounted on a heatsink. Note that the case of the LM320-12T must not be grounded. The bridge rectifier's rating is 100 PIV, 1 ampere; the transformer is 25.2 Vac, 1 ampere.

		+5 volts	ground	−8 to −9 volts
U1	SN74L00	14	7	
U2	SN74L74	14	7	
U3	SN74L42	16	8, 12	
U4	MCS-6504	4	2	
U5	SN74L04	14	7	
U6, 7	1702-A	12, 13, 15, 22, 23		16, 24
U8-11	2111	18	8	
U12	MCS-6530-005	7, 16, 20	1, 5, 6, 8, 10, 11, 12	
U13	SN74L74	14	7	
U14	NE555V	8	1	

course of program execution, and perform software generated control sequences solely from the keyboard. The control character is implemented by pressing the CONTROL key first and while keeping it depressed, the desired alpha-numeric key. The following control characters are programmed for use with this system:

1. BACKSPACE (CONTROL H)
2. CONTROL X
3. CONTROL S
4. CONTROL L
5. CONTROL T
6. RETURN (Carriage return)

BACKSPACE is used only for FIFO error correction. This key backs up the FIFO pointer to the last character entered; the keystroke that follows will replace that character. This feature literally makes it possible to send perfect code. Since the BACKSPACE key can be used to correct mistakes as they are made, a special key for the standard error signal (8 dits) was not included in the software.

CONTROL X is used for entering the code speed entry routine. For example, the code speed can be changed to 25 words per minute by the following sequence:

1. CONTROLX
2. 25
3. CONTROLS

CONTROL X allows the operator to enter the two-digit code speed into a buffer. CONTROL S then initiates a subroutine that programs the interval timer by calculating the equivalent time element for that code speed. The calculation was based on the ARRL rule that 12 wpm is analogous to 5 dits per second, or 10 Hz. Thus, from wpm, the time interval can be calculated. Since the interval timer is binary and not decimal, the timing interval must be converted to hexidecimal.

After all the conversion constants are computed, the final equation is

$$t_{16} = \frac{550}{wpm} \quad \text{milliseconds} \qquad (1)$$

The division is performed in the microprocessor by using a repeated subtraction technique that performs the subtraction in decimal, and counts the number of subtractions in hex (base 16).

CONTROL L is used to store call-up or CQ sequences in the 64-character auxiliary buffer. When CONTROL L is activated, the contents of this buffer are, in effect, erased. The auxiliary buffer can be loaded as follows:

1. CONTROL L
2. CQ CQ CQ CQ CQ CQ DE WB2DFA WB2DFA WB2DFA K
3. RETURN

The RETURN key jumps the program back into its normal flow, and the auxiliary buffer can be recalled by depressing CONTROL T. If it is desired to save the auxiliary buffer for later recall, the RETURN key is used so that the operator can go on with typing data into the FIFO.

There are many situations in which this buffer can be used to make the operator more efficient. For example, while your QSO partner is answering, you can load a signing sequence as follows:

1. CONTROL L
2. K2SMN K2SMN DE WB2DFA
3. RETURN

When he is finished, you go on the air instantly by hitting CONTROL T and start typing your QSO. After the auxiliary buffer is empty, the FIFO memory is read out, in the order in which characters were entered.

To finish your transmission, you can recall the auxiliary buffer data by using CONTROL T again.

1. CONTROLT
2. AR AR K

After the content of the transmission has been sent, the FIFO will send K2SMN K2SMN DE WB2DFA AR AR K.

If a mistake is made while entering data into the auxiliary buffer, you must start all over again by depressing CONTROL L. BACKSPACE will not correct auxiliary buffer errors. When using a data terminal or a TV type terminal to enter data, the capacity of the auxiliary buffer will be one line of text.

construction

Since construction of this project involves the handling of MOS devices, their handling precautions should be observed. It is best to leave the EPROMs,

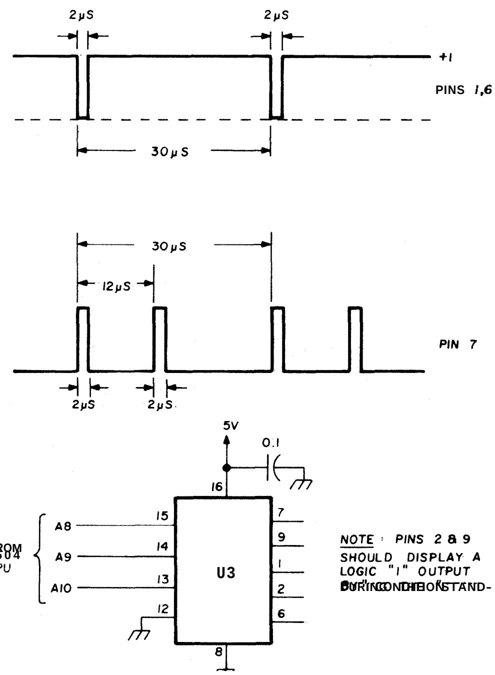


fig. 4. System timing pulses that are on U3, the chip select IC, during the halt loop.

CPU, PIA, and RAM chips in their protective carriers, installing them last. The use of sockets is highly recommended; a one-dollar socket is good insurance for the prevention of irreparable damage to a 20 dollar CPU or 18 dollar PIA.

The best place to start building is the power supply (**fig. 3**), measuring the output voltage from regulators U15 and U16, +5 and −12 volts respectively. The potential at the cathode side of CR2 should be −8 to −9 volts. After the voltage checks have been made, all capacitors, resistors, diodes, and transistors should be installed. Do not install any of the ICs before checking the pin voltages shown in **fig.** 3. When the voltage and continuity checks agree, it's safe to proceed installing the U1, U2, and U14 circuitry.

After you've re-applied power, check pin 8 of U1C for a 1-MHz signal, then pin 6 of U2 for a 500-kHz square wave. The tone oscillator, U14, should put out an 800-hertz square wave when pin 4 is tied to +5 volts by a clip lead. The tone should stop when pin 4 of the 555 oscillator is grounded.

The next step is the installation of the CPU, EPROMs, and RAM chips, along with their support devices, U3, U5, and U12. When power is re-applied, the tone oscillator should be running. Momentarily grounding pin 1 of U4 will stop the tone. This means that the microprocessor has stepped through the configuration software, and is executing the STANDBY routine in U6. Verification of this condition can be made by comparing the waveforms from U3's outputs, with those depicted in **fig. 4.** Also, the output pins of U12, pins 2, 23, 24, and 25 should be at logic 0.

The address and data lines of the CPU can only drive one standard TTL load. Therefore, it is necessary to use low-power TTL (SN74L00 series) devices so the microprocessor can reliably drive the address, R/W, and clock pins of the EPROMs, RAMs, and the PIA. SN74LS00 devices could also be substituted. Note that the 1-MHz oscillator uses an SN74L00 in a self-biasing scheme provided by the 4.7k resistors. A standard SN7400 will not work with the values shown in the schematic.

The system, as is, represents a minimal configuration. The CPU itself is capable of addressing a total of 8192 bytes of memory; of this amount, only 1024 bytes are used for RAM and PROM. Address locations 0200 to 04FF are not used.* Thus, the system can be expanded by adding RAMs or more I/O with software to supervise the expansion in the PROMs for locations above address 0800. However, expansion of the basic system means buffering the address and data bus by means of TTL inverter pairs on the address lines, and a transceiving buffer on the data bus,

*A copy of the memory map and programming information for the EPROMs is available by sending a self-addressed, stamped envelope to ham radio, Greenville, New Hampshire 03048.

references

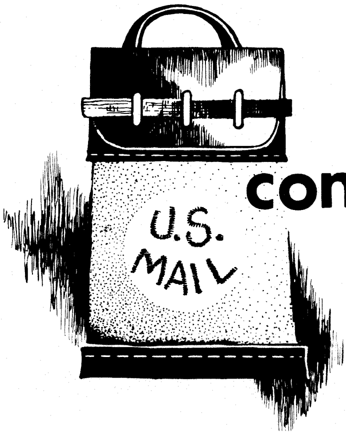
1. Brice, Ward, *Microprocessor/Microprogramming* Handbook, Tab Books, Blue Ridge Summit, Pennsylvania, 1975.

2. L. Krakauer, "Efficient Storage of Morse Character Codes," *BYTE*, October, 1976.

3. W. Sewell, WB5NYC, "If Sam Morse could see us now," *BYTE*, October, 1976.

4. J. Pollock, WB2DFA, "1000 WPM Morse Code Type," 73, January, 1977.

ham radio



comments

mobile color code

Dear HR:

I would like to propose a standard color-coding system for mobile antennas so that one mobiler could quickly tell visually what band another mobiler was operating on at that time. Using the standard resistor color code, the following colors would be used for each of the high-frequency amateur bands:

80 meters	gray-black
40 meters	yellow-black
20 meters	red-black
15 meters	brown-green
10 meters	red-gray

The red-gray code is chosen for 28 MHz because of its higher visibility over brown-black for 10 meters.

The two colors for each of the bands could be applied with colored tape (or paint) to the antenna loading coil, displayed on a pennant flying from the tip of the antenna, or shown with two colored tape strips on the rear bumper.

Ray Day, WB6JFD
Palos Verdes, California

fm repeater channel spacing

Dear HR:

I wish to thank Jerry Pulice, WB2CPA, for his fine article on direct synthesizers (August, 1977). The technical portions appear to be most needed in the fm community, especially for repeater usage (it's amazing what spurs on a mountain can do).

This letter is to comment on a statement made by Jerry in discussing the local-oscillator noise performance of his synthesizer. His statement, "In normal operation, crystal-controlled equipment would be able to maintain DX communications within 10 kHz of a repeater channel," is misleading and requires clarification.

As is well known, fm spectrum width is wider than the actual deviation. In fact, the bandwidth actually occupied by a narrowband fm signal is very close to twice the sum of the deviation and maximum modulation frequency. For voice operating a 5 kHz peak deviation (typical repeater operation), the bandwidth of the transmitted signal is 13 kHz or 6.5 kHz from the carrier or no-modulation frequency. A receiver with 15 kHz bandwidth would have interference when tuned within 7.5 +6.5 = 14 kHz of the transmitter. In actual practice a physical spacing of 40-50 miles (64-80km) is required for operation of 15-kHz channels, and 10-kHz spacing is totally impractical.

This condition is becoming a serious problem in the high density areas (like Southern California), not because of Jerry's statement, but primarily because of the transceiver manufacturers' insistence that their radios offer "400 channels" simply because they can be tuned to 400 discrete 5-kHz frequencies. In reality, there are only a theoretical maximum of some 100 channels from 146-148 MHz, but due to the repeater bandplans based on 15-kHz (or 30-kHz) spacing, only a theoretical maximum of some 80 channels are available. This is with *ideal* equipment. With real receivers and transmitters available today, this number is closer to 60 discrete simultaneous channels available from one location.

Robert O. Thornburg, WB6JPI
Studio City, California

300-Hz crystal filter for Collins receivers

Dear HR:

Referring to the article by W1DTY in September 1975 *ham radio,* I have found that the transistor impedance matching circuit, which has an input impedance of less than 15 thousand ohms, caused a 10 dB insertion loss when used with my 75S3C. To present a high impedance input, I modified the circuit as shown in fig. 1. This method gives the same output as the other filters in the receiver and properly matches the crystal filter.

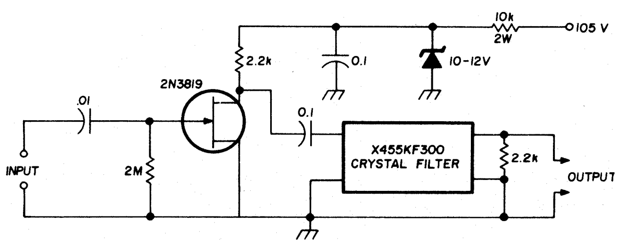


fig. 1. High-impedance matching circuit for the 75S3C.

The circuit is mounted on a small Veroboard held in place with nuts on the filter mounting screws. The whole unit fits nicely under the screening can without defacing the receiver in any way. I am very pleased with the performance of the filter. The attenuation is approximately 80 dB when 600 Hz from the center frequency.

C. H. Foulkes, G3UFZ
Herts, England

IC-crystal oscillator

Dear HR:

VK2ZTB is to be congratulated for his excellent article in March 1976 *ham radio.* However, it did have one omission in the section on IC oscillators. The Motorola MC12060 and MC12061 are specifically designed as series-mode crystal oscillators. The MC12060 covers 100 kHz to 2 MHz and the MC12061 2 MHz to 20 MHz. Both ICs produce sinusoidal, ECL and TTL outputs. The MC12061 has also been operated as an overtone oscillator by connecting the components as shown in **fig. 1.** Also, not mentioned in the data sheet* is the ability to get twice the oscillator frequency by tying the sinewave outputs (pins 2 and 3) together. This connection performs a full-wave rectification of the sinewaves to achieve the doubling. These two features allow VHF oscillator signals to be easily developed in one IC package with only one tank circuit.

*Motorola MTTL Phase-Locked Loop Components. MC12060 Data Sheet.

Ron Treadway, W7EKC
Scottsdale, Arizona

low-resistance measurements

Dear HR:

In the September, 1977, issue of *ham radio,* the accuracy of a low-resistance measurement method is indicated as being in the range of 1 or 2%.

This is not always the case, however, since an error of plus 100% will occur when the resistance being measured, R_x is equal to the resistance of the millivoltmeter, R_m. In fact, only when the resistance of the millivoltmeter is infinite is the method strictly accurate.

Since the millivoltmeter shunts the unknown resistor and the total line current is 100 mA

$$(V_m/R_m) + (V_m/R_x) = 0.1 \text{ ampere}$$

Hence $10\,V_m = R_m \times R_x/(R_m + R_x)$

where V_m is the voltage across the millivoltmeter. In other words, the method gives the resistance *not* of the unknown resistor, R_x but of the combined resistance of R_x and R_m in parallel.

When selecting the millivoltmeter, one should be sure that its resistance is at least 50 times (and preferably 100 times) the highest resistance to be measured to maintain the stated accuracy of this useful method. The percentage accuracy of the method, insofar as the resistance of the millivoltmeter is concerned, is limited to $100\,(R_x/R_m)$.

Ed Sampson, W1PT
Brockton. Massachusetts

magnetron development

Dear HR:

I have read with considerable interest W1HR's article on "Solid-State Microwave RF Generators" in the April, 1977, issue of *ham radio.*

In the opening paragraphs where W1HR briefly reviewed the history of devices used to generate microwave power leading up to the modern solid state devices, I felt that some reference should have been made to the considerable work done by Dr. Eric Megaw, G6MU, on the split and multi-segment anode magnetrons principally done for the Marconi Company. Eric Megaw worked at the General Electric Co. Ltd., Research Labs at Wembley. Also the development of the first high power cavity magnetron developed at Birmingham University in February, 1940, which gave 400 watts CW at 9.8cm, and later pulse types for our radar, might have been mentioned.

You are no doubt well aware that microwave activity in the UK is growing significantly, largely under the direction of Dain Evans, G3RPE, and his associates. We how have three 10-GHz beacons operating at the Isle of Wight, Alderney (Channel Islands), and Romford in Essex. Microwave Associates at the Luton factory are considering another beacon in this band, and one in Aberdeen is projected.

G.R. Jessop, G6JP
General Manager, RSGB
London, England

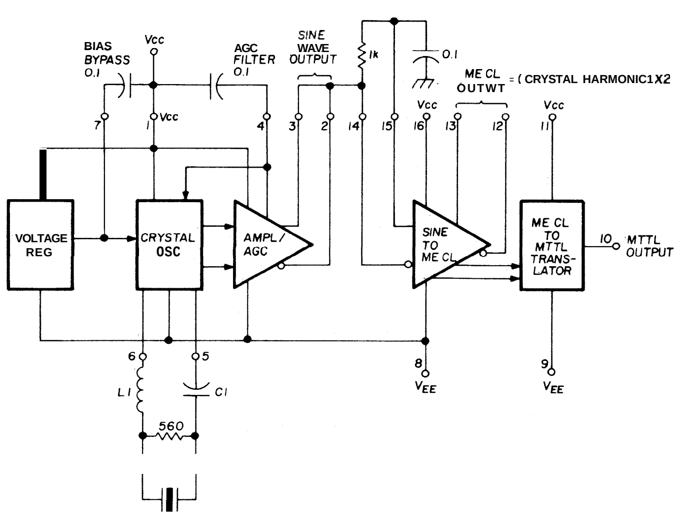
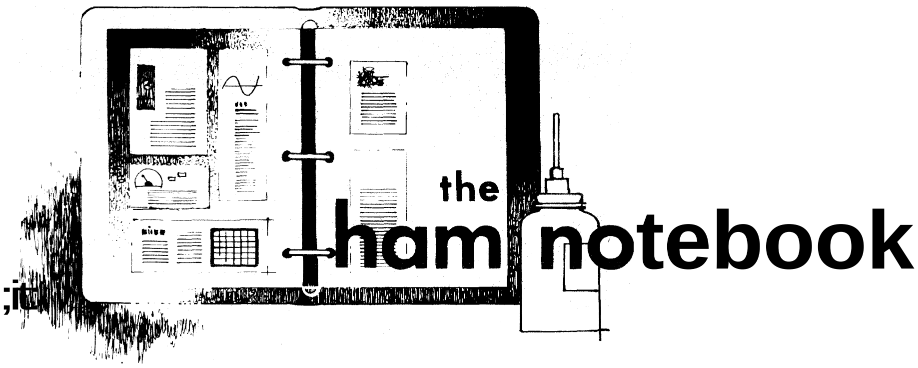


fig. 2. Pin connections for using the MC12061 as an overtone oscillator.



CMOS programmable divide-by-N counter

Most divide-by-N counters require either a complicated system of gates or pin strapping; changing the divisor is difficult for either method. If the speed of the circuit is such that a CMOS IC can be used, a single connection change will permit division by any integer between 2 and 10. This can be done with a simple, single-pole switch.

This divide-by-N counter uses an RCA CD4017A Johnson decade counter. In normal operation, the counter provides decimal outputs that are low and go high only at their respective time slots. For divide-by-N operation, the reset line of the counter is connected to one of these outputs, depending on the desired divisor. As the count progresses, each decimal output goes high and then low until the count reaches the one connected to the reset line. The counter is then reset, and the output of the divider appears on the 0 line.

Fig. 1 shows the circuit connected to divide by 7.

It should be noted that to divide by 10, no feedback is used and the internal gating of the counter is permitted to function in the normal manner. Division by 0 and 1 are not permitted. For division greater than 10, this circuit can be cascaded with other divider schemes.

Ken Stone, W7BZ

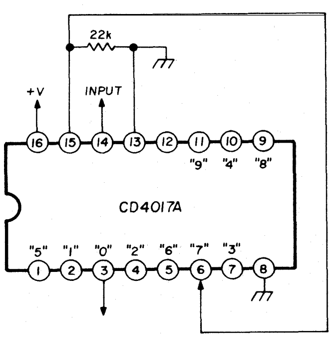
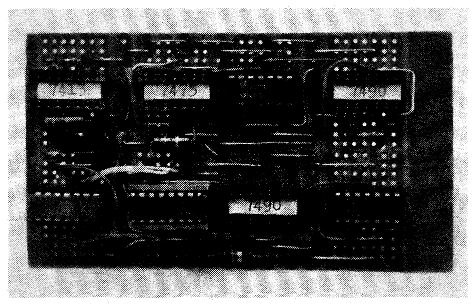


fig. 1. The CD4017A connected as a divide by *7* counter. The resistor is used to hold the reset line low. When the appropriate number is reached, that output and the reset line are driven high, resetting the counter. To divide by other integers, pin 15 should be connected to the desired output. For example, pin 1 for a divide by *5*, or pin 7 for a divide by 3.

socket label for integrated circuits

How many times have you borrowed integrated circuits from other boards and then when you went to return them, forgot which chip went in which socket? Well, it happened once too often for us. We decided that anything would be better than retracing circuits to determine which integrated circuit goes where.

Various labeling methods were tried. Decals were too hard to apply in cramped spaces; painting was out due to a total lack of manual dexterity. Finally, strip labeling worked and has been adopted for labeling old and new boards. This is done by typing,



or printing , the IC type on strips of gummed paper which are then trimmed to fit the groove between the rows of pins. The result is neat, accurate, and inexpensive. In addition, the labels are visible only when the integrated circuit is removed from its socket.

John M . Franke, WA4WDL
Norman J. Cohen, WB4LJM

using the National NCL-2000 with the Drake T-4XC

Af-ter finally moving to a QTH

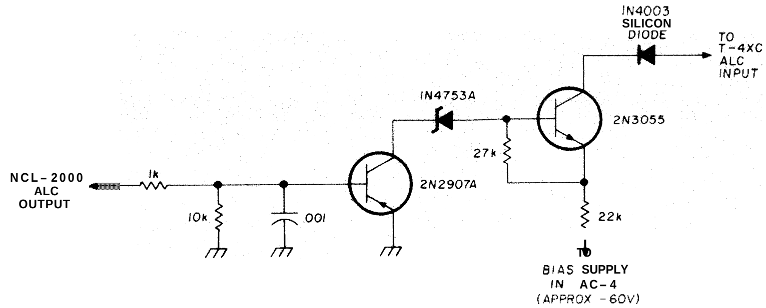


fig. 2. Schematic diagram of the amplifier for use between the NCL-2000 amplifier and a T-4XC. The zener diode has a 36-volt rating. A 1N4003 or equivalent diode is used in the collector of the 2N3055.

where I could use my NCL-2000, I found that the alc output level was insufficient to drive the T-4XC. To overcome this difficulty, I designed a simple amplifier to make the equipment compatible. No adjustments are necessary, and the circuitry can be conveniently built into the Drake AC-4 power supply.

Parts selection is not critical. If a higher voltage transistor is used in place of the 2N2907A, the zener diode can be eliminated. However, a very low-leakage transistor will be necessary to keep from turning the 2N3055 on. The 2N3055 was probably overkill for this application, but it was the only high-voltage transistor in the author's junkbox.

Without the modification, the NCL-2000 would severely flat top unless careful attention was paid to the GAIN control on the transmitter. After modification, no flat topping is evident on the monitor scope even if the GAIN control is operated wide open. Of course, the severe compression renders speech unintelligible.

Edwin R. Ranson, K5ER

a simple adjustable IC power supply

Last year, National Semiconductor introduced the first 3-terminal adjustable positive voltage regulator. When used with a few external components, the LM117 is capable of delivering voltages from 1.25 to 37 volts at 1.5 amps. Mounted in a TO-3 case, this device is also protected against current and thermal overloads.

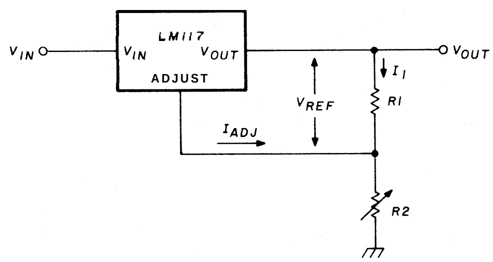


fig. 3. Basic configuration for the LM117 adjustable IC voltage regulator.

In normal operation,[1] the LM117 regulator develops a 1.25 V reference (V_{REF}) between its output and adjustment terminals, as shown in **fig. 3.** Since this reference voltage is constant across R1, a constant current flows through the output set resistor R2, so that the output voltage can be calculated from

$$V_{OUT} = V_{REF}\left(1 + \frac{R2}{R1}\right) + I_{ADJ}R2$$

Typically, $V_{REF} = 1.25$ V, and $I_{ADJ} = 50\ \mu A$, so that

$$V_{OUT} = 1.25\left(1 + \frac{R2}{R1}\right) + (50\ \mu A)(R2)\ \text{volts}$$

As shown in **fig. 4,** the input voltage is provided by a simple bridge rectifier and capacitor input filter arrangement delivering approximately 37 Vdc. The 0.1 μF disc bypass capacitor C2 is strongly recommended if the regulator is physically located some distance from C1. C3 (10-μF tantalum) is added to improve the ripple rejection.

Although the LM117 is capable of good load regulation, typically 0.3 per cent at constant junction temperature, the 240-ohm current-set resistor R1 should be connected directly to the regulator's output terminal, rather than near the load. When external capacitors are used with any IC regulator, it is wise to add protective diodes to prevent the capacitors from discharging through low current points in the regulator. Therefore, CR1 protects against C4, and CR2 protects against C3.

reference

1. **Linear Data Book, National Semiconductor Corporation, Santa Clara, California 95051, June,** 1976.

Howard Berlin, W3HB

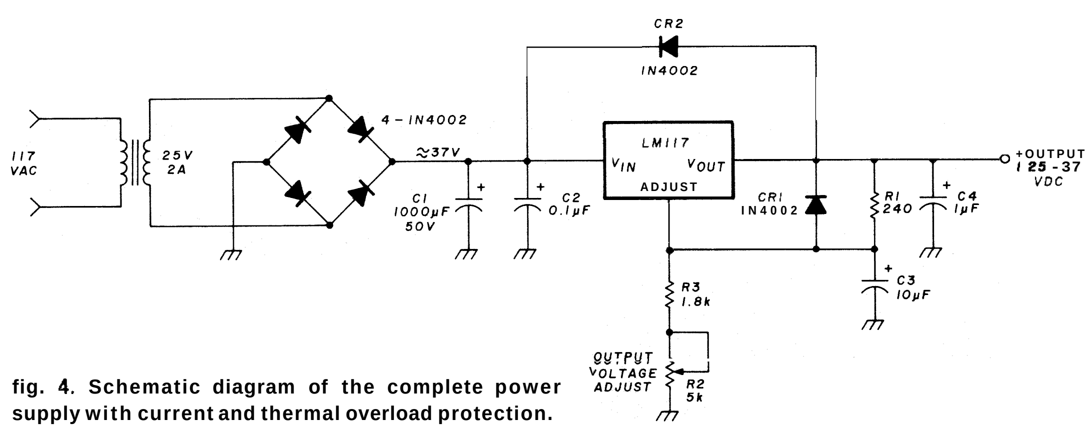
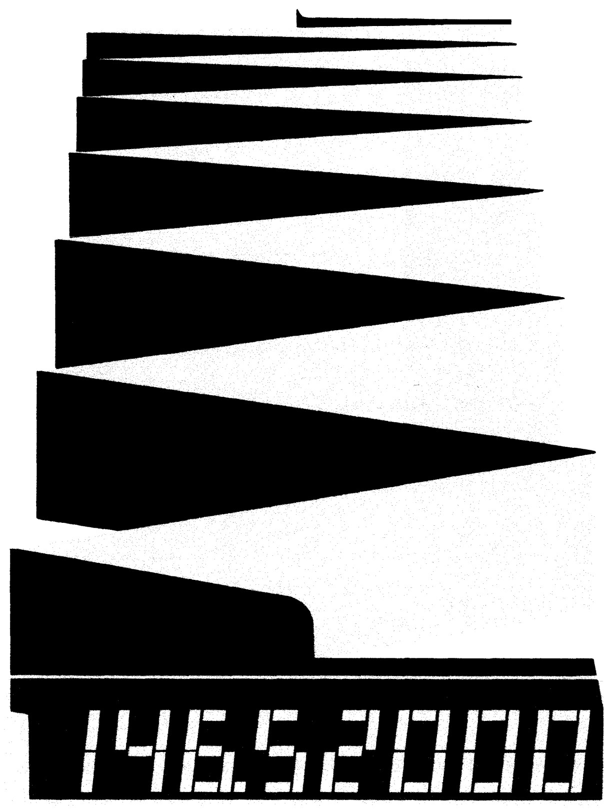


fig. 4. Schematic diagram of the complete power supply with current and thermal overload protection.

$1.50

radio

FEBRUARY 1978



ham radio magazine

FEBRUARY 1978

volume 11, number 2

T. H. Tenney, Jr., W1NLB
publisher

James R. Fisk, W1HR
editor-in-chief

editorial staff

Charles J. Carroll, K1XX
Alfred Wilson, W6NIF
assistant editors

Patricia A. Hawes, WA1WPM
Thomas F. McMullen, Jr., W1SL
Joseph J. Schroeder, W9JUV
associate editors

Wayne T. Pierce, K3SUK
cover

publishing staff

Harold P. Kent, WA1WPP
assistant publisher

Fred D. Moller, Jr., WA1USO
advertising manager

James H. Gray, W1XU
assistant advertising manager

Therese R. Bourgault
circulation manager

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

U.S. and Canada: one year, $12.00
two years, $22.00
three years, $30.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

Foreign subscription agents are
listed on page 101

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents

a second look

by Jim Fisk

As this issue goes to press, it appears that the launch of the next amateur communications satellite, AMSAT-OSCAR D, is imminent (after it is in orbit this satellite will be known as OSCAR 8). Like its famous predecessors, OSCARs 5, 6, and 7, this new "bird" has two transponders: a two-to-ten meter unit similar to that used in OSCAR 7 called *Mode* A, and a two-meter, 70-centimeter transponder designated *Mode* J. The Mode J transponder was built by members of the Japan AMSAT Association in Tokyo; a similar combination of input/output frequencies was used in the short-lived OSCAR IV spacecraft back in 1966.

The new OSCAR will be launched from NASA's Western Test Range in California as a "piggyback" payload aboard the second stage of the two-stage Thor-Delta launch vehicle which will carry NASA's Landsat-C earth resources technology satellite into orbit. Because of scheduling at the Western Test Range, and the complex pre-launch checkout of Landsat-C, it's impossible to pinpoint the exact launch date, but late reports from AMSAT indicate it will be sometime in early March.

The new spacecraft is a 38 cm rectangular solid 33 cm high, weighs 27 kg, and is solar powered. The solar cells, combined with the 12-cell rechargeable nickel-cadmium battery, should be adequate to power the satellite in Mode A for several years. The receiving antenna for both modes is a turnstile comprised of four 48 cm lengths of 12 mm carpenter's rule. Four permanent magnets located inside the spacecraft provide stabilization; this is the same technique used in OSCARs 6 and 7. The polarity of the magnets is such that the top of the satellite always points toward the earth's magnetic north pole. Permalloy damping rods mounted behind the solar panels are designed to reduce the spin of the spacecraft; their operation is similar to a shorted transformer turn as it cuts the lines of flux of the earth's magnetic field. OSCAR 7 used the same system with good success.

The spacecraft will be automatically powered up upon ejection from the Thor-Delta launch vehicle over northern Greenland. It is designed to come on in Mode J; the Mode A transponder will not be turned on until the satellite is almost completely stabilized in orbit, which may take as long as a week. This is because the 10-meter dipole antenna cannot be deployed until the spacecraft's spin rate is less than 1 revolution per minute; otherwise the antenna may be severely damaged. The deployment process takes about 15 seconds and cannot be reversed — the elements can't be retracted once they are extended — so correct deployment is crucial.

OSCAR 8's orbit is planned to be sun-synchronous, with passes repeating at approximately the same time each day on a one-day cycle (as opposed to the two-day cycle of OSCARs 6 and 7). Since the altitude of OSCAR 8's orbit, at 900 km, is just over half the altitude of OSCARs 6 and 7, the maximum communications range will be slightly shorter. The usable time on an overhead pass will be about 18 minutes instead of the 22 minutes provided by OSCAR 7, and the horizon range will be 3220 km (down slightly from the 3940 km horizon range of OSCAR 7). In practical terms this means that transatlantic communications will still be possible with OSCAR 8, but not as often as with OSCAR 7.

One of the big advantages of the 900 km sun-synchronous orbit is that keeping track of OSCAR 8 is going to be much simpler than it was for earlier amateur satellites; it will come into range at nearly the same time every day — the overhead descending node pass is planned for 9:30 AM local time. The satellite's anticipated useful operating lifetime is three years.

Since the prime mission of the OSCAR 8 spacecraft is to use the Mode A transponder for the ARRL OSCAR educational program in schools, the spacecraft may be left in Mode A during weekdays and put into Mode J on weekends. Because of the relatively high current drain of the Mode J transponder, however, the power budget may not support the Mode J transponder for continuous full-time operation over an entire weekend. The spacecraft may also be switched to Mode J during the evening hours in the Western Hemisphere, depending on the burden to the command stations and the condition of the on-board batteries.

The Mode A transponder on the new spacecraft has the same frequency passband as OSCAR 7 (input between 145.85 and 145.95 MHz, output between 29.40 and 29.50 MHz). Approximately −95 dBm is required at the transponder input terminals for an output of one watt; this corresponds to an effective radiated power from the ground of about 80 watts. The 250 mW telemetry beacon operates at 29.402 MHz.

The Mode J transponder operates with an input frequency passband between 145.90 and 146.00 MHz — the output is between 435.10 and 435.20 MHz. Power output is 1 to 2 watts PEP, and the output is inverted (upper-sideband uplink signals become lower-sideband downlink signals). Uplink sensitivity for 1 watt output is −105 dBm which corresponds to an effective radiated power from the ground of about 8 watts (note the greatly improved sensitivity of this mode, and keep your power down). A 100 mW beacon at 435.095 will carry telemetry information.

Jim Fisk, **W1HR**
editor-in-chief

understanding and using electronic counters

Only **a few short years ago,** it was extremely rare to see an electronic counter outside of a laboratory or a specialized service installation. Today counters can be found in ham shacks all over the world. Of course the reason for this proliferation is obvious — the integrated circuit, and in particular, medium- and large-scale integration. In the early sixties, a typical 10-MHz counter weighed nearly 120 pounds (55kg), occupied about 5.75 cubic feet (165,000 cubic centimeters), and dissipated approximately 600 watts as heat. By way of contrast, a 520-MHz counter currently produced by the same manufacturer weighs 4.75 pounds (2.16kg), has a volume of approximately 213 cubic inches (3890 cubic centimeters), and dissipates less than 20 watts.

As size has decreased, so has cost. That antediluvian counter cost $2600 in ''1966 dollars;'' today you can buy a counter for under $100 if you want a bare-bones instrument, and can get a 250-MHz multifunction counter for less than $400. Because of today's relatively low costs, counters have become a versatile tool in the ham station and on the work bench. If you have one, this article may help you make better use of it. If you are planning to buy one, it may help you to decide what to look for.

You may have noticed that the title of the article uses the term *electronic,* rather than *frequency,* counter. This was not a pedantic choice; electronic describes the type of counter and is inclusive of all functions that the counter may perform, only one of which may be the measurement of frequency. We shall discuss these various functions, although emphasis will be placed on frequency measurement, which is of primary interest to the average ham.

Before discussing the applications and limitations of the frequency counter, it is important to cover the method by which frequency is measured by the counter. Regardless of the type and complexity of the instrument, all counters measure frequency by comparing the frequency of the input signal with a known frequency or time period. **Fig. 1** shows the basic functional blocks of a typical counter. The main function of the signal conditioner is to convert the input signal to one whose amplitude and waveshape are compatible with the internal circuitry or logic of the counter. It generally includes an amplifier to increase the amplitude of the incoming signal, and may also contain an attenuator for input signals of high amplitude, trigger level and slope selection circuits, and so on. No matter how the signal is processed, the output of the signal conditioner is a pulse train in which each pulse corresponds to one cycle or event of the input signal.

The conditioned signal is applied to a gating circuit, which is shown symbolically as a single logic gate, but which is actually a more complex circuit. The gate is opened for a predetermined, accurate time interval, during which the signal passes through to the decade counters. These counters count the number of pulses which are gated through, and transfer the count to the display. The number of decade counters determines the number of digits which are displayed, one counter being required for each digit. The display can utilize any type of visual readout device, such as gas-discharge numeric tubes, light-emitting diode arrays, or liquid-crystal displays.

Since the decade counters count the number of pulses which pass thorugh the gate, it follows that the accuracy of the instrument is a function of the time that the gate is open. This interval is, in turn, a function of the time-base accuracy. The time-base oscillator in the modern counter is invariably a crystal-controlled oscillator operating at a frequency between 1 and 10 MHz, although there have been counters made in the past which used crystal frequencies as low as 100 kHz, or even used the ac line frequency as a time base. Even though the oscillator frequency must be divided, crystals in the 1- to 10-MHz range are used because they are inherently more stable than those which work at lower frequencies; the optimum range for stability is between 4 and 10 MHz for most types of crystals.

The divided time-base frequency drives the gate-

By Robert S. Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022.

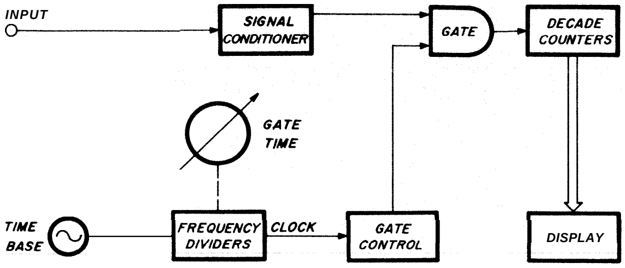


fig. 1. Functional block diagram of the basic frequency counter.

control circuit, which controls the gate-time interval in accordance with the divided time-base frequency. To explain the need for dividing the time-base frequency, we must at this time discuss resolution, or the smallest frequency increment which the counter displays.

Let us assume that the frequency of the time-base oscillator is 10 MHz, and that the gate control holds the gate open for exactly 10 clock pulses (clock being the term used to designate the time-base signal or the signal derived, through the dividers, from the time base). Since each cycle from the 10-MHz oscillator has a period of 0.1 microsecond, the gate will be open for 1 microsecond. If a 1-MHz input signal were being measured, only one pulse would be gated through, and the counter would display a **1.** Frequencies below 1 MHz might or might not produce a reading at all, and those above 1 MHz could be read only to within one digit of the nearest megahertz (more about this later). Thus, the resolution would be 1 MHz at best, an obviously unsatisfactory arrangement.

Suppose, instead, that the time-base frequency were divided down to 10 Hz, or a period of 0.1 second. The gate will now be open for 1 second, and 1 million pulses from a 1-MHz input will be counted. Now the counter will display 1000000, which provides us with a resolution of 1 Hz. Thus, the resolution is the reciprocal of the gate time, and in fact, some counters with selectable gate times have the switch positions designated by the gate time.

The limiting factors governing resolution are the number of digits in the display and the tolerable gate time. Usually 0.1 Hz is the smallest resolution practical, in that it involves a 10-second gate time and a 9-digit display up to 99.9999999 MHz. The gate time can be reduced in a computing counter, but that is outside the scope of this discussion.

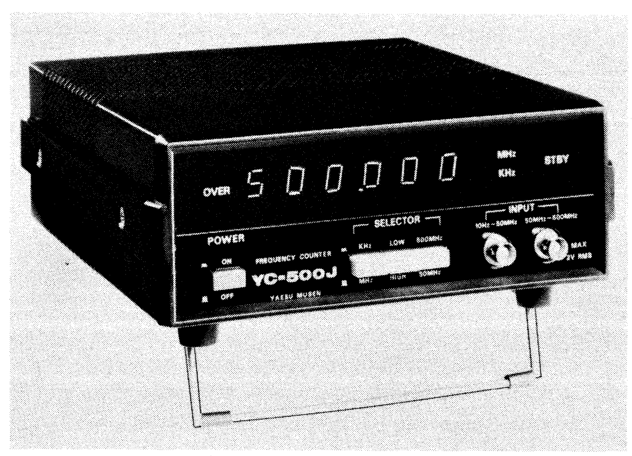
Of course, it is not always necessary to read frequency to a tenth of a hertz, nor is it particularly convenient to have to wait for a 10-second count, By selecting the appropriate output from the frequency-divider chain, you can reduce the gate time and resolution to values which may be more appropriate to the measurement. The normal range of gate times is typically between 1 millisecond and 10 seconds, corresponding to resolutions of 1 kHz to 0.1 Hz.

The number of digits in the display can be reduced by switching both the displays and gate times, a technique which is used in many low-priced counters having a 5-digit display. A 2-position switch is used to select gate times of 1 second (I-Hz resolution) and 1 millisecond (I-kHz resolution). When the gate time is 1 second, the five decade counters can produce a display up to 99,999 Hz; when the gate time is 1 millisecond, up to 99,999 kHz or the frequency limit of the counter can be displayed. Thus by switching the clock, the equivalent of eight digits is obtained, with overlap between the two readings. This is an economical, Gut oftentimes inconvenient, way of obtaining improved resolution.

It should be reiterated at this point that a counter displays a pulse count. Whether the display reads out in Hz, kHz, or MHz is simply one of convenience and the location of the display decimal point. The decimal point is either fixed, or is switch selected with the gate time, and its position is independent of the actual count.

time-base accuracy

It should be apparent from the preceding discussion that the time-base oscillator is the most critical part of the counter, in that it determines the overall accuracy of the instrument. Let us examine its effect, in terms of the specifications usually given for a counter.

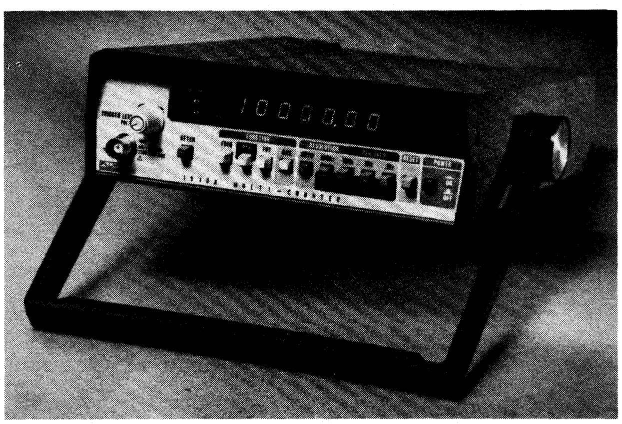


The Yaesu YC-500 Frequency Counter has a frequency range of 10 Hz to over 500 MHz. Its six-digit display provides the equivalent of eight digits when the gate time is switched. A room-temperature crystal, TCXO, or ovenized crystal time base may be ordered *(photo courtesy Yaesu Electronic Corporation).*

First of all, the accuracy of the time base, either in per cent or parts per total, is directly translatable to the measurement of frequency, period, interval, or any other function which the counter may be capable of measuring and which utilizes the time base. This holds true, regardless of the magnitude of measurement. For example, if a I-MHz time-base oscillator is off frequency by 2 Hz, that represents an error of 0.0002 percent. The gate interval, therefore, will also be in error by the same percentage, and the displayed count will have the same error. If the frequency being displayed is 50.000000 MHz, the error will be 100 Hz. If the time-base frequency is high, the displayed count will be low, since the higher the frequency, the shorter the gate time. If the time-base frequency is low, the opposite will hold true.

Time-base accuracy specifications should include the parameters listed in **table 1,** although most lower-priced instruments may omit one or more. Typical values for the various types of oscillators are included as examples. It can be seen that temperature change has the greatest effect on frequency. In the examples listed, the specification for temperature stability can be improved by one order of magnitude by using a TCXO (temperature-compensated crystal oscillator) instead of a room-temperature crystal. In the real world, however, a good room-temperature crystal may be better than a poor TCXO; you must compare the specifications.

The oscillator aging rate is not as important, since this will manifest itself as a gradual change in frequency, and is predicated on the oscillator running continuously. If the counter is designed so that the oscillator circuit is powered as long as the instrument is connected to the primary power source, the specified aging rate is valid. If the



The Fluke 1910-A Multi-Counter is one of a series which provides frequency, period, period-average, ratio, and totalize functions. The 1910A is rated to 125 MHz; the 1911A and 1912A are similar in appearance and will measure frequencies to 250 and 520 MHz, respectively *(photo courtesy John Fluke Manufacturing Company).*

table 1. typical specifications for time-base oscillators.

	room temperature crystal	TXCO*	oven oscillator
Aging rate (long term stability	5×10^{-7}/mo	3×10^{-7}/mo	5×10^{-10}†
Temperature, 0-50°C	5×10^{-6}	5×10^{-7}	7×10^{-9}
Line voltage, ± 10%	1×10^{7}	5×10^{8}	5×10^{9}
Short-term stability per day			1×10^{-10}

*Temperature-compensated crystal oscillator
†After 24-hour warm-up
‡rms/sec

oscillator is deenergized along with the rest of the counter when the instrument is turned off, however, the aging specification means little or nothing.

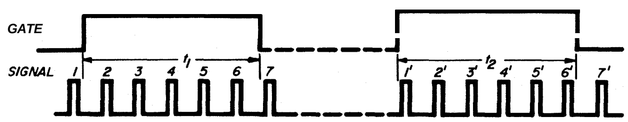


fig. 2. Gate time and signal pulse train, showing ±1 count ambiguity.

Short-term stability is generally specified only for very stable, ovenized oscillators and is pertinent only to laboratory-type measurements.

Time-base errors can be corrected by recalibrating the oscillator against a known standard or against WWV. Virtually all counters incorporate an adjustment control for this purpose. The techniques used in recalibrating the oscillator will be covered later in this article.

frequency-measurement accuracy

Although the preceding discussion of time-base

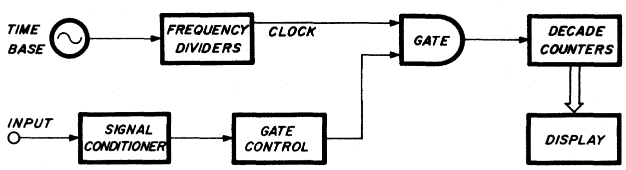


fig. 3. Functional block diagram of an electronic counter configured for period measurement.

accuracy would appear to account for any inaccuracies in the measurement of frequency, such is not the case. The specification for the frequency-measurement accuracy of all electronic counters is invariably stated as ± time-base accuracy ± 1 digit. The last term of that statement is known as the 1-count ambiguity — but what does this mean?

Fig. 2 shows the signal under measurement and its relationship to the gate. Although the successive gate times, t_1 and t_2, are equal in duration, the gate is not synchronized with the signal.

Therefore it is possible, during gate time t_1, for five signal pulses (numbered 2 through 6) to be gated, while during time t_2, six signal pulses (numbered 1' through 6') may be gated. Thus there is always an irreducible ±1-count ambiguity in the least significant digit of the display.

The per cent of error due to the 1-count ambiguity is reduced as the measured frequency increases, since it becomes increasingly less significant compared to the total count. The maximum error is an inverse function of the frequency being measured and the number of pulses being counted, *i.e.* the gate time, and is expressed as

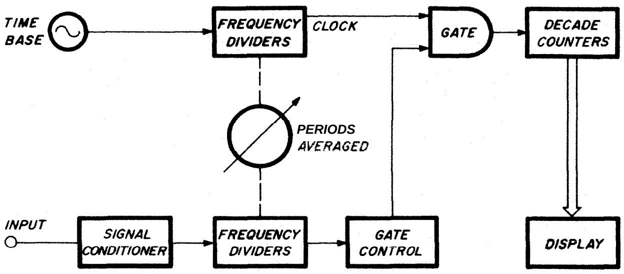


fig. 4. Functional block diagram of a period-averaging counter.

$$\%\ Error = \pm \frac{100}{f \cdot t}$$

where f is the frequency in Hz, and t is the gate time in seconds.

From the above equation, it can be seen that measuring a 10-MHz signal with a 1-second gate time will be subject to a ±0.001 per cent error. However, measuring a 20-Hz signal with the same gate time may result in a counter display between 19 and 21 Hz, a ±5 per cent error. This would not be very satisfactory if you were attempting to calibrate the low-frequency end of an audio oscillator, and must be taken into account.

period and period-averaging measurements

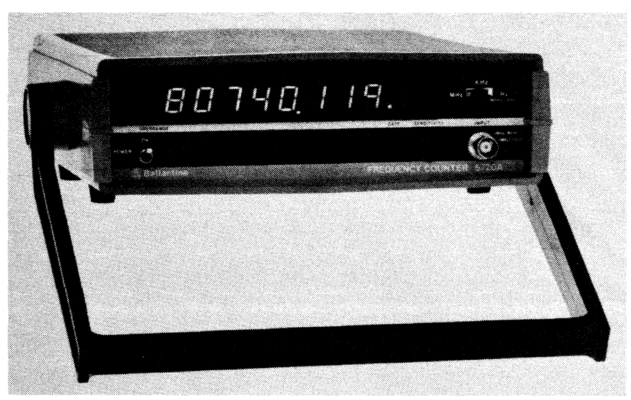
One of the ways in which accurate low-frequency measurements may be made is to measure the period of the signal, rather than the frequency. Since the period of a signal is the reciprocal of its frequency, the frequency can be calculated accordingly. It might also be expected that a simple reciprocal arrangement of the functional blocks of an electronic counter would provide a measurement of period, which turns out to be true.

In **fig. 3**, the time-base and signal inputs have been interchanged. If the gate-control circuit is configured so that the gate is open for one period of the input signal, and the 1-MHz clock is applied to the gate input, a series of pulses having a period of 1 microsecond will be gated through to the decade counters. Therefore the counter will indicate the period of the input signal in microseconds. If the clock frequency were reduced to 1 kHz, the counter would display the signal period in milliseconds.

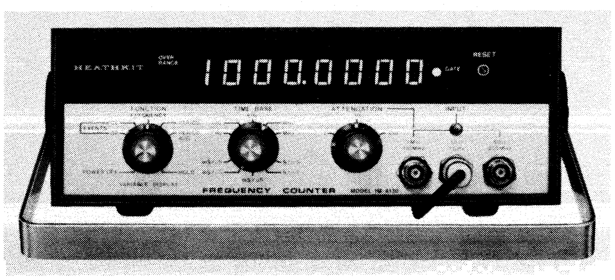
Let us now reconsider the frequency measurement of the aforementioned 20-Hz signal to see how we can improve the possible ±5 per cent error which can occur with a I-second gate time. The 20-Hz signal period is 0.05 second, so that if the counter were configured to measure period, it would display 50,000 microseconds. Thus, the number of significant digits in the display has been increased from two to five, and the gate time reduced from 1.0 to 0.05 second. Now if you were calibrating the audio oscillator and wanted a 1 per cent dial accuracy, you could accept any reading between 49,500 and 50,500 microseconds, subject to the correction for period-measurement accuracy.

Period measurements are inherently less accurate than frequency measurements because it is the signal, rather than the time base, which controls the gate time. Noise on the input signal, regardless of the measurement mode, causes an uncertainty in the point at which the trigger circuit in the signal conditioner switches. (It is the trigger circuit which converts the input signal to a waveform which is compatibie with the counter's circuitry.) It the noise is not great enough to cause false triggering which would result in more or less output pulses than correspond to the input, no significant error is introduced in a frequency measurement.

For period measurements, however, this uncertainty results in an error in the gate time, since the in-



Ballantine's model 5720A Frequency Counter covers the range from 10 Hz to more than 80 MHz and provides frequency and ratio measurements. This counter also includes an audio multiplier circuit for input frequencies from 50 Hz to 1 kHz which provides resolution of 0.01 Hz with only 1-second measurement time (*photo courtesy* Ballantine Laboratories).



Heath's model IM-4130 is capable of measuring period, period average, events (totalizing). and frequency over a 5-Hz to 1-GHz range. Since it has provisions for connecting an external time base, ratio measurements can also be made, as explained in the text *(photo courtesy Heath Company).*

put signal controls the gate time. This error is known as trigger error, and is part of the instrument specification for period measurement, usually expressed as ± time-base error * trigger error ± 1 count. Notice that the trigger error has been added to the previously discussed expression for frequency-measurement error. For low-frequency noise on a sine-wave input, the approximate worst-case errors are ± 3 per cent for a 20-dB signal-to-noise ratio, ± 0.3 per cent for a 40 dB signal-to-noise ratio, and ± 0.03 per cent for a 60-dB signal-to-noise ratio.

In addition to the trigger error caused by noise, the stability of the input signal may be such that successive gate times are of differing durations. Even though the differences may be minute, they will manifest themselves as a continuously changing display on the counter, especially at high resolutions. This is not to be considered a counter error, since it does not occur with a stable input signal.

Period errors may be minimized by averaging the readings over several periods of the input signal. If the input-signal frequency is divided to a lower frequency. the gate will remain open for a multiple of the input-signal period, so that the counter will display the number of clock pulses for 10, 100, 1000, or more periods. A typical counter configuration for the period-averaging mode appears in **fig. 4.** The frequency-divider chain is split so that both the time-base oscillator and/or signal frequencies are divided to obtain the desired resolution and number of periods which are to be averaged. The counter will display the period measurement, regardless of the number of periods averaged, simply by having the

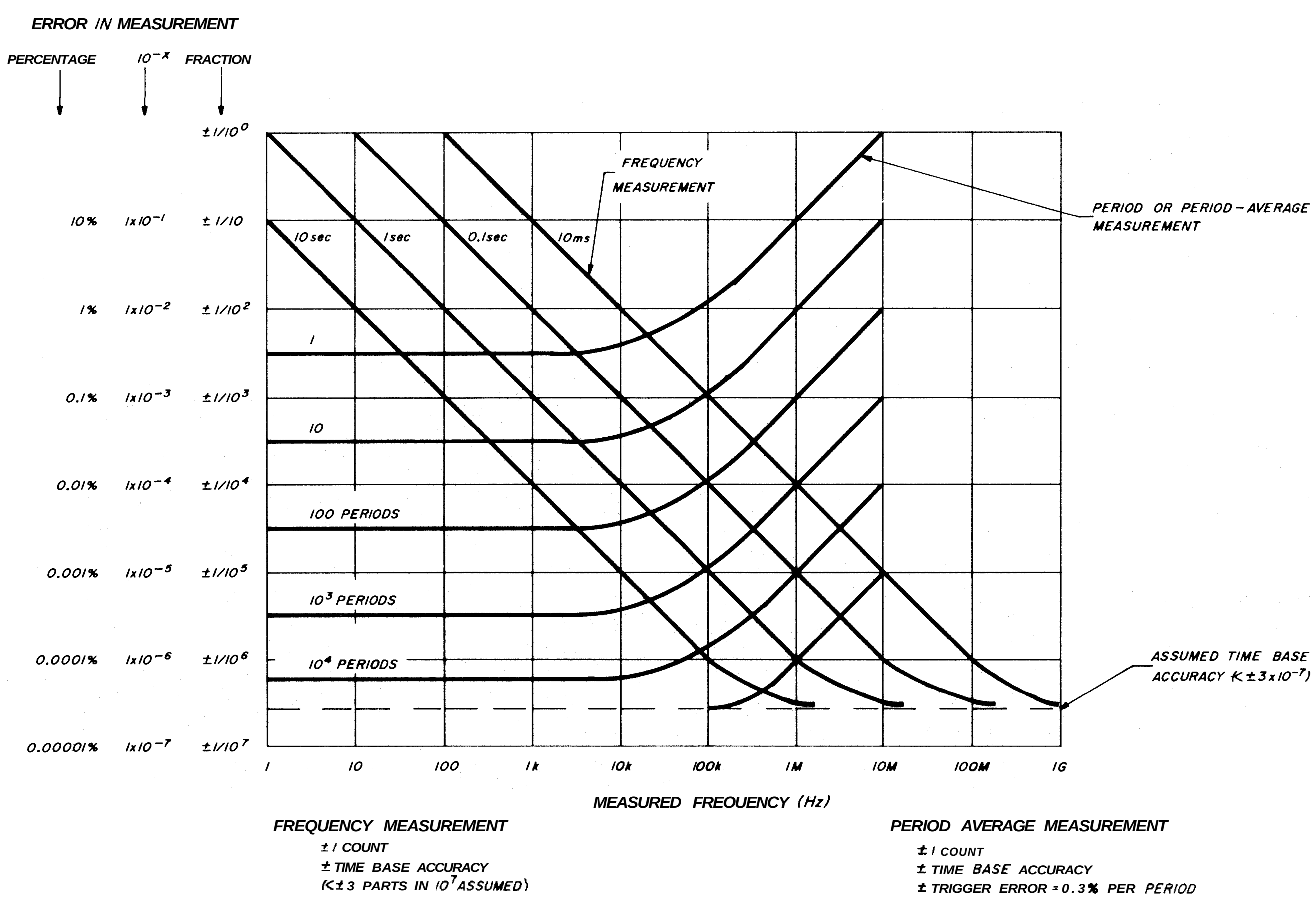


fig. 5. Measurement accuracy of a counter having a 10-MHz time base with an assumed accuracy of better than 3 x 10^{-7}.

display decimal point moved as the periods-averaged switch is changed.

Period averaging reduces the possible trigger error by a factor equal to N, the number of periods averaged, so that the error for this mode is $\pm$ time-base error $\pm$ (trigger error)$/N \pm 1$ count. If enough periods are averaged, the trigger error can be reduced to a value which may be of little significance. It must be remembered, however, that the gate time increases by the same factor, which may make the measurement time quite long. For example, a 20-Hz signal has a period of 0.05 second; averaging 100 cycles results in a gate time of 5 seconds; averaging 1000 cycles entails a 50-second gate time, which is normally too long for convenient measurements.

From the preceding discussion, we can deduce that there is a point below which period or period-averaging measurements provide a more accurate reading than a correspondng frequency measurement. This can be calculated, taking into account the I-count ambiguity, time-base error, trigger error, and gate time. More conveniently, it can be plotted, as shown in **fig. 5.** These curves apply to a counter having a 10-MHz time base of the accuracy specified, and indicate which measurement mode should be used for the desired measurement accuracy.

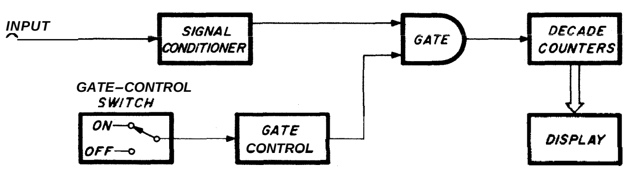


fig. 6. Functional block diagram of a totalizing counter. The gate-control switch can be either a manual switch or an internal switching circuit.

In many instances, measuring the ratio of two frequencies is a time-saving procedure. A typical case might involve designing or troubleshooting a phase-locked loop, where the output frequency is a discrete multiple of a reference oscillator. Since the output frequency may be divided by a factor of up to several thousand within the loop, an error or glitch causing a one-count error in this division may not be readily apparent unless a ratio measurement is made.

In conjunction with **fig.** 1, we discussed the method by which frequency is measured. Another way of defining this measurement is to state that the counter displays the ratio of the input frequency to the clock frequency. By using an internal clock whose frequency is known, the ratio can be displayed in megahertz, kilohertz, or hertz. If an external signal were used in place of the time-base oscillator,

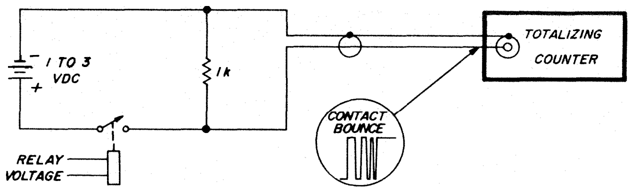


fig. 7. Using a totalizing counter to measure contact bounce. When relay voltage is applied, the counter will display the number of contact bounces.

the counter would still display the ratio of the two frequencies, except that it would no longer be in hertz or a multiple thereof (unless the external frequency were the same as the time-base frequency). Counters which provide specifically for ratio measurements incorporate provisions for changing the display to a dimensionless number, and position the decimal point accordingly.

Many counters do not have an apparent capability of measuring ratio, but can actually be used in this mode. If the counter has provisions for using an external time-base oscillator, the reference signal against which ratio is to be measured can be introduced into the external time-base connector. It is necessary that the amplitude of this external reference signal be as specified for the counter being used, that its frequency be within the range that the counter will accept as an external time base, and that the internai time-base oscillator frequency be known.

The ratio of the input signal frequency to the external reference frequency is determined from the expression

$$\frac{f_{sig}}{f_{ref}} = \frac{f_{ctr}}{f_{int}}$$

where

f_{sig} is the input frequency
f_{ref} is the external reference frequency
f_{ctr} is the frequency displayed on the counter
f_{int} is the internal time-base oscillator frequency.

totalizing

Perhaps the simplest function of which an electronic counter is capable is that of totalizing, or accumulating, a count of input events. Because this mode does not require a time base, as indicated in **fig. 6,** it probably should have been covered previously as the most basic counter circuit. However, totalizing is not usually a function of low-priced counters, nor does it have major applications in amateur work; therefore I have delayed discussing it until the modes of greater interest were described.

The gate-control switch shown in **fig. 6** can be either a manual switch or an internal switching circuit actuated by the input signal. Switching the gate-control switch to *on* resets the decade counters to zero and allows the processed input signal to pass through the gate for the length of time that the switch is held on. When the switch is turned *off,* the count stops and the number of input events which has occurred is displayed on the counter.

An application which is of interest in these days of digital logic circuits is that of measuring contact bounce. **Fig. 7** shows a simple circuit which permits such a measurement for either a relay or a manually actuated switch. When voltage is applied to a relay coil (or a manual switch is operated), the contacts will usually open one or more times after the initial

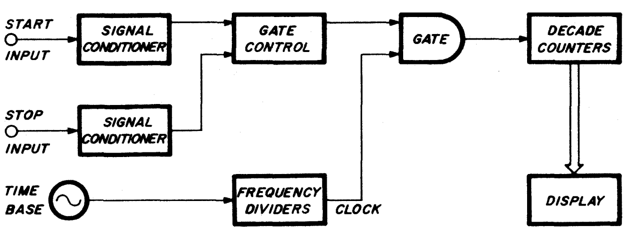


fig. 8. Functional block diagram of a time-interval counter, which counts the number of clock pulses between the time the gate is opened by the START **input and closed by the** STOP **input.**

closure because of the elasticity of the switch materials. This results in the waveform shown in the illustration, which is applied to the counter. The counter will then totalize and display the number of bounces.

time interval measurements

A counter may be used to measure the time interval between two input events, but this mode of operation requires two input-signal-conditioning circuits and a more complicated gate-control circuit; it is therefore found only in the more expensive professional instruments. As shown in **fig. 8,** the gate control has two inputs, one from each of the signal conditioners. The gate is opened by the processed *start* input, allowing the accurate clock pulses to pass through to the decade counters until the *stop* input closes the gate. Thus the counter will display the time interval between the two input signals.

The start and stop points are determined by the triggering levels and slopes selected by circuits in the signal conditioners. The time-interval resolution is limited by the clock frequency, and is subject to the same ±l-count ambiguity as all other measurements. As with period averaging, this ambiguity can be reduced for repetetive signals by averaging the time-interval measurements. When averaging, the

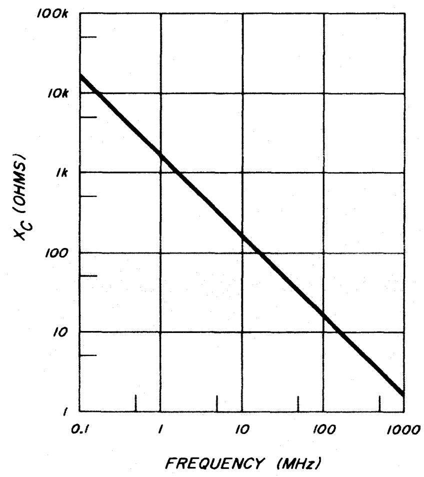


fig. 9. Reactance of 120 pF (typical for a high-impedance counter input with a 3-foot or 1-meter cable) plotted against frequency.

ambiguity becomes $\pm 1\ count \div \sqrt{N},$ where N is the number of time intervals averaged.

An important application of time-interval measurement is the accurate determination of pulse width. The signal under measurement is applied to both the *start* and *stop* inputs. If the *start* channel is set to trigger on the positive slope, and the *stop* channel on the negative slope (or vice versa), the counter will indicate the time interval between the leading and trailing edges of the input signal. Adjustment of the triggering levels will permit the measurement to be made between the desired points on the edges.

The upper frequency limit of the modern basic counter is dependent on the type of digital logic devices used in the signal conditioner and the first decade counter. This frequency limit may be as high as 50 MHz for conventional TTL, 120 MHz for Schottky TTL, and 250 MHz for ECL. Above those frequencies, prescaling is generally used to increase the frequency range, up to about 1300 MHz.

Prescaling simply means that the input frequency is scaled, or divided, down to one which is within the basic range of, and is measured by, the basic counter. The divisor may be any integral number. If the prescaler is external to the counter, it will usually divide by 10 or 100, so that the frequency can be read directly from the counter after you have mentally multiplied the counter reading by 10 or 100, as applicable. If the prescaler is built into the counter, it may scale by any integral factor.

The advantage of using an external prescaler is obvious — it permits extending the frequency range of an existing counter at relatively low cost. Its disadvantages become equally obvious after it has been used. First, there is the necessity of mentally moving the decimal point, since the counter is actually displaying the divided input frequency. Second, one digit of resolution is lost for every decade of scaling. For example, a 145,600.0-kHz signal measured with a scale-by-ten prescaler will read 14560.0 kHz on a counter having a 0.01-second gate time (0.1-kHz resolution). Multiplying by ten yields a frequency of 145,600 kHz; the 0.1-kHz resolution is lost by scaling. It can be re-established only by increasing the gate time by a factor of ten, provided the counter has that capability.

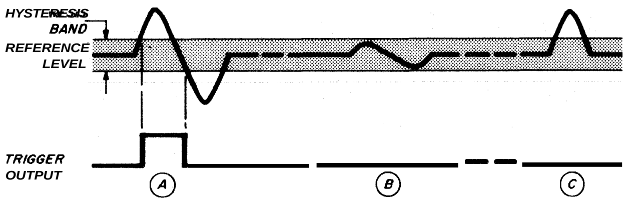


fig. 10. The effect of triggering hysteresis. The waveform at (A) will result in output from the trigger circuit, while those at (B) and (C) will not because neither crosses both limits of the hysteresis band.

If the prescaler is an integral part of the counter, mentally scaling and moving the decimal point is eliminated, since this will be accomplished in the counter when the mode is changed from direct count to prescaled count. Nevertheless, the loss of resolution remains. It can be minimized, however, by scaling by a factor less than ten, and simultaneously increasing the gate time by the same factor.

Suppose that the internal prescaler divides the input frequency by four. If the gate time is increased by the same factor, there will be no change in the number of signal pulses gated through to the decade counters, and the display will read out the correct frequency. Consequently, prescaling is accomplished with only a fourfold increase in gate time, which is generally acceptable.

Switching from direct to scaied operation may be carried out in one of three ways. If a single input connector is used, the counter mode is generally switched manually. If two separate input connectors are employed, one for low-frequency signals and the other for high-frequency inputs, the counter mode may be switched manually or automatically when the input signal is present at the high-frequency input.

input impedance

Counters which measure frequencies below 250 MHz or so usually present a high input impedance — typically 1 megohm shunted by 30 to 40 picofarads. Above that frequency, the input impedance is generally a nominal 50 ohms, although the vswr may be as high as 2.5:1. At audio and low radio frequencies, a high input impedance is normally desirable, since it minimizes the load on the circuit under test. But just how high in frequency is this true?

Consider a counter with an input impedance of 1 megohm shunted by 32 pF, which is used with a three-foot (91cm) cable made from RG-58C/U coax. The capacitance of RG-58C/U is 29.5 pF per foot (96.8 pF per meter), so that the total shunt capacitance presented to the circuit under test is approximately 120 pF. The reactance of this shunt capacitance, plotted against frequency, is shown in **fig. 9.** It can be seen that the reactance drops to approximately 1300 ohms at 1 MHz, and is only about 130 ohms at 10 MHz. So the input impedance can no longer be considered high. On the other hand, if the counter had a nominal 50-ohm input, you would know the loading effect, within the limits defined by the specified vswr.

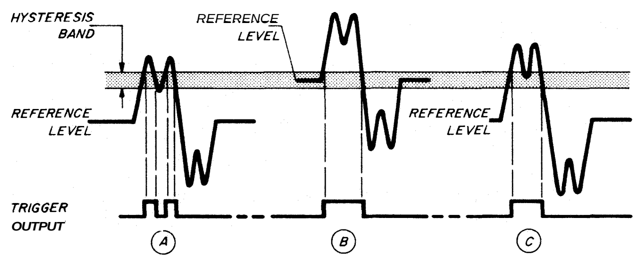


fig. 11. Erroneous counting caused by harmonic distortion is shown at (A). The false count can be eliminated by adjusting the level control, as indicated at (B), or by increasing the signal amplitude, as shown at (C).

Suppose that you had to check the frequency of a 70-MHz crystal oscillator which was designed to feed a 50-ohm load. If your counter has a 50-ohm input, all is well. However, if it has only a high impedance input, the shunt capacitance of the counter plus a cable will more than likely load down the oscillator and change the frequency, if it continues to oscillate at all. Fortunately, a relatively inexpensive accessory will solve the problem. By using a 50-ohm feedthrough termination* at the counter connector, a 50-ohm interconnecting cable will be reasonably well terminated, and will present a load close to 50 ohms at the oscillator.

The same thing may be accomplished by using a 20-dB loss pad at the input connector of the counter, provided that there is enough signal to trigger the counter after having been attenuated by the pad.

Even at low frequencies, the shunt capacitance may be too high for certain applications, such as checking filters. Capacitive loading can be reduced by using a 10X oscilloscope probe. Such probes typi-

*Such as the Heath SU 511-50, Hewlett-Packard 10100C, Tektronix 011-0049-01, Systron-Donner 454, and other similar types.

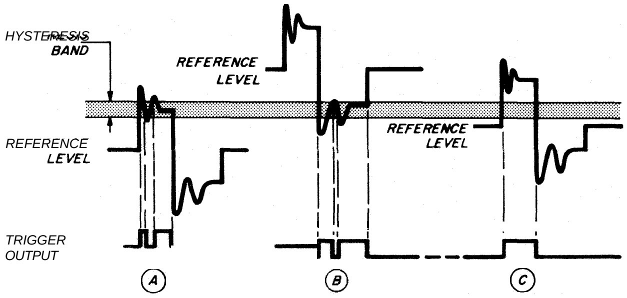


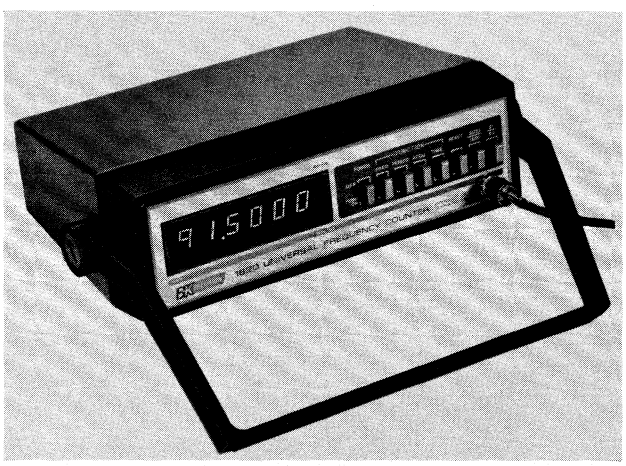
fig. 12. False counting caused by ringing is shown at (A) and (B). Proper adjustment of the reference level and/or amplitude, as shown at (C), corrects the fault.

cally present a 10-megohm resistive ioad shunted by 5 to 15 pF, but of course attenuate the signal by a factor of 10.

input signal levels

One of the parameters invariably specified for a counter is sensitivity, generally in millivolts, but often in dBm for 50-ohm inputs. This indicates the minimum signal needed at the *counter input* to ensure reliable triggering. Of equal, and possibly more importance, however, is the maximum signal which may be applied to the input without damaging the instrument.

For high-impedance inputs, the maximum signal voltage is usually specified as the sum of a dc value plus a peak ac value. The peak ac value may vary with frequency, going down as the frequency increases. The sum of ac plus dc is limited by the input blocking capacitor; the limiting ac value alone is a function of the input device in the signal conditioner. To be safe, when measuring at any point in a circuit where dc is present, always use an external blocking



B&K Precision 1820 Universal Frequency Counter will measure frequency from 5 to 520 MHz, and permits high-resolution period measurements from 5 Hz to 1 MHz. Decimal point position and unit-of-measure display is selected automatically for best resolution *(photo courtesy B&K Precision).*

capacitor of the smallest value which will permit reliable triggering. And if there is any possibility of the ac signal exceeding the specified maximum for the counter, use an external attenuator or dividing probe.

For low (50-ohm) impedance inputs, the maximum signal level is limited by the input circuit of the signal conditioner. This level is generally much lower than

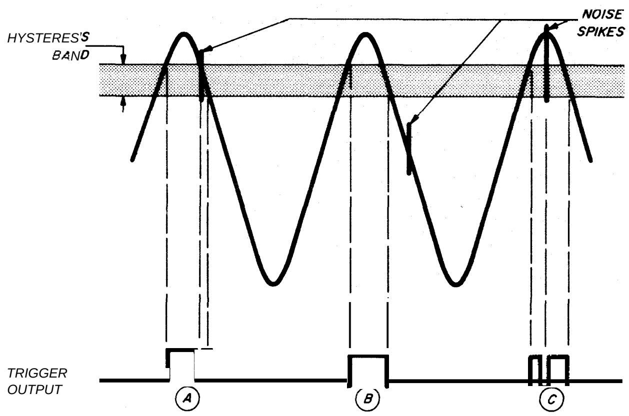


fig. 13. Spurious counts can result from noise on a signal when the noise is of sufficient amplitude to cross the hysteresis band.

for high-impedance inputs, and is typically between +19 and +27 dBm (2 to 5 volts rms across 50 ohms). Because of the relatively high cost of the high-frequency input device and the possibility of applying excessive power from a transmitter, the 50-ohm inputs are fuse-protected in many counters which can function at 500 MHz and higher.

Although it should be obvious, the following warning must be included: *Never connect a counter directly to a transmitter or any other high-power signal source!* Use a short length of unshielded wire as an antenna at the counter input connector, an inductive coupling loop at the end of a shielded cable, and/or an attenuator of sufficient power rating. The counter you save may be your own!

If the counter is battery-powered, and there is no direct connection between it and the circuit or generator under test, the counter should be grounded. This will reduce noise pick-up, especially when using a counter with a high input impedance.

triggering

The signal-conditioning circuits in all counters include a trigger circuit which, as previously stated, provides output pulses whose amplitude and waveshape are compatible with the counter circuitry which follows. The sensitivity of the counter depends on the threshold level of the trigger input and the amplification between it and the input of the counter. If the amplified input signal has insufficient amplitude to reach the threshold level, the instrument will not count or will perform erratically.

All trigger circuits have a hysteresis band, through the limits of which the input signal must pass in order

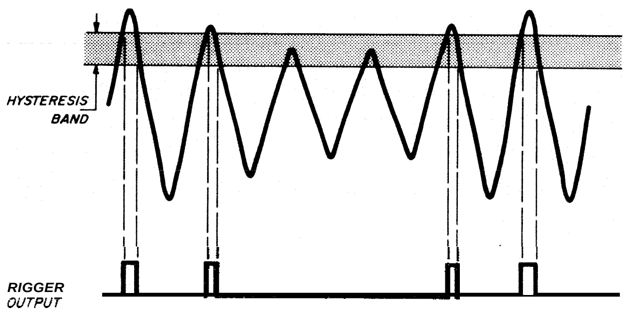


fig. 14. Amplitude modulation of the input signal can cause missing counts when the signal amplitude is too low.

to result in an output pulse. **Fig. 10** shows three input signals in relationship to the hysteresis band. Sine wave **A** crosses both the upper and lower limits of the band, and will actuate the trigger circuit; the amplitude of sine wave **B** is too small, so triggering will not occur; and waveform **C** crosses the upper threshold, but not the lower, so again no output will be produced by the trigger circuit.

It is the action of this hysteresis effect which can result in erroneous counting which is so confusing to a relatively inexperienced operator. Suppose that the input to the counter were a sine wave with considerable second-harmonic distortion, a not uncommon situation. In **fig. 11A**, the amplitude of the signal is such that the positive half-cycle crosses the hysteresis-band limits twice, instead of once. The trigger circuit will generate two output pulses for each input cycle, and the counter will display twice the fundamental frequency of the signal. If the counter has a level control, which adjusts the reference level at the input of the trigger circuit, it can be adjusted to eliminate the false count, as shown in **fig. 11B**. If there is no level control, as is the case with most low-priced counters, the problem can be eliminated by increasing the amplitude of the input signal, as depicted in **fig. 11C**.

A similar problem may arise when measuring the frequency or period of a signal comprised of fast pulses. If the interconnecting cable is not terminated in its characteristic impedance, or if other impedance discontinuities exist, ringing will occur on the pulses. If the ringing traverses the hysteresis band, as shown in **fig. 12A** and **12B** for two different reference levels, a false count will result. Proper adjustment of the signal amplitude and reference level, indicated in **fig. 12C**, will provide the correct count.

Another way of solving the ringing problem, which is useful when the reference level andior arnpiitude cannot be changed, is to use a low-value resistor (100 to 1000 ohms) between the circuit point under test and the counter cable. This resistor, in conjunction with the cable and counter input capacitance, integrates the pulse and minimizes the pulse aberrations which reach the counter.

Figs. 13 and **14** illustrate two other conditions which can result in false counts. The noise transients on the signal shown in **fig. 13** will cause additional counts, while amplitude modulation may result in missing counts, as shown in **fig. 14,** if the amplitude of the input signal is too small. In either case, the solution is the same as previously prescribed – change the reference level and/or the signal amplitude.

In our earlier discussion of period and period-averaging measurements, it was stated that the trigger error resulting from noise on the input signal contributed to the measurement error. This is shown in **fig. 15,** in which a sawtooth wave is used to demonstrate the effect of slope, or slew rate, on the trigger error. It can be seen that the noise voltage on the relatively slow rise-time can cause a much greater trigger error than that which occurs on the fast fall-time. Thus we can see that the trigger error can be minimized by triggering on the steepest portion of the input signal to the counter. For a sine wave, this will be that part of the waveform at the zero axis, leading to the conclusion that a signal of the maximum possible amplitude should be used.

It should be apparent from the preceding discussion that an input attenuator on the counter can be of considerable help in establishing the correct input level to the trigger circuit. In many of the lower-priced counters an attenuator has been omitted because of cost and because it was felt that limiting diodes at the input of the signal conditioner would protect the input device. The latter reason is valid

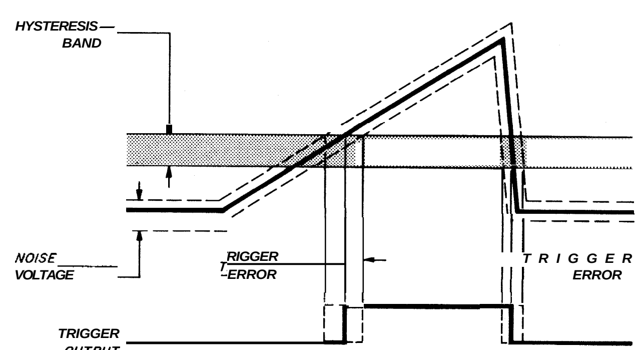


fig. 15. Trigger error in period and period-averaging measurements, caused by noise in the input signal. The error is minimized by a fast slew rate through the hysteresis band.

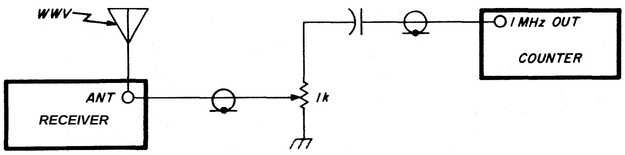


fig. 16. Test setup for calibrating the time-base frequency against WWV.

only where overload is considered, for even a two-position attenuator can be extremely valuable in eliminating false counting.

time-base calibration

Unless an oven oscillator or TCXO is used as the time base in a counter whose oscillator circuit is energized continuously, the oscillator frequency should be checked, and adjusted if necessary, whenever accurate measurements are to be made. Be sure, however, that the counter is fully warmed up before checking or recalibrating the oscillator.

In order to calibrate the time-base frequency, either a standard of known accuracy or a communications receiver capable of receiving WWV is required. If the standard is at least five times more accurate than the best resolution of the counter at the time-base frequency, it can be applied directly to the input of the counter. Then adjust the time-base oscillator frequency control for the correct frequency read-out on the counter.

A more accurate adjustment may be made if the counter has an output connector from which a I-MHz time-base signal can be obtained. Connect this output to the vertical input of an oscilloscope, and connect the output of the frequency standard to the horizontal input." The scope will display a Lissajou pattern, which will probably be moving. Adjust the time-base frequency control until a stationary pattern is obtained.

If the counter is to be calibrated against WWV, a time must be chosen during which transmissions are received with an absolute minimum of fading. Select the highest receiving frequency possible (*e.g.* 15 MHz) to achieve the greatest calibration accuracy. The calibration technique involves obtaining a visual beat indication on the receiver S-meter, and adjusting the time-base oscillator frequency for as close to a zero-beat as possible.

In order to obtain a good beat null, the time-base signal which is applied to the receiver must be of the correct amplitude relative to the signal level from WWV. Since we cannot control the latter, we have to be able to vary the signal level from the counter. If the counter has a 1-MHz output from the time-base circuit, make the connections shown in **fig. 16.** If a time-base signal is not brought out to a connector on the counter, substitute an insulated wire for the coax shown connected to the receiver antenna terminal and place it near the counter time-base oscillator or frequency-divider chain. In either case, the harmonic of the 1-MHz signal should result in a low-frequency beat with WWV. (This will not be an audible beat unless the time-base oscillator is very far off frequency; more likely it will be observed as a rythmic variation in the S-meter reading.) Adjust the potentiometer shown in **fig. 16,** or change the position of the insulated wire, to obtain the deepest beat null on the receiver S-meter.

Leader LDC-822 Frequency Counter measures frequency to 80 MHz and features selectable gate time and input attenuation *(photocourtesy Leader Instruments).*

It should be possible to adjust the time-base frequency so that the beat-frequency period is several seconds, which corresponds to a remarkably accurate short-term frequency setting. To demonstrate this, assume that eight beats are observed on the S-meter in a 60-second period. The beat frequency is therefore equal to 8160, or 0.133 Hz. If the beat is measured at 15 MHz, the error is 0.133115 x 106, or 8.9 x 10^{-9}. Of course, this degree of accuracy may hold only for a short period of time, because the stability of the counter time-base oscillator, unless it is an oven type, is nowhere near that good. Nevertheless, highly accurate measurements may be made until the counter is turned off or a temperature change affects the time-base oscillator.

ham radio

"These connections are based on the assumption that both the horizontal and vertical amplifiers in the oscilloscope will pass a 1-MHz signal. If the counter has a time-base output of higher frequency, it can also be used, provided that it is within the frequency range of the scope. The lower of the two frequencies (the standard and the counter time base) should be connected to the horizontal input, since the horizontal frequency limit is usually lower than the vertical.

simplifying the digital frequency counter

Some innovative ideas for high-resolution counters using CMOS — TTL devices

Radio amateurs have had a long history of pioneering in electronics. My experiments with the new IC technology resulted in the following article. I decided to find out just how much the digital frequency counter could be simplified by using some of the newer CMOS combination chips.

Today's literature shows that much can be done. Almost everyone seems to be using TTL technology; for example, the 749017447 combination dating from the 1960s: an 8-digit counter using 7490 devices requires 24 chips for the main counter alone, plus 12 or more (typically) for a crystal-controlled time base and pulse conditioning. Power required is 5 volts at 1.5-2 amps or more! Only one example was found using CMOS. It still used 21 chips and was limited to 4 MHz.

A number of the newer CMOS combinations are available from which to choose for simplifying the counter and decreasing power requirements; the Intersil 7208/7207A combination seemed to be the most promising, so it was chosen for this project.

The counter is shown in **figs. 1** and **2.** Both circuits comprise a complete frequency counter with 1-Hz resolution from below 20 Hz to above 50 MHz, and with 10-Hz resolution to above 300 MHz. Nine ICs and four transistors are used including the power supply and prescaler. Current drain is 200 mA for frequencies below 50 MHz; an additional 130 mA is required for higher frequencies.

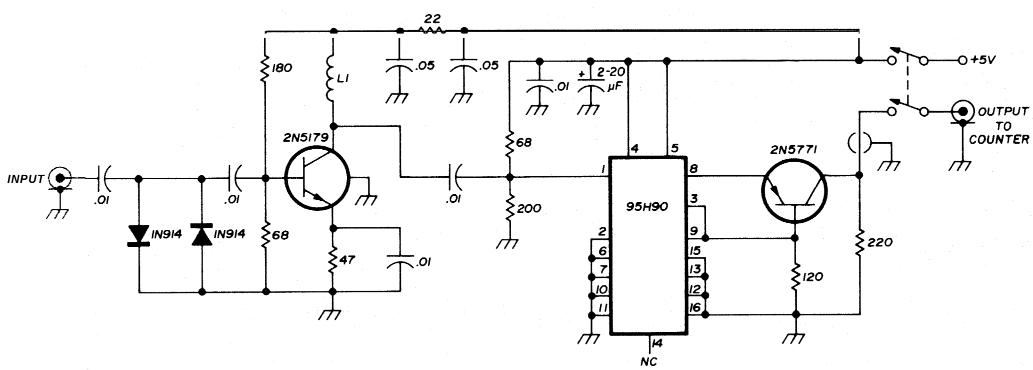


fig. 1. 300-MHz prescaler for the high-resolution counter.

Device description. The heart of the counter is the 7208 CMOS chip. This device contains a 7-decade counter, multiplexer for the display, 7-segment decoder, digit and segment drivers, and the logic required for blanking, reset, input inhibit, and display on-off — a regular one-man band!

The 7207A is another CMOS chip that teams up

By Holton E. Harris, W1WP, 5 Newtown Turnpike, Westport, Connecticut 06880

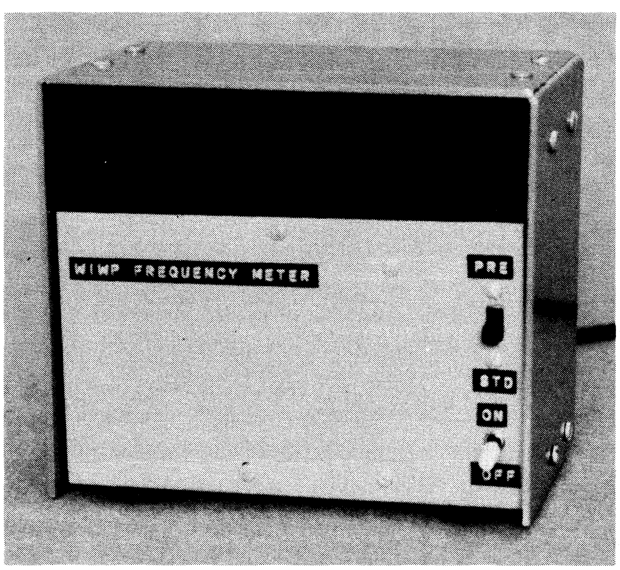
with the 7208. It contains a high-stability oscillator and a frequency divider for dividing the 5.24288-MHz crystal frequency to obtain the l-second gate required for counting. It also provides outputs to synchronize the mutliplexer for the displays as well as short pulses for latching and resetting the counters.

CMOS and TTL combination. The one real deficiency of CMOS is that it is slow – i.e., low-frequency response. The 7208/7207A combination alone, with power supply, LED displays, and crystal can be used to make a complete counter (as shown later) but it won't count above about 6-7 MHz. Almost all of the remaining circuitry in this counter is the old workhorse TTL, purely and simply to extend the frequency range.

The 74196 counter is used to get to 50 MHz. This device is similar to the 7490 but has a higher frequency range. A prescaler could have been used instead, of course, but a synchronized counter was preferred because it retains accuracy in the least-significant digit. The 9368 performs the decoder/driver functions for the least-significant digit of the LED display.

When Intersil first announced the 7208, the companion driver was the 7207. This combination gave a gate signal of 0.1 second, so the resolution was 10 Hz instead of 1 Hz. But the worst of it was that, if all seven decades of the 7208 were used, the most-significant digit was 10 MHz, which was above the frequency range of the counters in the 7208. In effect, therefore, all seven decades could never be used.

The 7207A, announced in late 1976, corrects the problem. It has a gating pulse of 1 second duration, permitting l-Hz resolution. Unfortunately, to make room for the added counter stage, the output buffers on the reset and enable lines had to be eliminated. These signals therefore can't be used to drive the



7432 TTL gate directly, and an extra buffer was required. The 4049 IC is a hex buffer/converter designed for this CMOS to TTL interface. The signal is passed through the buffer twice with a 4.7k pullup

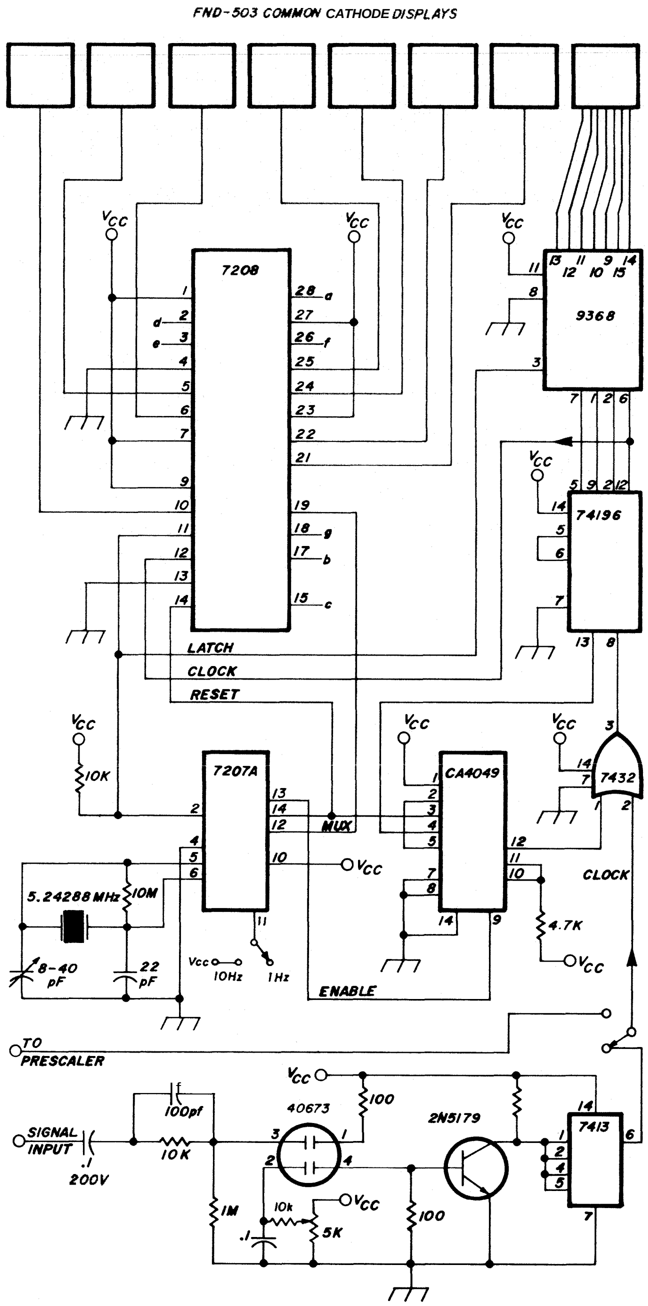
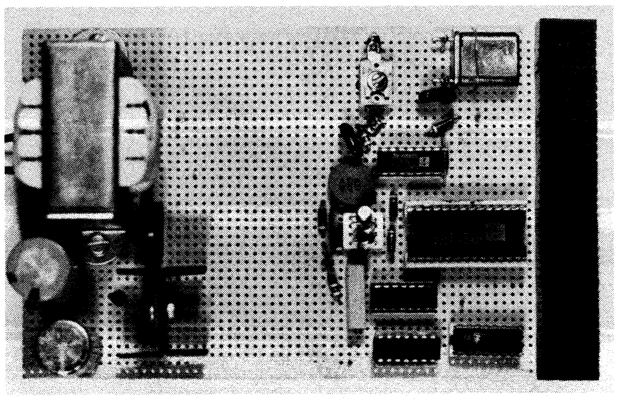


fig. 2. Schematic of a 50-MHz counter with 1-Hz resolution. In each counter example, the resolution is decreased by a factor of ten by taking pin 11 of the 7207A to V_{CC}.

resistor at the interstage in an effort to make the gating pulse rise and fall times equal. Equal rise and fall times are important, since the accuracy of the entire counter depends on the time the counting gate is open. The 7432 is an OR gate, which removes the



Pegboard construction of the counter shown in fig. 3. The power supply is shown at left. The two sections were cut apart for mounting into a cabinet.

input signal from the counter input except during the counting interval.

Circuit description. The 5.24288-MHz crystal and the oscillator portion of the 7207A form a high-precision time reference that determines counter accuracy. The crystal is a low drift, 5 ppm type. If higher precision is desired, an oven should be used.

The 7207A has a binary divider chain that divides the oscillator frequency by 2^{20} to yield a 0.5-Hz square wave that sets the counting interval to exactly 1 second. A higher frequency is picked off to synchronize the 7208 for multiplexing the output displays. Pulses are also generated to set the display latches and reset the counters at the appropriate times.

After passing through the 4049 buffer, which increases the power level to drive the TTL gates, the timing signal is applied to the 7432. For the 1 second during which the signal is low, the 7432 allows the input signal to pass through to the 74196 counter. The 74196 counts to 10, puts out a pulse to the 7208, then repeats. The 7208 has seven decade stages that similarly count successive decades.

After the l-second counting period, the gate to the 7432 goes high and the counters stop. The latch pulse from the 7207A transfers this count into the latches, and the decoder/drivers in the 7208 convert the count to 7-segment form and pass it to the display lines. Another pulse from the 7207A through the 4049 then resets the 74196 and the seven decades in the 7208 to zero. Meanwhile the multiplexer in the 7208 energizes each LED display in sequence.

The low-frequency preamplifier was cribbed from Stark[1]. An additional stage of amplification was added after the input fet to increase sensitivity. The circuit thus consists of a two-stage amplifier driving a Schmitt trigger. The Schmitt trigger turns on at one level and turns off at another, much lower, level so that slowly rising signals can't cause jitter, and false triggering is avoided.

The prescaler to extend the range is likewise conventional. It was lifted from the excellent article by Kitchens[2]. Why argue with success?

One point may appear puzzling. If you check the Intersil 7208 data, you'll find the multiplexer input brought into pin 16. This leaves two prior CMOS gates open, which is bad practice. I couldn't get the device to work at all with that connection. Bringing

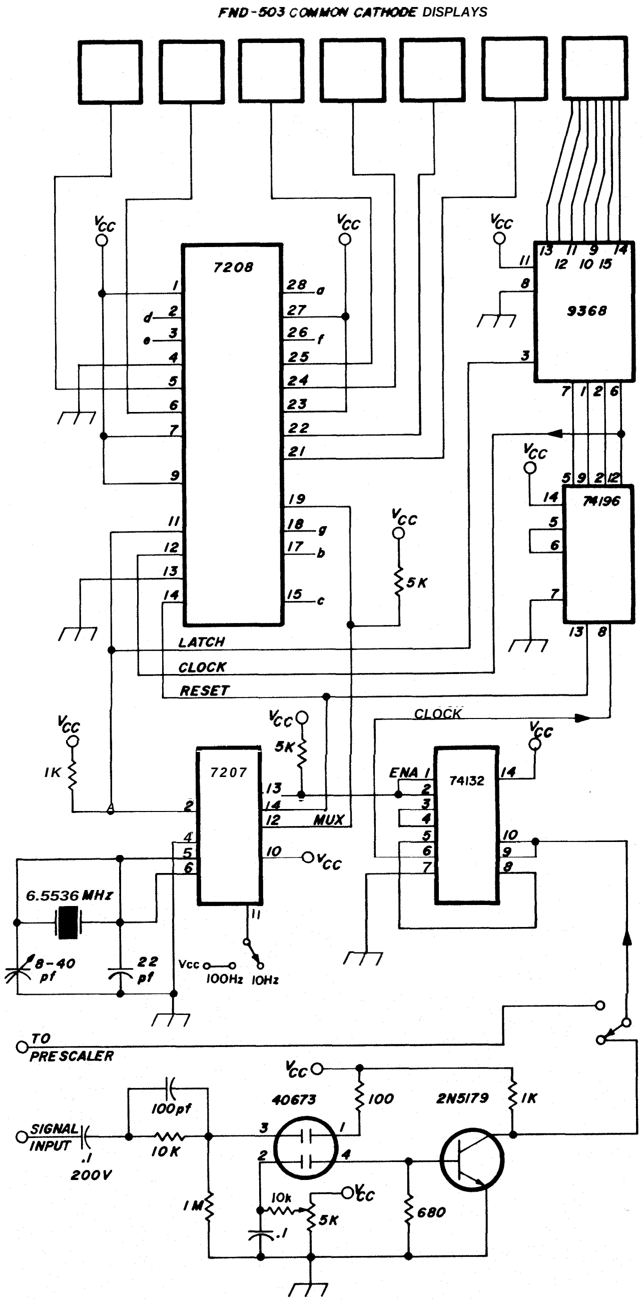


fig. 3. Schematic of a 30-MHz counter with 10-Hz resolution.

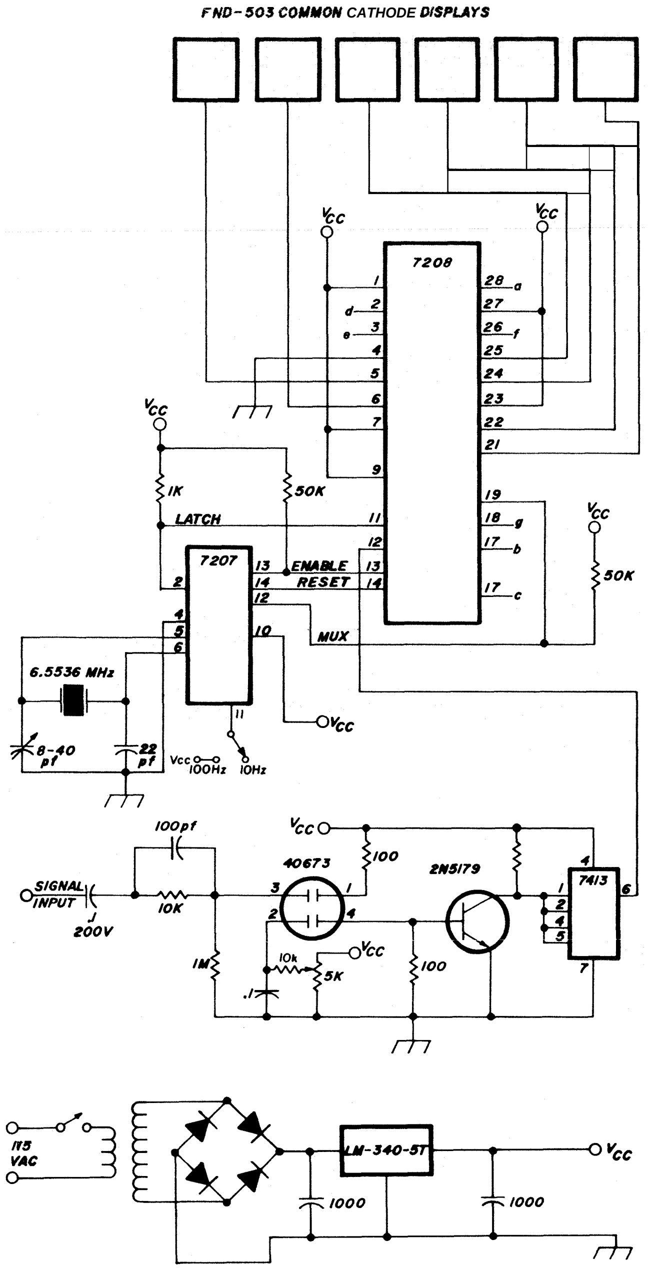


fig. 4. Simplified counter that operates up to 6 MHz with 10-Hz resolution.

the multiplexer input to pin 19, as shown in **fig 2,** ties everything down and cures the problem.

The new National 5881, 581, and 583 multiple display LEDs were used because when they first came out they were even cheaper than surplus standard units. If I had it to do over again, I'd use the conventional FND-507 even if more wiring is required.

construction

I built the circuit on pegboard using point-to-point wiring with a Vector wiring pencil. A more enterprising builder might wish to make a PC board. The only adjustments required are to tune the crystal to exact frequency using the 8-40 pF trimmer and to adjust the 5k pot in gate 2 of the 40673 prescaled transistor **(fig. 1)**. This latter adjustment is easily made by connecting a voitmeter between the 7413 pin 1 and ground and adjusting the trimmer pot until the dc level is about 1.3 volt.

other forms of the counter

The one big disadvantage of the counter described above is that its sampling time is quite slow. Obviously, if a resolution of 1 Hz is required, the counting gate must be open for one full second. Another second is allowed for reset and latch, so that the total time to update the display is 2 seconds. Under some circumstances, this can seem like forever!

For most applications, a 1-Hz resolution isn't really necessary, and the circuit of **fig.** 3 can be used. Here the 7207 is used as the oscillator/timer. This device gives a counting interval of 0.1 second and a total period of 0.2 second. The counter thus updates 5 times per second instead of once every 2 seconds, and action seems much more normal. The resolution is, of course, only 10 Hz instead of 1 Hz, which is sufficient for most purposes.

The counter shown in **fig.** 3 includes further simplifications. The functions of the 7413 Schmitt trigger and the 7432 OR gate are combined in a single IC, the 74132. This device is a quad 2-input NAND Schmitt trigger, which does both jobs. This change could, of course, be made in the circuit of **fig. 2.**

Fairchild FND-503 displays are used here. The 100-ohm limiting resistors were eliminated for a brighter display, which increased total counter current from 200 to 300 mA.

A further simplification is possible if the highest frequency to be counted can be limited to 6-7 MHz. In this case the 74196, together with the 9368 and 7432, can be eliminated. The circuit is shown in **fig. 4.** The 7208 IC provides the counting function. Resolution is again 10 Hz, and the seventh digit in the readout is omitted, since it can never be used in this instance. However, the circuit makes a mighty simple counter.

I'd like to express my appreciation to my co-worker, Josh Schwartz, without whose excellent and timely suggestions this project couldn't have been completed.

references

1. Peter A. Stark, K2OAW, "A Modern VHF Frequency Counter," 73, July, 1972, page 5.
2. Marion D. Kitchens, Jr., K4GOK, "Vhf prescaler for Digital Frequency Counters," *ham radio*, February, 1976.

ham radio

how to modify your frequency counter for direct counting to 100 MHz

Simple IC circuit can be used in an existing frequency counter to extend its direct counting range to 100 MHz

When the prices of TTL integrated circuits first dropped to the point where the average amateur could use them to build a frequency counter, the maximum operating frequency was about 25 MHz. Then came the Fairchild Semiconductor 95H90 prescaler with its 350 MHz capability. This promised reliable measurements at 220 MHz, but some of the earlier homebrew counters were frequency limited and could not use the prescaler to its full advantage. Next came the Fairchild 11C90, which was rated at 650 MHz and the 1300 MHz counter from Hewlett-Packard. It will only be a matter of time before an inexpensive 1 GHz prescaler reaches the market. It hasn't arrived yet, but I decided to redesign my counter to adapt to such a prescaler, when it comes.

There has been a flood of second-generation TTL ICs appearing in the past few years that would both increase the speed and diminish the size of a modern frequency counter over one using the standard 7490 decade counter, 7475 latch, and 7447 LED decoderldriver. I decided that a complete rebuilding of my counter, while interesting, could not be justified. In this project described in this article a single board, containing a gate circuit and the first decade counter, is substituted for the original. This board can be used in any counter which has a positive gate-enable pulse, a positive or negative reset pulse, and enough room to sandwich in the modification.

circuit operation

The gate function is performed by one gate of a 74S00. When a positive enable pulse is applied to one NAND gate input (during the counting period), the gate will act like an inverter to a square wave ap-

By **Ian MacFarlane, WA1SNG,** 102 Columbus Avenue, Greenfield, Massachusetts 01301

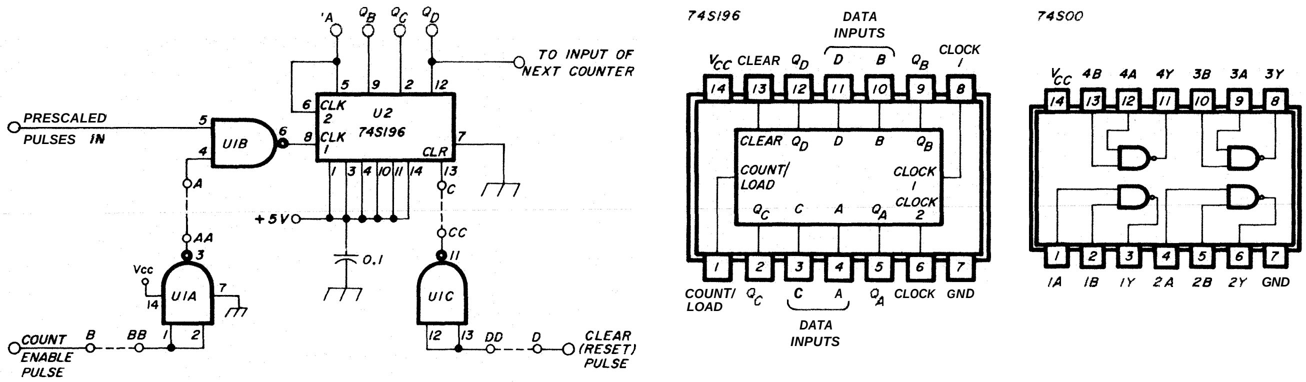


fig. 1. Schematic diagram of the 100 MHz counter. If your existing counter has a positive reset pulse, jumper C to CC and D to DD as shown here; if the reset pulse is negative, jumper C to D. If the count enable pulse is negative, jumper A to AA and B to BB as shown; if positive, jumper A to B. Not shown here but included on the circuit board are two bypass capacitors from the power supply to ground, C2 and C3.

plied at the second input. As the gate enable pulse falls to a logic zero, the NAND output rises to one, and is not affected by state changes at the second input. Consequently, the gate operates as an on/off switch to control the flow of the signal to be counted into the first decade counter. If your counter uses a negative enable signal, provisions have been made on the circuit board to use an extra NAND gate to invert the count-enable [illegible]lse

[illegible] 74S196 t[illegible] decade counter is used to extend the guaranteed count frequency to 100 MHz. The typical frequency limit is 140 MHz. To disable the preset feature, all presettable data inputs must be held at a logic one, along with the *Count/Load* input. Binary-coded-decimal counting is provided by connecting out Q_A to *Clock Input* 2, and injecting the signal to be counted into *Clock Input 1*. Counting occurs on the negative transitions of the count pulse. The reset *(clear)* pulse must also be negative, unlike that used for 7490 decade counters, so one NAND gate is used as an inverter. If your counter also uses a negative reset pulse, it is possible to directly drive the *clear* input, so long as the fanout of your clear line is not exceeded by the 74S196 requirements. Unlike a 7490, the clear input of the Schottky chip is two standard Schottky loads (or 2.5 standard 7400 series loads). If your counter has a maximum of eight digits, you should not encounter any difficulty. However, don't try to drive the reset line with an L or LS series device because it will not be able to supply sufficient current. The 74S196's four binary-coded-decimal outputs (Q_A through Q_D) are connected to the respective inputs of the latch.

Signetics Corporation manufactures a plug-in replacement for the 74S196 called the 82S90. This might be the easiest device to find for some people, but for me the difficulty of trying to locate a 74S196 was second only to finding an 82S90. Neither device is stocked by most suppliers, but I have been advised by Active Electronic Sales that they can supply SN74S196N ICs at $3.45 each." These are Texas

"Active Electronic Sales Corporation, Box 1035, Framingham, Massachusetts 01701. They have a minimum order requirement of $10.00, plus a $1.00 postage and handling charge.

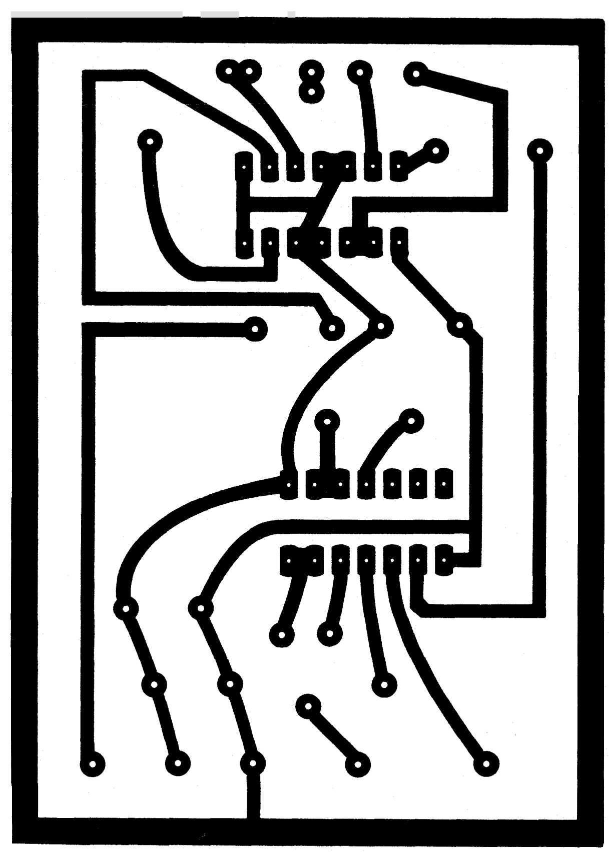


fig. 2. Full-size printed circuit board for the 100 MHz counter stage. Component layout is shown in fig. 3.

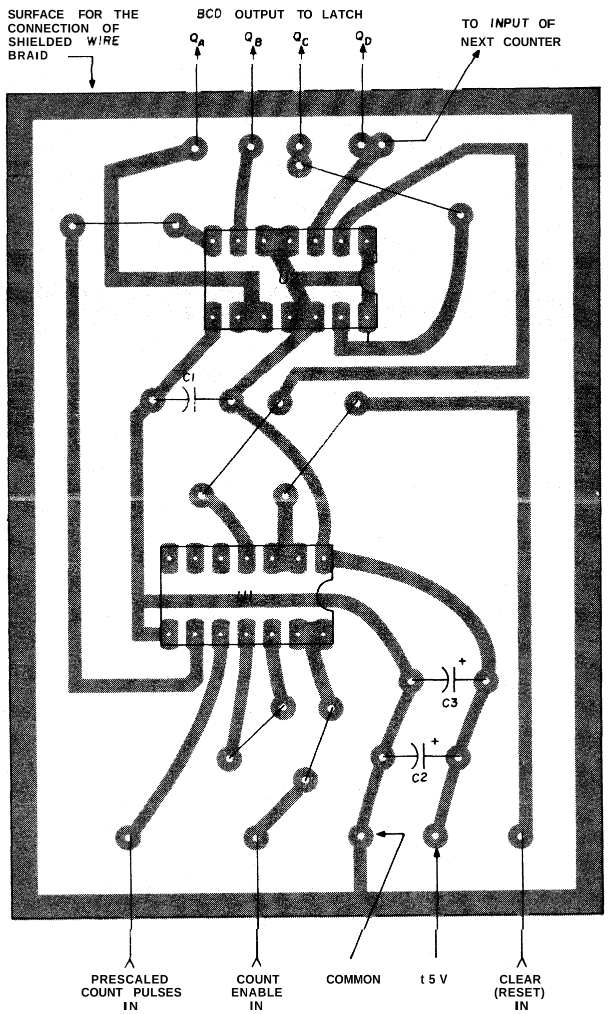


fig. 3. Component layout for the 100-MHz counter. Bypass capacitor C2 is a 0.1 μF or larger disc ceramic; C3 is a 47 μF, 10 volt, tantalum or electrolytic.

Instruments devices with current date code devices, but since they are not normally stocked, delivery time is three weeks.

construction and installation

Circuit details and IC pin-out diagrams are shown in **fig. 1.** The printed-circuit pattern and component placement information is given in **fig.** 2 and **3.** The power supply lead should be a separate shielded cable (audio type is fine) connected directly at the power supply. Connect the shield to the common power supply ground point and attach the other end to *common* on the PC board. To prevent ground ioops, do not ground the board where it is mounted.

Disc or rectangular ceramic bypass capacitors are used liberally to discourage transients but their values are not critical – the larger the better. One large tantalum or electrolytic capacitor is used to eliminate low-frequency transients. Shielded cable may also be used to bring the *Gate-Enable, Reset,* Q_A through Q_D, *Count Pulses In,* and *Count Pulses Out* signals into and out of the board. Shielded cable is not really needed in practice, but it will reduce the amount of rf floating around the counter, and may prevent jamming of the input circuit. If shielding is used, ground the shield at the board and trim back the braid at the other end.

modifying the input circuit

Extending the range of the input circuit to 100 MHz is not absolutely necessary because a prescaler will extend the counting range anyway. However, direct counting to 100 MHz is easy to accomplish in counters which use the two most popular input circuits.

One popular input circuit consists of a fet amplifier, often an RCA 40673, driving a Schmitt trigger; a typical circuit is shown in **fig. 4.**[1] Almost any dual-gate mosfet will operate well above 100 MHz, so no change is required there. If the Schmitt trigger is a 7413, it can be replaced directly with a 74S13. Discrete Schmitt triggers built up from 7400 or 7404 gates can usually be replaced with Schottky devices without changing any external resistors.

Another popular counter input circuit first appeared in *QST*[2] during a review of the HUA Electronics 1BC-1a frequencv counter. This circuit **(fig. 5)** has been widely duplicated over the years with varying success. Input sensitivity estimates have been reported from 10 to 300 mV, so a few suggestions are in order for anyone having difficulty with the HUA circuit. Problems stem from two sources: the extreme sensitivity of the unit, and the substitu-

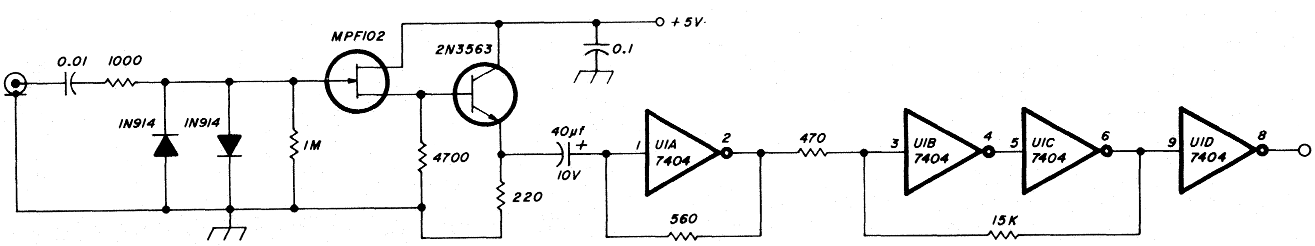


fig. 4. Counter input and shaper stage used in many homebrew frequency counters was originally used in a commercial instrument; it can be modified for use at 100 MHz as discussed in the text.

tion of one TTL sub-series for another (such as a 74LS04 for the 7404). Sensitivity can be so great that leakage from other counter circuitry jams the input, establishing a threshold that must be exceeded to trigger the counter. The solution is to shield the entire input circuit in a minibox, using coax to bring the signal in, and shielded wire, bypassed on both ends with 0.1 μF and 47 μF capacitors, for the +5 volt supply. The shield should be grounded at the power supply and connected as the common return on the circuit board.

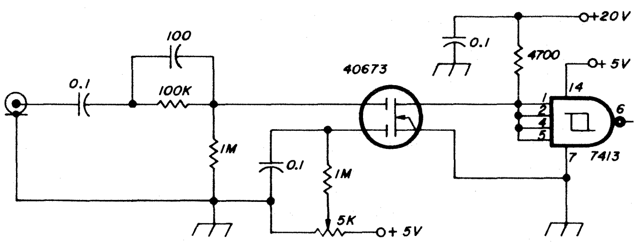


fig. 5. Popular counter input stage based on the 40673 mosfet which can be modified for operation to 100 MHz as described in the text.

The second problem concerns differing bias requirements between the standard and LS series. When substituting a 74LS04 for a 7404, it is impossible to simply interchange one for the other without changing the external resistors. This problem is not as acute when replacing the 7404 with a 74S04, because the difference in input current between the Schottky and standard series is slight. Only one resistor seems to be at all critical, and this is the 560-ohm feedback resistor across the first hex inverter. A sure way of obtaining the most performance from your particular combination of devices is to optimize this resistor. Connect a 1000-ohm potentiometer in place of the 560-ohm resistor and adjust it for maximum sensitivity. Using an ohmmeter, measure the resistance (after carefully removing the pot from the circuit) and replace it with a fixed value.

conclusion

For a total cost of less than $20 for all materials, including printed-circuit drafting and etching supplies, this circuit can put new life into an otherwise outmoded counter. The 74S196 will typically extend counting to 125 or even 140 MHz, which should accommodate the output of prescalers for many years to come.

references

1. Peter Stark, K2OAW, "A Modern VHF Frequency Counter," 73. July. 1972. page 5.
2. Gerald Hall, K1PLP, "HUA Electronics Frequency Counter Model 1BC-Ia," QST, April, 1972, page 60.

ham radio

simple front-ends for a 500-MHz frequency counter

Basic front-end design is stressed in this adaptation of the Intersil seven digit CMOS frequency counter

The advent of the Intersil CMOS counter chip pair, the ICM 7208 and 7207A, has resulted in many designs which take advantage of the low cost and simplicity these chips make possible. I wanted a complete, 500-MHz counter, but with no bells and whistles. Since most hams do not need more than 100 Hz resolution at 500 MHz, the counter was restricted to seven digits. With most of the basic work already documented in Intersil application notes, the only real design problem was to fabricate a suitable front-end that interfaced between the prescaler and the CMOS integrated circuit.

As an engineer I have fought many debugging wars, enough to know that sure things don't always work the way they are designed. Accordingly, I researched the literature for the tried and true. As a result, what follows is not entirely original, but it has the redeeming virtue that it works.

50-MHz front end

The front end is composed of Q1 through Q7, with a sensitivity of 300 mV at 30 MHz, falling to 1 volt at 50 MHz (see **fig. 1**). Most of the credit for the design goes to Marvin Moss, W4UXJ, who adapted it from several other similar designs.

As a single device, the fet used for Q1 does not give satisfactory performance. Therefore, Q2 is used to bootstrap the voltage at the source of Q1 to more closely equal the voltage at the gate of Q1. This also greatly reduces the effect of Q1's input capacitance at high frequencies, thus maintaining the input impedance with increasing frequency. Capacitor C2 compensates for the small amount of rolloff that will never-the-less occur. R5 allows quiescent point adjustment for maximum sensitivity. Since the remainder of the front end is dc coupled, R5 also sets the operating point for the rest of the amplifier.

Q3 has a fixed gain of approximately 6.8, the ratio of R7 to R8. R8 also raises the input impedance of this stage to minimize loading on Q1-Q2. C3 tends to raise the gain of Q3 to compensate for the rolloff in gain brought about by the output capacitance of Q3 being in parallel with R7.

Q5 and Q6 form a high-gain amplifier with hysteresis. Basically, Q5 is an emitter follower driving the common base amplifier, 06. The base of Q6 is held at approximately 6 volts by R12 and CR4. Because of the high voltage gain of the common base configuration, the signal alternately drives Q6 close to saturation and cutoff. However, R11 creates a small amount of positive feedback, or about a 0.6 volt hysteresis. This is necessary to avoid extraneous counts on low-frequency signals.

Transistor Q7 is used as an emitter follower to drive the TTL counter circuitry that follows. The low value of R14 is necessary, since TTL likes to see a low source impedance in the low state. A common-emitter stage might work here, but would require more components and would draw just as much current.

By John H. Bordelon, K4JIU, 1132 Research Drive, Marietta, Georgia 30066

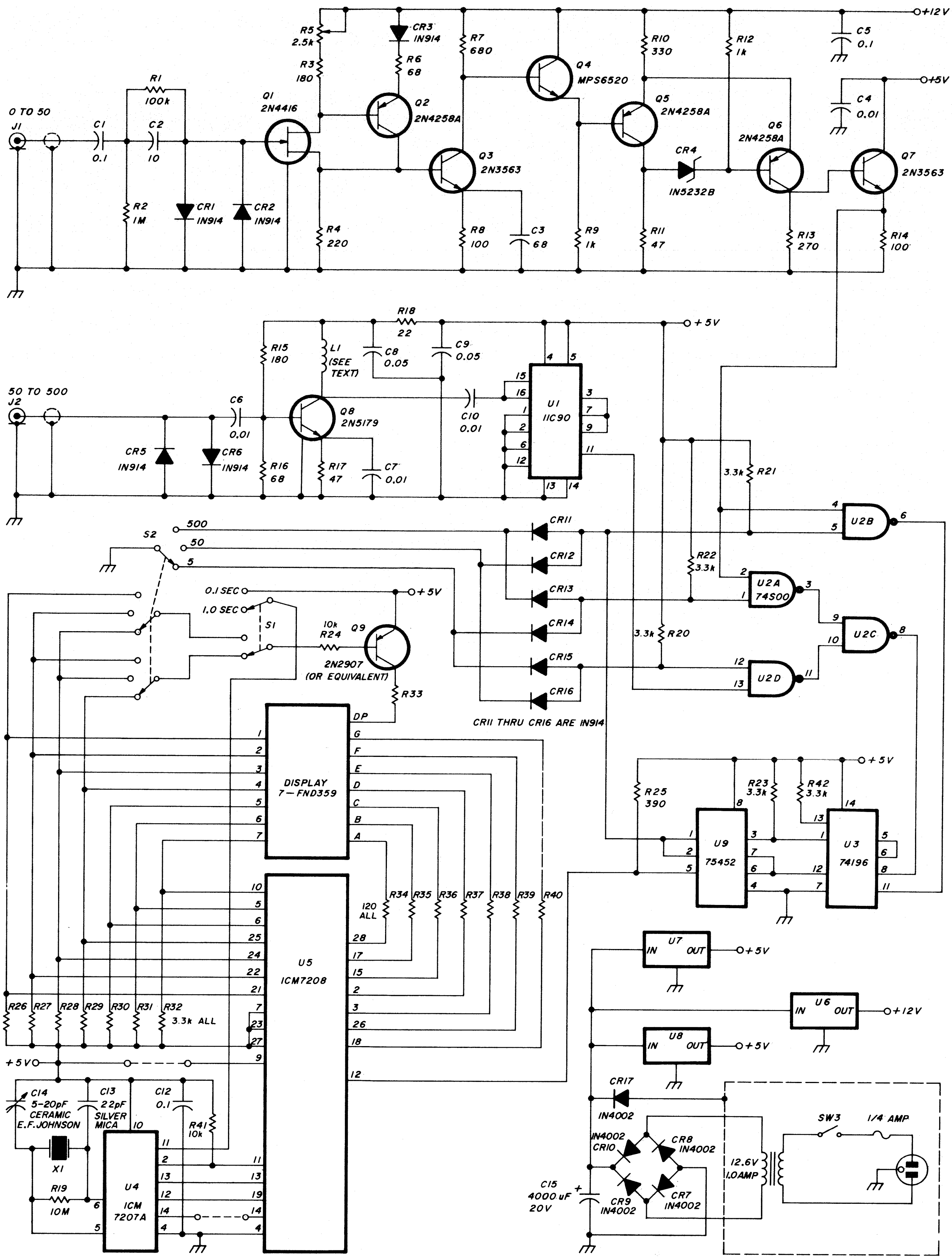


fig. 1. Schematic diagram of the complete frequency counter including the low-frequency (0-50 MHz) and high-frequency (50-500 MHz) preamplifiers. The displays have been wirewrapped into a set of sockets and are shown here as a multiplexed assembly. Other displays can be used but they must be common-cathode types. The crystal frequency is 5.242880 MHz (available from International Crystal). All resistors are ¼ watt, 10 per cent tolerance. The 5 and 12 volt regulators are MC7805 and MC7812 ICs.

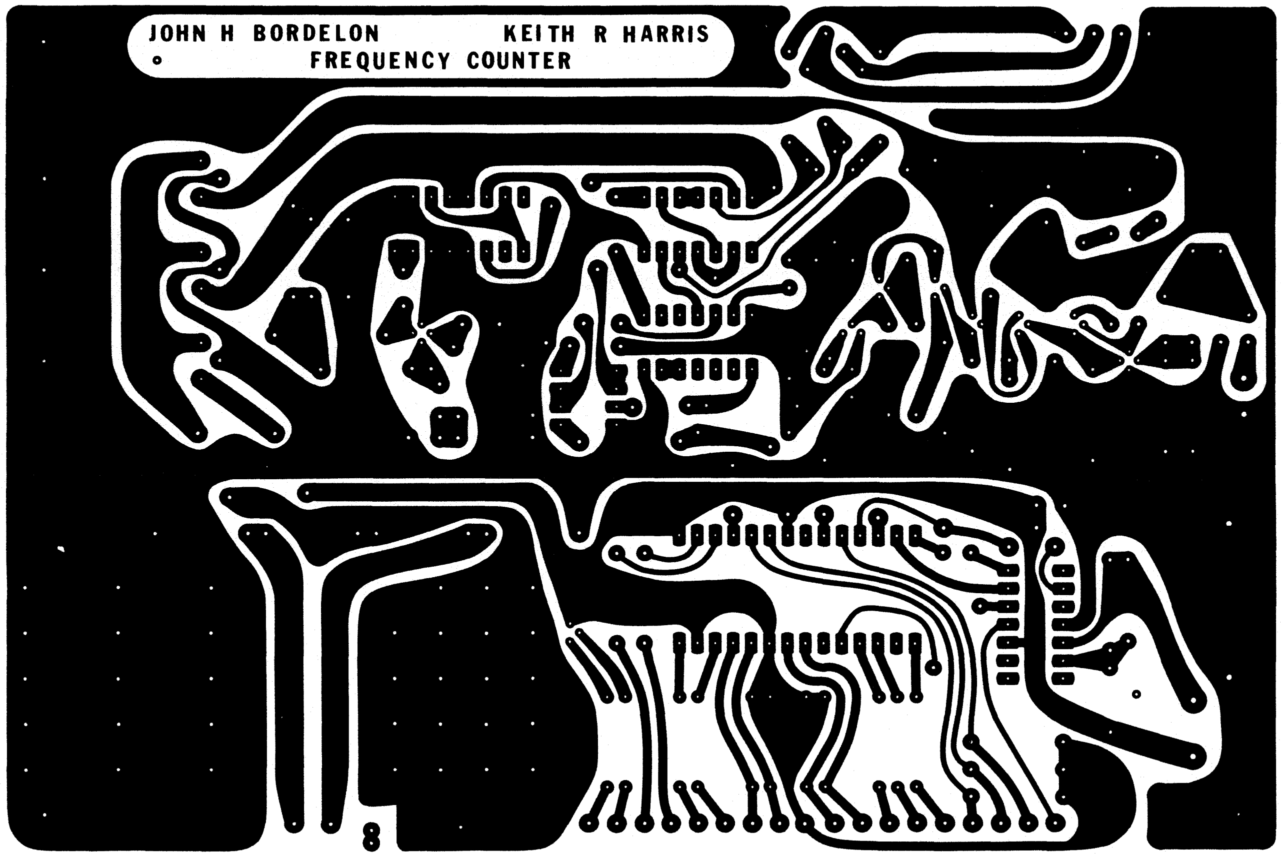


fig. 2. Foil pattern for the 500-MHz frequency counter.

The 500-MHz prescaler uses the popular and readily available Fairchild 11C90. This ECL device not only has a minimum count rate of 500 MHz, it directly drives TTL, and has built-in biasing networks to insure optimum sensitivity. It does not have the older 95H90's reputation for being picky about supply voltage.

The 2N5179 has got to be one of the most bang-for-the-buck transistors around. Reviewing articles for using it to drive 95H90s to 300 MHz led me to think it would almost certainly work well at 500 MHz, especially in view of its f_T of over 1 GHz. Such is the case, and the sensitivity at 500 MHz is 200 mV RMS, and 25 mV at 50 MHz. A 0.3m (1-foot) antenna yields solid counts on a l-watt, 2-meter mobile 6.1m (20 feet) away. The low-input impedance is not a handicap in most cases, and a virtue in some. L1, in combination with the 2N5179's output capacitance, serves as a low-*Q* resonant circuit.

the complete counter

Referring again to the schematic, I chose a very common IC, the 74196, for the 50-MHz prescaler (U3). U2, a 74S00, routes the signals as directed by the range switch (S2). This also eliminates the need for any front-panel rf switching. Q9 drives the appropriate display decimal point, under command of S1 and S2, to provide a display that always reads in kHz. The CMOS counter will not count to its specified maximum frequency when driven by TTL; the output voltage swing of TTL is too low. U9 cures this problem by providing a signal that swings from nearly ground to very near the supply voltage. Observant readers may notice that the signal goes through the 50-MHz prescaler, even when the 5-MHz range is in use. I've applied the signal to the Q_D data input; with the count/load control (pin 1) low, the Q_D signal will follow the data input. When pin 1 is high, inputs on pin 8 are accepted and prescaled by 10. Taking advantage of the architecture of the 74196 in this way eliminated the need for any extra gates.

Resistors R33 through R40 are current limiting resistors for the FND359 displays which I plugged into wire-wrap sockets. R26 through R32 are recommended by Intersil to preclude digit driver leakage from causing "ghosting" on the display.

The power supply is conventional. CR17 allows use of external battery power and protects the battery pack should ac be applied to the counter power supply. It should be noted that U8 supplies U4, U5, and U9. Otherwise, U5 may be damaged by any voltage difference between the two regulators. I designed my circuit board (**fig. 2**) to take care of this.

construction

A printed-circuit board seemed the only reasonable way to construct the counter." **Fig.** 2 shows the layout of the board. The two front ends use good high-frequency layout practices. Although two-sided boards are generally thought to be *de rigueur* for such situations, they are not always necessary. My counter does not talk to itself. One reason for this is the use of broad ground planes dividing and encircling some circuits. Note especially the layout of the 500-MHz preamp and prescaler.

The chassis is homemade, black anodized aluminum. Transfer letters were used for labeling, with clear spray enamel used to protect them. Enamel does not seem to cause smearing of the letters, but caution was exercised.

There are no adjustments to the 500-MHz front end. In use, a no-count condition may just as likely be an overload as well as indication of not enough signal.

The 50-MHz front end requires R5 to be adjusted for maximum sensitivity. The best way to adjust it is with the aid of a signal generator having a variable output attenuator. This way the input signal may be reduced in small steps, touching up the adjustment of R5, until maximum sensitivity is obtained.

Adjustment is most critical at 50 MHz, so finai alignment should be done there. Because my 74196 was of questionable origin, the maximum count rate was just shy of 50 MHz. If operation at 50 MHz seems impossible, drop down in frequency and come up in small steps, touching up R5 each time. Don't forget that if the 74196 is not up to snuff, the 11C90 will be handicapped as a result. Check the former before blaming the latter.

conclusion

Two of these counters have been built and are operational. About twenty-two more are in various stages of construction around the southeastern United States. Minimum cost, exclusive of the circuit board, is about $50, with a maximum cost of about $85.

"Etched, drilled, and plated circuit boards, of G-10 epoxy fiberglass, with 7 pages of documentation, are available for $15, postpaid, from the author.

reference

1. Marion D. Kitchens, Jr., K4GOK, "VHF Prescaler for Digital Frequency Counters," *ham radio,* February, 1976, page 32.

ham radio

precision temperature control for crystal ovens

The National LM3911 temperature-controller IC is featured in a circuit for precision temperature measurement and control in commercially available crystal ovens

Many amateur stations are equipped with frequency synthesizers and digital counters. These circuits require a highly stable frequency source. The most commonly used frequency standard uses the 100-kHz quartz crystal. Such standards exhibit drift with changes in temperature. More recently, the trend has been toward the use of high-accuracy AT-cut crystals operating in the 4- to 10-MHz range. These crystals also exhibit some temperature drift. They must be used in a *good* crystal oven to obtain a stability of better than one part per million over extended periods.

Crystal ovens fall into two categories: on-off (or bang-bang) and proportional control. In the former, a bimetallic strip makes and breaks the heater circuit. Most inexpensive ovens used by amateurs are of this type. Temperature variations frequently exceed 9°F (5°C). These ovens are somewhat erratic in operation, especially the older units that have worn and pitted strip contacts. With proportional control, the heater current is continuously and automatically varied to maintain constant temperature.

This article shows how the National Semiconductor LM3911N temperature controller IC can be used for precision temperature measurement and control. This marvelous little device, selling for under $2.00, makes it possible to maintain oven temperatures to better than 0.18°F (0.1°C). The LM3911N can be used to replace the thermostat in an existing oven or

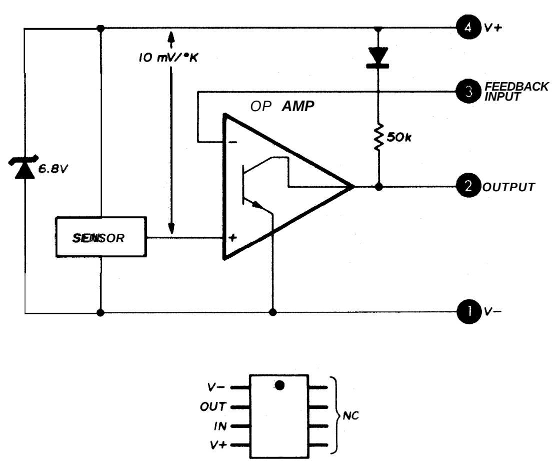


fig. 1. Block diagram of the LM3911N temperature controller. The pinout diagram is shown as a top view.

By Fred Schmidt, K4VA, 3848 Parkcrest Drive, N.E., Atlanta, Georgia 30319

can be incorporated into an easily and inexpensively built oven.

The LM3911N* consists of an operational amplifier, zener, and a temperature sensor in an 8-pin plastic DIP. The block diagram, taken from the data booklet is shown in **fig. 1.** The sensor develops 10 millivolts per degree Kelvin between the positive supply terminal on the op amp chip and its noninverting input.

on-off and proportional control

Figs. 2 and **3,** also from the device instruction booklet, show how the LM3911N can be used for simple on-off and proportional control respectively. In on-off control, the operational amplifier is used as a comparator. When the sensor temperature increases to the value at which the noninverting input voltage is equal to the inverting input voltage, the output switches from about 6 volts to a fraction of a volt. The output can be used to control oven heater current by means of a suitable power transistor. The voltage at the inverting input terminal determines the temperature at which the output switches.

The proportional-control circuit **(fig. 3)** generates a square wave at the output terminal. The duty cycle

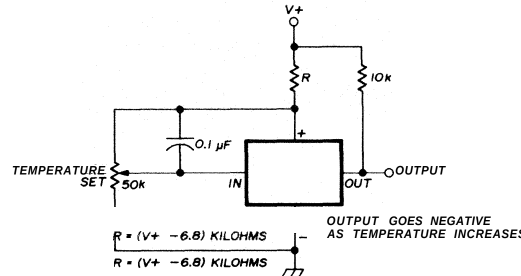


fig. 2. Circuit for on-off temperature control.

(ratio of OFF to ON time) is determined by the sensor temperature and the voltage at the inverting input terminal. Any departure of temperature from the desired value causes the duty cycle to change. This action is used to change the average heater current in such a way as to bring the temperature back toward the desired value. Proportioning bandwidth refers to the temperature range over which the output is a square wave. When the temperature is above

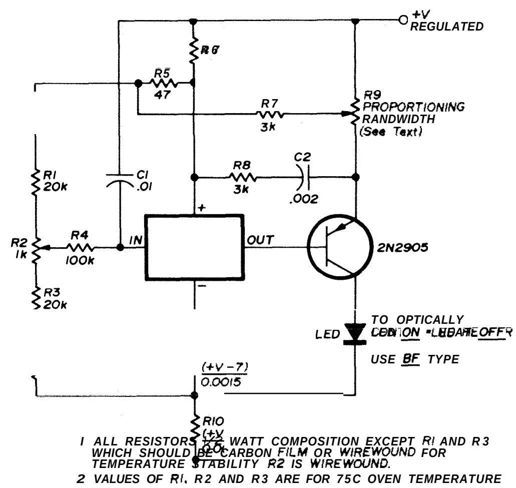


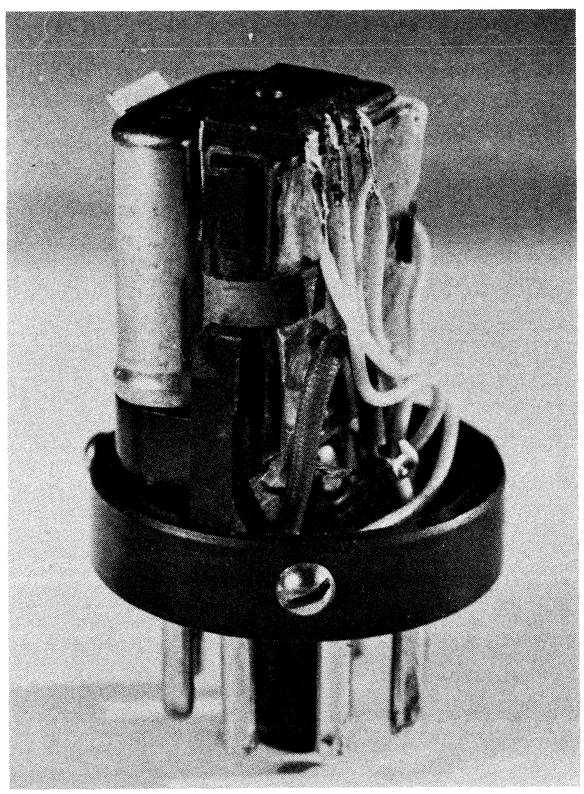
fig. 3. Circuit for proportional temperature control.

the bandwidth the output will remain low, and when the temperature is below the bandwidth output will remain high. The square-wave frequency is determined by the time constant R4C1. The frequency varies somewhat over the bandwidth and is maximum at the center.

The complete circuit for proportional control of an oven is shown in **fig. 4.** The 4N30 is a 6-pin dual inline IC containing a light-emitting diode, which is optically coupled to a photo-sensitive darlington transistor. It drives a power transistor. The oven heater is in the power-transistor collector circuit. During the ON intervals of the square wave, the power transistor is driven to saturation. During the OFF intervals it is cut off.

The photo shows how the LM3911N is installed in an oven of the ON-OFF variety. The oven is a Bliley

*Available from Tri-Tek, Inc., 6522 North 43 Avenue, Glendale, Arizona 85301, Specifications and Applications Booklet, $.80 (also, Radio Shack RS3911, Catalog 276-1706, $2.19).



Installation of the LM3911 IC in a commercial oven.

TC0-1A surplus unit with a 6.3-volt, 0.85-A heater element. The thermostat was unsoldered and discarded.

The LM3911N was cemented to the inverted-U copper strip, which is used to conduct heat from the heating element to the crystal. The LM3911N die is on the base of the package. Therefore, it's important that the base be coupled as closely as possible to the heat source.

Pins 4-8 are not used electrically and should be bent in such a way as to make contact with the copper strip to help conduct heat into the package. Pins 1-4 are connected to pins in the oven socket.

testing

Performance can be checked by connecting a voltmeter across the heating element. When first turned on, the heater voltage will nearly equal the supply voltage. As the oven heats up, the heater voltage will go through several damped oscillations about a final value of about three volts. (The damped oscillation is caused by the thermal lag between the heating element and the copper heat conducting strip.)

The maximum temperature overshoot is about 1 degree C. Bandwidth control R9 should be set about two-thirds of the way down from the end connected to the supply line. R1 and R3 should be wirewound or metal-film resistors. (Composition resistors have a large temperature coefficient and would cause the oven temperature to change somewhat with changes in ambient temperature.)

R1 and R3 values can be changed if it's desired to operate the oven at a temperature other than 75°C. The maximum operating temperature of the LM3911N is 85°C.

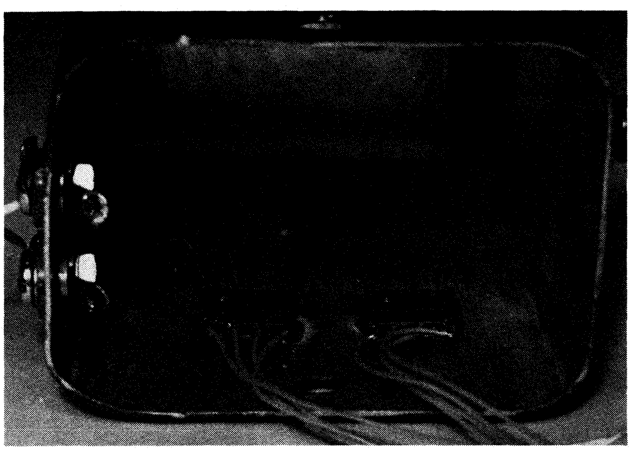
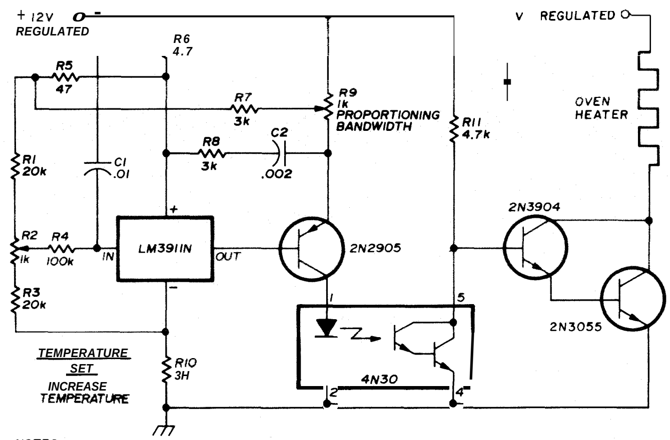


Photo of can inside showing the placement of the LM3911 ICs — one for temperature control and the other for temperature measurement.

Temperature calibration can be performed by drilling a small hole in the metal cover just above the crystal socket and inserting a thermometer into the



NOTES:
1. ALL RESISTORS 1/2 WATT COMPOSITION. EXCEPT R1 AND R3. WHICH SHOULD BE METAL FILM OR WIREWOUND.
2. VALUES OF R1, R2 AND R3 ARE FOR OVEN TEMPERATURE OF 75C
3. +V SHOULD BE 5-6 VOLTS FOR 6.3 VOLT SURPLUS OVEN: 9-12 VOLTS FOR HOME BREW OVEN.

fig. 4. Complete circuit for a proportional temperature-control crystal oven.

oven. The thermometer will take a minute or more to reach final temperature.

homebrew oven

If a surplus oven isn't available you can make one easily and inexpensively. A homebrew oven offers several advantages:

1. It can be made large enough to include the complete oscillator circuit for enhanced stability.

2. There is room for an additional LM3911N for temperature measurement.

3. It can be constructed with very close thermal coupling between the heating element and the controller to avoid temperature oscillations (hunting).

The photo shows how an oven can be made with a surplus i-f transformer can. The unit measures 1-1/2 by 2 by 3-1/2 inches (38x51x89mm).

The heating element consists of no. 28 (0.3mm) enamelled copper wire wound directly onto the aluminum can. The wire should be close wound over a length of about 2-1/2 inches (64mm). After winding, check for short circuits to the can then paint with Red X Corona Dope (General Cement Catalog no. 50-2), or with Dipping Varnish (General Cement Catalog no. 56-2).

Connections are brought out to two screws, which are insulated from the can by fiber shoulder washers. The oscillator circuit is built onto a PC board, which is held in place by two metal brackets. At the top of the board is a ceramic trimmer capacitor for frequency adjustment. A hole in the top of the can allows screwdriver adjustment for frequency trimming.

Two LM3911N chips are cemented to the inner surface of the can as shown in the photo. The second chip is used as an electronic thermometer. After assembly, the oven is insulated with two layers of styrofoam. A convenient source of this material is the 16-ounce (472ml) styrofoam drinking cup.

In the unit shown, the cold resistance of the heater winding is about 6 ohms. The regulated supply for the heater should deliver between 9 and 12 volts. For a smaller-size can it may be necessary to use smaller wire for the heater. A cold resistance of 6 to 8 ohms is about right.

thermometer circuit

The thermometer circuit is shown in **fig. 5.** Resistance values are for a 0-1 mA meter and a temperature range of 70° - 80°C (158-176°F). Other temperature and meter ranges can be obtained by changing the resistance values according to the equations in the appendix. It's not necessary to install a milliammeter permanently in the unit: a pair of terminal posts can be instaiied to allow temperature to be monitored with your multimeter.

Many AT cut crystals exhibit a turning point in the neighborhood of 60°C (140°F). Advantage should be taken of this by operating the oven at that temperature. The turning point can be found by observing the crystal frequency as the oven warms up. If the frequency initially drops then starts to rise, adjust the oven temperature to obtain the minimum frequency. The oven temperature should be at least 5°C above maximum ambient in the equipment in which it will be installed.

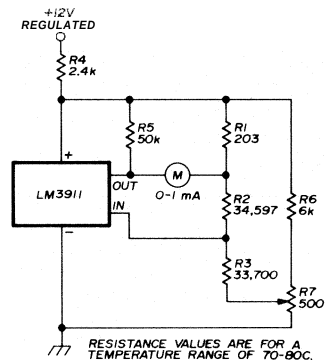


fig. 5. Circuit of the LM3911N IC as an electronic thermometer.

The performance of either oven leaves little to be desired. Temperature stability is reached in about 5 minutes. Temperature variation is too small to be detected on the thermometer calibrated from 70°-80°C (158°-176°F).

acknowledgement

I wish to thank Jim Bell, K4FUP, for suggesting the use of the LM3911N in this project.

appendix

The resistances for the thermometer circuit are calculated by the following equations:

$$R_1 = \frac{(6.85)(0.01)(\Delta T)}{I_M\,(6.85 - 0.01T_0)} \qquad \textbf{(1)}$$

$$R_2 = \frac{0.01\,T_0 - I_0 R_1}{I_0} \qquad \textbf{(2)}$$

$$R_3 = \frac{6.85}{I_0} - R_1 - R_2 \qquad \textbf{(3)}$$

$$R_4 = \frac{V^+ - 6.85}{0.001 + I_M + I_0} \qquad \textbf{(4)}$$

$$R_5 = 50K \qquad \textbf{(5)}$$

$$R_6 = 6k$$

$$R_7 = 500 \text{ ohms} \qquad \textbf{(6)}$$

where

A T =Meter temperature span (degrees C)

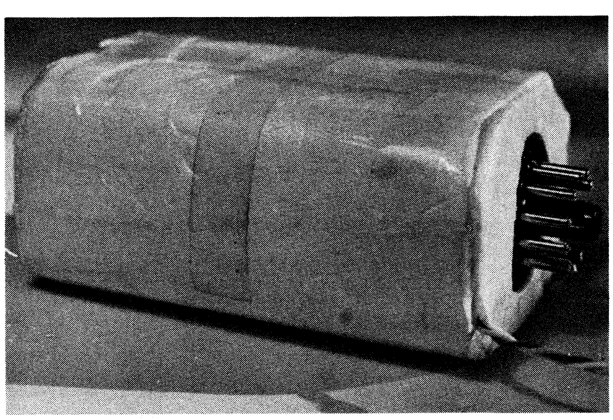
I,, = Meter full scale current (amperes)

T_0 = 5 + meter zero temperature (degrees K)

I_0 =Current through R_1, R_2, and R_3 at zero meter current (use0.0001 amperes)

Notes:
1. °K = °C + 273.
2. Resistors should be metal film.
3. If no thermometer is available for calibration set potentiometer to its midpoint.

ham radio



The assembled oven in its styrofoam jacket.

satellite tracking — pointing and range with a pocket calculator

How to use a pocket calculator to calculate antenna pointing angles and range for an earth-bound or satellite station

A frequent question in long-distance radio communications is where to point the beam. This article contains algorithms for RPN pocket calculators to calculate distances and headings between stations on the earth, and pointing angles and slant ranges to an earth satellite or the moon.

earth stations

Given the longitude λ_1 (west longitudes are +) and latitude ϕ_1 (north latitudes are +) of station 1 (home) and the longitude λ_2 and latitude ϕ_2 of station 2; the algorithm below* calculates A, the initial heading or pointing angle (*north* reference clockwise azimuth) from station 1 toward station 2, and D, the great-circle distance between stations.

Alternatively, if λ_2 and ϕ_2 are the coordinates of the sub-satellite point and h is the height of a satellite over the surface of the earth, then the algorithm also gives E, the elevation look angle and r, the slant range or straight-line distance from station 1 to the satellite. If ***E*** comes out negative, then the satellite is invisible below the horizon.

On an HP-45 calculator, the longitudes and latitudes can be keyed in the format DD.MMSS (degrees, minutes, and seconds of arc):

(ϕ_2: DD.MMSS) [G] {D.MS→} 1 [G] {– R}
(λ_1: DD.MMSS) [G] {D.MS→}
(λ_2: DD.MMSS) [G] {D.MS→} [–] [x↔y]
[G] {→R} [x–y] [↑] [R↓][R↓] [→P]
[x–y]
(ϕ_1: DD.MMSS) [G] {D.MS→} [–] [x↔y]
[G] {→R} [R↓] [x↔y] [R↓] [→P] [x–y]
(if negative: 360 [+]; see A in degrees)
[R↓] [x↔y] [→P] [x–y]

At this point choose one **of** the following **two** options:

A. 69.1 [x] (see D in miles).
B. (h, miles) [↑] 3958 [+] [G] {→R} 3958
[–] [x↔y] [→P] (see r in miles) [x–y]
(see E in degrees).

For an HP-21 calculator, the longitudes and latitudes should first be converted to decimal degrees. Select DEG mode, then:

(ϕ_2, degrees) [↑] 1 [B] {– R}
(λ_1, degrees) [↑]
(λ_2, degrees) [–] [x↔y] [B] {→R} [x↔y] [↑]
[R↓] [R↓] [B] {→P} [x↔y]
(ϕ_1, degrees) [–] [x–y] [B] {– R) [R↓] [x–y]
[R↓] [B] {– P} [x↔y] (if negative: 360
[+]; see A in degrees) [R↓] [x↔y]
[B] {– P} [x–y]

By John A. Ball, Oak Hill Road, Harvard, Massachusetts 01451 (Mr. Ball is a radio astronomer at the Center for Astrophysics in Cambridge, Massachusetts)

*These algorithms are based on similar algorithms in my "Algorithms for the HP-45 and HP-35," Center for Astrophysics Technical Report, Cambridge, Massachusetts, March 9, 1975; and in Appendix A.7 of Algorithms for *RPN Calculators*, to be published by John Wiley & Sons, New York.

At this point choose one of the following two options:

A. 69.1 [x I (see D in miles).
B. (h, miles) [↑] 3958 [STO] [+ I [B] {− R) [RCL] [−] [x− yI[B] {→P} (see r in miles) [x − yI (see E in degrees).

For a Corvus-500 calculator:

(ϕ_2, degrees) [ENT] [SIN] [$y \leftrightarrow x$] [COS]
(λ_1, degrees) [ENT]
(λ_2, degrees) [− I [$y \leftrightarrow x$] [INV] [G] { − POL} [$y \leftrightarrow x$] [ENT] [R↓] [R↓] [G] {−-POL) [$y \leftrightarrow x$]
(ϕ_1, degrees) [−] [$y \leftrightarrow x$] [INV] [G] {− POL) [R↓] [$y \leftrightarrow x$] [R↓] [G] {− POL) [$y \leftrightarrow x$] (if negative: 360 [+ I; see A in degrees) [R1I [$y \leftrightarrow x$] [G] { → POL} [$y \leftrightarrow x$]

At this point choose one of the following two options:

A. 69.1 [x] (see D in miles).
B. (h, miles) [ENT] 3958 [+] [INV] [G] {− POL}3958 [−] [$y \leftrightarrow x$] [G] { → POL} (see r in miles) [$y \leftrightarrow x$] (see E in degrees).

On some Corvus-500 calculators, [R↓I is just [1I, and [$y \leftrightarrow x$] is [$x \leftrightarrow y$].

If you prefer to work in kilometers (km), change the constant 69.1 miles/° to 111.2 km/° (this is the length of 1° on the earth's surface) and change 3958 miles to 6370 km in two places (this is the radius of the earth). Also change the units of D, h, and r. To work in decimal degrees on the HP-45, replace the first two appearances of [G] {D.MS→} by [↑], and drop the last two [G] {D.MS→}. On an HP-25 calculator, use the HP-45 algorithm but change [G] {D.MS→} to [g] {→H}, [G] {→R} to [f] {→R}, and [→P] to [g] {→P}. This algorithm is approximate because it uses a spherical earth and E is not corrected for refraction (which can be as much as ½°).

Example. $\lambda_1 = 71°03' = 71.°05$, $\phi_1 = 42°22' = 42.0367$ (Boston), $\lambda_2 = 74°$, $\phi_2 = 40°42' = 40.°7$ (New York City); get $A = 234°$ (southwest by west), $D = 191$ miles away. If a satellite is 900 miles directly over New York City, then $r = 925$ miles and $E = 75°$ as seen from Boston. If $\lambda_2 = 70°40' = 70.°667$, $\phi_2 = -33°$ 25' = − 33.417' (Santiago de Chile on the west coast of South America); get A = 179.°7 (slightly east of south) and D = 5237 miles from Boston.

As an exercise, compare the distance from Los Angeles to London with the distance from Los Angeles to Rio de Janeiro.

the moon and other celestial objects

This algorithm, slightly modified, also gives pointing angles for the moon. Substitute the local sidereal time T for λ_2, the moon's right ascension a for λ_1, and the moon's declination 6 for ϕ_2. Multiply by 15 to change the units of a and T from hours to degrees. For the moon, h should be about 235,000 miles, but an error of less than 1° in E comes from taking $h = \infty$. With these changes, the HP-45 algorithm becomes:

(δ: DD.MMSS) [G] {D.MS→} 1 [G] {− R}
a : HH.MMSS) [G] {D.MS→}
(T: HH.MMSS) [G] {D.MS→} [−] 15 [x] [$x \leftrightarrow y$] [G] {→R} [$x \leftrightarrow y$] [↑] [R↓] [R↓] [→P] [x− yI
(ϕ_1: DD.MMSS) [G] {D.MS→} [−] [$x \leftrightarrow y$] [G] {− R) [R↓] [x− yI [R↓] [→P] [x− yI (if negative: 360 [+ 1, see A in degrees) [R↓] [→P] [$x \leftrightarrow y$] (see E in degrees).

For an HP-21, α and T should first be converted to decimal hours, and 6 and ϕ_1 to decimal degrees, then:

(δ, degrees) [↑] 1 [B] {→R}
(a, hours) [↑]
(T, hours) [− I 15 [x] [x− yI[B] {→R} [x− yI [↑] [R↓] [R↓] [B] {→P} [$x \leftrightarrow y$]
(ϕ_1, degrees) [−] [x− yI[B] {− R) [R↓] [x− yI [R↓] [B] {→P} [$x \leftrightarrow y$] (if negative: 360 [+]; see A in degrees) [R↓] [B] {→P} [$x \leftrightarrow y$] (see E in degrees).

For a Corvus 500:

(δ, degrees) [ENT] [SIN] [$y \leftrightarrow x$] [COS]
(a, hours) [ENT]
(T, hours) [− I 15 [x] [$y \leftrightarrow x$] [INV] [G] {− POL}[$y \leftrightarrow x$] [ENT] [R↓] [R↓] [G] {− POL) [$y \leftrightarrow x$]
(ϕ_1, degrees) [−] [$y \leftrightarrow x$] [INV] [G] {− POL) [R↓] [$y \leftrightarrow x$] [R11 [G] {− POL}[$y \leftrightarrow x$] (if

note on notation

Keystroke symbols in brackets (e.g., [+]) are printed on the top of the key, those in braces (e.g., { → R}) on the side of the key or on the land area above or below the key. [G] represents an unlabelled gold-colored key, [B] an unlabelled blue key, and [↑] stands for [ENTER↑I. The symbol :| is analagous to a musical repeat symbol and means loop back to the last preceding colon (:) not in parentheses. Parameters to be keyed or read and comments or optional sequences are in parentheses.

The symbol DD.MMSS means degrees, minutes, and seconds of arc, with two digit locations for each. The decimal point after DD must be keyed. Any digits following SS will be taken for a decimal fraction of a second. Similarly HH.MMSS means hours, minutes, and seconds of time. Use [CHS] for negative numbers. For example − 5.420202 would mean − 5°42′02.″02 with DD.MMSS or − 5h42m 02s02 with HH.MMSS. For details, see the instruction booklet with the calculator.

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM.

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00			GMT				in HH.MMSS
01	15 00	g →H	GMT				in hours
02	23 00	STO 0	GMT				
03	15 01	g FRAC	*				* GMT mod 1
04	24 02	RCL 2	$\Delta\alpha$	*			
05	61	×	**				** change in α
06	24 01	RCL 1	α	**			α on the hour
07	51	+	α				α at GMT
08	24 00	RCL 0	GMT	α			
09	24 07	RCL 7	r_7	GMT	α		
10	51	+	GMT+r_7	α			
11	24 06	RCL 6	15.041	GMT+r_7	α		
12	61	×	T	α			T in degrees
13	41	-	$\alpha-T$				in degrees
14	24 00	RCL 0	GMT	$\alpha-T$			
15	15 01	g FRAC	*	$\alpha-T$			* GMT mod 1
16	24 04	RCL 4	$\Delta\delta$	*	$\alpha-T$		
17	61	×	$\dagger\dagger$	$\alpha-T$			$\dagger\dagger$ change in δ
18	24 03	RCL 3	δ	$\dagger\dagger$	$\alpha-T$		δ on the hour
19	51	+	δ	$\alpha-T$			δ at GMT
20	01	1	1	δ	$\alpha-T$		unit vector
21	14 09	f →R	$\cos\delta$	$\sin\delta$	$\alpha-T$		
22	21	$x\leftrightarrow y$	$\sin\delta$	$\cos\delta$	$\alpha-T$		
23	22	R↓	$\cos\delta$	$\alpha-T$		$\sin\delta$	
24	14 09	f →R	§	$\dagger$		$\sin\delta$	§$\cos\delta\cos(\alpha-T)$
25	22	R↓	$\dagger$		$\sin\delta$	§	$\dagger\cos\delta\sin(\alpha-T)$
26	22	R↓		$\sin\delta$	§	$\dagger$	
27	22	R↓	$\sin\delta$	§	$\dagger$		
28	15 09	g →P	—	—	$\dagger$		
29	21	$x\leftrightarrow y$	—	—	$\dagger$		rotate
30	24 05	RCL 5	ϕ_1	—	—	$\dagger$	clockwise
31	51	+	—	—	$\dagger$	$\dagger$	by ϕ_1
32	21	$x\leftrightarrow y$	—	—	$\dagger$	$\dagger$	
33	14 09	f →R	—	—	$\dagger$	$\dagger$	
34	21	$x\leftrightarrow y$	—	—	$\dagger$	$\dagger$	
35	22	R↓	—	$\dagger$	$\dagger$	—	
36	15 09	g →P	—	A	$\dagger$	—	
37	22	R↓	A	$\dagger$	—	—	convert
38	22	R↓	$\dagger$	—	—	A	rectangular
39	22	R↓	—	—	A	$\dagger$	to
40	21	$x\leftrightarrow y$	—	—	A	$\dagger$	spherical
41	15 09	g →P	1	E	A	$\dagger$	coordinates
42	22	R↓	E	A	$\dagger$	1	
43	13 00	GTO 00	E	A	$\dagger$	1	
44							
45							
46							
47							
48							
49							

MEMORY REGISTERS
R0 GMT,
hours
R1 α,
degrees
R2 $\Delta\alpha$,
degrees/hour
R3 δ,
degrees
R4 $\Delta\delta$,
degrees/hour
R5 ϕ_1,
degrees
R6
15.04106864
R7 r_7=
$(S-\lambda_1)/15.041$
+24×Day-Number

fig. 1. HP-25 program to calculate antenna pointing angles for the moon or other celestial objects.

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM.

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00			τ				in minutes
01	23 07	STO 7	τ				save τ
02	04	4	4	τ			
03	23 71 07	STO ÷7	4	τ			$\tau/4$ to R7
04	24 04	RCL 4	$180-\iota$	4	τ		
05	14 21	f $\bar{x}$	$360\tau/\rho$	$180-\iota$	4	τ	
06	14 04	f sin	*	$180-\iota$	4	τ	* $\sin(360\tau/\rho)$
07	14 09	f →R	—	$\sin\phi_2$	4	τ	
08	14 21	f $\bar{x}$	$360\tau/\rho$	—	$\sin\phi_2$	4	
09	14 05	f cos	$\dagger$	—	$\sin\phi_2$	4	$\dagger$ $\cos(360\tau/\rho)$
10	15 09	g →P	$\cos\phi_2$	$\Delta\lambda-\tau/4$	$\sin\phi_2$	4	
11	21	$x\leftrightarrow y$	$\Delta\lambda-\tau/4$	$\cos\phi_2$	$\sin\phi_2$	4	
12	24 07	RCL 7	$\tau/4$	$\Delta\lambda-\tau/4$	$\cos\phi_2$	$\sin\phi_2$	
13	51	+	$\Delta\lambda$	$\cos\phi_2$	$\sin\phi_2$	$\sin\phi_2$	
14	24 01	RCL 1	—	$\Delta\lambda$	$\cos\phi_2$	$\sin\phi_2$	
15	51	+	**	$\cos\phi_2$	$\sin\phi_2$	$\sin\phi_2$	** $\lambda_2-\lambda_1+90$
16	21	$x\leftrightarrow y$	$\cos\phi_2$	**	$\sin\phi_2$	$\sin\phi_2$	
17	14 09	f →R	§	—	$\sin\phi_2$	$\sin\phi_2$	§ =
18	31	↑	§	§	—	$\sin\phi_2$	$\cos\phi_2\sin(\lambda_1-\lambda_2)$
19	22	R↓	§	—	$\sin\phi_2$	§	
20	22	R↓	—	$\sin\phi_2$	§	§	
21	15 09	g →P	—	—	§	§	
22	21	$x\leftrightarrow y$	—	—	§	§	
23	24 02	RCL 2	ϕ_1	—	—	§	rotate
24	41	-	—	—	§	§	clockwise
25	21	$x\leftrightarrow y$	—	—	§	§	by $-\phi_1$
26	14 09	f →R	—	—	§	§	
27	22	R↓	—	§	§	—	
28	21	$x\leftrightarrow y$	§	—	§	—	
29	22	R↓	—	§	—	§	
30	15 09	g →P	—	A	—	§	convert
31	21	$x\leftrightarrow y$	A	—	—	§	rectangular
32	23 00	STO 0	A	—	—	§	to
33	22	R↓	—	—	§	A	spherical
34	21	$x\leftrightarrow y$	—	—	§	A	coordinates
35	15 09	g →P	—	§§	§	A	
36	21	$x\leftrightarrow y$	§§	—	§	A	§§ zenith angle
37	24 06	RCL 6	$3958+h$	§§	—	§	
38	14 09	f →R	—	—	—	§	
39	24 05	RCL 5	3958	—	—	—	parallax
40	41	-	—	—	—	—	algorithm
41	21	$x\leftrightarrow y$	—	—	—	—	
42	15 09	g →P	r	E	—	—	
43	24 00	RCL 0	A	r	E	—	
44	13 00	GTO 00	A	r	E	—	
45							
46							
47							
48							
49							

MEMORY REGISTERS
R0 A,
degrees
R1 $\lambda_0-\lambda_1+90$,
degrees
degrees
R3 ρ,
days
R4 $180-\iota$,
degrees
R5 3958,
miles
R6 $3958+h$,
miles
R7 τ or $\tau/4$,
minutes or
degrees

fig. 2. HP-25 program to calculate antenna pointing angles and range for earth satellites.

negative: 360 [+]; see A in degrees) [R↓]
[G] {→POL} [$y\leftrightarrow x$] (see E in degrees).

This is a general pointing algorithm for any distant celestial object. The α and δ of the moon and other objects are in *The American Ephemeris and Nautical Almanac* or *AENA.*[1]

sidereal time

If your shack is not equipped with a sidereal clock, the *AENA* shows how to calculate T, or on an HP-45:

(Day-Number) [↑] 24 [x] [↑] [↑]
(GMT: HH.MMSS) [G] {D.MS→} [+]
1.0027379 [x] [$x\leftrightarrow y$] [−]
(λ_1: DD.MMSS) [G] {D.MS→} 15 [÷] [−]
(S, hours) [+] (if negative: 24 [+], if greater than 24: 24 [−]) [G] {→D.MS} (see T in HH.MMSS).

Day-Number means day of the year, GMT is Greenwich mean time or universal time (UT), and S is from the table below.

Year, AD	S, hours
1976	6.5865
1977	6.6363
1978	6.6204
1979	6.6044
1980	6.5885
1981	6.6383
1982	6.6224
1983	6.6065
1984	6.5906

S is the Greenwich sidereal time on January 0.0 of the indicated year, and is in the *AENA.* For an HP-21, key GMT in decimal hours and delete the following: [G] {D.MS→}; key λ_1 in decimal degrees and replace the following: [G] {D.MS→} by [↑]; and delete the final [G] {→D.MS} and see T in decimal hours. For a Corvus 500, follow the HP-21 scheme but replace [↑] by [ENT] and [$x\leftrightarrow y$] by [$y\leftrightarrow x$]. The precision of this algorithm is about ±1 second. A sidereal clock runs one day per year or 3^m57^s per day faster than an ordinary clock.

Example. On 1976 December 25 (Day-Number = 360) at GMT = 21^h, the moon was at $\alpha = 22^h25^m37^s.564$, $\delta = -5°21'33.''33$ (from the 1976 *AENA,* page 187). The local sidereal time at Boston was $T = 22^h33^m46^s$, and the moon was visible at $A = 182.°7$ (almost south), and approximately $E = 42°$ (just under halfway from horizon to zenith).

The program in **fig. 1** combines the celestial pointing and sidereal-time algorithms for the HP-25.

Key in the program, then initialize:

(α: HH.MMSS) [g] {→H} 15 [x] [STO] 1
($\Delta\alpha$, seconds/hour) [↑] 240 [÷] [STO] 2
(δ: DD.MMSS) [g] {→H} [STO] 3
($\Delta\delta$, arcseconds/hour) [↑] 3600 [÷] [STO] 4
(ϕ_1: DD.MMSS) [g] {→H} [STO] 5
(S, hours) [↑] 15 [x]
(λ_1: DD.MMSS) [g] {→H} [−] 15.04106864
[STO] 6 [÷]

(Day-Number) [↑] 24 1x1 [+] [STO] 7 [f] {PRGM}
Calculate: (GMT: HH.MMSS) [R/S] (see E in degrees) [$x \leftrightarrow y$] (see A in degrees) :|.

If A is negative, you may want to add 360°. For this program, a is the right ascension on the hour, $\Delta\alpha$ is the change in a in one hour, 6 is the declination on the hour, and A6 is the change in 6 in one hour. Don't forget minus signs [CHS] where needed. The rates $\Delta\alpha$ and $\Delta\delta$ are called first differences and are tabulated for the moon in the *AENA.*

This program interpolates linearly through the hour; α and δ should be for the preceding hour. For a planet such as Jupiter, tabulated daily rather than hourly in the *AENA,* change the [g] {FRAC} in lines 03 and 15 both to [g] {NOP) and key α and 6 for Oh, The tabulated first differences in this case need to be divided by 24 to get the right units (per hour rather than per day). Jupiter is an interesting object for amateur radio astronomy. For a star with constant a and 6, set $\Delta\alpha$ and A6 to zero.

Example. Repeat the preceding test case on the moon but for GMT = 21^h20^m. From the 1976 AENA, $\Delta\alpha = 122^s.767$ and $\Delta\delta = 615''.44$. Get A = 189.2° (A = − 170.78 + 360 = 189.2° and E = 41.9°).

satellites

Given Q, a satellite's period in minutes of time; ι, the inclination in degrees of the satellite's orbit to the equator; τ, the time since an ascending node (northbound equator crossing) in minutes; and λ_0, the longitude of the ascending node ($\tau = 0$) in degrees;" the HP-45 algorithm below[2,3] calculates λ_2 and ϕ_2 for the satellite:

(τ, minutes) [↑] [↑] (ϱ, minutes) [÷] 360 [x] [↑] [SIN]
(ι, degrees) [x – yl [G] {– R}[x – yl [G] {SIN – ^{1}} (see ϕ_2 in degrees) [R↓] [x – yl [COSI [→P] [R↓] [$x \leftrightarrow y$] 4 [÷] [–]
(λ_0, degrees) [$x \leftrightarrow y$] [– **I** (see λ_2 in degrees).

This algorithm is approximate because it uses a circular orbit; h is taken to be constant.

For an HP-21, select DEG mode, change [G] to [B] and change [→P] to [B] {→P}. For an HP-25, change [G] to [g] or [f], [SIN] to [f] {sin), [COS] to [f] {cos}, and [→P] to [g] {→P}. For a Corvus 500, change [↑] to [ENT], [$x \leftrightarrow y$] to [$y \leftrightarrow x$], [G] {SIN^{-1}} to [INV] [SIN], [→P] to [G] {– POL), and [G] {– R) to [INV] [G] { – POL}. These algorithms are also easy to convert for the HP-19, HP-27, HP-29, HP-46, HP-55, HP-65, HP-67, HP-91, HP-97, HP-9815,

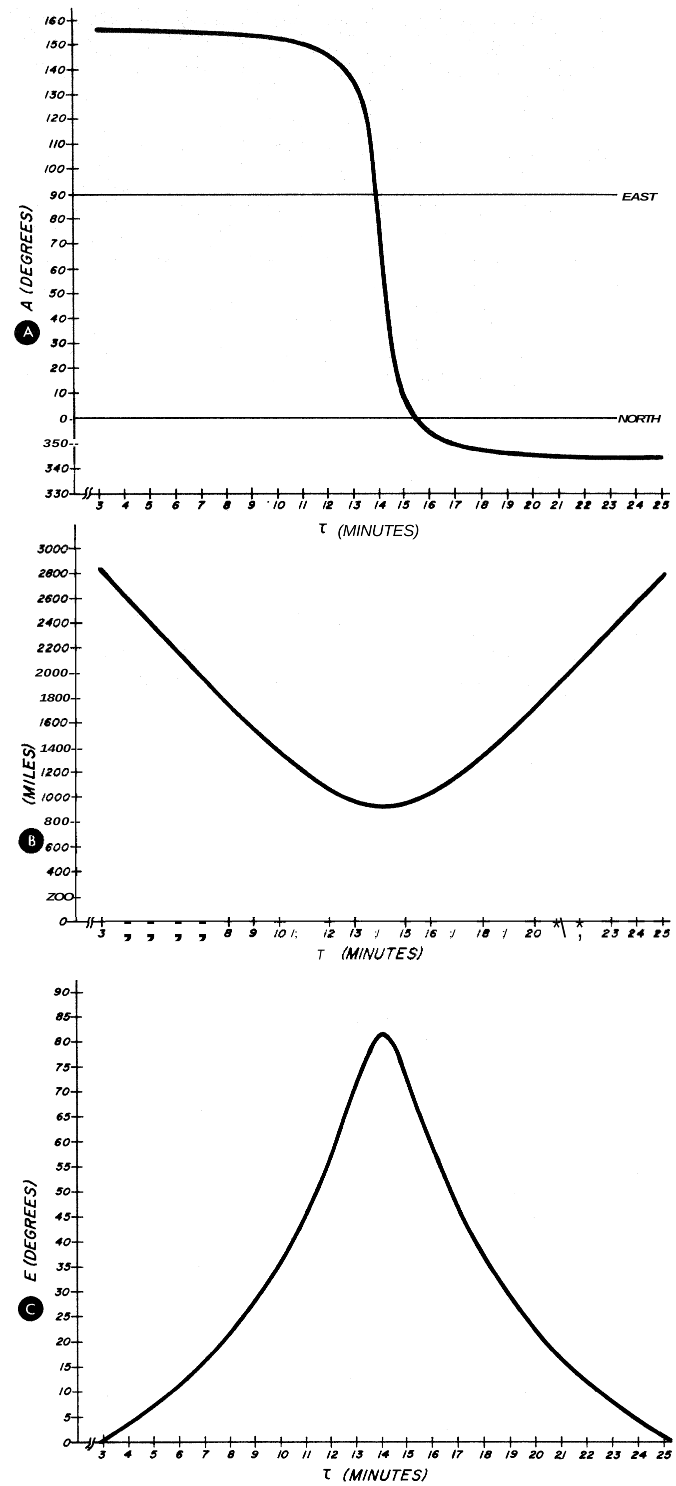


fig. 3. Azimuth, range, and elevation for Oscar 7 orbit 10590 as seen from Boston.

"The orbital elements of Oscar satellites are available from AMSAT, Post Office Box 27, Washington, D.C. 20044. For Oscar 7: ϱ = 114.95 minutes, ι = 101.7 degrees, h = 908 miles, e = 0.001; but these numbers change slowly with time. Also see the *Satellable* available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048. For other satellites, see *Satellite News,* 12 Barn Croft, Preston PR1 OSX, England.

summary of symbols

Station *1* is normally home; station *2* is sometimes a subsatellite point.

	heading (north reference clockwise azimuth) from station *1* toward station *2* or toward a celestial object. $A=0$ is north, $A=90°$ is east, $A=180°$ is south, and $A=270°$ is west. Negative A means counterclockwise, so $A=-90°$ is also west.
c	speed of light; $c = 11,177,000$ miles/minute.
D	great-circle distance from station 1 to station *2*.
Day-Number	day of the year (GMT) available on some desk calendars.
	elevation look angle from station 1 toward a satellite or celestial object. $E=90°$ is straight up toward the zenith, $E=0$ is the horizon, and negative E means that the object is invisible below the horizon.
e	eccentricity of a satellite's orbit; $e=0$ is a circle.
	frequency
GMT	Greenwich mean time or universal time (UT or UTC) Ephemeris time differs from GMT by about 48 seconds.
h	height of a satellite over the surface of the earth.
h_a	the average h over an orbit.
i	inclination angle of a satellite's orbit to the equator.
r	slant range or straight-line distance from station *1* to a satellite.

S	Greenwich sidereal time on January *0.0* (tabulated in the article).
7	local sidereal time at station *1*.
v	Doppler velocity of a satellite relative to station *1*.
	right ascension of a celestial object.
δ	declination of a celestial object.
Δf	change in f due to a Doppler shift.
$\Delta\alpha$	change in α per unit time (first difference).
$\Delta\delta$	change in δ per unit time (first difference!
λ_0	longitude of the ascending node at $\tau=0$ in a satellite's orbit. West longitudes are positive (+), east longitudes are either negative (−) or greater than 180°.
λ_1	longitude of station *1*
λ_2	longitude of station *2* or a subsatellite point
π	*3 1415926536*
ϱ	period (time for one complete orbit) of a satellite
τ	time since an ascending node (northbound equator crossing) in a satellite's orbit.
ϕ_1	latitude of station 1. North latitudes are positive (+), south latitudes are negative (−).
ϕ_2	latitude of station *2* or a subsatellite point.

National Semiconductor 4640, APF Mark 55, and Omron 12-SR calculators. Converting for non-RPN calculators is more difficult.

The program shown in **fig.** 3 combines the satellite and pointing algorithms for the HP-25. Key the program, then initialize:

(λ_0, degrees) [↑] (λ_1, degrees) [−] 90 [+] [STO] 1 (ϕ_1, degrees) [STO] 2 (ϱ, minutes) [↑] 1440 [÷] [STO] 3 180 [↑] (ι, degrees) [−] [STO] 4 3958 [STO] 5 (h, miles) [+] [STO] 6 [f] {PRGM}

Calculate: (τ, minutes) [R/S] (see A in degrees) [R↓] (see r in miles) [R↓] (see E in degrees) :|.

If A is negative, you may want to add 360°. The constant 1440 is the number of minutes in a day.

Example. Oscar-7 orbit 10590 on March 10, 1977, $\tau=14^m$ (which corresponds to $GMT=00^h08^m + 14^m = 00^h22^m$), $\varrho=114^m945$, $\iota=101.°7$, $\lambda_0=5405$, h = 908 miles; get $\phi_2=42.°71$, $\lambda_2=69.°02$, $A=76°$, $r=916$ miles, and $E=82°$, almost overhead as seen from Boston.

Fig. 2 shows A, *r*, and E as a function of τ for this orbit over Boston. For a pass almost overhead such as this, A is nearly constant except for a couple minutes around closest approach. The slope v of the *r* curve is the Doppler velocity (in miles per minute) and can be calculated by subtracting two ys a minute apart. The largest possible v is $2\pi(h+3958 \text{ miles})/\varrho$ or about 266 miles/minute for Oscar 7. Convert v to a Doppler frequency shift using $\Delta f=-fv/c$, where f is the frequency, Δf is the shift in frequency, and c is the speed of light ($c=11,177,000$ miles/minute). The Oscar-7 beacon at 145.975 MHz, for example, shifts as much as 0.00347 MHz and so appears somewhere between 145.972 and 145.978 MHz.

When the satellite is approaching, v is negative, Δf is positive, and the beacon appears at a higher frequency; when the satellite is receding, v is positive, Δf is negative, and the beacon appears at a lower frequency on the dial. If you are working through a repeater on a satellite, then two different Doppler shifts need to be calculated and added.

Kepler's third law relates ϱ and h_a, the average h,

(ϱ, minutes) [↑] [x] 8720351 [x] 3 [1/x] [G] {y^x} 3958 [−] (see ha in miles).

The first number is Kepler's constant for the earth, 8,720,351 miles^3/minute^2. With a circular orbit, h_a and h are the same. For Oscar 7 with $\varrho=114^m945$, get $h_a=908$ miles.

accuracy

These algorithms employ several approximations that cause inaccuracies in the answers. The earth's eccentricity causes an error up to 0.2 per cent (2 miles in 1000 miles) in D. The eccentricity of a satellite orbit has two effects: first, h is not constant. The maximum variation in h is $e(3958 \text{ miles} + h_a)$ where e is the eccentricity and h_a is the average h. So if $e=0.01$ and $h_a=925$ miles, then the variation in h is 49 miles and h varies from 876 to 974 miles.

Another effect of e is on the speed of the satellite, which can be as much as about $e\varrho/\pi$ ahead or behind in the orbit compared to a circular orbit. If $e=0.01$ and $\varrho=115$ minutes, then the satellite can be up to 0.37 minutes early or late. Also the "4" in the orbital algorithm, which converts τ from minutes of time to degrees, should be 411.0027379 = 3.9890783 because the earth's rotation speed is one degree per 4 *sidereal* minutes. The error due to using 4 is noticeable only for large τ.

Finally, the orbital elements of earth satellites change slowly due to influences such as the sun and moon, air resistance, and the uneven distribution of mass in the earth. For Oscar 7, the orbit regresses just about enough to cancel the error caused by using 4 minutes per degree; such orbits are called sun-synchronous,

I wish to thank George Rybicki for showing me the spherical-trigonometry trick used in these algorithms; Dick Ellis, W5YCK, for helping research articles on Oscar satellites; Tom Bates and Fritz Mansvelt-Beck for loaning me their calculators; and R.C. Vanderburgh for sending me copies of his programs.

references

1 *The American Ephemeris and Nautical Almanac* U S Government Printing Office. Washington. D C 70402. published each year

2 P D Thompson Jr . "A General Technique for Satellite Tracking," *QST*, November, 1975, page 29

3. *Specialized Techniques for the Radio Amateur, ARRL.* Newington, Connecticut, 1975, page208

appendix
RPN portable calculators

The table below lists all the available or recently available RPN portable calculators. The algorithms in this article can be easily converted to work on any calculator that has "yes" in the R – P column in this table. Non-RPN calculators are not listed; they are awkward for this type of problem

The Corvus 500, the APF Mark 55, and the Omron 12-SR are internally identical, but the Corvus 500 has a better keyboard and case. The instruction booklet with the Corvus 500 is very poor; if you have this calculator, you should also get the book *Everything You've Always Wanted to Know About RPN but were Afraid to Pursue — Comprehensive Manual for Scientific Calculators* available for $7.50 plus postage from T. K. Enterprises, 16611 Hawthorne Boulevard, Lawndale, California 90260. This book does not live up to its name and is not as good as the instruction books with HP calculators, but is the best book availabie for the Corvus 500, APF Mark 55, and Omron 12-SR.

For an HP calculator, you should get the appropriate HP applications book. Some dealers will throw these in with the calculator; otherwise they are about $10 from Hewlett-Packard, 19310 Pruneridge Avenue, Cupertino, California 95014. Some HP calculators have several such books.

Generally speaking, HP calculators are better made and will probably last longer than the others in this table. Don't overlook the possibility of finding a used or surplus calculator of a model listed as no longer available.

"Start extravagant, and you'll never finish. Get the cheap tool first, see if it feeds your life. If it does, then get a better one. Once you use i: all the rime, get the best. You can only grow into quality. You can't buy it."

– from *The Last Whole Earth Catalog*

Table **1.** List of portable calculators which use RPN architecture

manufacturer	model	stack of	scientific notation	number of storage registers	number of program steps	R – P	D.MS
Hewlett-Packard	HP-19C	4	yes	30	98	yes	yes
Hewlett-Packard	HP-21	4	yes	1	0	yes	no
Hewlett-Packard	HP-22	4	yes	10 + 5	0	no	no
Hewlett-Packard	HP-25	4	yes	8	49	yes	yes
Hewlett-Packard	HP-25C	4	yes	8	49	yes	yes
Hewlett-Packard	HP-27	4	yes	10 + 5	0	yes	yes
Hewlett-Packard	HP-29C	4	yes	30	98	yes	yes
Hewlett-Packard	HP-35*	4	yes	1	0	no	no
Hewlett-Packard	HP-45*	4	yes	9	0	yes	yes
Hewlett-Packard	HP-55*	4	yes	20	49	yes	yes
Hewlett-Packard	HP-65"	4	yes	9	100	yes	yes
Hewlett-Packard	HP-67	4	yes	26	224	yes	yes
Hewlett-Packard	HP-70*	4	yes	2 + 5	0	no	no
Hewlett-Packard	HP-80	4	yes	1	0	no	no
Hewlett-Packard	HP-91	4	yes	16	0	yes	yes
Hewlett-Packard	HP-92	4	yes	30	0	no	no
Hewlett-Packard	HP-97	4	yes	26	224	yes	yes
National Semiconductor	Novus 4510	3	no	1	0	no	no
National Semiconductor	Novus PR4515/4615	3	no	1	100	no	no
National Semiconductor	Novus 4520	4	yes	1	0	no	no
National Semiconductor	Novus PR4525	4	yes	1	100	no	no
National Semiconductor	NS 4640	4	yes	3	0	yes	yes
Corvus	500	4	yes	9	0	yes	no
APF	Mark 55	4	yes	9	0	yes	no
Omron	12-SR	4	yes	9	0	yes	no

*No longer available

ham radio

high-impedance preamp and pulse shaper for frequency counters

Simple circuit for improving your counter from dc to over 60 MHz using readily available devices

Many **of** the inexpensive frequency counters on today's market, as well as some of those built by homebrew enthusiasts, could use some improvement in the preamp and pulse-shaper circuit (sometimes called the trigger circuit). This is the circuit that brings the input signal waveform to TTL level (3.5 volts peak-to-peak). Often counters potentially capable of counting to 40-50 MHz don't produce best results because of an inefficient input circuit. On the other hand, if the trigger works fine at high frequencies, it shows some limitation in squaring low-frequency signals. Kritter[1] solved this problem by using two pulse shapers in his frequency counter. Although this approach is quite satisfactory, it can be impractical and expensive.

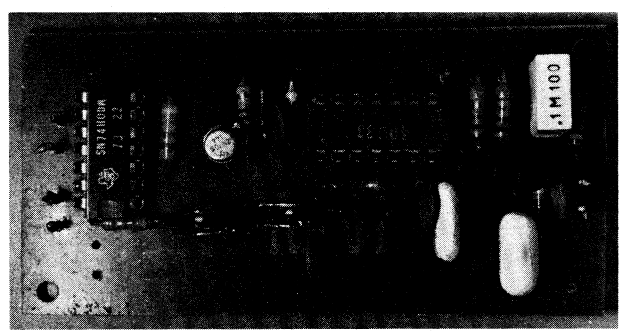
The circuit we're going to examine can process any frequency between dc to over 60 MHz, providing at the output a perfect square wave at TTL level. Furthermore, it's not very expensive, because it uses transistors and ICs available at most discount outlets.

circuit description

The input circuit (fig. 1) is a balanced amplifier using two field-effect transistors. For dc stabililty, the stages are dc coupled. These fets must be selected for the same ID_{SS}.[2] The absolute D, value isn't critical but must be nearly equal for the two devices.

The first stages are source followers, so the input impedance is extremely high. At low-input signal levels, when the two back-to-back protective diodes don't conduct, overload protection is provided only by the 2.2-meg polarization resistor. It would be possible to increase the input impedance even more by increasing this resistor value or by using bootstrap polarization. However, I felt that this value was more than adequate for amateur purposes.

The input stage drives a 733 IC which, according to reference 3, is a differential video amplifier with a bandwidth of over 100 MHz. The gain of this amplifier is selectable by proper connection of pins 3, 4, 11, and 12 (the DIP package).



Top view of the counter preamp PC board showing component layout.

Three different gain values are possible without adding external components: x10, x100, and x400. The last has been chosen for this application. Even if, in this case, the amplifier bandwidth is reduced

By **Giuseppe Beltrami, I4YAF**, Via Meloni 39, 1-41012, Carpi, Italy

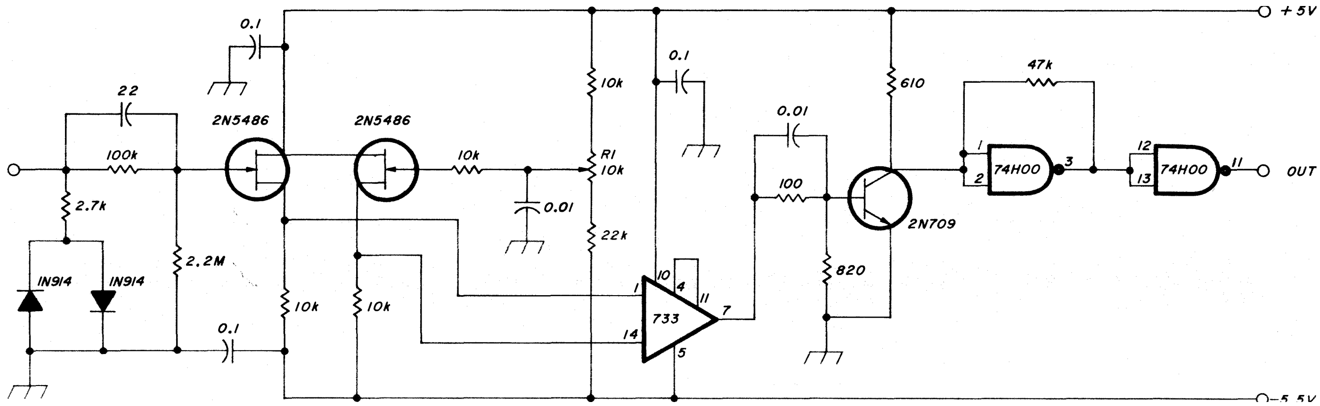


fig. 1. Input circuit for improving inexpensive counters. Circuit requires a dual-polarity supply that delivers at least 63 mA.

slightly, it remains well over 50 MHz — more than sufficient for our use.

A 2N709 switching transistor follows the preamplifier and squares the signal. If you want to reach the maximum frequency that the pulse shaper can handle, this transistor is highly recommended. It's fast. Some attempts that have been made to use other types of transistors (for example, 2N914, 2N2368) for processing 60-MHz signals haven't been very satisfactory. So I highly recommend the 2N709 if you don't want problems. Of course, I haven't tried all the available switching transistors, so there are many substitution possibilities if you like to experiment. If you have some computer transistors, try them; you might obtain even better results.

The last stage is quite conventional. It is a TTL translator using two sections of a high-speed quadruple NAND gate (type 74H00). For this application I've tried several ICs made by different manufacturers. Those that gave the best results were a Texas Instruments SN74H00 and a National Semiconductor DM74H00. I also tried some Schottky devices (74S00). All performed well, even at higher frequencies.

The circuit requires a supply that delivers plus and minus 5 volts. Without an input signal, the input transistor drain current is 18 mA from the negative side and 45 mA from the positive side, which increases to 63 mA with a strong input signal. Tests showed that the circuit was not too sensitive to an unregulated supply, so the filtering doesn't have to be elaborate. Of course devices such as the 320- and 340-series voltage regulator ICs can be used to solve a power-supply problem at a reasonable price. The only recommendation is to avoid any possibility of false counting by bypassing the positive supply as closely as possible to pin 14 of the TTL IC.

construction and alignment

The prototype was constructed on a 3 x 1.5-inch (8.2x4cm) board **(fig. 2).** The component layout shown in **fig. 2B** is slightly different from that in the photo because of some improvements made after the photo was taken. The second trimmer resistor on the PC board was a former polarization control for the level translator, which was disconnected after some tests.

The only alignment required for the circuit is the regulation of the trimmer pot, R1. With the counter connected at the output and a signal source (grid-dip meter or signal generator) at the input, adjust R1 for maximum sensitivity at the highest measurable frequency. With proper alignment, the input sensitivity must be better than 100 mV in the range from a few

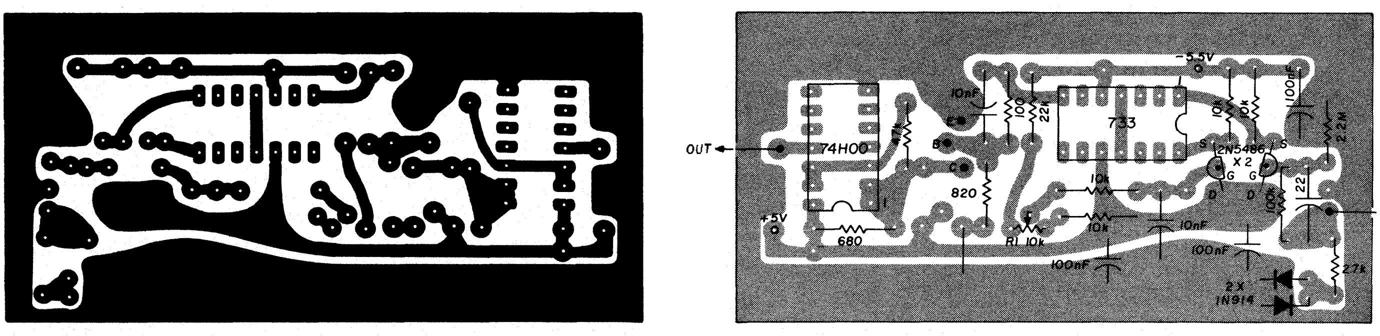


fig. 2. Circuit-board layout and component arrangement for the counter preamp.

MHz to at least 40 MHz, decreasing to about 150 mV at the lower (DC) and upper limit (66 MHz in the prototype).

counter improvements

Some counters using 7400-series ICs may not cover frequencies higher than 30 MHz. It's now time to update them using faster ICs provided by modern technology. Substitute the input gate of the counter (it may be a 7400 or a 7410, perhaps a 7420) with the Schottky TTL equivalent: 74S00, 74S10, or 74S20. The substitution was direct and required no wiring change. Then substitute the first decade divider (7490) with the faster type 74196 or 74S196. Here the connections are different, so you must have a little patience and, following the schematic of **fig.** 3, make the modifications needed on the PC board. The 74196 IC requires a reset pulse inverted with respect to that required by the 7490. If you have an unused NAND or inverter circuit in one of the other ICs, use it to invert the reset pulse; otherwise use one of the two NAND gates not used in the pulse-shaper circuit. With these simple modifications, your frequency counter can now display at least 50 or 60 MHz. Don't forget that some ICs are now available that permit scalers for GHz frequencies. With a pair of these ICs

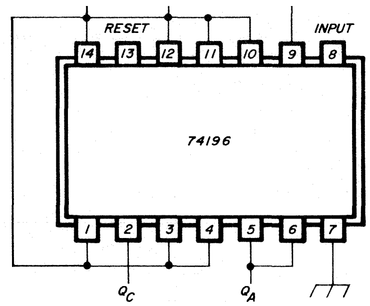


fig. 3. Pin layout for the 74196 decade divider, which may be used in your counter for faster response. Modifications shown must be made on the PC board.

and your improved counter, it will be possible to make precise measurements well into the microwave region.

references

1. W. R. Kritter, DL8TM, "A Dual-Input Preamplifier with 2:1 Prescaler for Frequency Counters from 1 Hz to a Minimum of 100 MHz," VHF Communications, May. 1973, page 91.

2 C. Hall. WA5SNZ, "New FETs Simplify Bias Problems," ham radio, March, 1974, page 50.

3. "Linear Integrated Circuits," National Semiconductors, August, 1973, pages 5-57.

ham radio

wide-range capacitance meter

A portable test instrument that combines three modes of capacitance testing using just a few components

Here's an instrument for the experimenter that combines three modes of capacitor testing into a portable unit, which uses readily available devices: two 555 timer ICs, a 2N5484 fet, and a CA3140 operational amplifier.

features

With the instrument described you can apply a polarized voltage of a few volts to the capacitor under test, one side of which is grounded. A single unregulated 9-volt power supply is used. Capacitance readout is linear.

Three testing modes are available: low capacitance (to 1 μF), high capacitance (to 2500 μF), and a logarithmic indication of the test-capacitor leakage current with up to 8 volts applied. Let's take a look at the circuit **(fig. 1)**.

low-capacitance measurement mode (to 1 μF)

U1, an NE555, operates as a clock, which provides negative-going pulses at about 350 per second to trigger U2, also an NE555. This action unclamps the test capacitor, allowing it to charge through a switch-selected resistor, until it reaches half the supply voltage. At this point U2 resets, discharging the capacitor through pin 7. During the charging period, U2 pin 3 is high (about 8 volts), and the duration of this high state is directly proportional to the test capacitance.

The resulting rectangular waveform can be used to drive a 1-mA meter directly through a 5k trimpot for a simplified circuit. In this instrument, the high signal is attenuated to 0.6 volt across silicon diode CR1 at the noninverting input of U3, a CA3140 op amp with mos input. U3 operates as a unity-gain buffer, which feeds the meter through calibrating trimpot R6. Meter deflection is proportional to the average value of the rectangular waveform output from U3 and is therefore proportional to the capacitance.

Supply voltage is noncritical because:

1. U1 clock frequency is, for practical purposes, independent of voltage.

2. The reset level of U2 pin 6 is at one-half the supply voltage, which compensates for voltage-change effects on the charge rate of the test capacitor through the switched resistors.

3. CR1 operates as a simple regulator, limiting the high signal input to U3 to 0.6 volt.

With no test capacitor applied to the circuit, about 30 pF of internal capacitance exists at U2 pins 6 and 7 (plus strays). To prevent this capacitance from causing a substantial residual reading on the lower ranges, a *negative* capacitance is used to cancel the internal capacitance. This unlikely device is simulated by C3, a 100-pF capacitor, which is connected not to

By P. H. Mathieson, M.I.E. Australia, JAARS, Box 248, Waxhaw, North Carolina 28173

ground but to Q1/Q2 output, which has a small positive voltage gain and a low output impedance. The current through C2 is equal and opposite to that through the residual 30-pF capacitance.

To calibrate, use a known accurate capacitor giving a high deflection on the 0.01 or 0.1 μF ranges. First adjust negative-capacitance trimpot R1 so that the source of Q1 is at the top end of R1. Then calibrate by adjusting R6. There may be a small residual zero error caused by the minimum pulse width from U2, which can be offset by the addition of R7, a 470 k in my instrument. Now switch to the 100-pF range with a known capacitor of 10-20 pF connected and adjust negative cap trimpot R1 until the meter reads the correct value, showing that the strays have been cancelled. The readings are then accurate to a few pF.

high-capacitance measurement mode (to 2500 μF)

This mode uses a single-shot method. U1 is not required. When +9 volts is switched on, U2 is triggered by the momentary low on pins 2 and 4 caused by the uncharged capacitor, C2. As in the previous mode, a high of 0.6 volts is applied to the noninverting input of U3 until the capacitor under test is charged to half the supply voltage. During this period, U3 behaves as an accurate integrator using low-leakage capacitor C4 in the feedback loop. At the end of the high input period, U3 output voltage will be proportional to the duration of that period and therefore to the test capacitance.

Accurate high-value capacitors are difficult to find, so calibration using trimpot R5 is best done with values around 1 μF. Leakage resistance of the test capacitor extends the charging time, causing a false high reading.

After the integrating period the meter should remain stationary while the reading is taken. If drift is a problem (assuming feedback capacitor C4 is not leaky) it may be minimized by correcting the offset in the U3 input stage. Try a 10k resistor R3 from U3 pin 1 or 5 to ground and adjust for minimum drift.

leakage mode

This mode produces a logarithmic indication of the test capacitor leakage current with up to 8 volts applied. The lower end of the capacitor is disconnected from the supply minus, and the leakage current now flows through limiting resistor R8 and diode CR1. The voltage across CR1 bears an approximately logarithmic relationship to the current flowing through it. U3 is again used as a unity-gain buffer, and trimpot R4 is set to produce full-scale deflection with a short circuit across the test terminals. U1, U2, and the negative capacitance amplifier are disconnected from the negative supply line.

Electrolytic capacitors that have remained unused for some time can be reformed in the leakage test mode before their capacitance is measured.

The logarithmic readout can be interpreted by observing the readings obtained with known resistors across the test terminals, ranging from a few kilohms to hundreds of megohms.

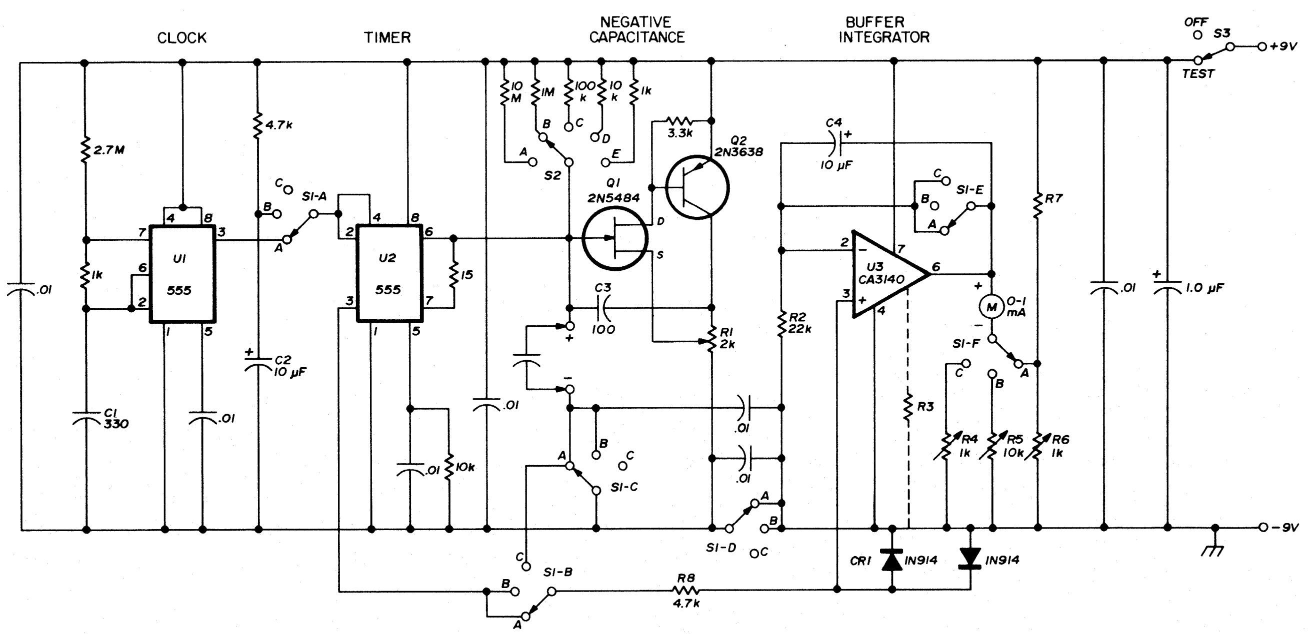


fig. 1. The three-mode capacitor tester. Circuit features linear readout and common devices. Author's original was built on perf board but with a little ingenuity an etched board could be used. See *table* 1 for switch configurations and positions.

table 1. Switch arrangement for the portable capacitance meter.

switch	configuration	position		
S1	3-pole. 3-position, 2 section	A (low-capacitance rnodel B (high-capacitance model C (leakage model		
		mode A μF	**mode B** μF	**mode C**
S2	1-pole, 5-position	A 0.0001	0.25	
		B 0.001	2.5	
		C 0.01	25.0	
		D 0.1	250.0	
		E 1.0	2500.0	leakage
S3	SPST (test)			

The capacitance ranges are in steps of 1:10. Intermediate ranges are most economically obtained as follows. In mode A, **(table 1)** switching to a larger clock timing capacitor, C1, will permit larger capacitances to be read on scale. In mode B, switch resistor R2 to a higher value, which slows the integrating rate and thus allows a greater full-scale capacitance reading. Ten-thousand μF is probably a practical limit because of leakage.

Use accurately scaled values for the switched range resistors connected to S2 for best accuracy.

construction notes

The CA3140 op amp was chosen because of its very high input impedance and because both inputs and the output can be swung down to the negative supply line, eliminating the need for a separate negative supply. The positive supply may be varied from +6 to +12 volts with little effect on calibration. Current drain is 20-30 mA for capacitance measurements and a few mA for leakage measurements.

The original circuit was built on matrix board with the layout approximately following this schematic. Because fast rise times are involved, several 0.01 μF bypass capacitors are included on the board, and leads should be kept short and neat. The op amp is zener protected — keep a shorting ring around the pins while soldering! If a shorting-type wafer switch is used for S1, insert a 470-ohm current-limiting resistor in series with U2 pin 3.

The test points can be alligator clips on short flexible leads fed through grommets in the front panel. Take care to minimize stray capacitance to ground from the plus test point.

Finally, never apply reversed power-supply voltage to the circuit unless you want to buy three new ICs. A reverse-biased l-amp diode across the 9-volt supply will provide protection.

ham radio

solid-state vhf-uhv
transmitlreceive switch

New PIN diode device provides good isolation and low vswr at frequencies up to 1000 MHz

Microwave Associates recently introduced a solid-state T/R switch for vhf-uhf applications. The device, designated the MA8334, makes use of PIN diodes in a hybrid rf circuit which is small and easy to use. This spdt switch is rated at 50 watts CW and has a nominal 50-ohm impedance. Frequency of operation is from 20 to 1000 MHz. Specifications list typical insertion loss at 0.2 dB with 1.2:1 vswr from 20 to 500 MHz.

After evaluating the MA8334 on the test bench, I decided to replace the conventional relays in a solid-state 2-meter transverter I had recently built. The circuit of fig. **1** was used. To operate a switch path, approximately 50 mA forward bias is applied; removal of the bias releases the switch path. Capacitors C1 and C2 provide isolation of the dc bias and the rf source feeding the switch. These capacitors should have low rf loss and be able to handle the power used. I used button micas in my design. The inductance of the rf choke is not critical, and anything around 3 μH should work satisfactorily. Capacitors C3 and C4 are feedthrough bypass types. The 220-ohm resistors provide the correct bias current when using a 12-volt power supply; their values will have to be adjusted accordingly if another supply voitage is used.

Operation of the MA8334 has been completely satisfactory. With the circuit of fig. **1,** the measured insertion loss was 0.25 dB, and the swr is 1.23:1 when operated at 50 ohms. Isolation between ports has not been measured, but from observation I would judge it to be at least the 37 dB specified by the manufacturer.

To conclude, this device provides interesting possibilities for vhf-uhf switching at power levels up to

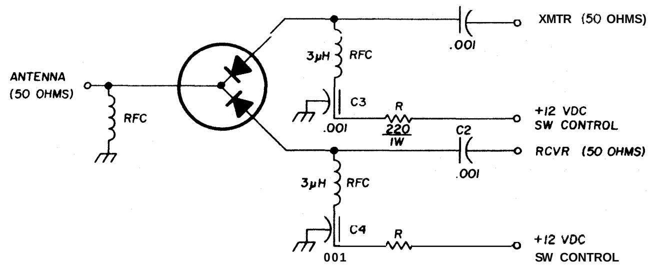


fig. 1. Circuit diagram of a solid-state T/R switch for 144 MHz. The MA8334 can be used at other frequencies up to 1000 MHz by proper choice of circuit constants. The lead marked with an asterisk is identified by a diagonally cut end; the other two leads may be used interchangeably for either transmitter or receiver.

50 watts CW. With proper component selection, the switch should perform well on 432 MHz. When compared with conventional switches, the MA8334's compact size, rugged construction, and reliability give it a definite advantage. The present single-quantity cost is $19.00, but with increased production the price is expected to drop significantly. It may be purchased through any dealer handling Microwave Associates products.

ham radio

By Tom Reddeck, W4NHH, Route 9, Box 515, Greensboro, North Carolina 27409

digital scanner

for 2-meter synthesizers

Complete construction details plus other unusual ideas for integrating a digital scanner with a 2-meter synthesizer

After operating a synthesizer, you begin to realize the number of 2-meter repeater channels that exist and also how long it takes to turn the switches through all channels. The digital scanner presented in this article was designed to permit easy, hands-off, monitoring of the 2-meter fm band. It can also serve as a good indicator of 2-meter band conditions by listening for repeaters outside your local area.

The features incorporated in the scanner were based on several months of on-the-air operation of a prototype in an area heavily populated with repeaters. It scans all 2-meter repeater input and output frequencies between 146.01 MHz and 147.99 MHz, in 30 kHz steps; all 67 frequencies are scanned in about 6 to 8 seconds. The frequency is also read out directly by five, 7-segment LED displays. Operating features include three modes of scanning, A, B, and C. Mode A scans until a signal is received, at which time the scanner stops, listens for about 3 seconds, and then continues. This mode allows rapid scanning of all channels to determine activity. Mode B scans until a signal is received and waits until the signal is gone before continuing the scan. Mode C is the same as Mode B except that a 3-second delay occurs before continuing the scan. Mode C allows monitoring of repeaters that require the repeater carrier to drop between transmissions. To prevent the scanner from locking-up on a very active repeater, a timer is incorporated to ensure that the scanner does not stay on any frequency more than 3 minutes.

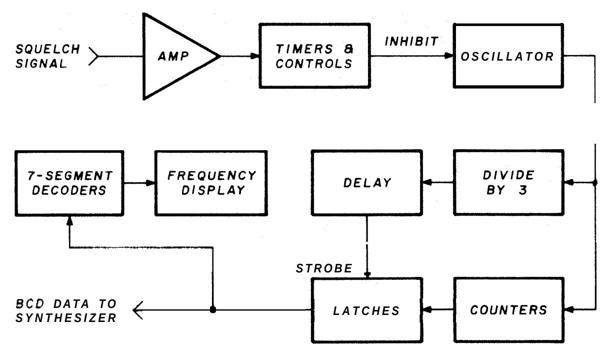


fig. 1. Functional block diagram of the digital scanner.

A latch function is provided to permit locking a synthesizer receiver to the scanner's receive frequency by command. If the synthesizer has automatic transmit offset capability, this feature can be used to good advantage. In my case, I've wired the function to a momentary type toggle switch but it's readily adaptable to a push-to-talk (PTT) mike switch. Wire

By Marion D. Kitchens, K4GOK, 7100 Mercury Avenue, Haymarket, Virginia 22069

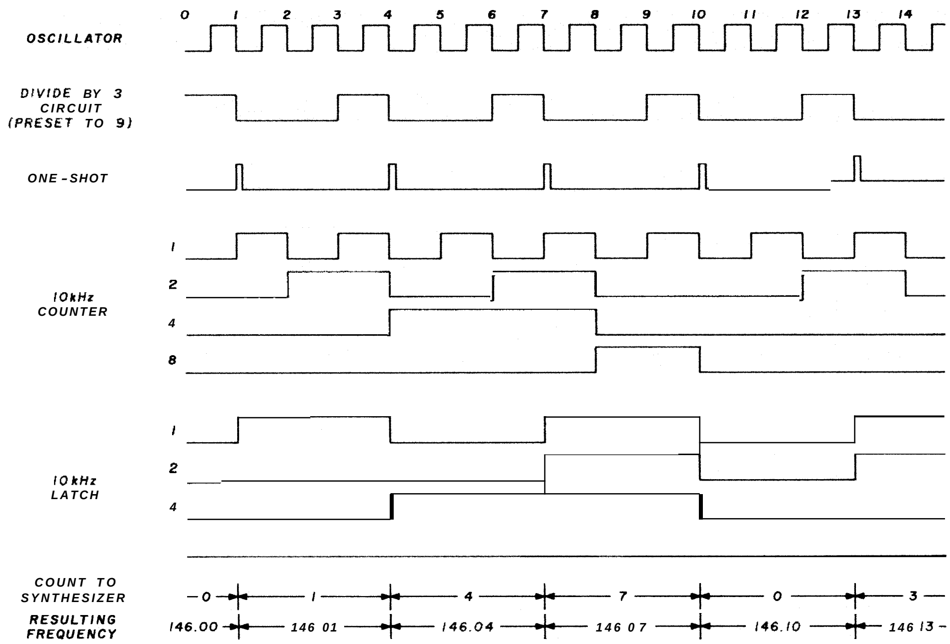


fig. 2. Timing diagram of the scanner. The divide-by-three circuit sets the channel spacing at 30 kHz.

the latch function to the PTT line and when a QSO is to be joined, just push the mike button and the synthesizer is on frequency. The scanner is latched or locked on this receive/transmit frequency, even when the PTT is released. Scanning will continue when manually commanded; the continue command will cause the scanner to run continuously, even though a signal may be present, until the command is removed. If the scanner has stopped on an undesired frequency, a quick tap of the continue switch will move the scanner one or two channels up frequency to where it will resume scanning.

circuit description

The circuit diagram of the scanner is shown in **fig. 1**. A 555 timer is used as an oscillator to drive three 7490 decade counters. The BCD outputs of the counters are applied to three 7475 latches. The oscillator output is also applied to another 7490, connected as a divide-by-three circuit, which provides a strobe to the latches on every third count. A time delay is provided by a 74121 one-shot multivibrator to insure that the inputs to the latches have settled.

The basic timing diagram is shown in **fig.** 2 that indicates the signal relationship between the oscillator, the first decade counter, the divide-by-three counter, and the one-shot delay circuit. The output from the 7475s is a BCD output occurring in steps of three; 1, 4, 7, 10, 13, etc. The latch outputs are also applied to 7446 BCD to seven-segment decoders that drive the displays, and to the synthesizer inputs in place of the frequency control switches. This section of the scanner is standard for TTL counting and display circuits.

Two things are sufficiently different, however, that

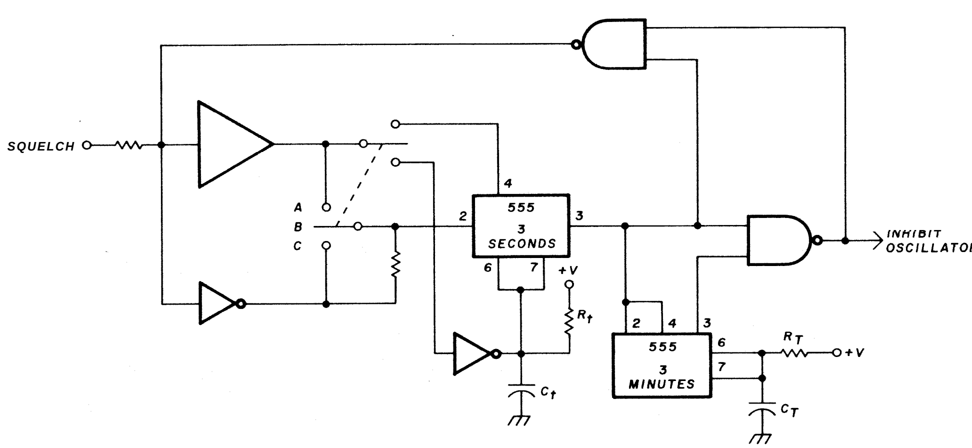


fig. 3. Functional diagram of the 3-minute and 3-second timers. They control the time that the oscillator runs and when it is inhibited.

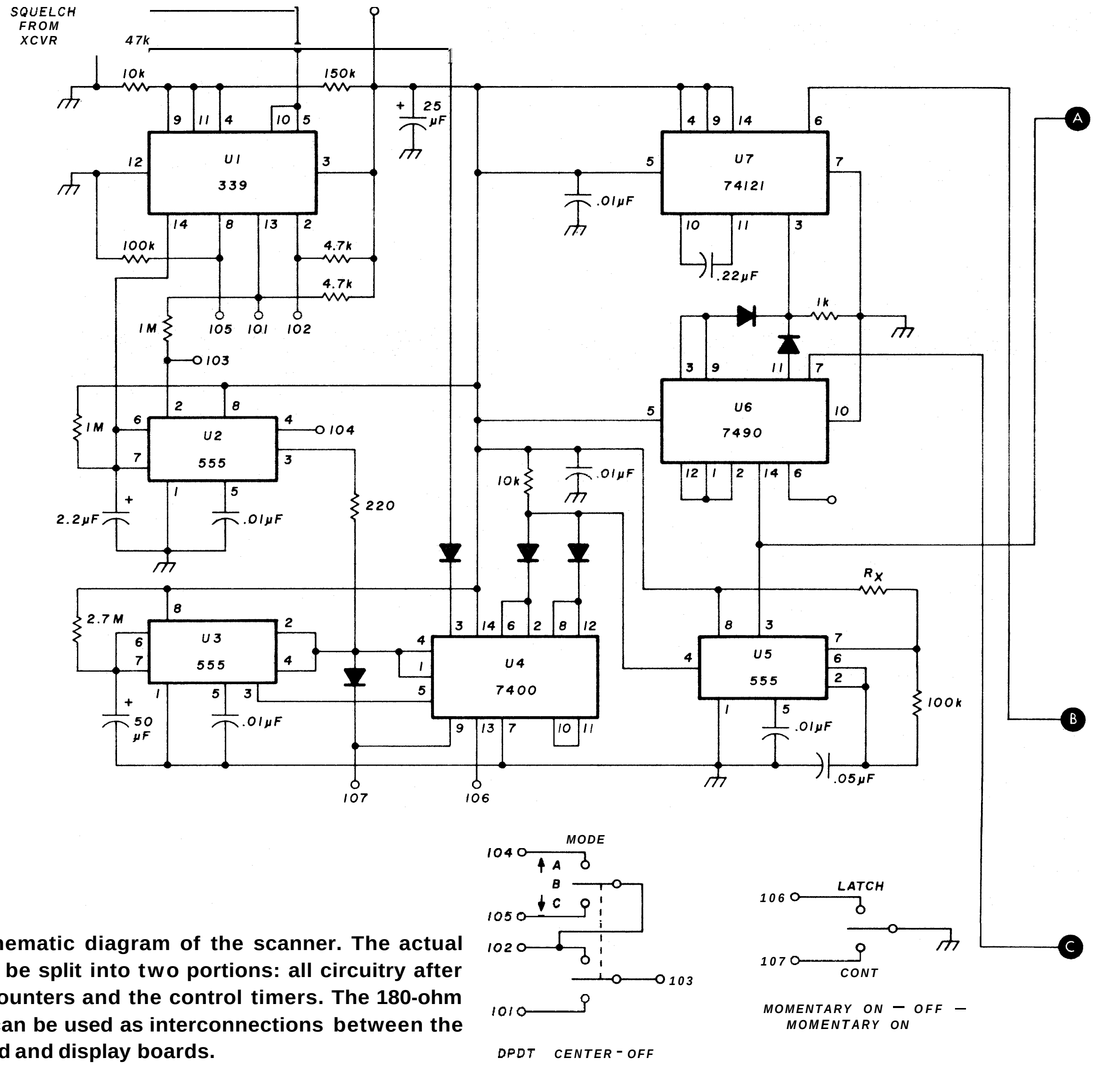


fig. 4. Schematic diagram of the scanner. The actual board can be split into two portions: all circuitry after the 7490 counters and the control timers. The 180-ohm resistors can be used as interconnections between the main board and display boards.

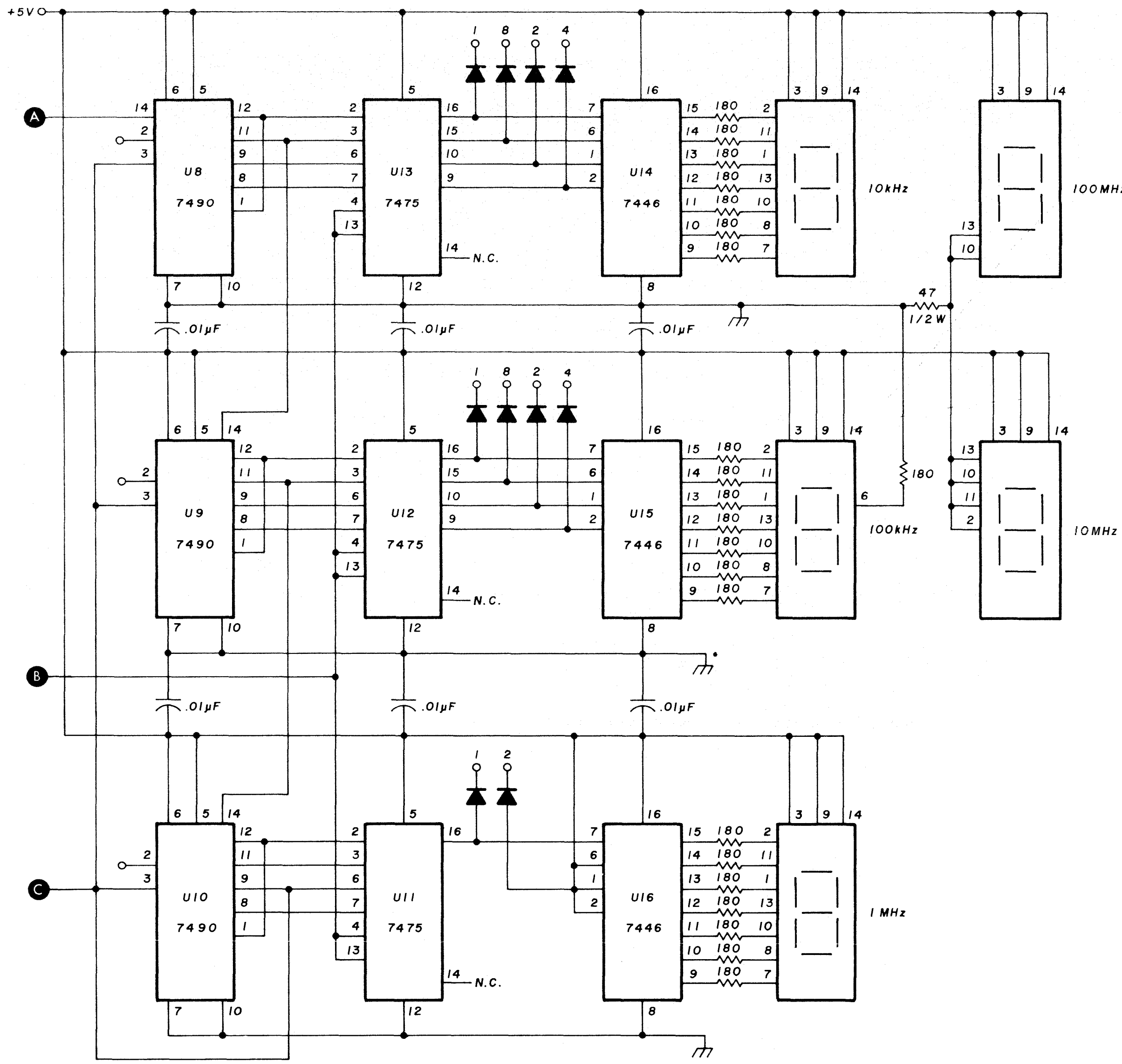
they should be explained. First, the GLB type synthesizers do not utilize a complete BCD signal from the frequency switch; a portion of the BCD signal is hard wired within the synthesizer while the remaining signal lines are controlled by the frequency switches. **Table 1** shows the BCD inputs required at the synthesizer divider chain to produce the frequencies indicated. Examination of the table shows that some of the data does not change for frequencies between 146 and 147.99 MHz. For example, to scan from 146.01 MHz to 147.99 MHz requires only the appropriate change between BCD 6 and BCD 7. The 7490 decade counter supplying the MHz data then has to supply the 0 and 1 count data of column d for the MHz BCD data as shown in **table 1.** The 2 count of this decade counter then represents 148.00 MHz and is used to reset all three 7490 decade counters and the 7490 divide-by-three circuit.

table 1. Required BCD information for the synthesizer input.

frequency column	a	b	c	d	e	f	g	h	i	j	k	l
	bcd data: MHz				100 kHz				10 kHz			
146.01	0	1	1	0	0	0	0	0	0	0	0	1
146.04	0	1	1	0	0	0	0	0	0	1	0	0
146.07	0	1	1	0	0	0	0	0	0	1	1	1
147.00	0	1	1	1	0	0	0	0	0	0	0	0
147.03	0	1	1	1	0	0	0	0	0	0	1	1
147.06	0	1	1	1	0	0	0	0	0	1	1	0
147.99	0	1	1	1	1	0	0	1	1	0	0	1

The second major difference is the manner in which the divide-by-three counter is reset and its resulting operation. Outputs B and D of U6 are OR-wired via the two diodes to the input of the 74121. The A and B outputs of U6 are used to reset this counter. The resulting circuit then functions according to **table 2.** The 148-MHz reset signal from the third decade counter (U8) presets the divide-by-three circuit to count nine. Upon receipt of the first pulse from the oscillator, the circuit goes to count 10, or zero since no carry circuit is used. At count zero the output goes low and the 74121 sends a signal to latch the 7475s. The correct BCD data for 146.01 MHz is now stored in the latches.

The fourth pulse from the oscillator causes outputs



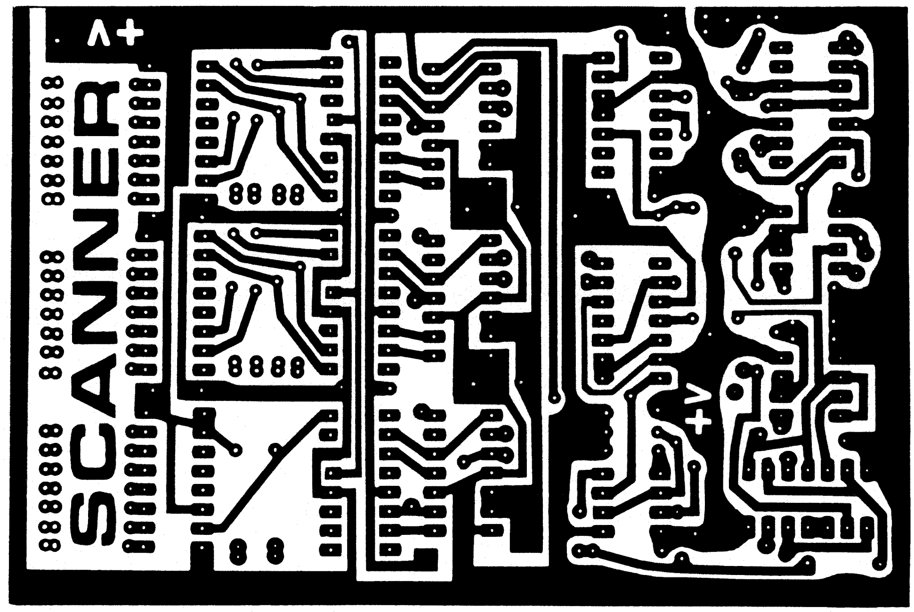
A and B of U6 to be high and the 7490 resets itself to zero. At this point, the output from the diode OR circuit goes low and strobes the latches. The 7475s now have BCD data for 146.04 MHz stored in them.

Three more pulses from the oscillator will again trigger the 74121 and the 7475s, providing BCD data for 146.07 MHz. The sequence continues for frequencies in 0.03 MHz steps until 148.00 MHz is reached, whereupon the divide-by-three circuit is reset to nine and the sequence starts over.

Two timing circuits are used to provide the three operating modes described earlier. One 555 timer provides a 3-second delay and another timer provides about 3-minutes delay. A receiver squelch circuit is sensed by the 339 voltage comparator whose output controls the 3-second timer. Both true and inverted outputs are obtained from the 339. **Fig.** 3 shows a partial schematic of the timer circuitry. With the dpdt switch in position A, pins 2 and 4 of U2 are connected to the voltage comparator. An output pulse will be generated if the pins receive a positive going signal from the comparator. The duration of the pulse from U2 is determined by the values of R_t and C_t. The output pulse will terminate after this time even though the input is still present. If the input signal is less than the R_tC_t determined pulse dura-

table 2. **Outputs from the 7490 showing the divide-by-three operation.**

inputs		7490 outputs d	c	b	a	circuit output
Reset to 9		1	0	0	1	1
Oscillator input count	1	0	0	0	0	0
	2	0	0	0	1	0
	3	0	0	1	0	1
	4	0	0	0	0	0
	5	0	0	0	1	0
	6	0	0	1	0	1
	7	0	0	0	0	0



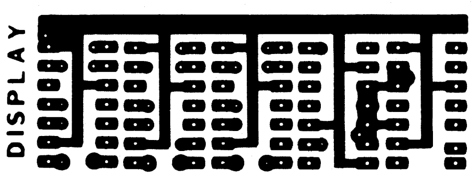


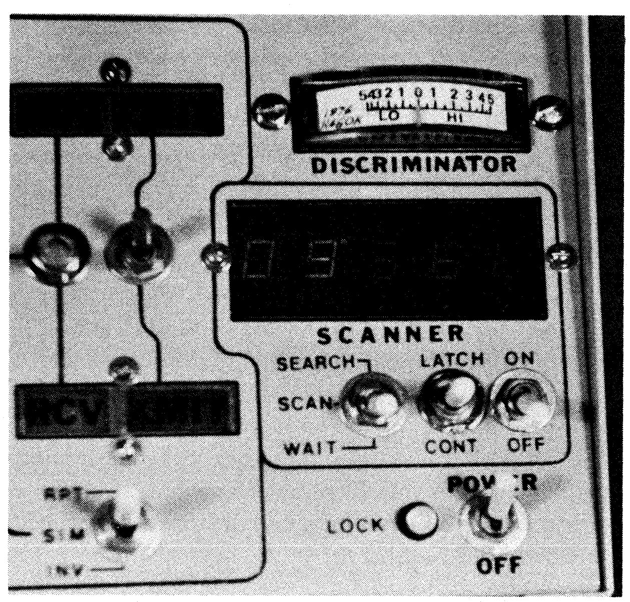
fig. *5.* Full-size layout of the printed-circuit boards.

tion, the output will terminate with the input. The output from the timer is used to inhibit the oscillator; this causes the scanner to stop on the frequency that opened the transceiver squelch. In this mode the scanner stops only for a duration determined by R_t and C_t, in this case 3 seconds (or less if the signal is present for less than 3 seconds).

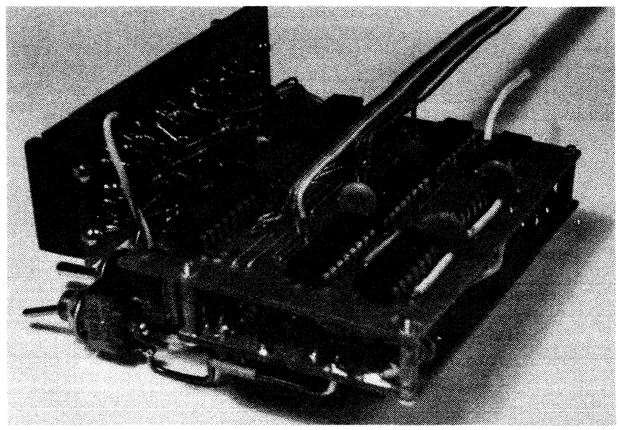
With the dpdt switch in position B, the inverted squelch signal is applied through the 1-meg resistor to only pin 2 of the 555. In this mode, the 555 output follows the squelch signal, without delays, as long as the squelch signal is longer than 3 seconds. The 555 output exists for at least 3 seconds even though the squelch input may be shorter. This is the mode used when it is not desired to wait for a return call after a repeater drops.

One section of the 339 voltage comparer is used as an amplifier and connected across the timing capacitor of U2. When the dpdt switch is in position C, the amplifier has an input and becomes active. A signal from the squelch drives the amplifier output to ground potential, thereby maintaining C_t in a discharged state until the squelch is present, and for a period afterwards that is determined by C_t and R_t. The receiver listens as long as the squelch is open and then for an additional period, 3 seconds in this case, waiting for a return call. This mode is used with repeaters that require a repeater carrier drop between transmissions. The timing circuits used in the scanner were based on reference1 and the data in the Signetics data books.2

The 3-minute timer is used to prevent the scanner from locking up indefinitely on very active repeaters. The 3-minute 555 timer (U3) senses the output from the 3-second timer, as shown in **fig. 4.** It operates just as the 3-second timer does when in mode A. A



Close-up view of the scanner portion of the synthesizer showing the operating controls.



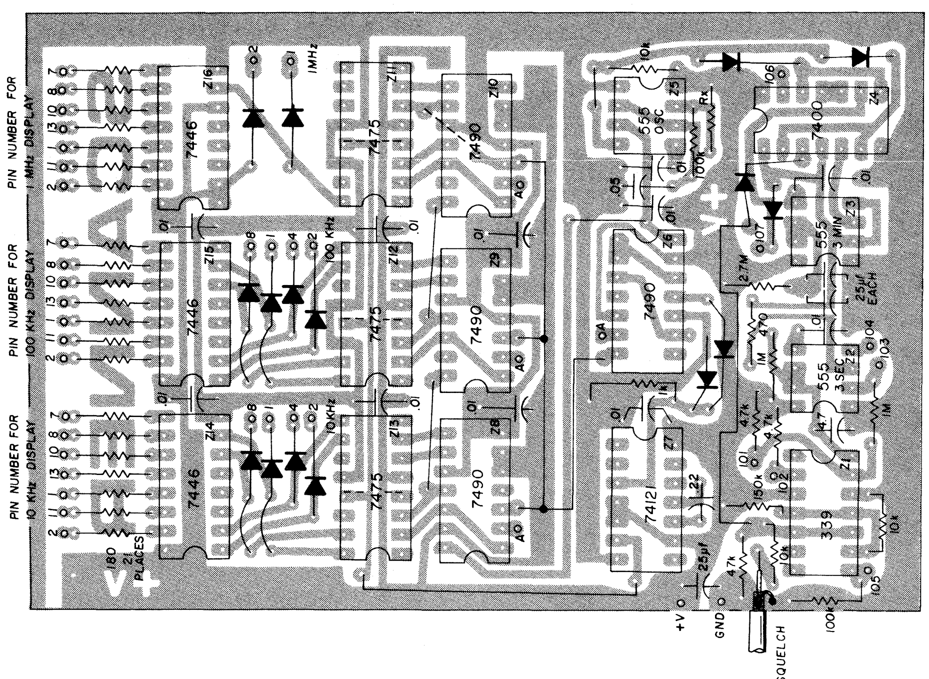
The main circuit board has been split into two separate portions with the display board connected by the 180-ohm current-limiting resistors.

positive-going signal at its input (pins 2 and 4 tied together) causes initiation of its 3-minute output pulse. The 3-minute timer's output is terminated when the input is removed or after 3 minutes, whichever is less. The output of the 3-minute timer is applied to one input of a 2-input NAND (U4B) gate to control passing or inhabiting of the 3-second timer output, which is connected to the second input of the NAND gate. The output from the NAND gate (pin 6) controls the scanner oscillator. The squelch input to the amplifiers is inhibited by a second NAND *gate* (U4A) when the 3-minute timer has expired and the 3-second timer's output still exists, thereby resetting both timers.

construction

Construction of the scanner is rather straightforward with use of the printed circuit board. A full-size circuit-board layout and parts placement diagram are shown in **figs. 5** and ***6,*** respectively. The circuit board can be built as a single unit, or can be cut and assembled into a compact unit as shown in the photographs. Since IC sockets usually cost more than the ICs themselves, soldering directly to the circuit board is recommended.

The circuit board should be built and tested in sections. The recommended sequence is to install the squelch amplifier, timers, and latch/continue chain, and then follow with the 555 oscillator, divide-by-three counter, and one-shot chain. A 1-meg pot should be temporarily installed in place of R_x. This pot will later be adjusted to suit the builder's transceiver and then replaced with a fixed-value



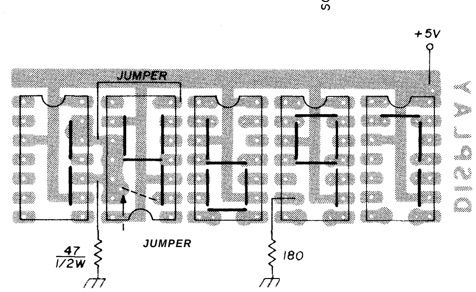


fig. 6. Parts placement diagram for the printed-circuit boards.

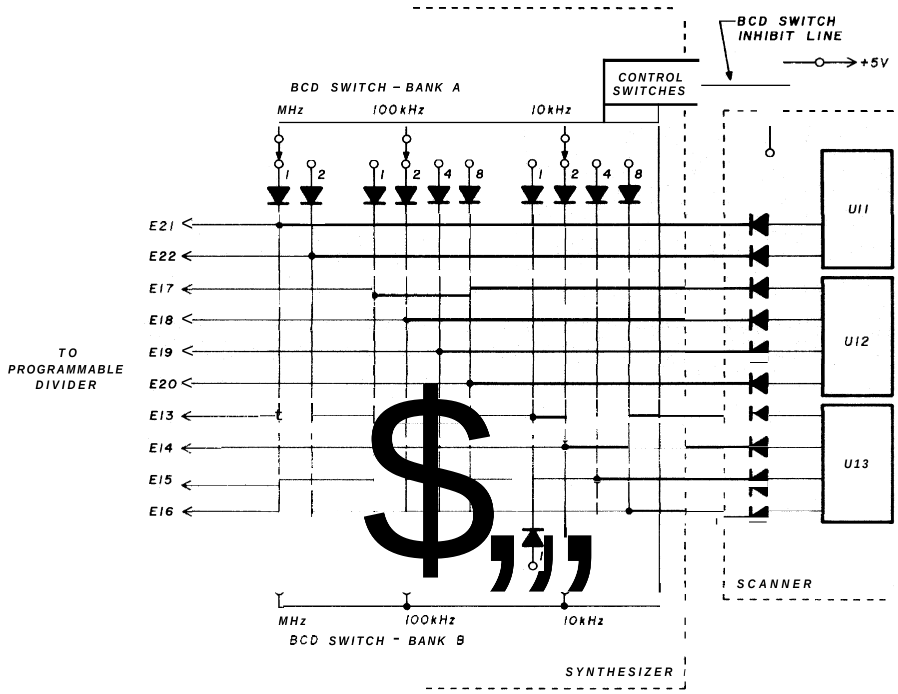


fig. 7. Scanner wiring to synthesizer. The control switches must be disabled when the scanner is in use.

resistor. At this point in construction, a 30-Hz signal should be observed at the oscillator (U5, pin 3) and a 10-Hz signal at the output from the one shot (U7, pin 6). The oscillator should also respond to squelch input signals. The 7490 counters and the 7475 latches can now be installed. Pin 14 of the 7475s must be removed from the package. The pin was removed to facilitate circuit board layout; jumpers on the foil side of the board connect pins 4 and 13 of the latches and between pins 3 and 9 of U10. Check that the 7490s are counting down and data is being transmitted through the latches. Install the decoders and verify the correct information exists at the outputs.

The seven-segment LED displays are mounted on a small, separate printed-circuit board. Note that the 10-MHz display (the number 4) is mounted inverted. The display circuit board will accommodate most of the SLA-1, MAN-7, or 707 displays with a common anode. Be sure to install the jumper wires on the foil side of this circuit board. The displays should first be mounted to the circuit board, then make the connections between the display circuit board and the main circuit board. The main circuit board can be cut near the 7476s and the 180-ohm resistors used to mount the display board to the main circuit board. Make these connections to suit your particular installation.

The control switches should now be wired and the completed scanner checked out before connection to the synthesizer. Use a regulated 5-volt power supply capable of providing about 750 mA. An independent LM309K or similar regulator, supplying power only to the scanner, is recommended.

The scanner outputs from the 7475s are wired to the synthesizer as shown in **fig. 7.** The synthesizer BCD transmit, BCD receive, and scanner outputs are OR wired through the diodes to the synthesizer programmable divider. When the scanner is in use, neither the synthesizer BCD transmit or receive diodes can have voltage applied; their anodes must be either open or grounded. **Fig. 8** shows the circuit used in my homebrew 2-meter TTL synthesizer.

Another method is to replace the synthesizer

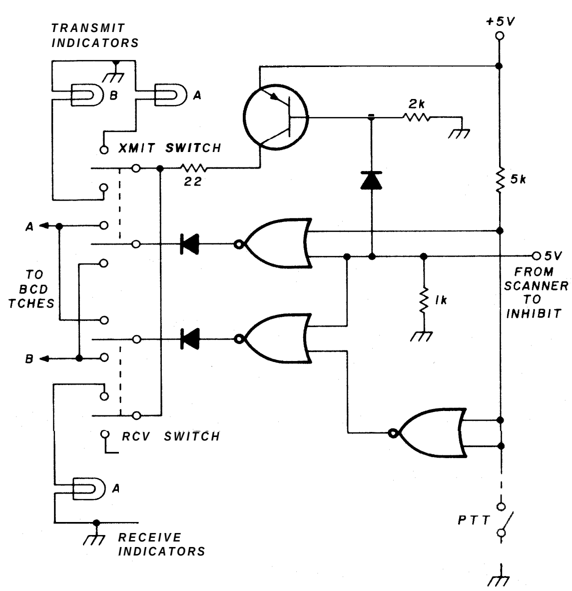


fig. 8. Control switching for the BCD lines in the synthesizer. The normal switches on the synthesizer are automatically inhibited when using the scanner.

receive select switch (used for selecting between the two sets of BCD switches) with a double-throw, center-off switch. This switch *must* be in the off position when the scanner is in use. An alternative to both of the above methods is to set the BCD switches for all zero outputs, usually 144.00 MHz. The squelch input to the 339 comparator (U1) should be connected, with shielded cable, to a point in the transceiver where the voltage goes high when the squelch opens.

Power can be applied to all portions of the scanner and correct operation of the control switches, display, etc. should be verified. The 1-meg pot temporarily installed in place of R_x should initially be set at its maximum resistance. It should then be adjusted for the maximum scan rate, as dictated by the lock-up time of the external synthesizer. After satisfactory operation is obtained, the pot should be replaced by the next largest, fixed value resistor. A 470k resistor can be used for R_x with most transceivers and synthesizers if maximum scan rate is not of particular importance.

circuit variations and additions

A number of practical and interesting circuit and functional variations are possible. This section will present several variations and additions that have occurred to me. Some have been tried, while others are only ideas that you may want to develop to suit your own particular needs.

One variation is to have the digital display indicate both the scanner frequency and the synthesizer BCD switch frequency. This can be accomplished by taking the BCD inputs to the 7446 seven-segment decoders from the synthesizer at the programmable divider inputs. The display will indicate the scan frequency when the scanner is in operation; when off, the display will indicate the synthesizer receive frequency when receiving and the transmit frequency when transmitting. Pull-down resistors may have to be added at the 7446 inputs; the 7446s and displays must, of course, have power when the scanner is off.

Scanning in 10-kHz steps can be obtained by inhibiting or bypassing the divide-by-three circuit. This feature will provide nearly continuous coverage of the 2-meter band.

Fig. 9 shows simple circuit changes that will eliminate scanning of the 147-MHz repeater input frequencies. This change will reduce the scan time by about 2 seconds. No comparably simple way was found to eliminate scanning of the 146-MHz repeater input frequencies.

Other variations include elimination of the scan mode control switch and the latch/continue switch. The scanner will be in mode B (does not wait for a return call) if the mode switch is simply omitted. Connecting a jumper between points 102 and 105 results in mode C operation (wait for a return call). The latchlcontinue function is useful and its elimination is not recommended unless minimization of panel space is desired.

Those builders needing the absolute minimum panel space could eliminate the control switches and use 0.1-inch (2.5mm) high displays. Only the 100 kHz and 10 kHz digits need displays. A discrete LED could be used to indicate 147 MHz. The synthesizer ON/OFF switch could be a center-off dpdt switch supplying power only to the synthesizer when in one on position, and power to the synthesizer and scanner when in the other on position.

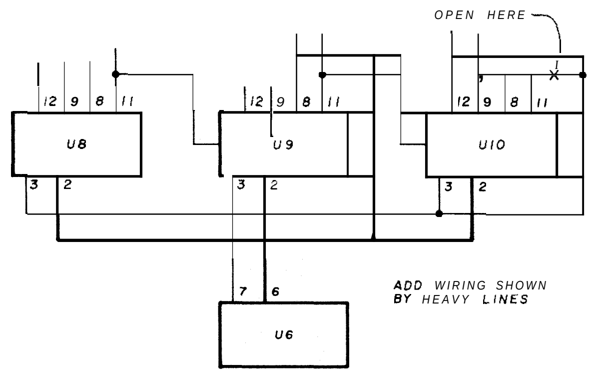


fig. 9. Circuit changes to prevent the counters from covering the 147-MHz repeater receive frequencies.

Another addition that could lead to some interesting possibilities is an automatic transmit frequency offset feature. Digital subtraction of 600 kHz could be provided for 146 MHz repeater frequencies and 600 kHz addition provided for 147 MHz repeater frequencies. A center-off, double-throw switch, marked REPEAT-SIMPLEX-INVERT could be used. The latchlcontinue control could then be used to place the synthesizer on frequency and latch it there. The bulky BCD transmit and receive switches could be eliminated, and a very compact, highly functional synthesizer/scanner could be built.

These variations and additions are presented to encourage building and experimenting among amateurs. I hope that others will build upon and add to these efforts, and eventually present the results for the benefit of all amateurs.

references

1. Mike Connor, WABBMP, Bob Henson, WBØJHS, "Super COR," 73, June, 1976, page 16.
2. *Signetics Digital, Linear, MOS Data and Applications Book,* 1974, Signetics Corporation, Sunnyvale, California

ham radio

modifying the Collins 51J receiver for ssb reception

If you're lucky enough to have one of the 51J-series receivers, here's an easy way to update it for single-sideband reception

One of the most popular surplus receivers is the Collins 51J series, available in limited quantities through MARS and some surplus stores. Designed in the mid 1950s, the stability, readout accuracy, and general excellence of this receiver literally revolutionized receiver design, setting the trend for most of the modern ssb receivers and transceivers. The immediate fallout from the 51J design was the well-known Collins 75A series of amateur-band-only receivers, followed by the present S-line.

The many virtues of the 51J series receivers do not include good ssb reception. An important modification is the inclusion of a product detector and alteration of the automatic gain-control loop to accommodate ssb signals. This article covers these modifications as well as other minor changes that make the 51J into a first-class receiver suitable for amateur service, including general-coverage operation.

Five models of the 51J receiver are available. The 51J-1 is quite rare; probably the quantity made was small. The 51J-2 and 51J-3 are fairly common on the surplus market; differences between the receivers are minor. The military R-388/URR is similar to the 51J-3. The 51J-4 was the latest production model and incorporates mechanical filters in the i-f system. A choice of three filters may be made with a panel switch.

At one time Collins made an adapter (Collins part number 354A-1) for the 51J-2 and 51J-3 that would modify the receivers for inclusion of crystal filters. The adapter is no longer in production.

The first job for the owner of a 51J is to align it correctly and test all the tubes. Complete alignment information is included in the Collins receiver manual and also in the military technical manual, *Radio Receiver R-388/URR*, TM-11-854, sometimes obtainable through MARS or surplus dealers.

receiver sensitivity

A common fault in most 51J receivers I've inspected is that overall gain is low and the receiver seems dead above about 15 MHz. Investigation has shown that receiver gain is reduced because of an uncommonly high bias voltage applied to the rf tubes. Bias is obtained from a voltage divider in the negative side of the high-voltage power supply (**fig. 1**). Normal bias voltage is −1.4 volts and, in the receivers tested, has usually run from −1.6 to −3.0 volts. This high

By William I. Orr, W6SAI, EIMAC, **301** Industrial Way, San Carlos, California 94070

negative voltage lowers the gain of the rf stages, leaving the receiver lifeless. Bias voltage is developed across resistor R149, which is 820 ohms, 1/2 watt. In many receivers, this resistor looks to be overheated or measures abnormally high in resistance. The cure is to remove R149 (which is located on a terminal strip on the inside wall of the receiver, near the line cord) and replace it with a 2-watt resistor of the proper resistance, which will develop a voltage drop of 1.4 volts across it. You'll find the value will run between 700 and 1000 ohms, depending upon your receiver.

receiver PTO

On occasion a 51J may be picked up for a song because the PTO (permeability tuned oscillator) "doesn't work." The usual cause of malfunction is a collection of matchstick capacitors in the PTO (C005, C006 and C008), which tend to short circuit after a few years of service. These are 0.01-µF, 400-volt capacitors of a design no longer made. Replacing these capacitors with 0.01-µF, 600-volt disc ceramic capacitors will usually restore the PTO to operation.'

the new product detector

Once the 51J has been aligned and is operational, the ssb modification may be added. The circuitry to be modified is shown in **figs.** 2, 3, and **4.** The major alteration is in the beat-frequency oscillator **(fig. 2)**, which is changed to perform as a product detector. To make this change, the receiver panel may have to be removed, as a new beat oscillator switch (S112) may be required. A 6BE6* is substituted for the 6BA6 BFO tube, and various circuit changes are made beneath the chassis. The final circuitry, after modification, is shown in **fig. 4.**

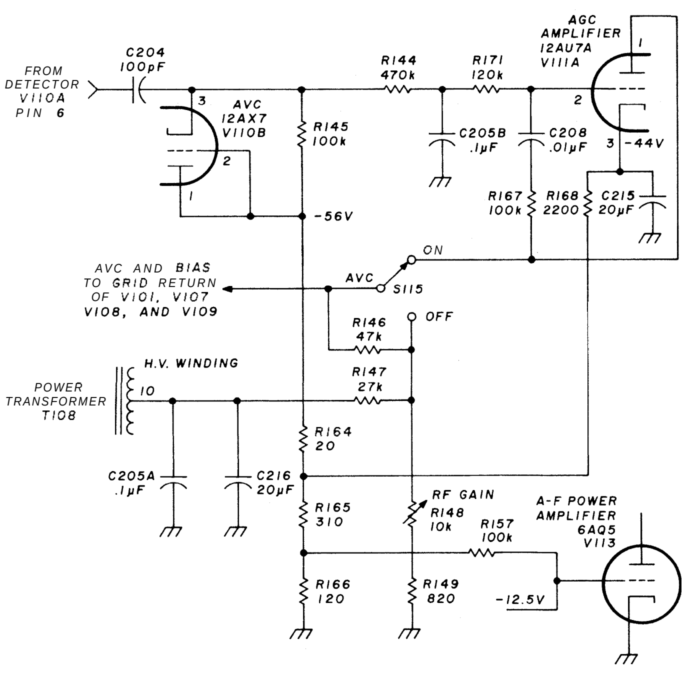


fig. 1. Agc and bias control portion of the 51J receiver. R149 establishes control-bias level. For a negative control voltage V110B and V111A operate below ground. Agc time constant is determined by R144 and C250B. External cathode-to-grid circuit (V111A) should be below 2 megohms after modification to prevent stray "gas current" in the 12AU7 from blocking the agc action. Audio amplifier bias is obtained from the negative source across R166.

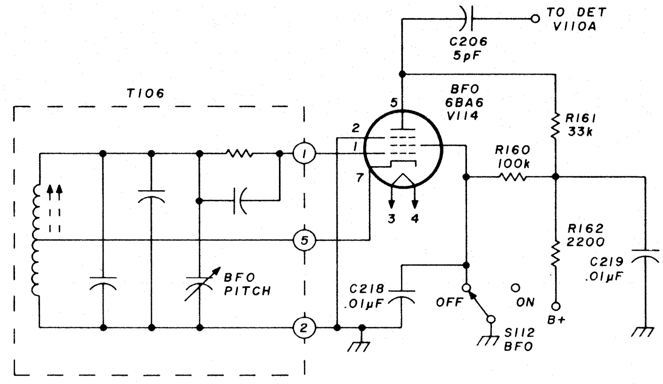


fig. 2. Original 51J BFO circuit. A 6BE6 tube is substituted for the original 6BA6 (V114) to provide a product detector. Tube is turned off by switch S112, which short circuits the screen voltage to ground. (See reference 2 for more details on the tube substitution.)

The first step is to start work on the BFO tube socket (XV114). Most Collins 51Js are wired with high-quality wire having a thin plastic coating, which can be easily damaged by a soldering iron. I suggest, therefore, that you use a small iron with a long, narrow point and proceed carefully so that you don't inadvertently burn any insulation on adjacent wires. Referring to **fig.** 2, remove the following components: R161 (33k), R160 (100k), R162 (2.2k), C218 (0.01µF) and C219 (0.01µF).

Next, capacitor C206 (5 pF) must be carefully disconnected from XV114 pin 5 and reconnected to pin 7. A 10k, 112-watt resistor is then connected between pin 7 and the adjacent ground lug. XV114 pin 2 is ungrounded and reconnected to the BFO transformer pin 5 (center pin) through the 220-ohm resistor and 0.01 µF combination.

The next step is to solder the 0.05-µF disc ceramic capacitors in place. One capacitor connects between pin 6 and the nearby socket ground post. The other, in the plate circuit, is attached to an existing terminal stud, which is screwed to the bolt holding the main filter capacitor socket. The 10k, 1-watt resistor is connected between the high-voltage terminal (pin 5 of C217B socket) and the terminal stud. The 47k, 112-watt resistor is placed between the stud and pin 5 of socket XV114.

'For additional information on the use of the 6BE6 product detector, see reference 2.

The final modifications at this point are to place the 0.1 μF filament bypass capacitor on the socket and revise the audio and agc circuitry.

audio-stage mods

The remainder of the modified circuitry is shown in fig. **4.** The plate circuit filter components (two 470-pF capacitors and a 47k, 1/2-watt resistor) are mounted on a two-terminal strip placed under one bolt of coaxial socket J104 (marked if **output).** The 0.05 μF coupling capacitor is connected between this assembly and XV114 pin 5.

The 51J panel must now be removed to get at selector switch S112 (BFO OFF-ON) (fig. **5**). If not, the switch will have to be replaced with the proper type (dpdt). The **A** section shorts the 6BE6 screen supply for am service. The B section switches the audio section of the receiver from the product detector to the diode detector, through limiter tube V112A. The audio takeoff point is XV112A pin 3.

To make the interconnections. three coax cables must be run from switch S112 to the rear of the receiver. For ease of wiring, the small-diameter RG-179/U is suggested. The outer braids of the three cables are grounded to the switch assembly on the panel. The cables are dressed into position and run to the respective termination points, at which place the shields are again grounded.

agc mods

To complete this step, capacitor C205A-B-C should be temporarily unbolted from the chassis and moved out of the way.

The agc loop in the receiver is designed to adjust the rf and i-f gain automatically for a-m signals. It must be modified for ssb reception. Pappenfus *et al*[3] recommends an attack time of about 0.002 second and a release time of 0.2 to 2 seconds. This time constant can be closely approximated within the limitations imposed by the 51J circuitry. The agc circuit is shown in fig. **1.** The agc time constant, as the receiver stands, is about 0.06 second, determined by capacitor C205B and resistor R144.

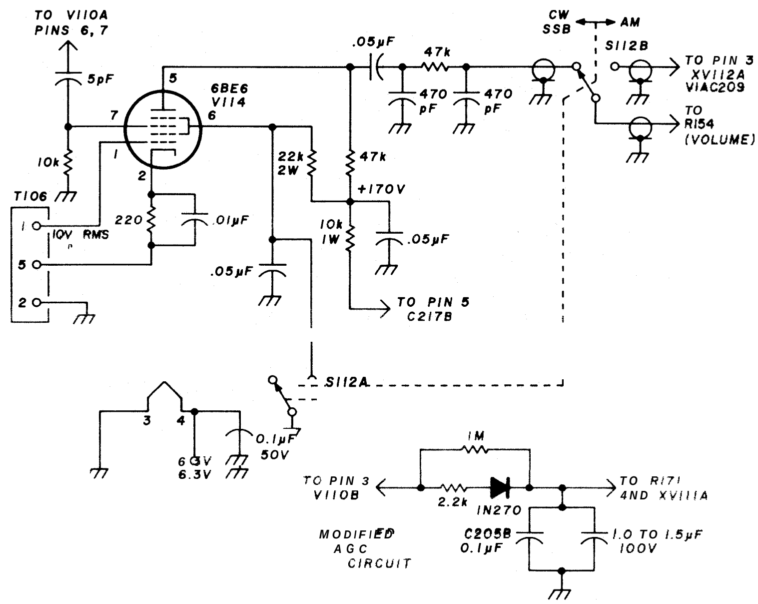


fig. 4. Revised product detector and agc circuit. Caps are ceramic except for the time-constant cap, which is Mylar (see text). BFO injection, measured at pin 1 XV114 socket, should not be more than 10V rms. Oscillator voltage can be set by varying the 22k, 2W screen resistor. Signal injection level is set by the value of the resistance between XV114 pin 7 to ground.

It's theoretically possible to increase the time constant by increasing R144; however, there's an upper limit to the value of this resistance, as pointed out by my friend and colleague, W6PO, who reminded me that oxide cathode tubes such as the 12AU7 are restricted as to the maximum value of grid resistance, which should run less than two megohms.

The reason for this restriction is that a combination

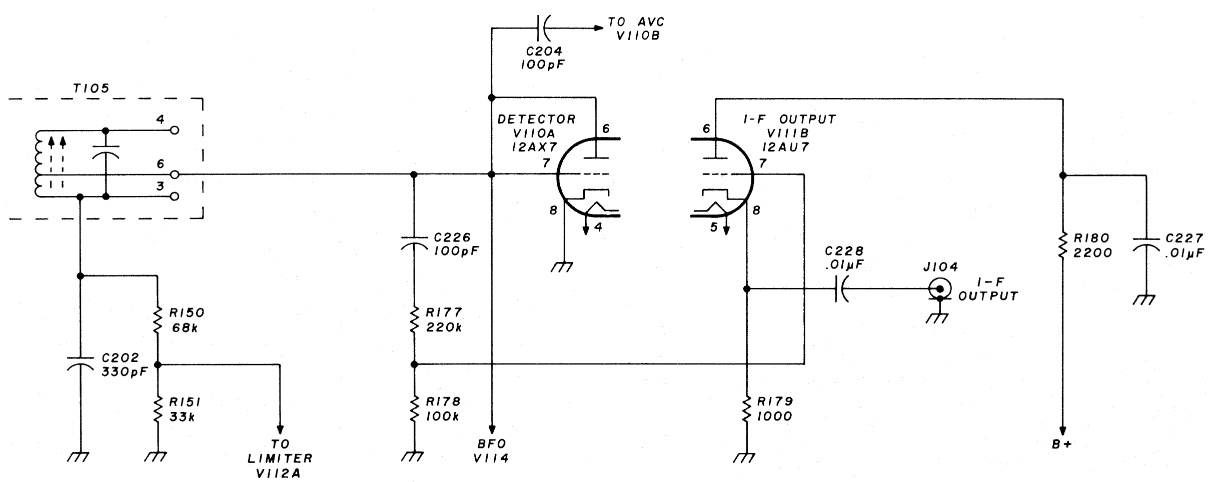


fig. 3. **Detector and i-f output amplifier schematic. V110A is connected as a diode detector. Audio is recovered across R151.**

of the Edison effect and the migration of oxide from the cathode to the grid as the tube ages can lead to grid emission. An electron flow caused by grid emission (even if only a microampere or so) can seriously disrupt the bias level when the grid resistor is an unreasonably high value. One microampere, for example, flowing through a 2-megohm resistor produces a 2-volt drop, enough to alter the operating characteristics of the 12AU7 agc amplifier tube. The flow of grid current can block the agc line, rendering the receiver inoperative. W6PO recommended that not more than 2 megohms, and preferably less, be used in the agc time constant circuit,

To achieve the desired results capacitor C205B (0.1 μF), which is part of the timing circuit, must be increased to at least 1 μF. The use of a low leakage, *Mylar* capacitor at this point is recommended. The capacitor can be placed directly from the center terminal of C205B to an adjacent ground lug. The resistive portion of the timing circuit is made up of a germanium diode and two resistors. The attack time is set by the 2.2k, 112-watt resistor and the release time by the 1 megohm, 1/2-watt resistor. The 1N270 diode disconnects the attack resistor during the discharge portion of the agc cycle. This tiny network is made up and then placed between pin 3 of socket XV110B and the adjacent terminal of capacitor C205B **(fig. 4)**.

testing

After the wiring is checked, the receiver should be tested on a-m to make sure that all original circuits are working. When the BFO switch is turned on, the BFO may be adjusted for good ssb reception. Once satisfied the receiver is working properly, you can check out ssb operation.

The first step is to check for BFO harmonics. With the antenna off, tune the receiver to 1 MHz, 1.5 MHz, and 2 MHz. The BFO harmonics should be heard weakly at the lower frequency and should be in the receiver noise level above 3 MHz. If the harmonics are loud enough to be troublesome, the BFO level should be reduced by increasing the value of the 22k, 2-watt screen resistor on the 6BE6. Once the BFO harmonics have been reduced to your satisfaction (about 2 or less divisions on the S meter at 2 MHz), you can check the product detector for signal overload.

With the constants shown, the signal from the product detector will be somewhat less than that from the a-m detector. The receiver has ample audio gain, so this presents no difficulty. You should be able to tune in a needle-banging ssb signal and receive it crisp and clean. If audio distortion shows up as a growl on speech, this indicates that the product detector is being driven too hard by the i-f signal. The remedy is to reduce the value of the 10k, 112-watt resistor in the rf input leg of the 6BE6 XV114 pin 7. In some cases, this resistor value will be as low as 1.2k for low intermodulation distortion.

The 51J receivers vary a bit from one production run to another, and changes in harness layout affect the oscillator level injection, oscillator harmonics, and intermodulation distortion. However, the values given in the schematic are representative and are a good place to start from.

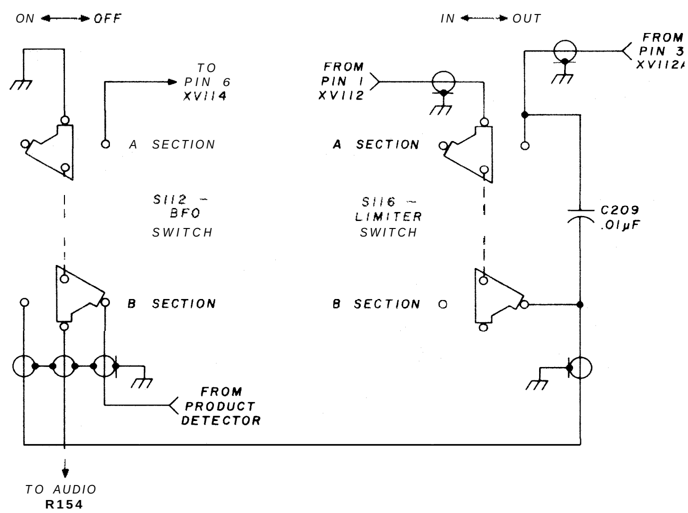


fig. 5. Many 51Js can be wired in this fashion for proper switching. If S112 and S116 are single-pole switches, they must be replaced with double-pole, 2-position, shorting switches. Note that section B of S116 is used only as a tie point for C209.

parting thoughts

One baffling 51J receiver, after modification, overloaded on even the weakest ssb signal. A painstaking check revealed that some previous owner, anxious of wringing every decibel of gain out of the receiver, had changed the detector tap on transformer T105 from pin 6 to pin 4 **(fig. 3)**. This upset the gain level of the receiver so that overload was inevitable. Changing the modification back to the original circuitry cured the trouble.

The modified 51J, especially if equipped with mechanical filters and a reduction tuning knob, is the equal of the best of today's ssb receivers. How many items of equipment, designed in the mid-1950s can equal that?

references

1. William I. Orr, W6SAI, "Collins 51J PTO Restoration," *ham radio,* December, 1969, page 36.
2. Lee, "The Single-Tube Pocket Detector," *CQ,* April, 1961, and Scherer, "More on Updated Improvements for the 51J Receivers," CQ, December, 1968.
3. Pappenfus, Bruene, and Schoenike, *Single Sideband Principles and Circuits,* McGraw-Hill Book Company, New York, 1964.

ham radio

discrete operational amplifier active filters

The functions of the hybrid active filter can be replaced by using individual operational amplifiers — the quad op amp now permits a package-by-package replacement

Two recent excellent articles have described the construction of active filter circuits for CW and ssb receiving applications, using the Kinetic Technology type FX-60 hybrid, integrated circuit.[1,2] As mentioned in K6SDX's article,[2] the KTI FX-60 "Universal Active Filter" is a basic building block, incorporating three micro-power op-amps with internal chip resistors and capacitors forming multi-loop negative feedback transfer functions. By the addition of external resistors and/or capacitors, the nominal center frequency may be changed, and the outputs modified to simulate a variety of classic filter characteristics.

However, the FX-60 is not always easy to come by since it is a "cull" or production reject of the commercial series FS-60. The FS-60 rejects have one or more tolerances out of limits, but are perfectly acceptable for experimental and amateur applications; they carry the designation FX-60. The FX-60 is only available directly from the manufacturer, the supply is limited due to a small rejection rate, and the commercial grade FS-60, at a five times higher price, is proportionately less attractive for amateur projects.

table 1. Comparison of multiple op amp integrated circuits.

manufacturer	type	number of op amps	supply voltage
RCA	CA3401E	four	single (+)
Motorola	MC3301P	four	single (+)
National	LM3900N	four	single (+)
National	LM-324	four	single (+)
RCA	CA3060E	three	dual (+ & –)
National	LM148	four	dual (+ & –)

Note: The CA3401E, MC330IP, and LM3900N are pin-for-pin compatible.

Fortunately, there are now a number of inexpensive, multiple op amp ICs which can be used to adequately simulate the basic functions of the FX-60. The multiple op amps can be substituted in most of the circuits for which the FX-60 is specified. A partial list of suitable ICs for this purpose is shown in table **1.**

Single supply voltage types require only a positive supply in the range of 5 to 25 volts, and have a built-in center-signal reference. Dual supply types more commonly require both a positive and negative voltage with respect to ground. Some may be found at bargain prices at surplus supply houses.

While all of the ICs listed in table **1** are suitable for active filter applications, I chose the LM324 for further consideration. Though not classified as "micropower," it has relatively low power drain (approximately 700 μA/amp), low internal noise (allowing use in low-level signal circuits), incorporates four independent op amps, and requires only a single positive supply voltage.

basic universal active filter

Fig. 1 illustrates the basic circuit of the FX-60 with its internal negative feedback loops, and connections for the DIP configuration (viewed from the bottom). The internal resistor and capacitors (R1A/R2A and C1A/C2A) set the nominal bandpass output center frequency of 230 Hz and also the cutoff frequency (f_c) of the lowpass and highpass outputs. This frequency (230 Hz) can be increased by connecting external shunt resistors, R1B/R2B, across pins 1 and 2, and pins 10 and 12. The external resistors are always of equal value for a specific frequency above 230 Hz, and can be calculated from the formula

$$R = \frac{455 \times 105}{f_c}$$

where f_c is the desired frequency above 230 Hz. If

By Peter A. Lovelock, K6JM, 1330 California Avenue, Santa Monica, California 90403

R1B/R2B are ganged variable units, the filter outputs can be made tunable.

For nominal center frequencies below 230 Hz, external capacitors C1B/C2B are connected across pins 2 and 12, and pins 7 and 10. These capacitors should also be of equal value to establish the desired center frequency below 230 Hz. In addition, the external resistors can be used, in conjunction with the external capacitors, to tune the filter outputs above the reduced nominal frequency.

External resistor R4, either fixed or variable, is connected between pin 8 and ground to trim the nominal Q and gain of the FX-60, in conjunction with R3, the external input resistor. Pin 14 is normally the signal input, with pin 6 used for special applications.

The basic circuit of the FX-60, which is called the "Bi-Quad Active Filter," can be adequately duplicated, for amateur applications, with an LM324 as shown in **fig. 2.** Since the LM324 has pin connections to each of the four independent internal op amps, the frequency-determining resistors and capacitors are combined into single components, R1/R2 and C1/C2. The Q is varied by appropriate values of a single resistor R3; increasing the value of R3 increases Q and vice versa. Since width of the bandpass output is related to Q, R3 can be adjusted experimentally for the desired bandpass characteristic.

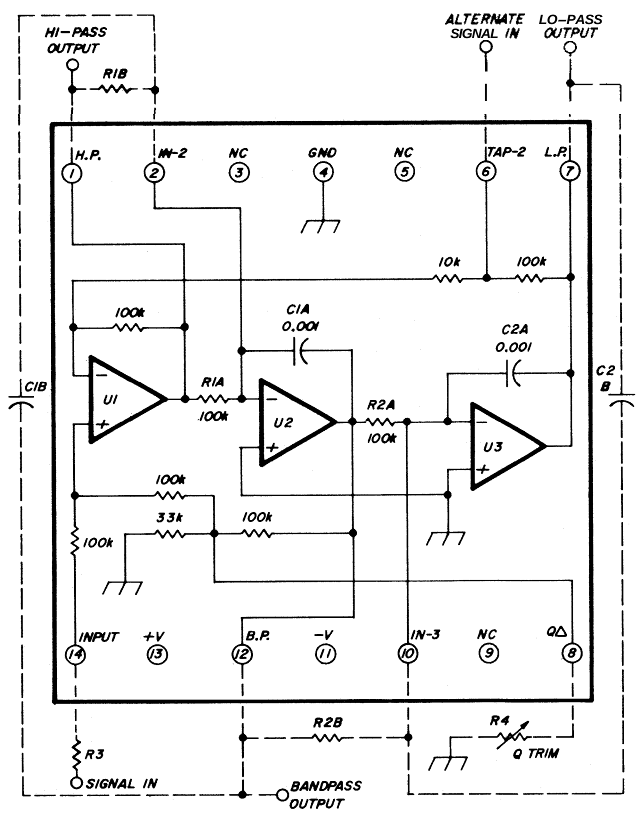


fig. 1. The internal configuration of the KTI FX-60 hybrid active filter. External resistors or capacitors can be added to either raise or lower the center frequency.

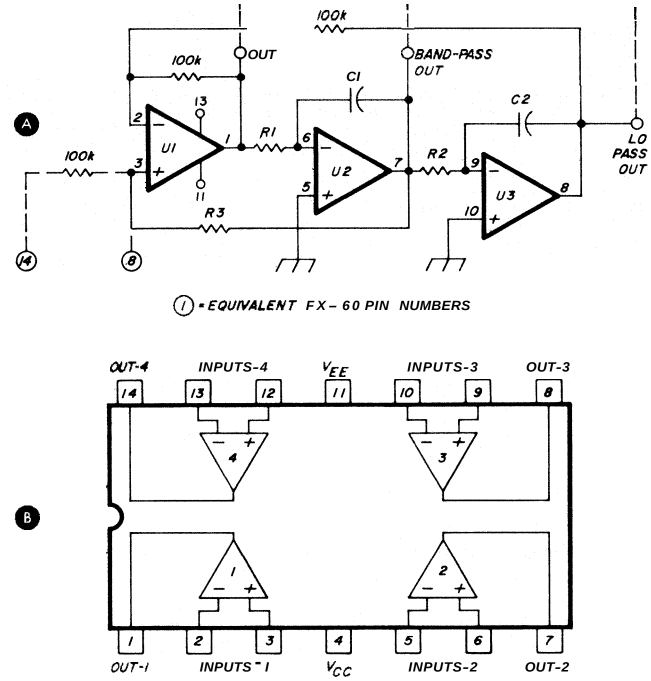


fig. 2. A basic bi-quad active filter using the individual op amps of an LM324(A). The equivalent pin numbers of an FX-60 show the LM324 can be used to replace the hybrid active filter. B shows the pin outs for the LM324.

Comparison of **figs. 1** and **2** show how the latter circuit, using the LM324, can be substituted for the FX-60 in previous articles. If the same resistance1 capacitance (100k and 0.001 μF) are used in the LM324 circuit, the same approximate nominal center frequency of the FX-60 (230 Hz) will result. For direct substitution, the user may want to configure R3 in **fig. 2** into the three resistor combination used in **fig. 1.**

In the Bi-Quad duplicated circuit, only three of the four available op amps are used. The fourth op amp may be used as an output amplifier in place of the 741 device required in some circuits,[2] or for summing the highpass and lowpass outputs.

Fig. 3 shows the complete circuit of an active filter using the LM324, with appropriate biasing for a single supply voltage of +5 to +25 Vdc. The R1IR2 value (150k) establishes f_c at 1000 Hz, and the value of R3 (10 meg) for a Q of 50. Values of R1/R2 for other bandpass center and f_c frequencies can be calculated from the formula

$$R = \frac{15 \times 10^7}{f_c}$$

The resistors should have a 1 per cent tolerance, but 5 or 10 per cent tolerance may be used, with some variation in resultant f_c. Variations of R1/R2, for values of C1/C2 other than 0.001 μF, are beyond the scope of this article; in general, the bandpass and f_c can be determined for values of R/C when $\boldsymbol{R} = X_c$.

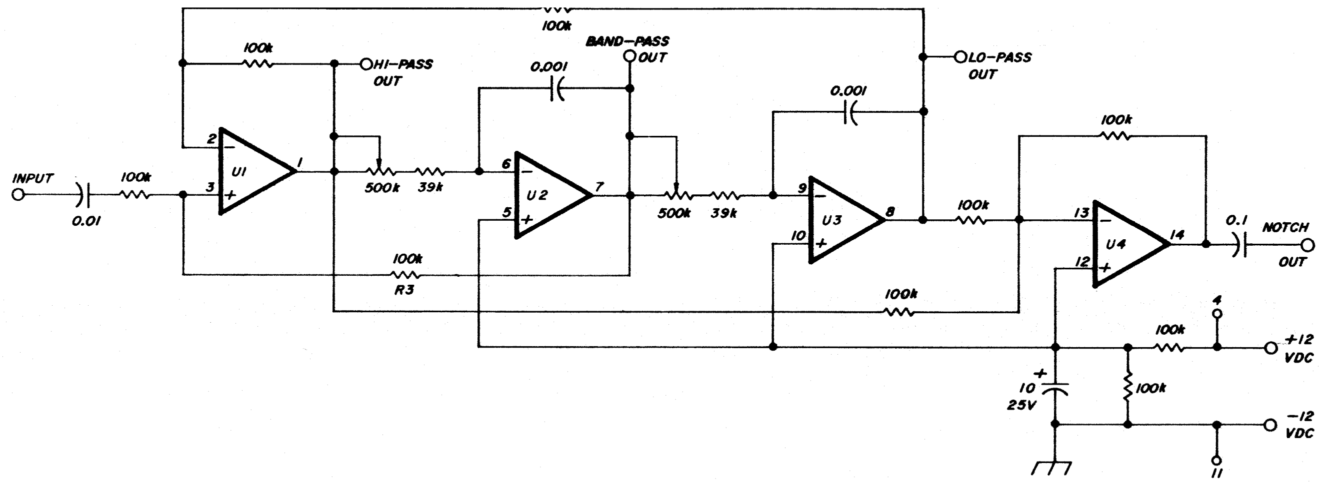


fig. 4. Schematic diagram of a tunable active filter. The highpass and lowpass outputs have been summed in the fourth op amp to provide a notch output. The potentiometers *must* have a reverse log taper.

A fully tunable active filter, covering the range of 300 Hz to 3000 Hz is shown in **fig. 4.** In addition to the previous highpass, bandpass, and lowpass outputs, the fourth op amp (U4) is used to sum the highpass/lowpass outputs which, being 180 degrees out of phase, result in a tunable notch at the output of U4. The tuning potentiometers are ganged, *reverse log taper,* 500k carbon, *2* watt units. Although exact tracking between the potentiometers is not critical, high quality components are recommended to minimize noise and frequency jumps. A notch of −35 dB can be attained using fixed components with 5 per cent tolerance. This circuit is similar to that used for audio notching in the new Atlas 350-XL transceiver, and is most useful for nulling out unwanted CW signals or broadcast hetrodynes in the 3.8 and 7 MHz bands. The low internal noise of the LM324 permits inserting this circuit between the product detector and first audio amplifier stages of a receiver.

Resistor R3 establishes the *Q* for a notch width of 200 Hz at the −3 dB points. While the notch may be narrowed by increasing the value of R3, tuning for maximum notch depth becomes increasingly difficult; 200 Hz is about optimum, for ease of adjustment. The notch output has unity gain with respect to the input signal, and any variations due to component tolerances can be adjusted for by trimming the value of the 100k resistor between pins 13 and 14 of U4.

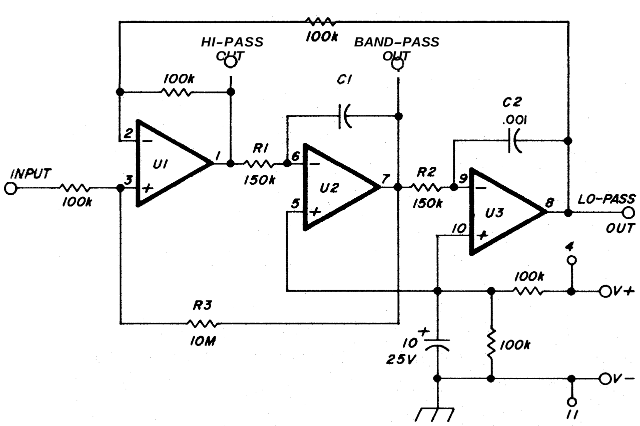


fig. 3. A practical fixed-frequency active filter using the LM324. The center frequency is 1 kHz, with a Q of 50.

This tunable active filter is generally useful for amateur receiver applications since the choice of high, low, bandpass, or notch outputs may be switched. It should be noted, however, that the highpass, lowpass, and bandpass outputs have gain with respect to the input. A resistive attenuator (minimum 1 megohm) coupled through a 0.1 μF capacitor to each of these outputs, is required to adjust the levels for unity gain.

An alternate, fixed-frequency notch filter, using only three op amps, is shown in **fig.** 5. Other than reduction of components, this circuit has no inherent advantage, but lends itself to triple op amp ICs. Not easily adaptable to tuning, this circuit is useful for discrete frequency notching.

general considerations

Reasonable care must be taken when laying out any circuit that uses multiple outputs and feedback loops. The LM324 is particularly well suited to minimizing stray coupling, since the output terminal of each op amp is located at the four corners of the DIP IC. Stray coupling between the input and output of the separate op amps must be avoided to prevent instability or performance degradation. This is particularly important in the notch filter circuits where stray coupling may limit the attainable notch depth.

If you wish to use one of the suggested devices other than the LM324, for a filter, I recommend that

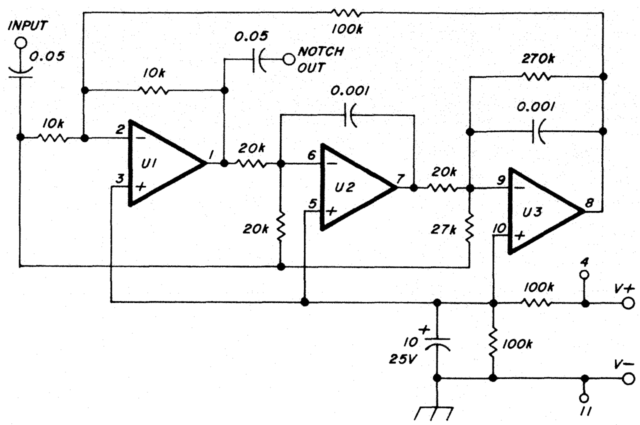


fig. 5. A fixed frequency notch circuit can be formed by using three sections of an **LM324**. This arrangement does not lend itself well to adjustable notch frequencies. The center frequency for this circuit is **3 kHz.**

you consult the manufacturer's specifications regarding supply voltages. For single-supply voltage types, the biasing requirements can be uniquely

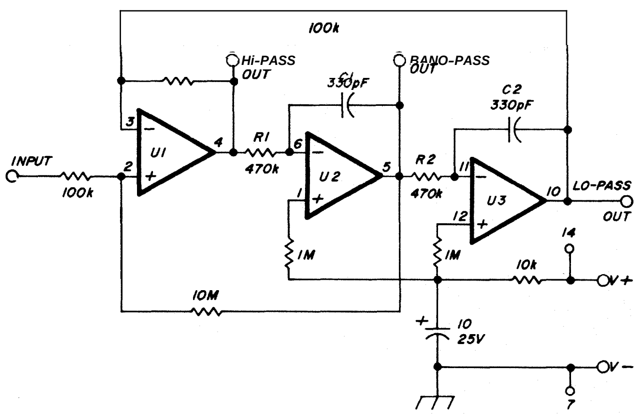


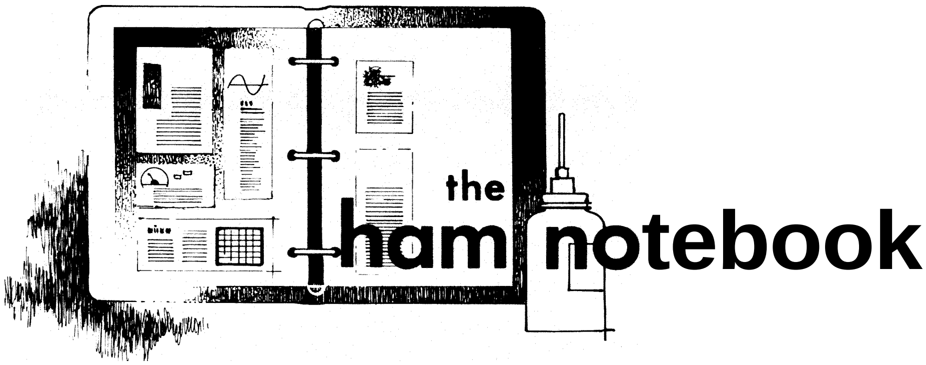
fig. 6. A basic fixed frequency active filter that uses the compatible **CA3401E, MC3301P,** and **LM3900N**. The center frequency is 1 **kHz.** The 1 megohm resistor in the non-inverting lead is used to limit the input current.

different. As an example, **fig. 6** shows a fixed frequency filter using the CA3401E, MC3301P, or LM3900. These pin-compatible devices employ an internal "current mirror" transistor for the single polarity supply. Compared with the previous circuits, you can see that these devices require a different biasing arrangement, including high-value series resistors for the non-inverting inputs to limit bias current to between 10 and 100 μA.

references

1. Ken Hollady, KGHCP, "Tunable Audio Filter for Weak-signal Communications,'' *ham radio,* November, 1975, page 28.
2. M. A. Chapman, KGSDZ, "Audio Filters for Improving SSB and CW Reception," *ham radio,* November, 1976, page 18.

ham radio



high voltage fuses in linear amplifiers

The addition of two short pieces of wire to many linear amplifiers will protect expensive components from damage in case of an arc-over or short in the high voltage circuit. Most high-voltage power supplies are fused in the primary circuit only, and a failure in the amplifier can destroy the rectifier string, grid, and plate current meters long before the primary fuse can open.

The partial circuit shown in **fig. 1** is a typical grounded-grid amplifier with the plate meter in the negative lead of the power supply. Note that the negative side of the power supply is not directly grounded. The ground path goes through both the grid and plate meters. If capacitor C2 shorts out, the short circuit current will go through both meters and, if the filter capacitor, C1, is large, this current can have an instantaneous peak of hundreds of amperes. Such a failure will surely destroy the meters and will very likely destroy the rectifier string. The meter coils will be vaporized and, if the meters are sealed, the glass faceplates may blow out.

The solution is to add high voltage fuses F1 and F2. Each fuse consists of a short piece of no. 40 AWG (0.08mm) copper wire, This wire has a fusing current of 1.75 amperes, high enough that it should never open up under normal circumstances, but low enough that it will blow in a hurry should a short or arc-over occur. Low voltage glass fuses must not be substituted in this application; they will explode when the internal element vaporizes. They'll also take longer to open up fully as the vaporized element will sustain an arc until the glass breaks and allows it to dissipate. This delay, while probably no more than a few milliseconds, may be long enough to damage the meters.

If resistors R1 and R2 are not present, they should also be added. Their purpose is to keep the negative lead of the power supply from going to a high negative potential with respect to ground should either of the meters or F2 open. They have no effect on normal circuit operation since they are in parallel with the meters, whose resistance is a fraction of an ohm.

In my homebrew 4-1000A linear, these fuse wires have blown twice due to arc-overs in the amplifier. On both occasions, they prevented damage to the power supply and meters, responding fast enough that the primary fuses did not blow at all. In seven years of heavy use, they have never failed during normal operation. I call that cheap insurance.

John Becker, K9MM

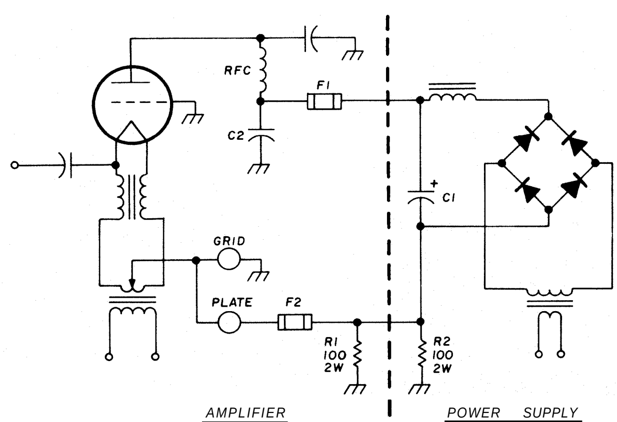


fig. 1. High-voltage fuse circuit for linear amplifier power supplies. Care should be exercised during the installation of the two fuses, F1 and F2. F1 is in the actual high voltage path.

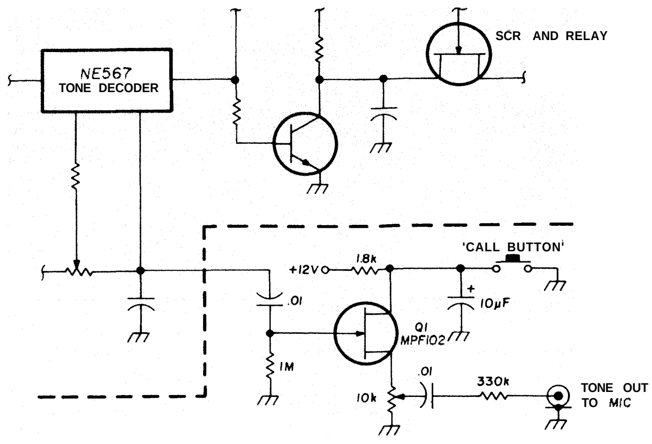


fig. 2. Schematic diagram of the source follower connecting the NE567 decoder and the transmitter. The tone is available, for the transmitter, when the CALL button is pushed.

dual-function integrated circuit

The article "private-call system for vhf fm," *ham radio,* September, 1977, required a separate tone oscillator be used by the initiating station. In reality, the NE567 tone decoder is actually already oscillating at the required frequency. **Fig.** 2 shows a method of using this IC for both originating the tone, and decoding it upon reception.

Cal Sondgeroth, W9ZTK

integrated-circuit oscillator

Many keyer circuits have appeared in amateur radio publications (W7BBX, *ham* ***radio,*** April, 1976; WA5KPG, QST, January, 1976). Most use transistors for the oscillator or clock. When using ICs for the keyer, why not go all the way? A keyed IC oscillator is shown in **fig.** 3. The clock will start when the key is closed and can be held until the dot, dash, or space is completed. The trick is to use a 74L04. If you use a regular TTL IC, you will get microsecond pulses, instead of millisecond. Diodes CR1 and CR2 prevent the first pulse from being different than the next; the 250pF capacitor on the output is necessary to prevent noise spikes from falsely triggering the keyer circuits.

J. T. Miller, WB6VZW

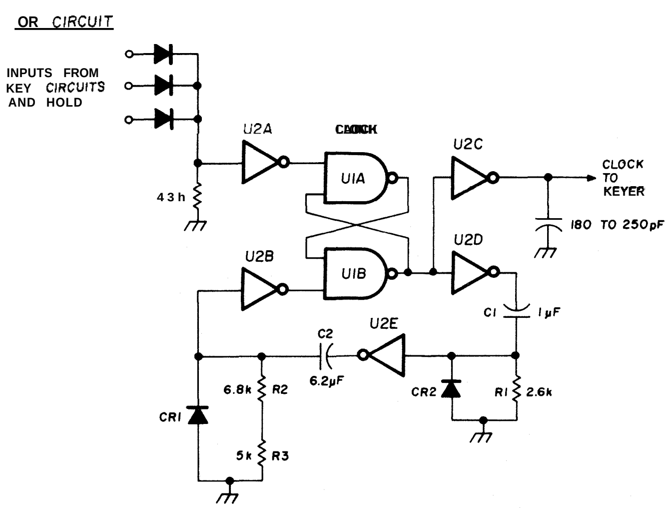


fig. 3. Schematic diagram of the keyer oscillator. U1 is an SN7400, while U2 is an SN74L04. The diodes on the input of U2A form an OR gate that controls the oscillator. These inputs can be used to keep the oscillator running, providing the self-completing feature. The time constant, as determined by C1 and R1, is 4 mS; this is the width of the clock pulse. The values for C2, R2, and R3 give a pulse repetition time of 50 to 95 mS, which equates to approximately 12 to 24 words per minute. For higher speeds, C2 and R2 can be reduced.

wire-wound potentiometer repair

Exact replacement units for those expensive wire-wound pots are often difficult to find. This factor makes repair of the defective control attractive.

The winding is repaired by bridging the opening with a small strip of thin-sheet metal as shown in **fig. 4.** One possible material is the metal from a tin can. Cut the strip slightly shorter than the element width, and wider than the break in the winding. Curve the metal strip to conform with the shape of the resistance element.

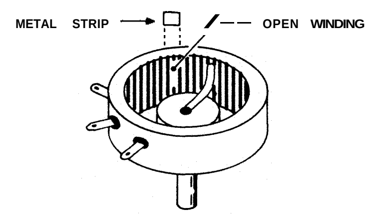


fig. 4. The open winding of a wire-wound potentiometer can be repaired by inserting a small metal strip between the winding and the outer insulation.

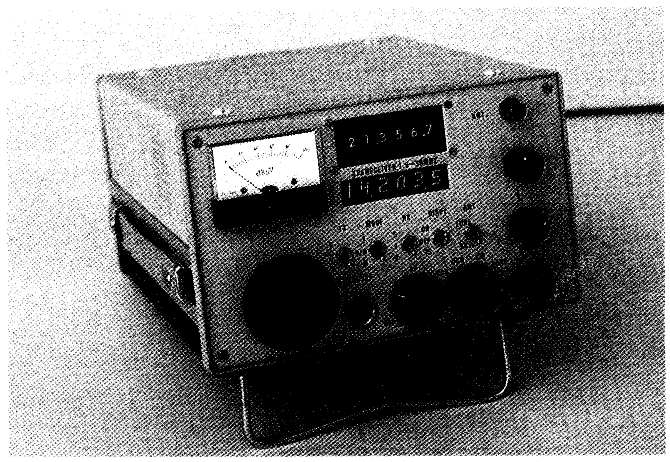
With power off, remove the rear cover to expose the wire resistance element. The opening in the winding is often evident by discoloration from overheating. Otherwise, it may be located by activating the equipment and adjusting the control knob to the setting where abnormal noise or other faulty performance occurs. The defect is now located directly under the movable wiper contact. Near the break, gently pry the resistance element away from the outer insulation using a thin screwdriver or knife point. This will permit starting the bridging strip into the opening. Now, press the strip behind the resistance element so that it does not interfere with free operation of the slider, or cover replacement.

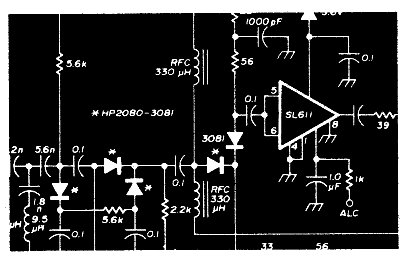
Gene Brizendine, W4ATE

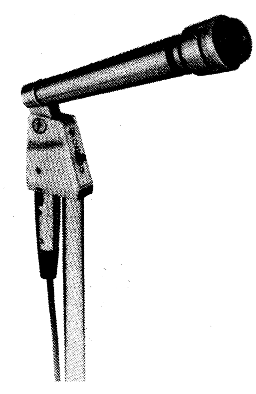
$1.50

radio

MARCH 1978







synthesized high-frequency ssblcw transceiver

10th *anniversary issue*

ham radio magazine

MARCH 1978

volume 11, number 3

T. H. Tenney, Jr., W1NLB
publisher

James R. Fisk, W1HR
editor-in-chief

editorial staff

Charles J. Carroll, K1XX
Patricia A. Hawes, WA1WPM
Alfred Wilson, W6NIF
assistant editors

Thomas F. McMullen, Jr., W1SL
Joseph J. Schroeder, W9JUV
associate editors

Wayne T. Pierce, K3SUK
cover

publishing staff

C. Edward Buffington, WB1AMU
assistant publisher

Fred D. Moller, Jr., WA1USO
advertising manager

James H. Gray, W1XU
assistant advertising manager

Therese R. Bourgault
circulation manager

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

U.S. and Canada: one year, $12.00
two years, $22.00
three years, $30.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

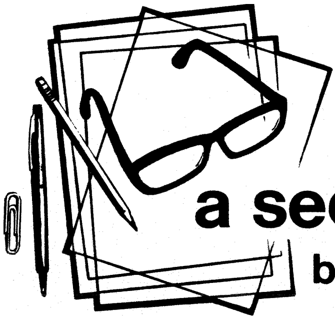
Foreign subscription agents are
listed on page 133

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

How time flies . . . it doesn't seem possible that it has been ten years since I put the final touches on the first issue of *ham radio* and sent it off to the printer. If you recall that very first issue, you'll remember that the cover featured a five-band ssb exciter designed by K1UKX. It was an all vacuum-tube layout, but ten years ago few hams were thinking in terms of solid-state station equipment, and the K1UKX exciter was as up to date as anything then in print.

To commemorate our tenth anniversary we're featuring a fully-synthesized high-frequency transceiver that reflects the current state of the art. A comparison of these two rigs graphically demonstrates the tremendous technological advances that have been made in the past few years. In 1968 there were a few rf small-signal transistors on the market, but practically none for rf power, and low-cost MOSFETs and integrated circuits were still a few years away. Less than half the active amateurs were still using a-m, but it was far from being a thing of the past, and the boom in portable vhf-fm equipment and repeaters was still far in the future. How things have changed!

You can still hear a few a-m stations on 75 and 160 (and a few on the vhf bands), but nearly everybody is now on sideband. In 1968 transceivers were starting to become popular but separate receivers and transmitters were still in widespread use. There were a few pioneers operating converted commercial fm gear on two meters but there were, perhaps, only a dozen repeaters in the entire country! Today there are hundreds of two-meter repeaters and the vhf wastelands of the late 1960s are crowded with signals.

Amateur radio in 1968 was attracting so few new members it was barely holding its own — in recent months it has shown more growth than ever in its history. Some oldtimers complain about the interference that comes with a larger amateur population, but only through larger numbers can we hope to attract the attention of manufacturers who will build equipment that incorporates the latest technological advances. Economics being what they are, you can't have one without the other — manufacturers aren't going to spend money developing state-of-art equipment for a dwindling number of consumers.

As amateur radio has changed during the past decade, so has *ham radio.* Although we weren't always the first to present the latest advancement in amateur radio state of the art, more often than not you read about it first in the pages of *ham radio.* As you read through this special 10th anniversary issue, you'll notice that we have reprinted a few articles from past issues which have long-term amateur interest. Those issues are long out of print, but since we receive so many requests for these particular articles, we thought it would be appropriate to reprint them for the benefit of readers who don't have a complete library of back Issues.

Amateur radio, by its nature, is a very diversified hobby. Each ham follows his own special interests whether it's home construction, vhf-fm, RTTY, moonbounce, slow-scan television, or any of a multitude of others. If you don't see a technical or construction article that covers your particular plane of interest, it's because no one has taken the time to write it. If you have some ideas for station accessories or other amateur equipment that you think others would be interested in, I'd like to hear about it; even if you don't have the time to develop the project yourself, perhaps I can plant the idea with an author who will bring it to fruition.

To paraphrase the closing paragraph of my editorial in the first edition of *ham radio,* we will not stand on our laurels, nor will we stand still. We will always be looking for ways to improve because amateur radio is a dynamic hobby, always on the move. As the equipment, techniques, and challenges of amateur radio change, so will we. We'll constantly try to make *ham radio* more useful to you as well as more interesting and stimulating. We will never become complacent — we will always try to make *ham radio* better. It has always been our goal to keep our readers informed of advances in electronic technology, and we will continue to do so in the future, but we will also make a bigger effort to present more simple projects that you can duplicate in your home workshop. Viewed from here, that's a bigger challenge for the decade ahead than it was for the decade past.

Jim Fisk, W1HR
editor-in-chief

general-coverage high-frequency transceiver with digital readout

Design of a ssb/CW transceiver with exceptional performance which features synthesized frequency control from 1.5 to 30 MHz

The high-performance ssb/CW transceiver described in this article is unusual in that it provides coverage of the entire high-frequency spectrum from 1.5 to 30 MHz. Although the present amateur bands represent only about 12 per cent of this spectrum space, amateur activities such as MARS require additional frequency coverage. In addition, it's uncertain whether the amateur bands in the 1980s will be the same as they are now, or whether they will be expanded or reduced. Regardless of the outcome of the World Administrative Radio Conference of 1979, this transceiver will provide exceptional performance on any of the high-frequency amateur bands – both now and in the future.

In addition to its unusually wide frequency coverage, this transceiver includes features which are not available in commercial amateur equipment such as the built-in antenna tuner and ac power supply, nicad battery pack, and charger. The transceiver is completely portable and is equivalent to the latest military

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

table **1**. Operating specifications for the general-coverage high-frequency transceiver.

Frequency range	1.5 to 30 MHz
Frequency accuracy	100 Hz
Operating modes	CW and ssb
receiver	
Sensitivity	0.3 μV for 10 dB S+N/N
Image suppression	greater than 80 dB
I-f feedthrough	greater than 80 dB
Intermodulation distortion	2nd order: better than 65 dB (independent of level) 3rd order: better than 60 dB (2 times 0 dBm input level)
I-f bandwidths	500 Hz and 2.3 kHz
I-f shape factor	1:2 (CW); 1:1.7 (ssb)
First i-f	41 MHz
Second i-f	9 MHz
AGC operation	Response: 2 milliseconds Attack time: 0.75 second Hold time: 100 ms decay Threshold: 0.2 μV
AGC characteristics	Less than 6 dB audio change for rf inputs 0.2 μV to 1 volt
transmitter	
Output power	20 watts
3rd order IMD	–33 dB with respect to PEP output
Harmonic radiation	–65 dB
Non-harmonic spurious	–70 dB
Other features	RF speech processing and ALC Built-in antenna tuner matches 12 to 200 ohms at all phase angles Split frequency operation, *any* separation Built-in power supply and nicad battery pack
digital synthesizer	
Increments	1 kHz (100 Hz resolution with VCXO)
Switching speed	6 milliseconds
Synthesizer control	thumbwheel switch or optical shaft encoder
Memory	for independent transmit/receive frequencies

manpack designs such as the Hughes Aircraft PRC104, Tadiran PRC174, or AEG-Telefunken S6861. A complete list of specifications is shown in **table 1;** a block diagram of the transceiver is shown in **fig. 1.**

general description

In the receive mode the signal from the antenna is fed into the receiver through the antenna tuner, directional coupler, and filters to the PIN diode attenuator. There are a total of seven lowpass filters and three highpass filters — the correct filter is selected by digital logic which is controlled by the frequency synthesizer. From the PIN diode attenuator the input signal passes through a 32-MHz lowpass filter to a high-level SRA3H double-balanced mixer (+17 dBm local oscillator drive). The 41-MHz output signal from the mixer is amplified 10 dB in a push-pull amplifier

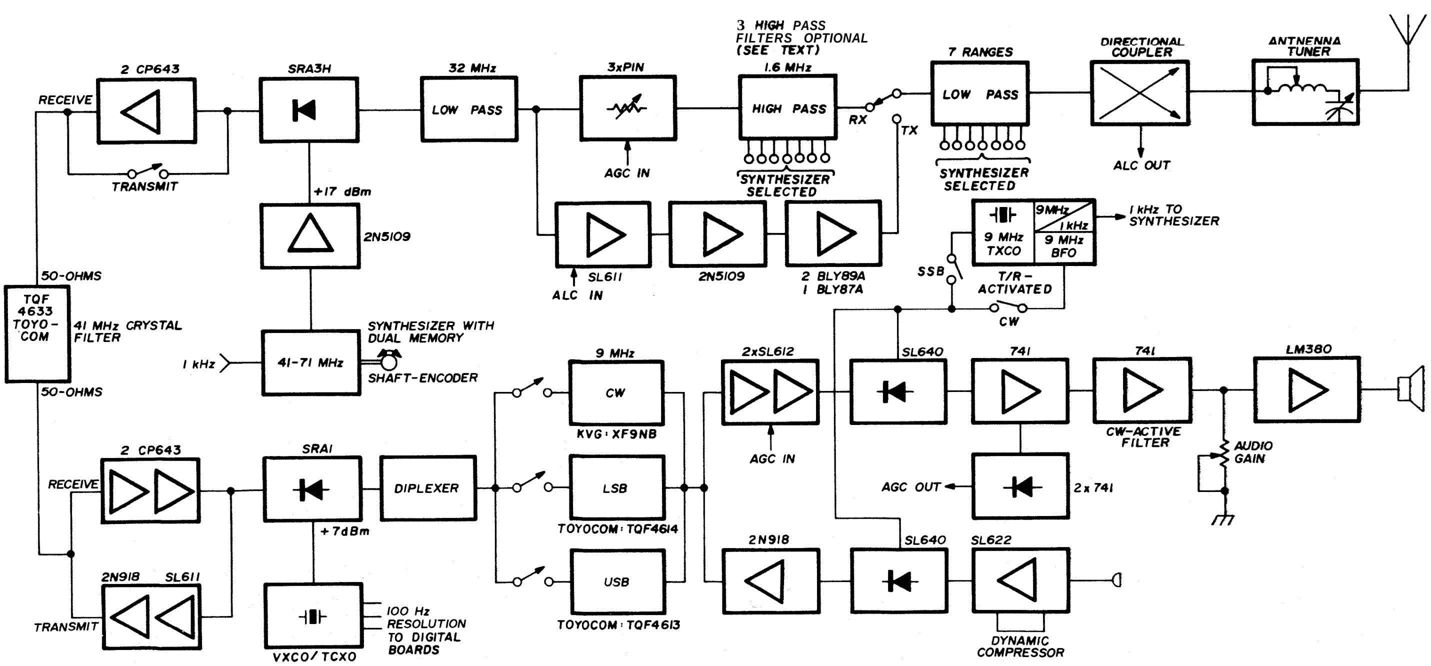


fig. **1**. Block diagram of the high-frequency transceiver. The unit features a frequency synthesizer with memory. and built-in power supply and antenna tuner. Complete specifications are listed in ***table 1***.

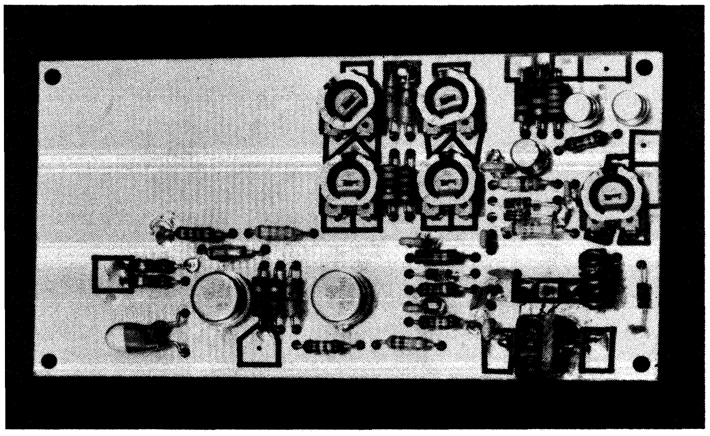


fig. 3. Photograph of the ALC board showing the directional coupler (lower right! and the potentiometer adjustments.

which uses two CP643 power fets. The push-pull amplifier provides an ideal wideband termination for the double-balanced mixer and compensates for losses through the following 41-MHz crystal filter; the crystal filter has a 3.5 kHz bandwidth with a shape factor of 1:2 (6 to 60 dB). The very low noise cascode amplifier following the crystal filter provides low noise and proper filter termination.

The amplified 41-MHz signal is down converted to 9 MHz with an SRAI double-balanced mixer. An attenuator between the cascode amplifier and mixer is adjusted for as little gain as necessary to maintain good overload performance. The output of the double-balanced mixer drives one of three crystal filters through a diplexer. The 9-MHz i-f amplifier has 60 dB gain and drives an active double-balanced mixer for CW/ssb operation.

The agc is provided by an audio agc generator, The audio power stage produces 2 watts of output power. For CW operation an active audio filter is available for greater selectivity and improved signal-to-noise ratio.

Transmitting mode. In the transmitting mode a dynamic microphone with 200 ohms impedance is required to drive the dynamic speech compressor. The compressor provides constant output level into the double-balanced mixer which produces the double-sideband signal. This signal is amplified by a 2N918 stage before being converted into ssb by one of two crystal filters. A dc offset is used to produce the CW carrier signal.

At the output of the 9-MHz crystal filter the ssb signal is up converted to 41 MHz in the SRA1 double-balanced mixer and amplified in a two-stage amplifier. The 41-MHz signal is fed through a crystal filter, and by proper selection of the ALC attack/decay time rf speech processing is accomplished; the har-

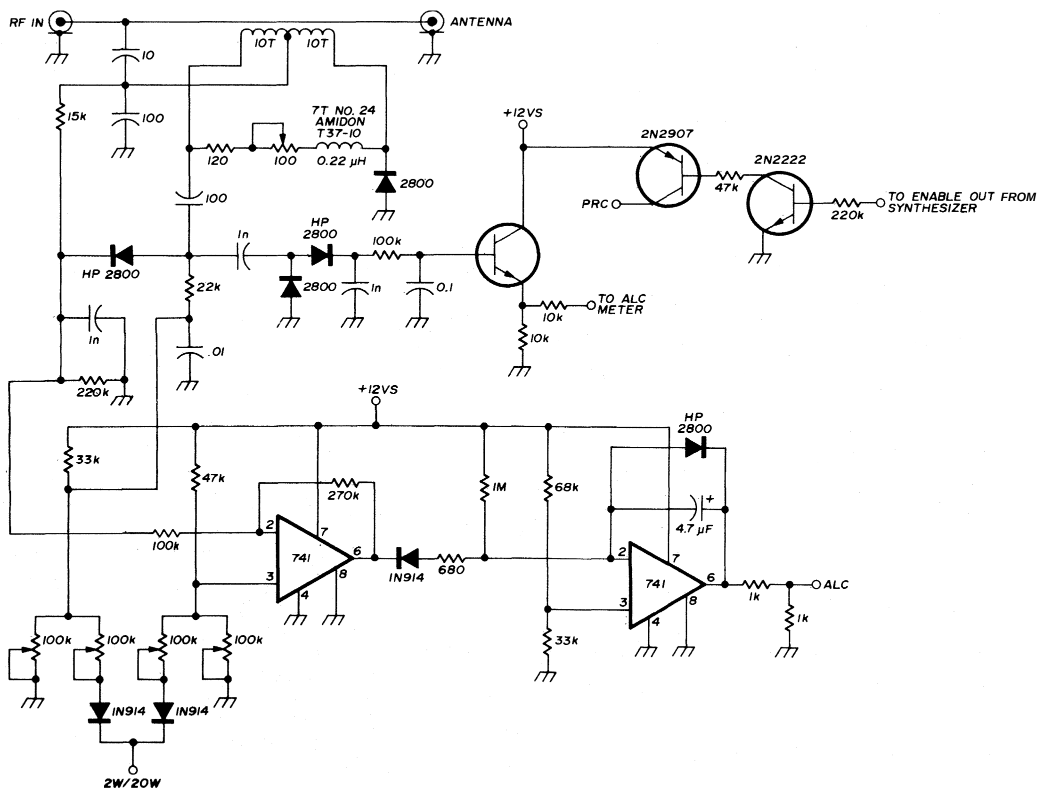


fig. 2. Schematic diagram of the ALC board which includes the directional coupler. In the tune position the rf power amplifier provides a constant two watts output for antenna tuning.

monics and spurious sidebands generated by the processing are kept under control by the 41-MHz crystal filter.

Following the 32-MHz lowpass filter the transmit signal is amplified to 20 milliwatts to drive the push-pull 20-watt rf power amplifier. The synthesizer automatically selects the proper lowpass filter (one out of seven) and the rf power is fed through the directional coupler and antenna tuning unit to the antenna.

CW **operation.** In CW operation the l-kHz signal which is required as a reference for the synthesizer (derived from the 9-MHz temperature-compensated crystal oscillator [TCXO]) is converted into a 1 kHz sine wave and fed into the audio amplifier as a sidetone. A dc voltage is used to offset the double-balanced mixer to generate the 9-MHz carrier (derived from the 9-MHz TCXO) which is passed through the 9-MHz crystal CW filter. The rest of the CW signal processing is identical to that used for single sideband.

circuit description

The quasi-continuous antenna tuning unit has

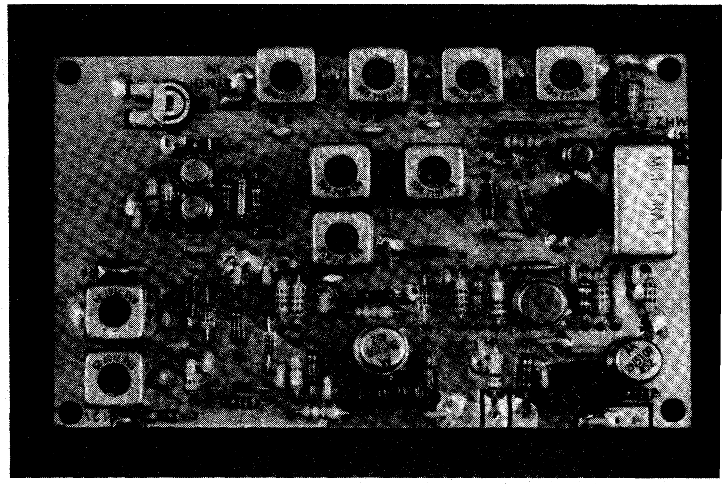


fig. 5. Rf input board showing the double-balanced mixer to the right and the lowpass filter for the synthesizer signal along the top edge of the board.

been previously described[1,2] and will provide a good match to a 6-meter (20 foot) whip at 1.25 MHz. The tapped inductors in the tuner are built with ferrite pot cores with an air gap; the tuning capacitors are sub-

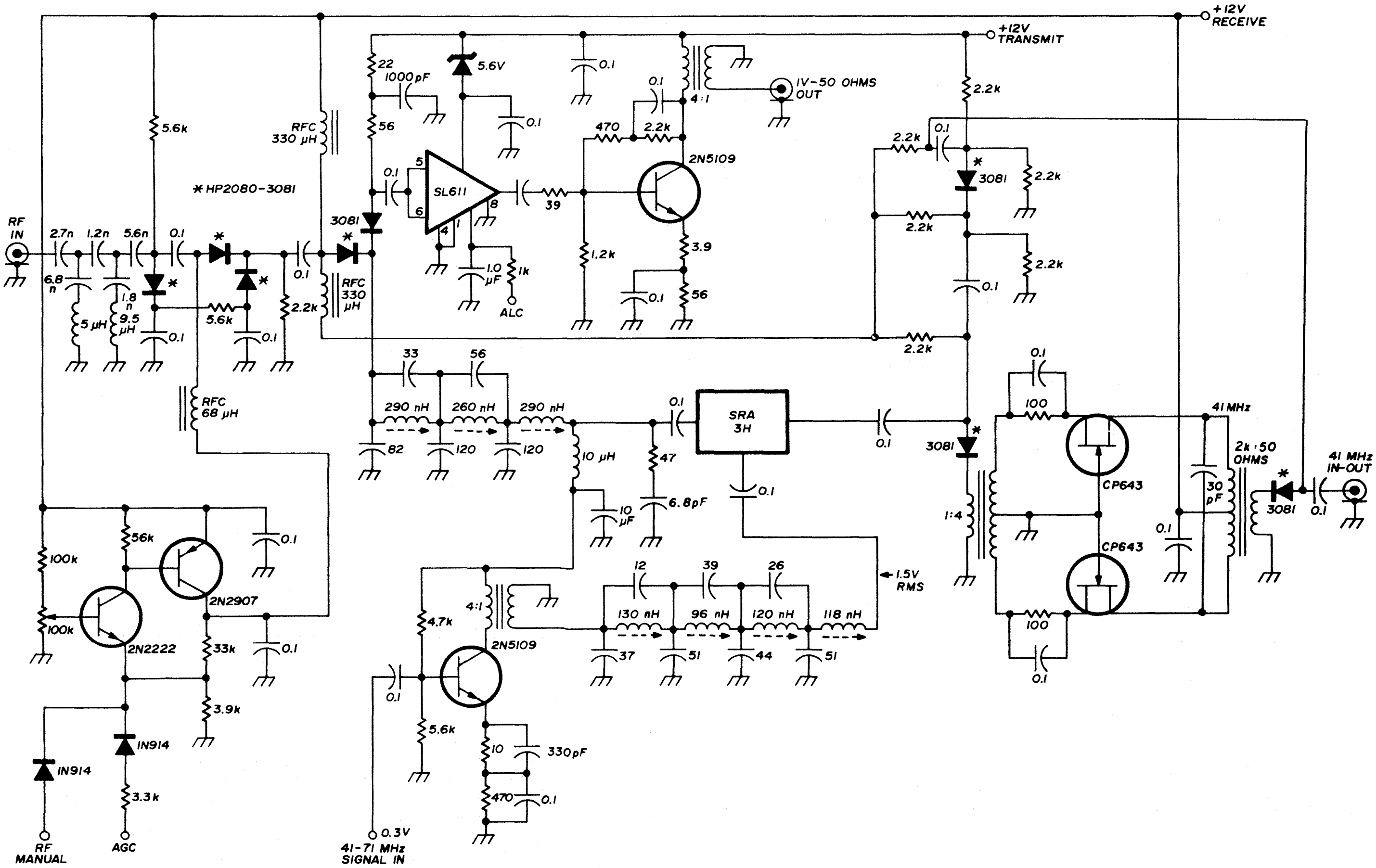


fig. 4. Schematic of the rf board with the input highpass-lowpass filter, PIN diode attenuator, and high-level transmit-receive mixer. The SL611 and 2N5109 at top center amplify the transmit signal.

miniature mica units which are switch selected from the front panel (knobs labelled L and C in the photograph of the transceiver). At the input of the antenna tuner is a 4:1 transformer which uses two 50-ohm coaxial cables wound on a 2.5cm (1 inch) ferrite toroid (TC9 material from Indiana General).

A schematic of the ALC board is shown in **fig. 2.** This board also includes the directional coupler and speech processor level adjustments. A photograph of the completed board is shown in **fig.** 3. Several new principles are used in this circuit; the voltages generated by forward arid reflected power are combined and used for two purposes:

1. Forward power peaks generate the ALC action. The first 741 IC in the circuit acts as a threshold amplifier while the second 741 is connected as a Miller integrator with fast attack and slow decay time. This is an ideal circuit for rf speech processing (clipping with a duration of a few milliseconds).

2. This circuit also detects reflected power and protects the power amplifier stage at full output while generating a constant output power of 2 watts in the *tune* position for optimum adjustment of the antenna tuning unit. The series combination of the 120-ohm resistor, 100-ohm adjustable resistor, and the 0.22 μH inductor are for frequency compensation.

As was mentioned in the general description of the transceiver, the bank of seven lowpass filters is addressed by the frequency synthesizer; the proper filter is selected by miniature relays packaged in T05 cans made by Teledyne (winding information for the lowpass filters is available upon request from *ham radio).*

A set of optional highpass input filters is recommended to improve the second-order intermodulation distortion products (e.g., 8 MHz + 6 MHz = 14 MHz, 8 MHz − 6 MHz − 2 MHz). One highpass filter has a cutoff frequency of 7 MHz while the other has a cutoff at 10 MHz. When listening to the amateur 20-meter band at night, the combination of 8 MHz + 6 MHz is totally suppressed by the second highpass filter; when listening to the 7-MHz band the combination of 3.5 MHz + 3.5 MHz is suppressed by the first highpass filter.

At frequencies below 7 MHz no additional highpass filter is required because the receiver input is provided with a 1.5 MHz highpass filter which is in the circuit at all times and eliminates problems with

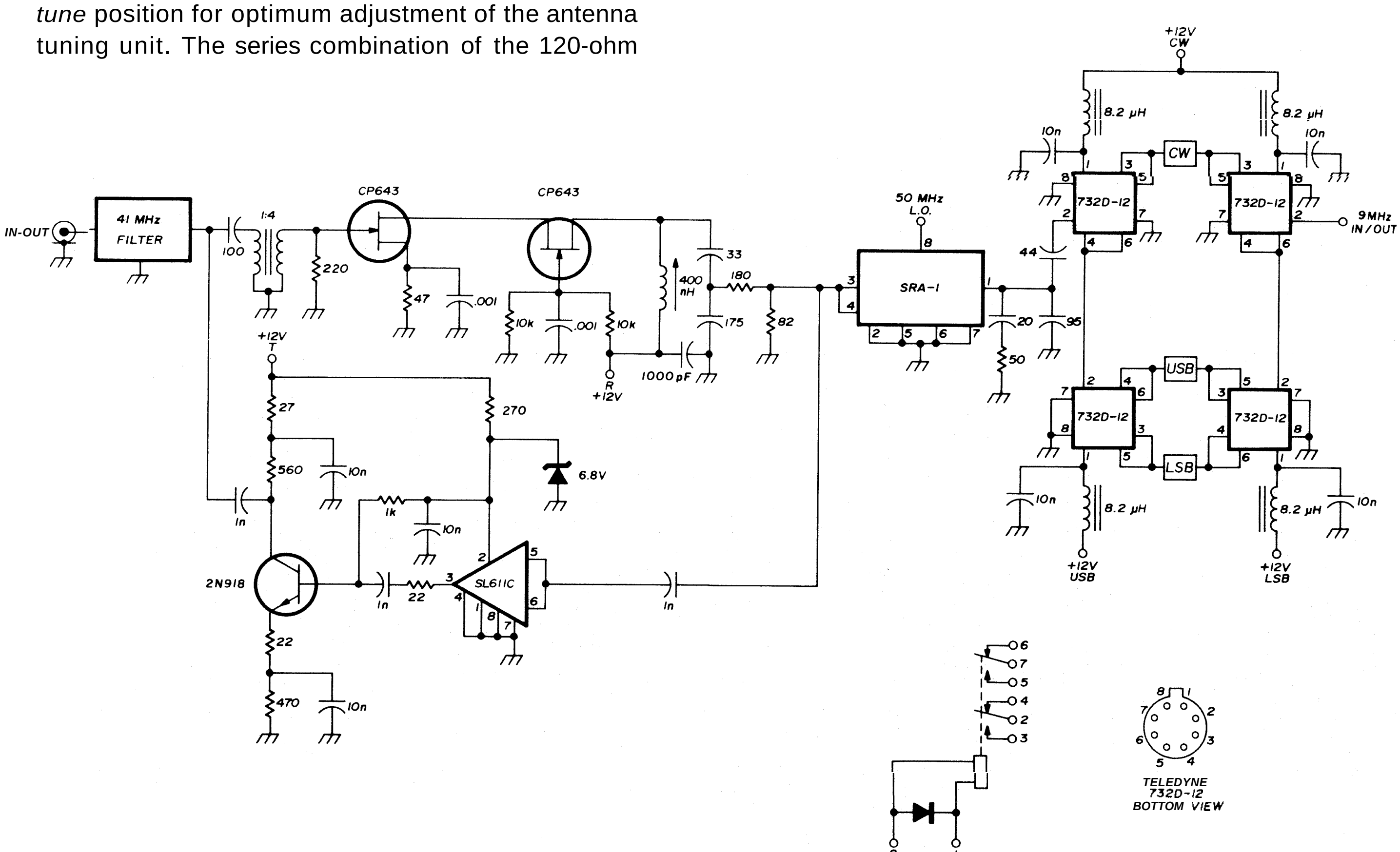


fig. 6. 41-MHz i-f and crystal-filter board. The overall gain of this stage is approximately 2 dB in the receive mode. The 9-MHz crystal filters are selected by miniature Teledyne relays which are housed in TO5 cans.

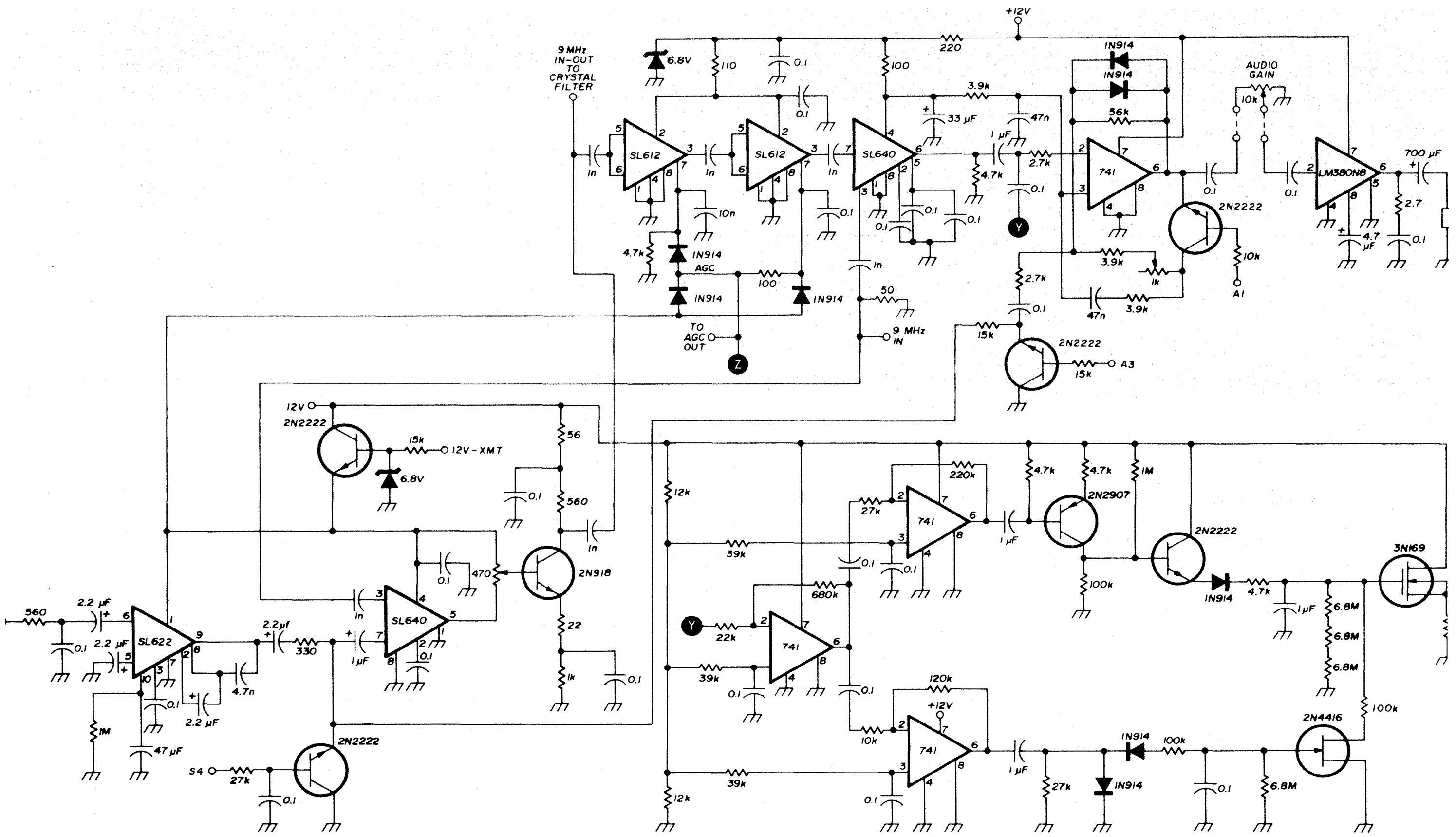


fig. 8. Schematic of the 9-MHz i-f and **agc** board. At the top are the two **SL612** i-f amplifier stages, followed by the **SL640** product detector, 741 audio preamp, and **LM380** audio power amplifier. The circuit on the bottom generates the **agc**; see text for **agc** operation.

strong broadcast stations (this filter can be retuned to 1.6 MHz if local conditions require it).

receiver input

A schematic of the receiver input section is shown

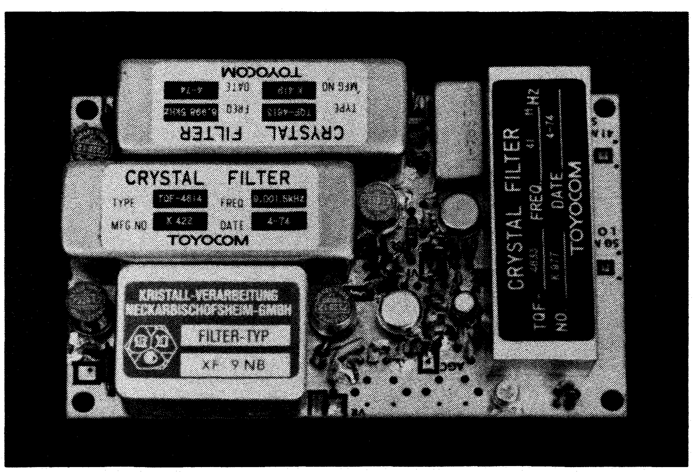


fig. 7. 41-MHz i-f and crystal-filter board. The 41-MHz crystal filter, and the two 9-MHz sideband filters, were specially made for **the** author by Toyocom in Japan. The 9-MHz CW filter is the latest design from KVG in West Germany.

in fig. 4; a photograph of the input section circuit board is shown in fig. 5. I have previously described many of the performance filters in this circuit (see references 3 and 4). Following 1.511.6-MHz highpass filter the received signal passes through a three PIN diode attenuator which has almost constant input and output impedance. The agc voltage derived from the audio agc generator feeds the dc amplifier for the PIN diode attenuator. The 100k resistor permits adjustment of the agc level which should be set for a 3 to 5 μV input signal. The available agc range is 60 dB.

The 32-MHz lowpass filter and the high-level double-balanced mixer following the attenuator are used in both the receive and transmit modes. In the transmit mode the 41-MHz input signal from the crystal filter is converted to the desired operating frequency in the high-level double-balanced mixer. The series combination of the 47-ohm resistor and the 6.8-pF capacitor provides adequate termination for the mixer in the transmit mode to keep the third-order IMD products below the distortion level of the output rf amplifier; the 32-MHz lowpass filter eliminates all unwanted harmonics.

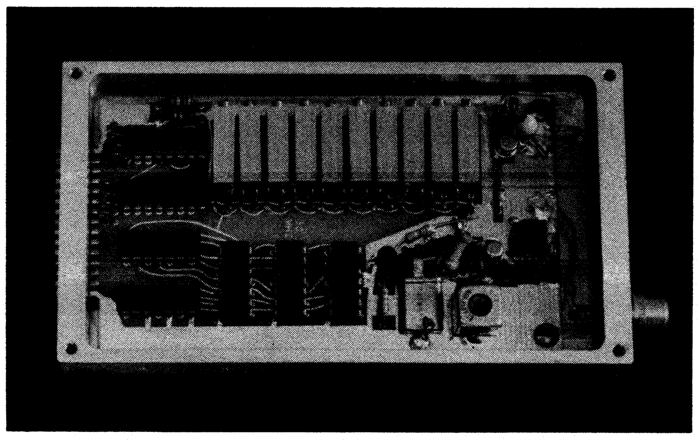


fig. 10. Photograph of the 50-MHz VCXO.

In the receive mode the push-pull fet amplifier provides the necessary wideband termination for the mixer and amplifies the received signal by about 10 dB. In the transmit mode this circuit is bypassed by switching diodes.

The 2N5109 stage amplifies the output from the frequency synthesizer and delivers +17 dBm of local-oscillator drive through a lowpass filter of elliptical design. The 2N5109 requires an input rf drive level of 300 mV.

Also on this board is an SL611 preamplifier IC and 2N5109 driver which boosts the 9-MHz transmit signal from the low-level mixer up to 1 volt. The SL611 also accepts the ALC voltage.

41-MHz i-f

Fig. 6 is the schematic of the first i-f board; a photograph of the board is shown in **fig. 7**. In the receive mode the signal from the 41-MHz crystal filter passes through a 1:4 step-up transformer which provides the necessary termination for the filter. The fet cascode circuit is a very low noise, unconditionally stable amplifier which feeds the SRA1 double-balanced mixer. The +7 dBm oscillator injection required by the mixer is provided by the TCXO.

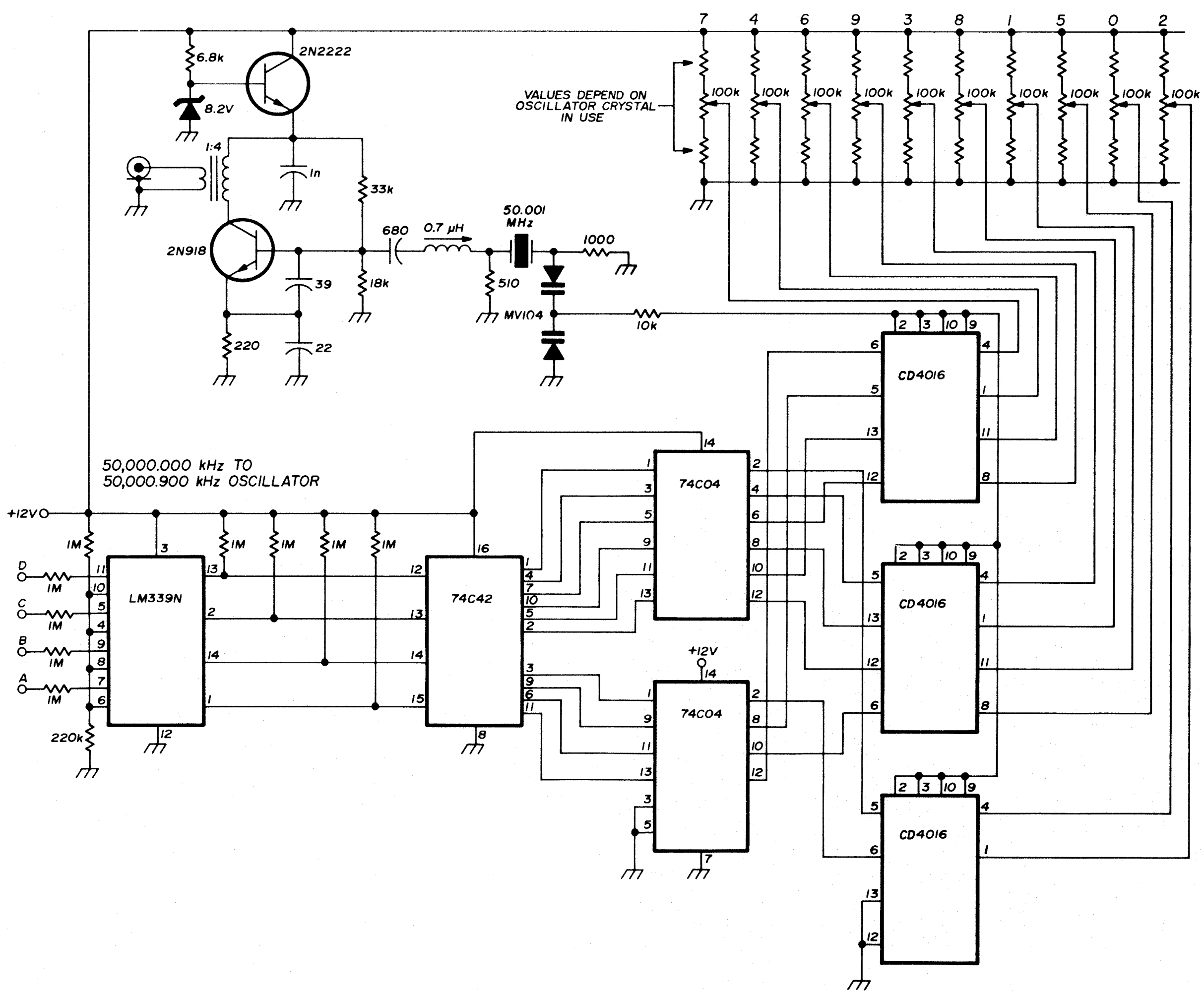


fig. 9. Schematic of the voltage-controlled crystal oscillator or VCXO. This circuit provides outputs in 100 Hz increments from 50.000 to 50.009 MHz, is stable and trouble free.

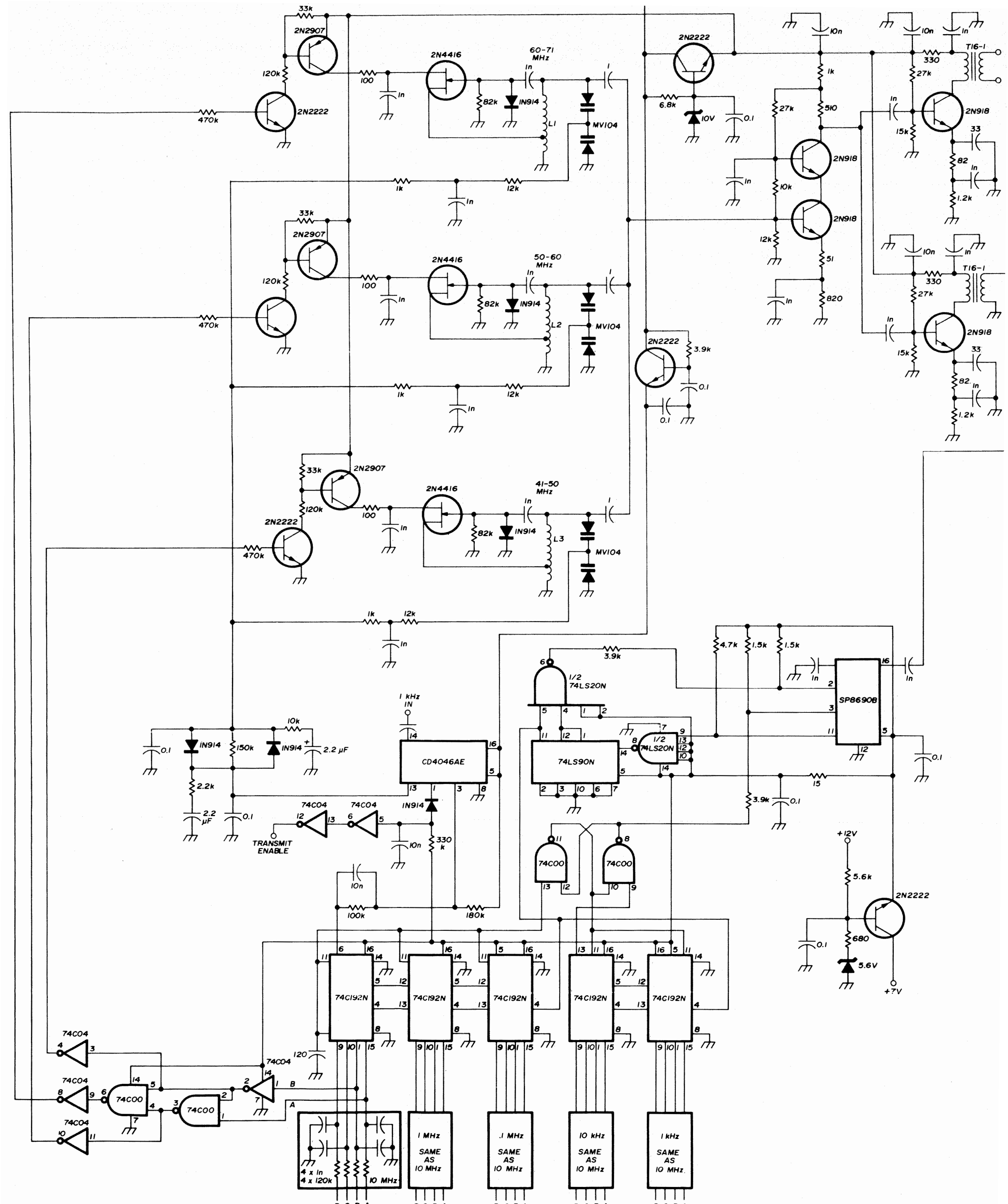


fig. 11. Schematic of the 41 to 71 MHz frequency synthesizer used in the general-coverage ssb/CW transceiver. Three VCOs are required: 41-50 MHz, 50-60 MHz, and 70-71 MHz. Input is either from thumbwheel switches or an optical shaft encoder.

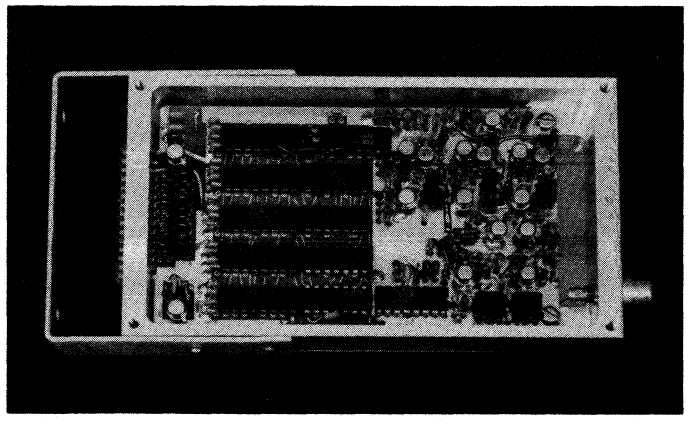


fig. 12. Photograph of the 41 to 71 MHz frequency synthesizer.

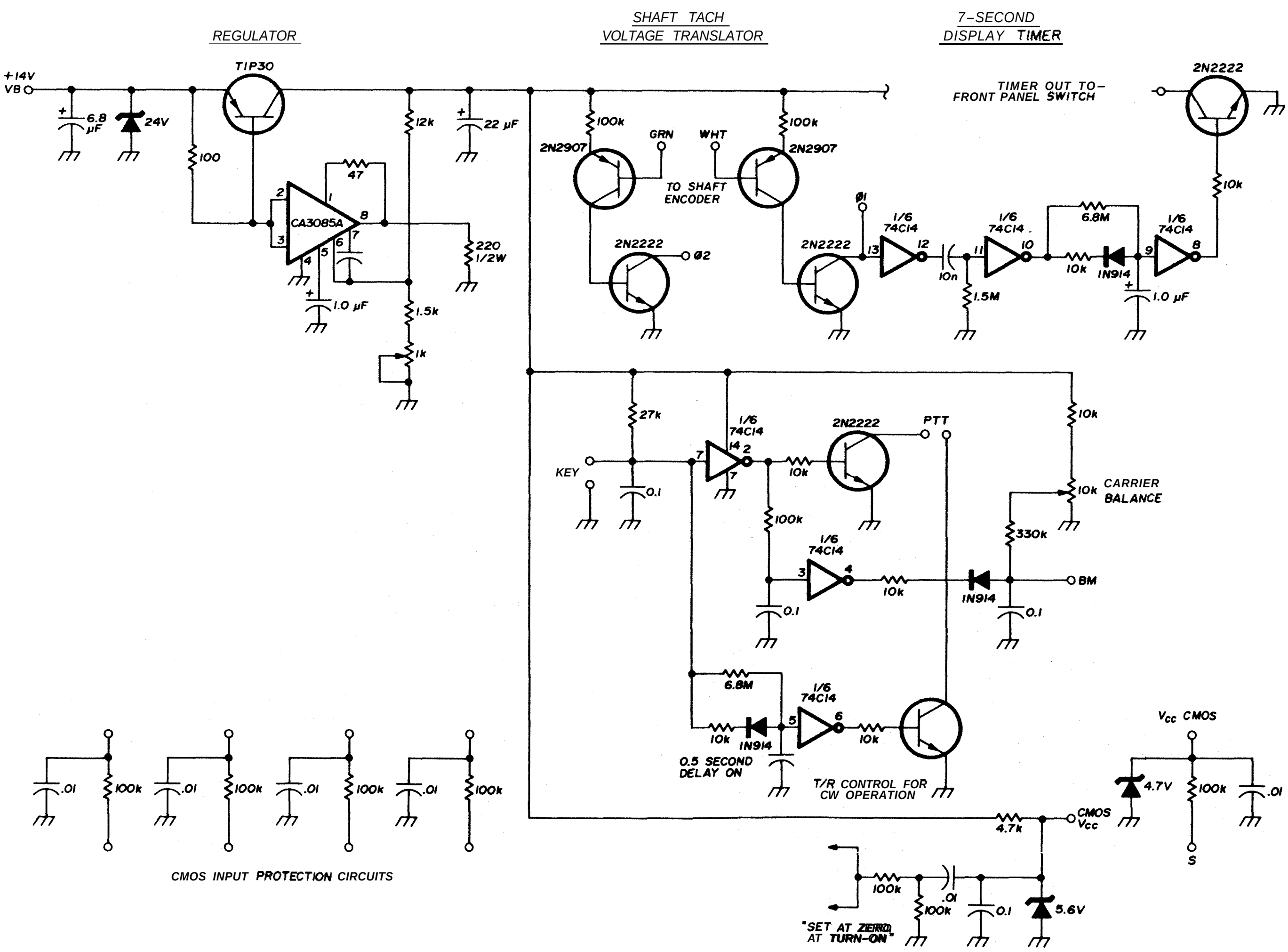
In the transmit mode the 9-MHz signal is filtered through one of three crystal filters, converted to 41 MHz by the SRA1 double-balanced mixer, amplified by the SL611 and 2N918, and passed on to the 41-MHz crystal filter. The 9-MHz crystal filters are selected by an arrangement of two miniature relays.

The 41-MHz crystal filter, and the two 9-MHz ssb filters, were made especially for me by Toyocom in Japan; the 9-MHz CW filter is the latest design from KVG in West Germany.

9-MHz i-f

The circuit of the main i-f/agc board is shown in **fig. 8;** a photograph of the unit is shown in **fig. 9,** In the receive mode the 9-MHz i-f input signal from the crystal filters is amplified by two Plessey SL612 i-f amplifier ICs and detected in the SL640 active product detector. The bfo signal is derived from the frequency synthesizer. At the output of the product detector the audio signal is amplified by the 741 operational amplifier and LM380 audio power amplifier. The audio signal from the product detector is also fed through the agc section (lower part of the schematic), amplified, and split into two channels. The 2N2907 at the output of the upper 741 amplifier is an audio detector; the following 2N2222 is a dc amplifier which charges the 1 μF capacitor.

The audio at the output of the lower 741 amplifier



is also detected and provides a negative voltage through the 2N4416 fet. The combination of the 0.1 μF capacitor and the 6.8 megohm resistor determines the agc hold time, while the 100k resistor in the drain circuit of the 2N4416, and the 1 μF capacitor, set the decay time. The 3N169 source follower provides the agc for the two SL612 i-f amplifiers and the PIN diode attenuator. (For more information on the operation of the various stages, see reference 5.)

In the transmit mode the audio from the microphone is fed into the SL622 which acts as a dynamic speech compressor and drives the SL640 double-balanced mixer IC. The 2N918 amplifier provides the necessary rf signal level for the up-conversion following the 9-MHz crystal filters.

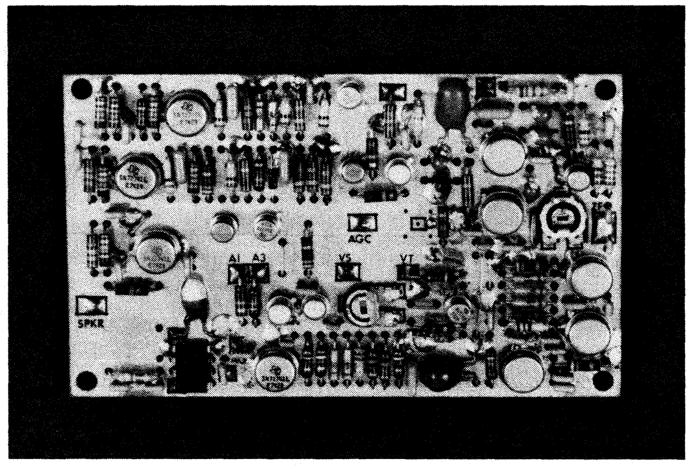


fig. 14. Photograph of the 9-MHz i-f and agc board. The agc circuitry is to the right, the i-f amplifier to the left.

crystal oscillator

Fig. **10** shows the schematic of the voltage-controlled crystal oscillator (VCXO) used in the transceiver; a photograph of the unit is shown in **fig. 11.** While analyzing frequency synthesis circuits for the transceiver, it was determined that it was not feasible to build a single-loop 100-Hz synthesizer because of the lack of loop gain. To do this would have required the use of very expensive diodes and coarse steering or presetting of the VCO to build a stable and low noise loop. Since this could not be done economically, the last digit (100 Hz) is achieved by pulling a 50-MHz crystal by the relatively small amount of 900 Hz, in 100-Hz increments from 50.0009 MHz down to 50.0000 MHz. The overall frequency stability of the transceiver is determined by this circuit, so the temperature coefficient of the 22 pF and 39 pF capacitors in the 2N918 oscillator circuit must be very carefully chosen. Since each crystal may require a different temperature coefficient capacitor for compensation, this is best determined by experiment.

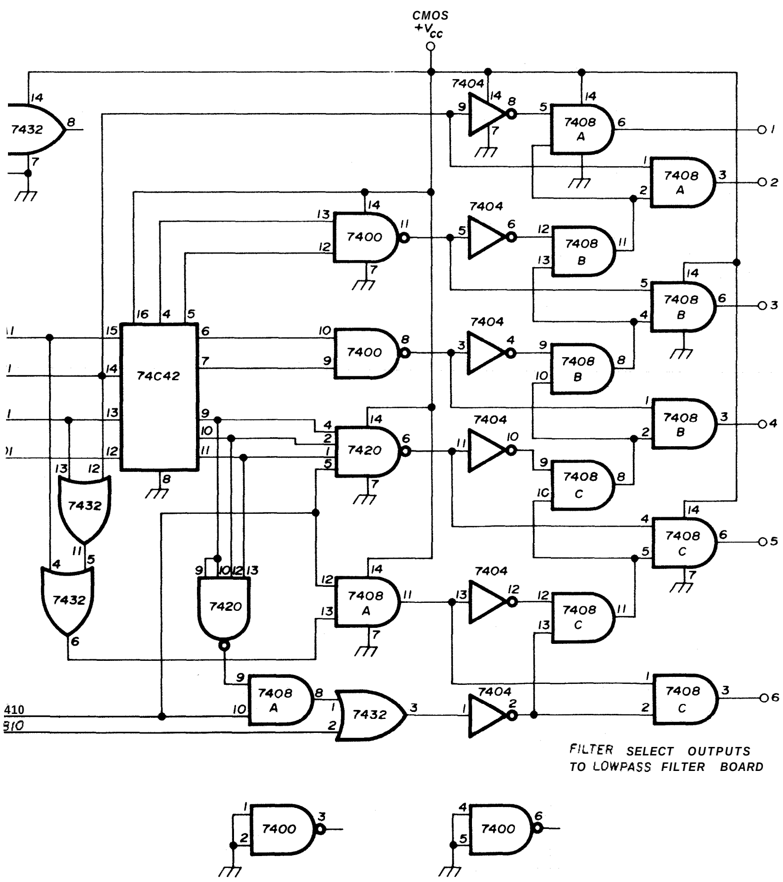


fig. 13. Schematic of the main dc control board. Included are the automatic lowpass filter selection circuits, right; transmit-receive control for CW operation, left; a 12 Vdc regulator; and 7-second display timer (see text).

To properly adjust the ten potentiometers in the VCXO requires the use of a good frequency counter. First enter 9 in BCD code at the input and set the 0.7 μH inductor and 100k potentiometer so the output is at 50.0009 MHz. Then enter 8 in BCD code and adjust the appropriate 100k pot for 50.0008 MHz. Continue in this fashion until all ten potentiometers have been set. If you run out of pulling range with the potentiometers, increase the size of the inductor and try again. Other than the adjustment procedure, which should give no problems if you use a counter, this oscillator is very simple and well behaved.

frequency synthesizer

The single-loop frequency synthesizer shown in **fig. 12** covers the range from 41 to 71 MHz in 1 kHz increments; a photograph of the synthesizer board is shown in **fig. 13.** Three VCOs are used in this circuit, each covering 10 MHz. The output from the VCOs is fed into a high isolation amplifier and two independent drivers. One driver feeds the 2N5109 amplifier on the receiver board **(fig. 4)**, while the other amplifier drives the SP8690B swallow counter which is used

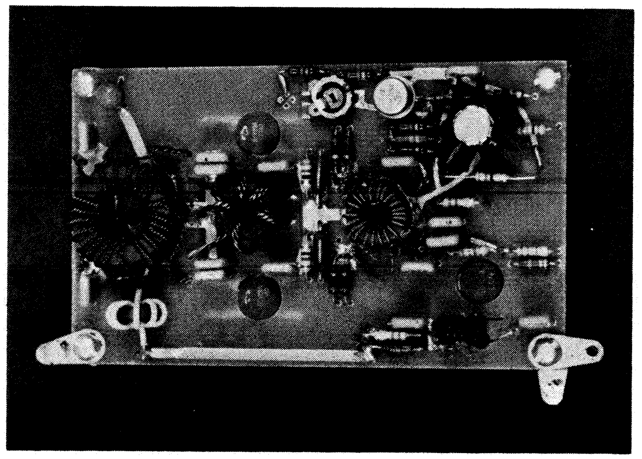


fig. 16. 20-watt rf power stage; the input is to the left, output to the right.

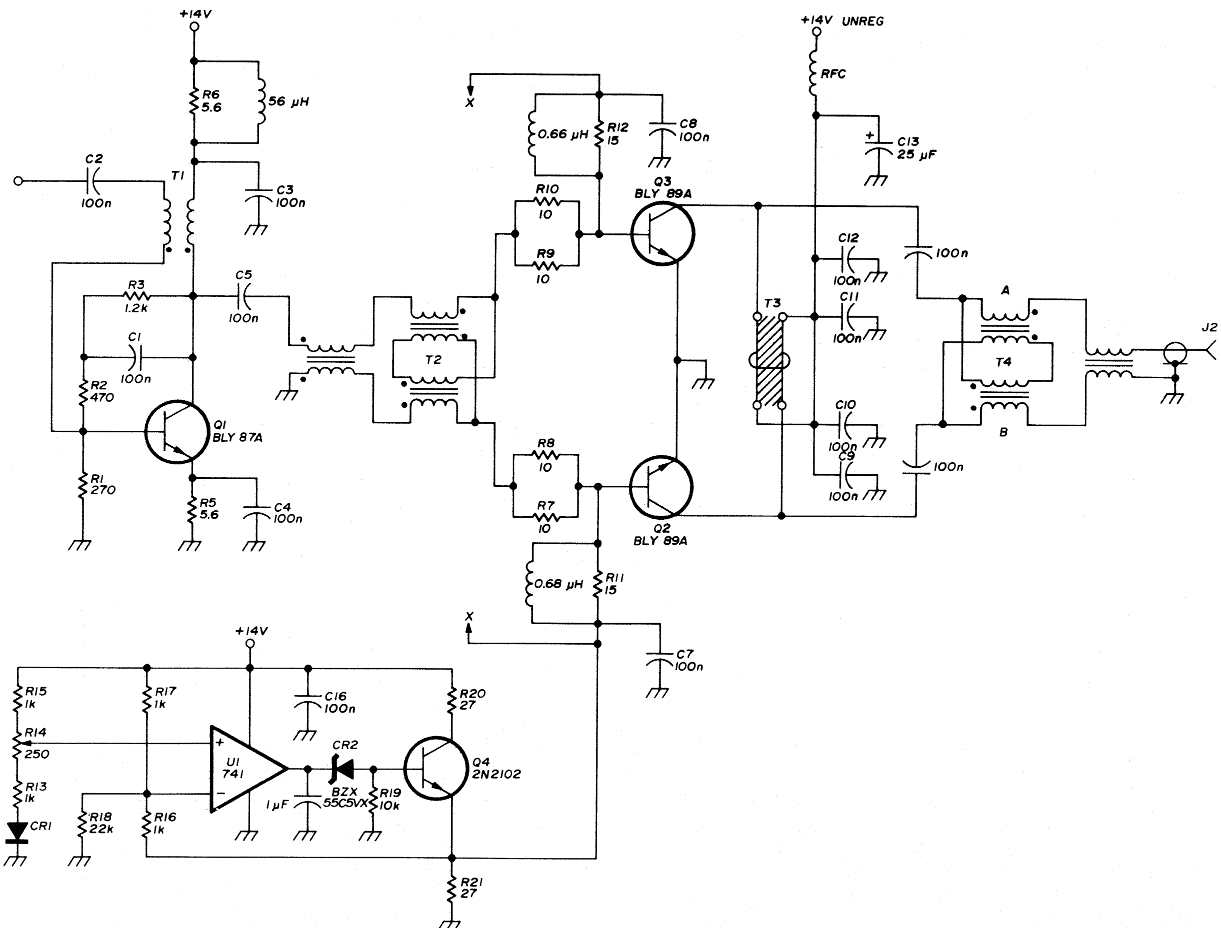
with the 74LS20 and 74LS90 to form the 100/101 synchronous counter. A more complete description of synthesizer operation, and the CD4046E phase detector, is contained in references 6 and 7.

A schematic of the 9-MHz reference oscillator is not shown; the TCXO I used in the transceiver was manufactured by McCoy.* The output of the TCXO feeds a divider chain which delivers the 1 kHz reference; the output of the TCXO is also used as the bfo. For CW reception a 9.001 MHz bfo signal is required; this is provided by a separate crystal oscillator circuit.

rf power amplifier

The 20-watt rf power amplifier used in the trans-

*McCoy par: number MC 163x2 070W 9 MHz, manufactured by McCoy.



T1 1:7 transformer. 1/7 turns of no. 26 AWG (0.4mm) enameled on Indiana General F625-9-TC9 toroidal core.

T2 2:1 transformer balun. 8 turns, 2 twisted pairs, no. 31 AWG (0.22mm) ernarneled wire; 5 turns, 2 twisted pairs, no. 31 AWG (0.22mm) emameled wire; all windings 5 twists per cm (12 twists per inch) on Indiana General F624-19-Q1 toroidal core.

T3 Collector choke, 4 turns, 2 twisted pairs, no. 22 AWG (0.6mm) enameled wire, 2-1/2 twists per cm (6 twists per inch), on Indiana General F624-19-Q1 toroidal core.

T4 1:4 transformer balun. Windings A and B: 5 turns, 2 twisted pairs, no. 26 AWG (0.4mm) enameled wire; winding C: 8 turns, 2 twisted pairs, no. 26 AWG (0.4mm) enarnaled wire; all windings 3-1/2 twists per cm (9 twists per inch); wound on Indiana General F617-8-Q1 toroidal core.

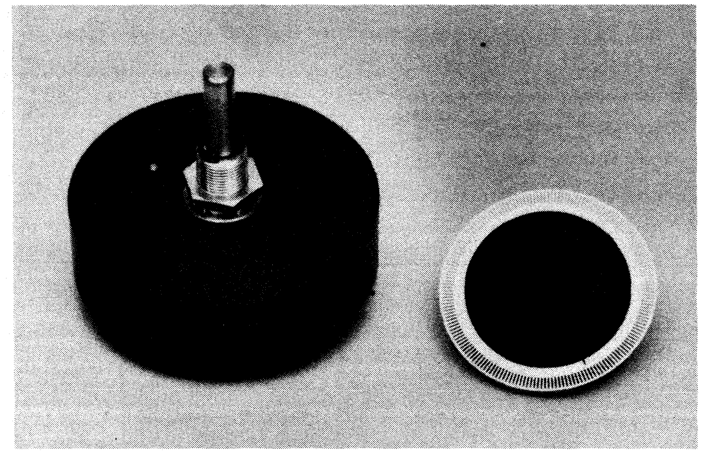
fig. 15. The 20-watt rf power amplifier in the high-frequency transceiver uses a BLY87A driver and push-pull BLY89As. The 741 op amp and 2N2102 provide constant dc bias; the 1N4448 is a temperature sensing device and should be mounted near the final power transistors.

ceiver is shown in **fig. 14;** a photograph of the assembly is shown in **fig. 15.** In this circuit the BLY87A driver transistor uses both voltage and transformer feedback to maintain flat gain and constant input impedance over the entire operating frequency range. Transistor Q3 in the push-pull final amplifier has an input stabilizing network consisting of R9 and R10 in series, and R12 in parallel for the rf path (R7 and R8 in series, R11 in parallel, make up the stabilizing network for transistor Q2). The voltage for the push-pull transistors is supplied through transformer T3; transformer T4 combines both phases and provides a single-ended 50-ohm output.

To maintain constant dc bias over a wide temperature range, a high-gain loop using a 741 operational amplifier is used along with a high-current 2N2102 transistor. The 1N4448 diode, which is the temperature sensing device, should be mounted very close to transistors Q2 and Q3.

control functions

The main dc control board for the transceiver is shown in **fig. 16.** The information from the thumbwheel switches on the front panel of the transceiver is fed in parallel to both this board and the 41-71 MHz



Photograph of an optical shaft encoder and its slotted encoding disc. The 180 slots are chemically etched in the disc. The disc is 1-5/8″ (41mm) in diameter.

synthesizer board **(fig. 12)**. In addition to controlling lowpass filter selection, this board includes a 12-volt regulator, transmit-receive control for CW operation, and circuitry for the optical shaft encoder.

The optical shaft encoder generates two quadrature square waves or pulses which are a function of shaft rotation – the waveforms are generated by a

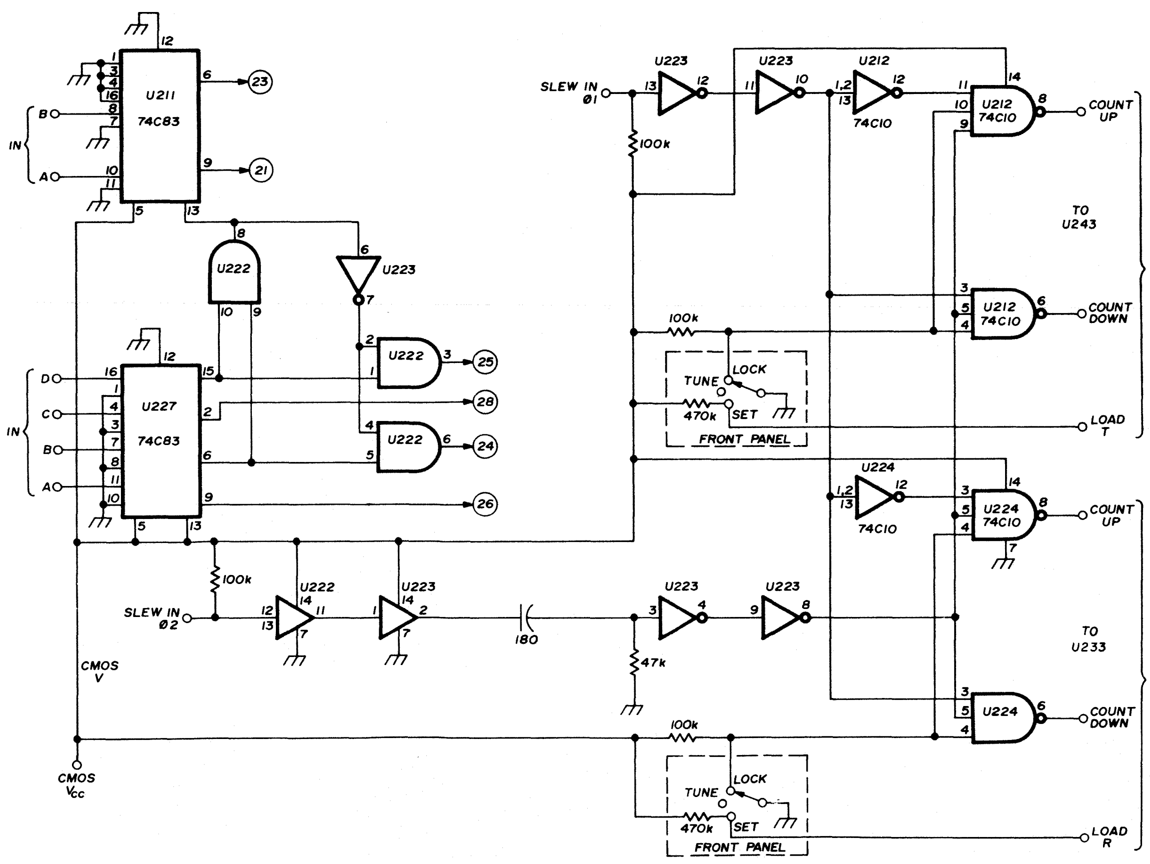


fig. 17. Tuner control logic provides for a digital readout of zero when the synthesizer is tuned to 41 MHz. Outputs to the right, labelled T and R, are connected to the lower and middle tuner boards (figs. 18 and 19).

slotted disc which interrupts the light path between two LEDs and two photo-detectors. The output from the shaft encoder is used to control the frequency synthesizer; since the shaft encoder used in this transceiver has 180 slots, the synthesizer tunes 18 kHz per complete dial revolution. For a more descriptive discussion of optical shaft encoders, see reference 8.

The two outputs from the shaft encoder are applied to the inputs of two 2N2222 transistors. The outputs – labeled phase 1 ($\phi 1$) and phase 2 ($\phi 2$) are used to feed the corresponding inputs labeled slew *in $\phi 1$* and *slew in $\phi 2$* in **fig. 17.**

The dc control board also includes a seven-second display timer so that whenever the frequency is changed and the display is not turned on, the LEDs will turn on for seven seconds to display the final frequency.

Since the thumbwheel switch and the LED display must show the digit zero when the synthesizer is set to **41** MHz, this is accomplished by the logic circuitry shown in **fig. 17.** The outputs labeled R and T in **fig. 17** are connected to the lower and middle tuner boards **(fig. 18** and **19,** respectively). These two boards have the necessary memory and display logic for the Hewlett-Packard 5082-7300 LED dot displays.

*The optical shaft encoder used in the transceiver was manufactured by Dr. Johannes Heidenhain GMBH in West Germany; it is available from their United States sales representative.

Internal construction of the optical shaft encoder. The LED light source is under the half-moon shaped shield on the left-hand side of the unit (right).

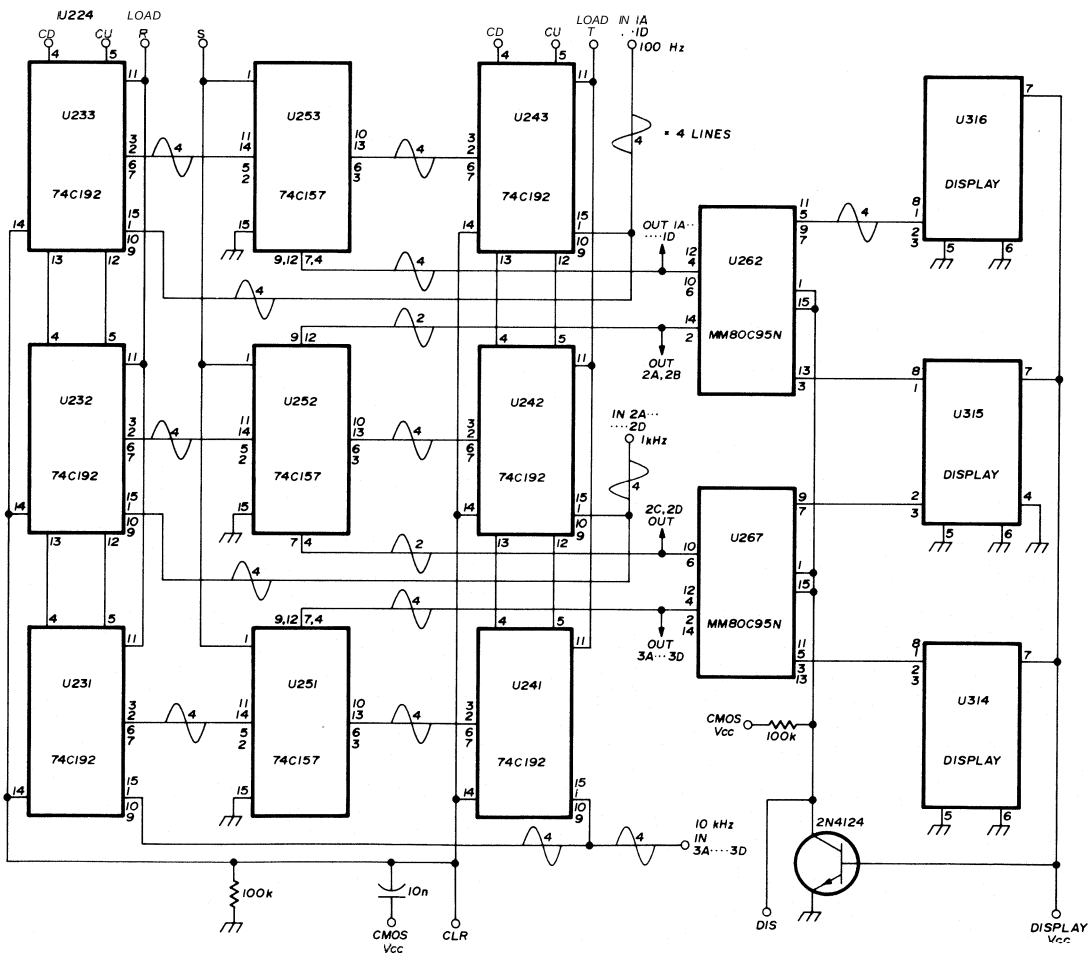
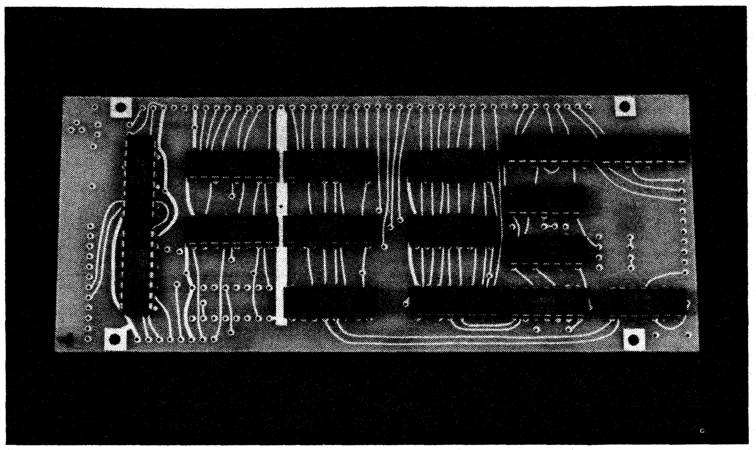


fig. 18. Middle tuner board with memories and decoding for the frequency synthesizer (100 Hz, 1 kHz, and 10 kHz digits).



Printed-circuit board for the middle tuner; other tuner board is similar.

The inputs from the thumbwheel switches to the memory are labeled 1A through 1D for the 100 Hz digit, 2A to 2D for the 1 kHz digit, and on to 6A through 6D for the 10 MHz digit.

summary

This transceiver was built more than a year ago, and since then it has been taken on a number of vacation trips, where it performed superbly, without difficulty, every time. The dual memory makes it ideal for DX operation because it allows for split frequency operation (receive and transmit can even be on separate amateur bands, if desired). The built-in power supply and nicad battery pack permit completely portable operation (this is the reason power input was limited to 20 watts); and the antenna tuner allows the use of a short whip or random wire antenna.

references

1. Ulrich L. Rohde, "Match Antennas over 1.5 to 30-MHz Range with only Two Adjustable Elements," *Electronic Design 19,* September 13, 1975, page 96.
2. Ulrich L Rohde, "Die Anpassung von kurzen Stabantennen fuer KW-Sender," *Funkschau* (Germany), Volume 7, 1974, page 673.
3. Ulrich L. Rohde, "Optimum Design for High-Frequency Communications Receivers: *ham radio,* October, 1976, page 10.
4. Ulrich L. Rohde, "Eight Ways to Better Radio Receiver Design," *Electronics,* February 20, 1975. page 87
5. Ulrich L Rohde, "I-F Amplifier Design," *ham radio,* March, *1977,* page 10.
6. Ulrich L. Rohde, "Aufbau und Anwendung digitaler Synthesizer," *Funkschau* (Germany), Volume 9, 1973, page 975.
7. Ulrich L. Rohde, "Modern Design of Frequency Synthesizers," *ham radio,* July, 1976, page 10.
8 Dennis D. Rasmussen, W6MCG, "A Tuning Control for Digital Frequency Synthesizers," QST, June, 1974. page 29.

ham radio

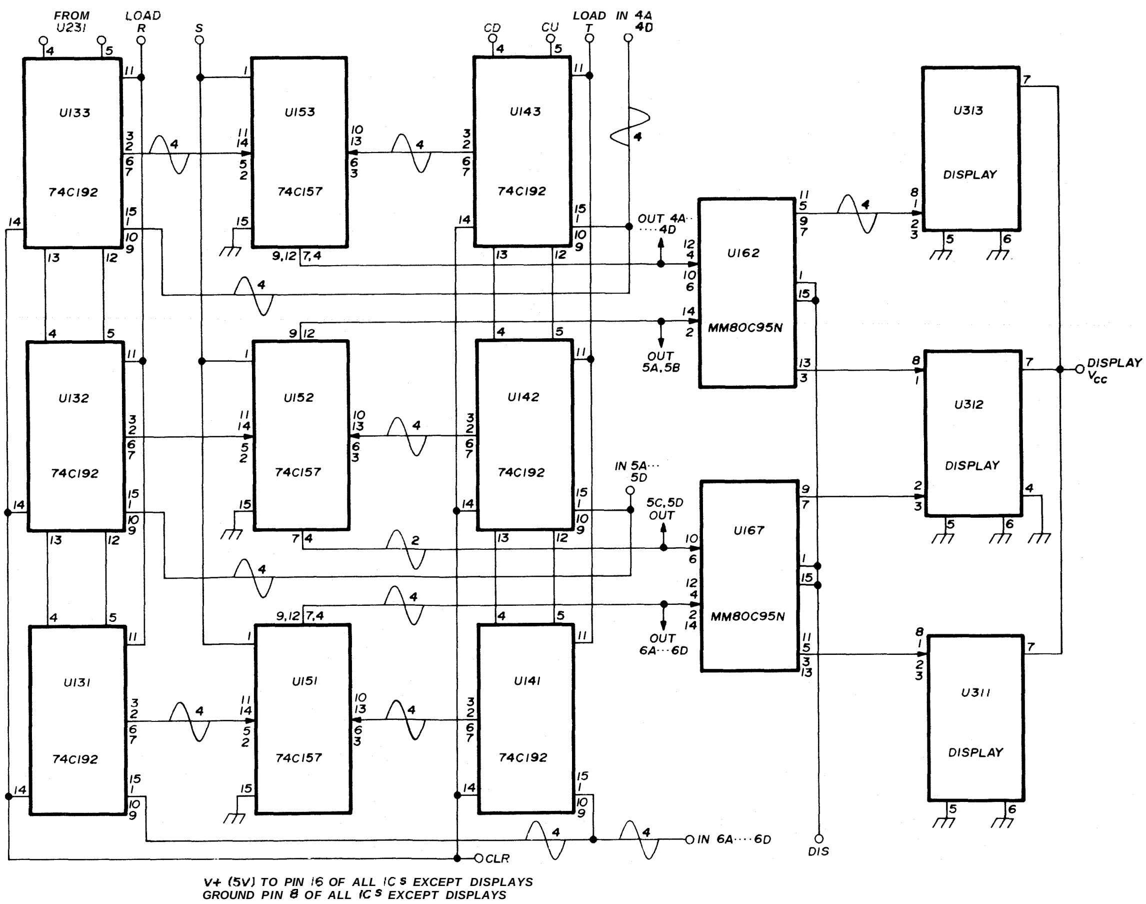


fig. 19. Lower tuner board with memories and decoders for the frequency synthesizer (100 kHz, 1 MHz, and 10 MHz digits)

a new approach to weak-signal communications

A new technique for obtaining solid copy during poor signal-to-noise conditions

For years, serious amateurs have been constantly improving their station capability. This is particularly true at vhf and uhf where noise figures have been honed to within a fraction of external noise, transmitter power and efficiency have been pushed to the limit, and antenna size now truly boggles the mind. We are now faced with the inevitable question of where does our next improvement come from, or are we at the end of the line for weak-signal work? This article will examine some of these questions and consider a possible approach which has not been previously applied to amateur work.

after the rf stage

In attempting to improve weak-signal capability the portion of the amateur station which has probably seen the least improvement, and possibly been the least understood, is that which follows the rf stages of the receiver. Let's start there.

Assume you are trying to copy a weak signal and can't quite hack it. What do you do? The most common idea is to narrow the selectivity to reduce the amount of noise coming through with the signal. This is perfectly valid and if you reduce the bandwidth by a factor of ten, you might expect to pick up 10 dB in signal to noise. Unfortunately, this is not the case – the improvement is considerably less.* There is some improvement, nevertheless, so the next question is how far can you go with this approach? Is there some limit, or can you continue to narrow bandwidth indefinitely and get as much weak-signal improvement as you desire?

bandwidth limits

There appear to be three possible limits when you narrow bandwidth to extreme values. The first is a practical limit that is set by the available state of the art (the matter of equipment stability and accuracy). It would be folly to design an i-f filter with a bandwidth of 2 Hz if the rest of the receiver (and the transmitter) couldn't set and maintain frequencies within that bandwidth!

The second limitation is built into the particular propagation mode you are using. In effect, propagation variations "modulate" the signal; in some cases propagation can cause a signal to have greater bandwidth than the narrow filter you are using. High-frequency signals are usually *narrower* than 2 Hz, but tropo scatter may easily be wider, and aurora is sometimes many hundreds of Hz wide.

The third limitation involves keying speed. If you narrow your filter far enough, you must also reduce keying speeds so that all the signal components will pass through the filter bandwidth. (Yes, CW has a frequency spectrum just like the newer, fancier models!) A good rule-of-thumb is that the keying speed in words per minute should be equal to the filter bandwidth in Hz. This also gives the best signal-to-noise ratio. If you were to narrow the filter any more (to cut out more noise) you would also begin to cut out signal components. If you increase band-

*If receiver bandwidth were the only filtering in the act, this would be perfectly true. The human ear provides considerable additional filtering, however, so the *effective* bandwidth is substantially narrower than that of the receiver alone. When you further narrow receiver bandwidth, you are trying to pick up an improvement you largely already have.

By Walter F. Bain, W4LTU, Route 2, Box 286, Lovettsville, Virginia 22080

width, you let in more noise for the same amount of signal.

Therefore, if you want to go down to bandwidths of 1 or 2 Hz to pick up the signal-to-noise improvement, you must go to some pretty slow keying speeds. This should not be considered a genuine limitation, however, for amateurs who want a contact badly enough will go to outlandish lengths; a typical vhf meteor scatter contact may average only a few words per hour!

These, then, are the limitations to improving signal-to-noise ratios by bandwidth narrowing techniques. Since there is not a sizeable improvement available before these limitations show up, where do we go from here?

a closer look

First, are we attacking the right problem? Let's closely examine just what happens when the signal-to-noise ratio is not adequate. Assume that bandwidth and keying rate have been optimized; the detected waveform (without noise) would look much like that of **fig. 1A** where the characters are fully rounded by the filter. For ease in handling let's square-up this detected wave to get back to a square-wave corresponding to the original keying waveform, **fig. 1B**. For the discussion that follows, it is essential that the signal be in binary form, corresponding either to key-up or to key-down.

I am also going to dispense with the idea that the signal is made up of dots and dashes. Instead, I will define it in terms of the shortest element it contains, which is called a *bit.*" A dot happens to be one bit, a dash is made up of three bits. The space between the dot and dash is also one bit although it happens to be a key-up bit. Normal spacing between letters is three bits (key-up). This may be somewhat different from the way you usually look at International Morse, but it allows each element to be treated separately.

Now let's get back to the effects of not having adequate signal to noise. As the signal-to-noise ratio is decreased, some of the bits are in error; that is, they are reversed from what they should be, indicating key-up when they should indicate key-down, or vice versa (see **fig. 2**). Note that the bits can't come out with in-between values if the squaring-up was done correctly — only key-up or key-down.

'This shortest element is basic because it sets the bandwidth of the optimum filter.

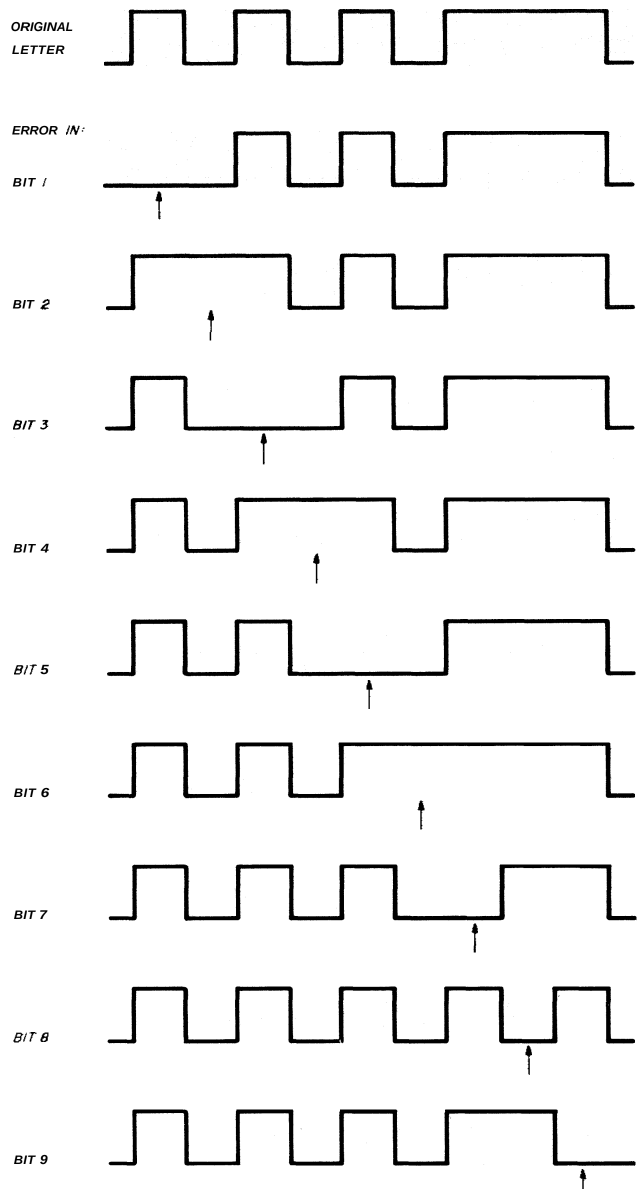


fig. 2. Effect of a single error on the Morse letter V; error in various positions.

Where, and how often do these errors appear when the signal-to-noise ratio gets too low? The errors are statistical in nature so we can state only the probability of error, or what per cent of a large sam-

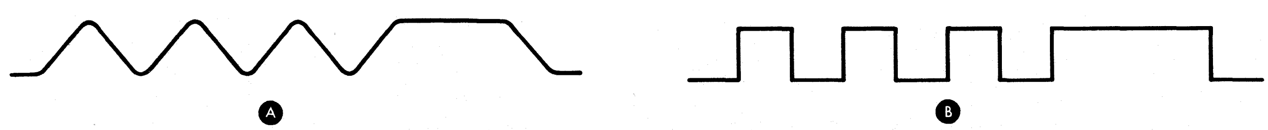


fig. 1. Detected waveform of the Morse letter V, optimized bandwidth (A). At (B) is the same waveform after being squared up.

ple of bits will be in error. This probability of error (Bit-Error Rate or BER) increases as the signal-to-noise ratio decreases, and it is possible to draw curves of BER *vs* signal-to-noise ratio as shown in **fig.** 3. Is this an actual measure of our weak-signal limitation rather than just signal to noise? Not quite; there's one more thing to be considered.

Assume the BER is 1 in 100 (a group of 100 received bits is likely to contain 1 bit error). This sounds pretty good, but remember that 100 bits make up about 8 letters or characters in International Morse.

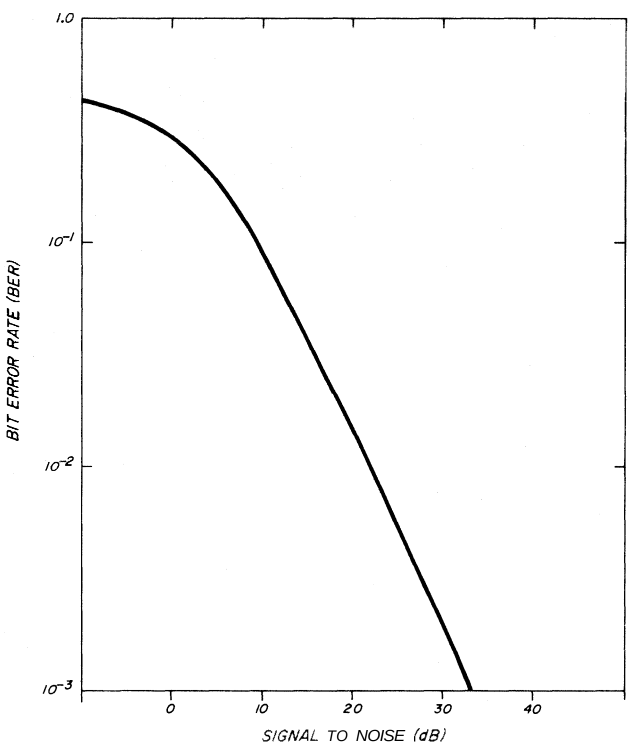


fig. 3. Bit error rate (BER) vs signal-to-noise ratio, fading signal.

Therefore, this one bad bit, wherever it may fall, is likely to foul up one whole character out of 8. This means that the Character-Error Rate (CER) is 1 in 8, which is not so good! Now we have arrived at a basic measure of weak-signal performance: the signal-to-noise ratio determines the BER, and the BER in turn sets the CER. It is the CER that actually limits our ability to communicate.

a new approach

Now that we have struggled through optimum bandwidth, bits, BER, and CER, are we any closer to a solution to the original question of how to further improve weak-signal capability? Perhaps. Instead of the perpetual struggle to improve signal-to-noise ratio, we now have a new handle on the problem and can wrestle with CER instead. Let's ask a new ques-

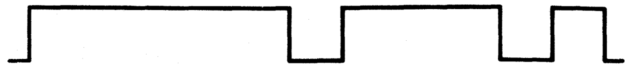


fig. 4. Received Morse letter with a single error.

tion: "How can we *improve* CER for the *same* signal-to-noise ratio?"

Fig. 4 shows a Morse character which has been mangled by a single error. Amateurs have always been skillful at salvaging and improvising, so let's see what can be done with this rather badly damaged character. Let's first eliminate any characters which it could *not* have been.

1. It can't be the letter Z because it differs from the garbled letter in bits 4, 6, and 8 (the character has only one error, and it can't be in three places at once).

2. It can't be the letter X either, as X differs at bits 4, 8, and 10.

3. Neither can it be a mangled O because of differences at bits 4, 6, 8, and 10. It would require *four* errors to change an O into the character of **fig. 4.**

4. It can't be any of the shorter letters such as U, R, D, W, G, K, V, F, L, B, M, H, A, N, S, I, T, or E, as the 3-bit spaces on either side of the letter accurately define its length. It can't be any of the longer letters, J, Y, or Q for the same reason. (The effect of errors on the spaces will be discussed later.)

Even though the letter was mangled, there was enough left of it that we were able to eliminate 24 letters which it could *not* have been! The only two re-

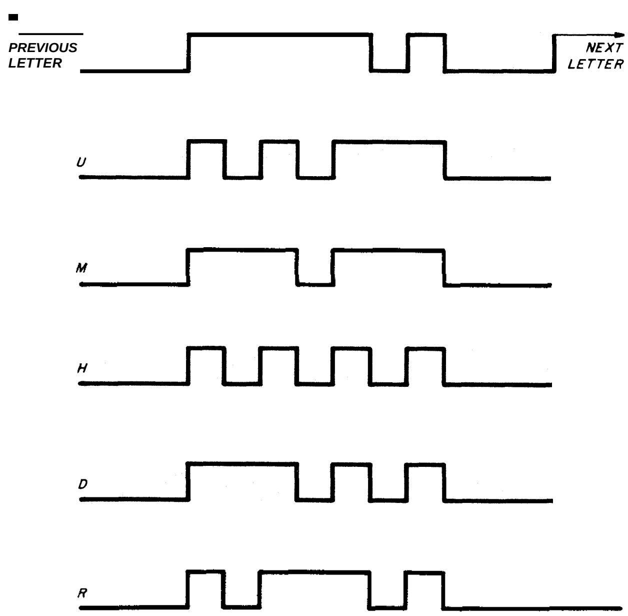


fig. 5. Received 7-bit Morse character with a single error (top), compared with 7-bit letters it could be.

maining letters are P or C; it could be a P with an error at bit 2, or a C with an error at bit 4.

What did this exercise accomplish? Instead of the letter being a total loss, enough has been recovered to narrow the choice down to one of only two letters. This is an improvement from about 4 per cent to 50 per cent.

a possible solution

This looks promising, even if we only partially recover copy which would otherwise be lost. And nothing has been done to improve the signal-to-noise ratio! But was this just a fluke? Is it possible to pull this off with other letters? Let's try another example and see.

Assume you receive the character shown in **fig. 5.** The error rate is such that the average is one error per character. The 3-bit spaces are intact so this is a 7-bit character; shown below it are the 7-bit characters it might be:

1. It can't be a U which differs in three separate bits.

2. It can't be an M either; this differs in two bits, too much to be caused by a single error.

3. H doesn't fill the bill — it differs in two spots.

4. D and R differ in only *one* position from our unidentified character and this *can* result from the single error.

So once again it has been possible to take a mangled Morse character and narrow it down to only two possibilities; in this case it is either a D or an R.

Do all International Morse characters have this capability? Let's check out a character like H. **Fig. 6** shows that an error in just one bit will not only garble the character, but will actually transform it into some other character which then appears flawless. This is particularly insidious, as there is no way to determine by simple examination that an error is present, much less correct it. (The occurrence of errors is statistical, so although an error may be probable, it is not guaranteed!) The letter H is so bad in this respect that a single error almost anywhere transforms it into another character(s) which appears valid (an error on bits 3 or 5 makes it look like two perfect letters, EI or IE).

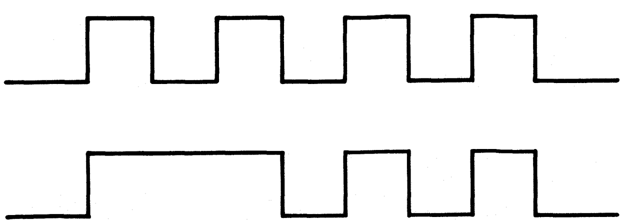


fig. 6. One effect of a single error on the Morse letter H.

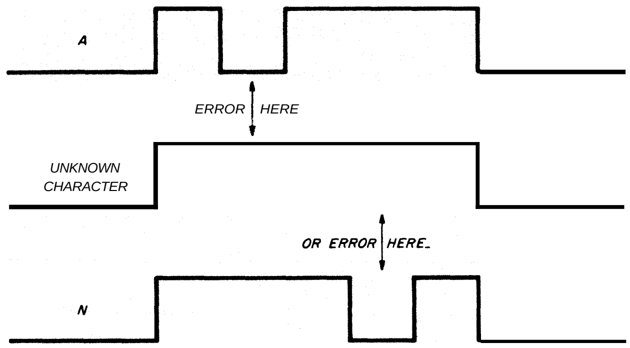
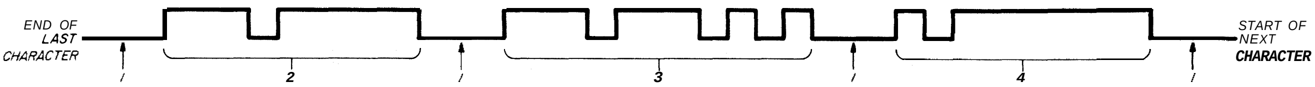


fig. 7. Possible ambiguity resulting from a single error when the Morse characters differ in only 2 bits.

Therefore, it appears that not all Morse characters can be narrowed down to a few possibilities after suffering an error. Is there some way of measuring or evaluating the ability of a character to do this? If you look at **fig. 5** and **fig. 6** you see that the "good" characters had a fair number of bits which differed from the bits in the corresponding positions in the other characters. The "bad" characters such as H have the fewest number of corresponding bits which are different. This is logical, for if a large number of bits differ, then a single error in any position cannot overcome the overall difference. If only a few bits differ, then an error is more likely to cause a change that cannot be resolved. It is this difference which allows us to narrow down the number of possibilities as to what the character actually was before the error. The greater the number of differing bits, the more distinctive the character is.

Just how much difference is necessary to survive the effect of a single error? Since a single error can reduce the difference by only a single bit, you might think that one additional bit-difference would suffice to identify the character. Unfortunately, the one error could just have easily hit the *other* character on the *other* difference bit to yield the same mangled character — you would never know the difference. Perhaps an example would clarify this. The characters A and N differ in two bits, 2 and 4. If you receive an unknown character as shown in the center of **fig. 7** you have no way of knowing whether it is an A with an error in the second bit, or an N with an error in the fourth bit. Therefore it is actually necessary for the Morse character to have differences in at least *three* bits to be sure of correcting a single error.

This provides the basis for a method of measuring the ability of a Morse character to resist an error. It must differ from other characters in at least three corresponding bits. Let's go through the Morse alphabet and pick out the error-resisting characters. We must compare each character only with those of equal length, since it has been assumed for now that character spacing is error-free.



1. The three-unit spacers appear intact; this defines character length and makes it easier to decypher the garbled message
2. Uniess there is more tnan one error in the first character, this can only be a K or a G. Since K is not used in ERMA it must be a G.
3. The second letter is a perfect Z, but Z is not an ERMA character. It might be a P with four errors, or an O with 1 error. The chances of 4 errors oer character are much less than 1 error per charactec, so it must be an O.
4. The third letter could be an L or a V with two errors, but since neither of these is an ERMA character, this can only be a W.

fig. 8. Short message showing the value of the Error Reducing Morse Alphabet (ERMA). The signal-to-noise ratio is so poor the copy averages 1 bit error per character. The correct copy is GOW.

It turns out that there is a unique alphabet of eleven Morse characters which are immune to a single bit-error. This Error Reducing Morse Alphabet (ERMA) is as follows:

E T S M R G W P O J $\overline{\text{J}}$*

Does ERMA really work? Let's try a simple example and see. Assume the copy shown in **fig. 8** has been received badly garbled but we know it consists only of ERMA letters. The signal-to-noise ratio is so poor we are averaging one bit-error per character — "no-copy" if the conventional alphabet were being used.

By using ERMA it has been possible to recover copy that would otherwise have been lost. How is it possible to do this at such a poor signal-to-noise ratio? The ERMA letters have a greater redundancy than other Morse letters, so they convey character identification in more than one way. If part of the character is mangled, the *remaining* part is still *sufficient* to identify the character. The net result is that you can now tolerate a much lower signal-to-noise ratio and poorer bit-error rate but still maintain a useable character error rate; and CER is the basic measure of communications ability.

How many dB can be picked up this way? This is not a fair question because you cannot directly equate CER and signal-to-noise ratio. Rephrase the question: "For the same CER, what is the difference in signal-to-noise needed with ERMA and the signal-to-noise needed with a conventional alphabet?" For a CER of about 1 in 8, ERMA can tolerate a BER close to 10^{-1}. A conventional alphabet would require a BER near 10 ?. Checking back to rhe curve of **fig.** 3, this corresponds to an advantage of about 12 dB.

*$\overline{\text{J}}$ (— — — •) is a perfectly good Morse character, although it's not used in English.

errors in character spacing

Now, what about that matter of the spacing between characters? I have been putting this off so far by always assuming that the error falls only within the characters and never within the space separating them. We should be so lucky! An error which falls in the center of a standard 3-bit space would neatly link two characters, making it impossible to sort out. Two characters would be lost by just one error!

This problem is greatly relieved, however, by the simple expedient of using an even number of bits between characters rather than the more traditional odd number of three bits. Four bits for spaces is probably the optimum value. Two bits would run a greater risk of linking errors, and six bits is probably more than necessary.

The use of an even number gives a significant advantage by a rather subtle means. The bit stream for each character is in effect *slipped* in phase relative to the previous character. This is best shown in **fig. 9** where a series of Hs and Ss are drawn, first with 3-bit spacing and then with 4-bit spacing. Note that with the 4-bit spacing the phase slip causes the bits of one character to occur in even-number positions and

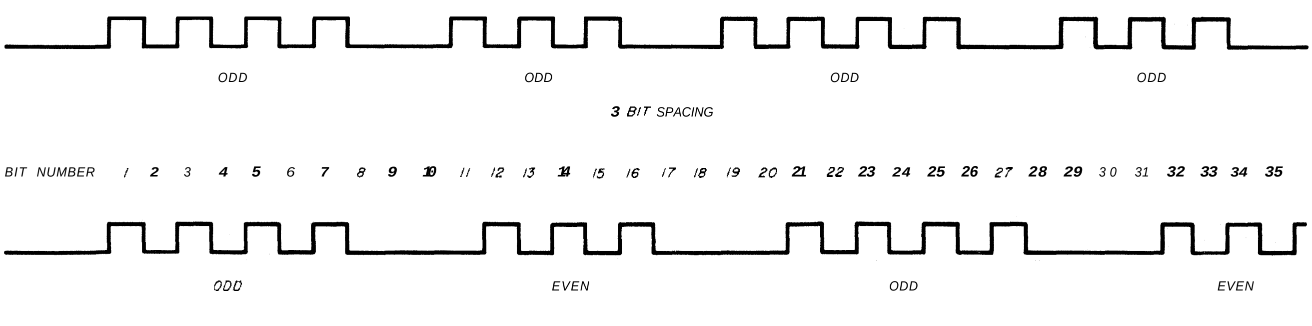


fig. 9. Comparison of 3- and 4-bit spacing, showing the odd-even "phase slip" that occurs between characters when 4-bit spacing is used.

those of the next character to fill odd-number positions, etc. This is a big help in sorting out characters during heavy error conditions. On the other hand, the 3-bit spacing makes this impossible.

Let's now give ERMA a real baptism of fire by letting the errors fall where they may. **Fig. 10** shows a set of ERMA letters, 132 bits in total length. To simulate a BER of 1 in 11, a total of 12 errors have been inserted at randomly determined positions. Character spacing is 4-bits throughout. (Before you tackle this you may want to look through the **appendix** which gives some hints on spotting errors.)

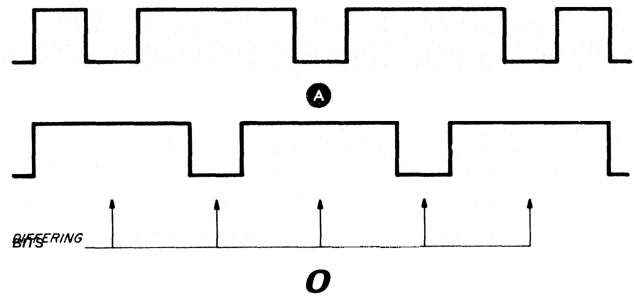


fig. **11.** Differing bits for the ERMA letters P and O.

using ERMA

Having struggled along this far, you are entitled to ask an obvious question; "How in the name of Samuel F. B. Morse are we supposed to actually communicate using this alphabet of only 11 letters?" If your call is W6PO you might have a chance, but if you have a combination like W4LTU you would be out of luck.

There is nothing sacred about any alphabet, including ERMA; it's just a set of symbols that stand for something else. Let's treat the ERMA alphabet as

bits	
1-11	No problem — clean 4-bit spacers with a T in between.
8-18	The second spacer is only 2 bits long so has an error. It's possible that bit 16 is in error, but then the letter would be an I (not ERMA) or another T with an additional error (2 total). More likely there is a single error in bit 19 and the letter is S.
17-42	This is badly garbled; there is obviously more than one letter and several errors. To sort it out start at bit 21 and work carefully several bits at a time.
21-27	Bits 21 to 23 could be a T, but it's followed by a 2-bit spacer. Perhaps bits 26 and 27 are errors (to give 4-bit spacing) but in that case bit 28 must also be an error if the next character starts there; requires 3 errors to justify the letter T which is unlikely. If bit 25 is an error, the letter is M and no other assumptions are required. (Note that the letter is not likely to be the letter G as the following bits are "out of phase," indicating a new letter has started.)
28-31	The spacer following is only 2 bits, indicating an error. If bits 21 to 27 are the letter M, then the error must be at bit 30 for a 4-bit spacer.
32-38	This is a perfect H but is not used in ERMA. It could be an M with 2 errors, or an R with 1 error. Following the rule of assuming the least number of errors, it must be an R.
39-55	There's obviously a problem here because of the 2-bit spacer in the middle. If it's assumed there are two characters with a 4-bit space, however, there are even more inconsistencies; any efforts to extend the spacer to 4 bits results in 2-bit dashes or an 8-bit spacer, both impossible. This is the result of the out-of-phase effect which indicates there are not two characters, but one. Therefore, the 2-bit gap is not a mangled 4-bit spacer but a 2-bit gap within a single character. This makes it easy; bit 45 is in error and the character is G.
52-68	Perfect spacers and a perfect L, but L isn't used in ERMA so there is an error. The simple solution is that there is a single error in bit 63 and the correct letter is W
65-83	This may look alright at first glance, but note that the last dash is 4 bits long and the following spacer is 5 bits, both impossible. If a 4-bit spacer is assumed by correcting bit 83, some impossibilities are generated in the following character. Therefore, it is corrected by assuming bit 79 is an error; this gives a 5-bit dash at the end of the letter, but neatly breaks down to a 3-bit and a 1-bit if bit 78 is also an error. The character is a P with errors at both bit 78 and bit 79.
80-98	No errors here; it's a perfect ERMA letter O. Assume no errors if there is no evidence of them
95-114	This might be two letters, but if it's corrected to something like RT there is only a 3-bit spacer between the letters. There is no way to insert a 4-bit spacer without causing other problems. This indicates that all the bits in this group are in phase and it must be a single character. There is a single error at bit 107 and the letter is J.
112-132	There are obvious problems here. Both spacers have only 3 bits; also within the group is an 8-bit dash and a 2-bit dash. Remember that the previous character was a J, ending at bit 111. Therefore, the 4-bit spacer must extend through bit 115, which must be an error. This leaves a 7-bit dash (bits 116 through 122). This might be an R with two errors, or an M with one error (these are the only two ERMA possibilities). It is followed by a flawless 3-bit dash, however, which must be a part of the character. The 2-bit dash which follows must also be taken into account. This is resolved by working back from the start of the following character (not shown) to get a 4-bit spacer; this requires that bit 129 is in error. Then if it falls into place — it's the lengthy letter J with errors at bits 115, 119, and 129.

fig. **10. A** message sent with **the ERMA** characters, total length of **132** bits. The bit-error rate is **1** in **11** (error positions randomly chosen). Four-bit character spacing is used throughout. The analysis shows that ERMA provides solid copy at a bit-error rate which would have given about 30 per cent copy with the conventional Morse alphabet.

just that, a set of symbols which can be used to spell out the normal alphabet. To do this simply arrange a 5x6 matrix with two ERMA letters serving to define each letter in the normal alphabet."

	G	W	P	O	J	$\overline{J}$
R	A	B	C	D	E	F
M	G	H	I	J	K	L
S	M	N	O	P	Q	R
T	S	T	U	V	W	X
E	Y	Z	1	2	3	4

For example, if you wanted to send the letter W, you would send the ERMA pair: TJ. This technique slows down the rate of communication, but remember that one of the assumptions made earlier was that you were willing to slow down if you could improve your weak-signal capability.

There is yet another scheme for using the ERMA advantage that is even simpler. Still thinking in terms of symbols, why not use only two ERMA letters, one representing a dot and the other a dash? This reduces speed even further, but if that bothers you, send the ERMA letters at a faster rate; you still come out ahead.

This last scheme leads to another point concerning ERMA. If you plan to use only two of the eleven letters, is there any best choice? You might assume the shortest characters are the best because they offer the least speed reduction. However, not all ERMA letters were created equal. In **fig. 10,** where ERMA was given its baptism of fire with noise, sharp-eyed readers may have noticed that certain ERMA letters were easier to salvage than others. A ponderous letter like J, for example, has enough redundancy that it can actually withstand two errors anywhere within the character and still be recognizable. On the other hand, the letter E survives only if its spacers are reasonably intact.

What governs this degree of error resistance? Again, it is determined by the number of bits that are different from the corresponding spots in other letters. Earlier I showed that a difference in 3 bits is necessary to survive a single error. It turns out that a difference in 5 bits gives full immunity to two errors. **Fig. 11** shows this in comparing the letter P and the letter O. Note that in five positions the bits are different in the two letters. This guarantees that no matter where the two errors strike you can still positively identify the letter.

Extending this idea, a difference in 7 bits is needed

'If you insist on having a full set of numbers available, you can use ten ERMA letters to form a three-dimensional matrix, 3x3x4, to give you a full **36** letters and numbers.

†Minimum, depends on spacers and adjacent letters. For other ERMA letters, values given are for the letter alone.

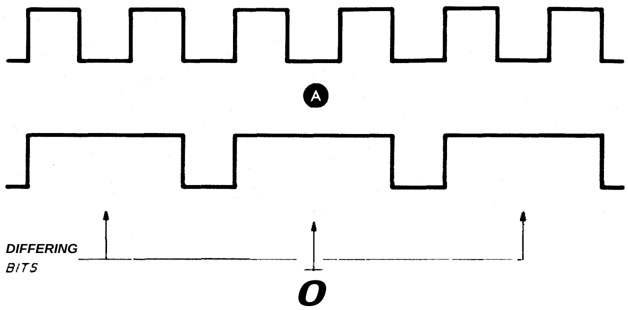


fig. 12. Differing bits for a simple repetition of dots (A) and dashes (B).

to correct three errors; a difference in 6 bits corrects two errors and in addition will detect a third error but not tell you where it is! The following tabulation clearly shows the advantage of using the longer ERMA letters:

letter	number of different bits	number errors it can withstand
E,T,S	3†	1
M,R	**3**	1
G,W	**4**	1
P,O	5	2
J,J	6	2

Why not simply use repeats? Well, if you are going to use two high-redundancy ERMA letters, such as P and O, to stand for dots and dashes in sending a conventional alphabet, there is another obvious question. If you take the time to send the entire letter P when you want to convey a dot, then why not just use that same amount of time to send a string of dots? Wouldn't this simple repetition of what you actually want to send be even better? Amateurs have been making good use of repeats for many years; isn't it just as good as this ERMA routine?

Let's take a look at both methods and compare them. Remember that the ability of a Morse character to survive errors depends on the number of bits in which it differs from other characters. The more bits that are different, the more errors **it** can survive and still be read correctly; the ERMA letters P and O differ in 5 bits so they can survive two errors and still be readable.

Now let's try the method of simple repetition. **Fig. 12** shows a string of dots of the same total duration as the letter P or O. Below the dots is a string of dashes; you must be able to tell the difference even with errors present. Note that the string of dots and the string of dashes differ in only 3 *bits,* so can survive only a *single* error. Thus the ERMA letters P and O show a greater error immunity than simple repetition of the same duration! This advantage also holds for other length Morse characters. The ERMA letters M and R can survive a single error, but a similar length of repeated dots and dashes cannot correct any errors. For longer lengths, the advantage of ERMA is even greater.

summary

Let's try to quickly recap what I've covered in this article; those of you who have stuck with it this far deserve every possible break!

1. Narrowing bandwidth helps, but there are limits.

2. Errors appear in individual bits — Bit-Error Rate or BER.

3. Each error can wreck an entire character — Character-Error Rate or CER.

4. CER limits ability to communicate; attack it, not signal-to-noise ratio.

5. Certain characters are more resistant to errors than others.

6. Evaluate error resistance and select only the best letters — ERMA.

7. Use ERMA to convey the conventional alphabet.

8. ERMA is better than simple repetition.

ERMA and similar techniques can allow amateurs to greatly improve weak-signal capability without the burden of staggering increases in hardware costs. These techniques can be either applied to existing bandwidths, or added to optimized bandwidths. As long as the detected signal is squared up to binary form, the method can be applied anywhere. Remember those immortal words, "To err is human, to correct is divine."

appendix

Following are some guidelines for correcting errors in ERMA copy. They are not all inclusive, but are designed to provide a starting point.

1. If several possibilities exist, choose the one which assumes the least number of errors.
2. A single error cannot change an ERMA letter into another ERMA letter.
3. A single error cannot divide an ERMA letter into two ERMA letters.
4. A single error can only distort an ERMA letter or change it into a non-ERMA letter.

error indicators

The following cannot exist in normal copy so are positive indicators that one or more errors are present, either within the given sequence or in a directly adjacent bit.

2-bit key-down	2-bit key-up
4-bit key-down	3-bit key-up
5-bit key-down	5-bit key-up
6-bit key-down	6-bit key-up
7-bit key-down	7-bit key-up

ham radio

pi network design

Pi networks are used extensively in amateur rf power amplifiers — here's the recipe for a pi

Pi networks have been used extensively for years, primarily for coupling final amplifiers to antennas. Useful for impedance matching, pi networks can also be used with mobile antennas and at receiver inputs. Since the pi network is a lowpass filter, it can reduce TVI caused by high-frequency transmitters. Component values make all the difference in application so the ingredients should be selected with care.

Fig. 1 shows the general circuit and component value equations for the pi network. The equations differ slightly from the usual methods but will provide the desired match. Allowance of parallel capacitance at each end of the network will give direct results and accommodate strays or stray-plus-circuit capacitances. Values of C, or C, must be finite and depend on the application; stray capacitance is always present.

The Q in the equations is a *design parameter, not* the component Q. It's value selection will determine higher frequency rejection, the resistive or *real* part of impedance presented to a source, as well as component values.

Fig. 1 gives two choices in calculation, T (a temporary value) and R_6. T *must* be positive or at least zero for solution. Results giving a negative T require a change either in end resistance or Q, and possibly both; increasing an end resistance is preferable and ways will be shown later on how to handle that situation. The value of R_6 is dependent on which end resistance is highest. It will be negative if R, is highest.

Reactances X_c and X_d must *both* be negative for a solution. If a positive value results, parallel end capacitance may have to be changed. In most cases a positive value comes from calculation error, so use care in handling terms. Some sources may present a parallel end inductance; this must be shunted with a fixed capacitor to cancel it at the operating frequency. Handling this situation is discussed later.

design example

Suppose a pi network is required for a 40-meter rf power amplifier. The amplifier has a tube type final; the center frequency is chosen as 7.15 MHz with design Q of 5. Initially

$R_o = 5000$ *ohms* (this will be the source end)
$C_o = 8$ *pF* (tube and socket stray capacitance)
R, $= 50$ *ohms* (load end)
$C_a = 3$ *pF* (stray capacitance in connector)
$X_o = -2782.43$ *ohms*
X, $= -7419.81$ *ohms*

Initial calculations result in $R_5 = 4950$ but $T = -18.25 \times 10^6$ so a solution cannot be done directly. The source end resistance is changed to 500 ohms but C, is kept the same. This gives

$R_5 = 450$ ohms	
$R_4 = 650$ ohms	
$R_6 = -1850$ ohms	
$X_L = 139.655$	$L = 3.1806\ \mu H$
$X_d = -127.181$	$C_d = 175.02\ pF$
$X_c = -56.6797$	$C_c = 392.72\ pF$

The design Q change was not done since this has other effects which are shown later. An R_o of 500 ohms might be alright for a transistor final but it won't fit the tube circuit. Fortunately, there's an alternative.

cascading and the virtual resistance

Two networks can be cascaded so the load end of one network becomes the source end of the following network. The latter has been calculated so we can try for the first with

$R_o' = 5000$ *ohms*
$R_a' = 500$ *ohms* (to match the example given)

By Leonard H. Anderson, 10048 Lanark Street, Sun Valley, California 91352

The same values of X, and X_a can be used again, assuming the same stray capacitance; it is always best to stay on the high side for strays. The new network with same frequency and design Q has values of

$R_5' = 4500$	$X_c' = -705.030$
$R_4' = 6500$	$L' = 31.086\ \mu H$
$R_6' = -18.5 \times 10^3$	$C_d' = 10.302\ pF$
$X_L' = 1396.55$	$C_c' = 31.572\ pF$
$X_d' = -2160.65$	

The manner of cascading is shown in **fig. 2.** The center resistance of **fig. 2A** is termed the *virtual resistance.* It's value is $R_a' = R_o$ but it does not appear physically in the circuit; virtual resistance is a mathematical technique which allows cascading of networks and must be the same at each network's design frequency.

Parallel capacitances at the midpoint can add. This also includes any assumed strays. Total C_m for both networks is 217.59 pF; you could assume a stray of 3 pF at this point since it is physically only one capacitor. Total network response calculations require the sum of all capacitors in parallel, including strays.

what happens in tuning?

Most pi network designs have a fixed inductor or at least a fixed value per band. The two capacitors nearest the antenna are made variable to accommodate impedance differences at the antenna. Each pi network design has different limits of equivalent parallel resistance that it will match. To find the resistance limits with a fixed inductor, the inductance equation can be solved for Q

$$Q = \frac{R_o + R_a + 2\sqrt{R_o R_a - X_L^2}}{X_L}$$

This presents two unknowns: Q and R,. Minimum R_a occurs when the square-root term reduces to zero. This also reduces the minimum Qequation to

$$Q_{min} = \frac{X_L^2 + R_o^2}{X_L R_o}$$

This brings up an interesting point when the inductor is held at a fixed value: matching to a lower load resistance will result in a lower Q than originally planned! Also, matching to a higher load resistance results in a higher Q. This case is found by letting $R_a = nR_o$ so that

$$Q_{high} = \frac{R_o(n+1) + 2\sqrt{nR_o^2 - X_L^2}}{X_L}$$

where n is an arbitrary fraction giving practical R,, but must be less than unity.

By now it is apparent that minimum load resistance is the critical factor in pi network design. It can be found by solving the first Q equation's square-root term group to yield

$$R_a(min) = X_L^2/R_o$$

Taking the first network example (500 to 50 ohms) with a fixed X_L and solving for minimum load resistance:

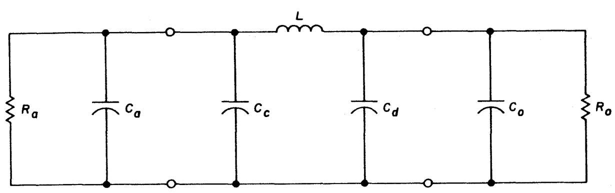
$$R_a(min) = 39.0070 \text{ ohms}$$
$$Q_{min} = 3.85956$$

Capacitor values are then found by recalculating everything except inductive reactance:

$R_5 = 460.993$	$X_c = -142.334$
$R_4 = 279.309$	$C_d = 151.30\ pF$
$R_6 = -1650.47$	$C_c = 156.39\ pF$
$X_d = -147.035$	

As a shortcut, the minimum R, case will result in total end capacitive reactance equal to the inductive reactance. The high end resistance, letting $n = 0.5$, gives

$R_a(high) = 250$ ohms	$X_d = -87.3164$
$Q_{high} = 10.0219$	$X_c = -61.2418$
$R_5 = 250$	$C_d = 254.93\ pF$
$R_4 = 3534.44$	$C_c = 363.47\ pF$
$R_6 = -1476.50$	



NETWORK

$X_a = -1/(2\pi F C_a)$ $\qquad$ $X_o = -1/(2\pi F C_o)$

F = CENTER FREQUENCY $\qquad$ $R_5 = R_o - R_a$

$T = R_o R_a Q^2 - R_5^2$ $\qquad$ SEE TEXT IF T IS NEGATIVE

$R_4 = \sqrt{T}$ $\qquad$ $R_6 = R_4 - R_o a$

$$X_L = \frac{Q R_o R_5^2}{R_5^2 + R_6^2}$$

CHOOSE

R_6 NEGATIVE

$$X_d = \frac{X_o R_o R_5}{X_o R_6 - R_o R_5}$$

$$X_c = -\left[\frac{R_a X_a R_5}{X_a(R_6 + R_5 Q) + R_5 R_a}\right]$$

R_6 POSITIVE

$$X_d = -\left[\frac{X_o R_o R_5}{X_o R_6 + R_o R_5}\right]$$

$$X_c = \frac{R_a X_a R_5}{X_a(R_6 - R_5 Q) - R_5 R_a}$$

fig. 1. Pi network and its design equations. The equations are from reference 4.

High resistance capacitor values *seem* wrong since they are higher than the original 50-ohm values. This is a result of keeping the inductor at a fixed value; each network will match the 500-ohm source at 7.15 MHz. Also, and important for harmonic suppression, the Q will change.

what to do about load reactance

No antenna is perfect so you can expect the reactance to change as well as the resistance. Capacitor C_c will compensate for antenna parallel reactance. If the antenna (or transmission line) reactance is capacitive, C_c is reduced; if it is inductive, C_c is increased. For practical variable capacitor values, the change should be around 50 per cent of C_c.

In this example the C_c range will be about 78 to 590 pF or 7.5:1, maximum to minimum. This is practical but we have found a definite lack of matching ability for the low resistance case. Maximum allowable vswr for that condition will be only 1.28:1 and an additional matching network may have to be added. Capacitor C_c alone will take care of parallel reactance but both capacitors must be changed for parallel resistance and reactance other than 50 ohms.

thinking in terms of admittance and the vswr circle

Antenna impedance can be represented by series equivalents of R $+jX$ or parallel equivalents G$+jB$ in admittance form. The admittance form is a bit easier to think of since the pi network has a load-end shunt variable.

The Smith chart is invaluable for seeing what happens with a network when the load changes. **Fig.** 3 is a normalized Smith chart showing the admittance matching range of the first example (dotted line). Normalization merely means that all conductance and susceptance values have been divided by 0.02 mho (50 ohms) since we are really concerned about the ratio of antenna admittance to a target value. Target admittance (or impedance) is usually the transmission line as well as design goal of the antenna.

Tranmission line length is usually arbitrary. Vswr is a ratio of actual admittance magnitude to a desired value or target. Neglecting line losses and assuming arbitrary line lengths, the antenna admittance will appear somewhere on a circle whose center is that of the chart with a radius determined by distance from center to antenna admittance. This is the vswr circle.

If you don't have an RX noise bridge but do have a fairly accurate vswr indicator, there is a shortcut to finding the radius; use a compass or divider and draw it using the normalized R or G line numbers higher than 1.0 for the vswr radius.

The matching area of the example is rather lopsided. You could match any vswr up to 5:1 by simply changing the transmission line length. This would move the matching area to wherever desired. The problem is that you have to measure the line end admittance to find out which way to move.

A better choice for a final amplifier network is to try for a symmetrical matching area on the chart. A 2:1 vswr is a reasonable design goal. This fits many antennas and line lengths don't have to be juggled.

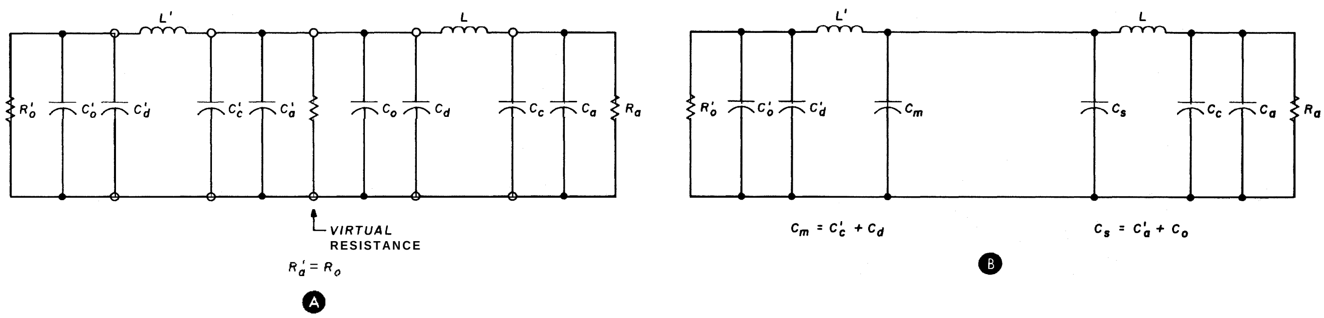


fig. 2. Two pi networks in cascade are shown at (A). Combined pi networks are shown in (B)

new component with an old circuit

A rule of thumb is that low R_o:R_a ratios yield lower possible $R_a(min)$ values with a fixed inductor. A variable inductor could be used, of course, but this adds another tuning element and possible mechanical problems. A pi network for a vacuum tube usually needs two networks in cascade so a low R_o:R_a ratio can be achieved by changing the virtual resistance, but there is another way.

A 4:1 toroidal balun with broadband characteristics will reflect an equal ratio of both conductance and susceptance. In addition to transforming impedance from 50 to 200 ohms, a well-insulated transformer is useful for draining off static charges during thunderstorms.

The modified cascaded network using the balun is shown in **fig. 4.** The admittance matching area of this circuit is shown by the solid outline in **fig.** 3 and is just short of meeting the 2:1 vswr goal. Calculation is done with a load resistance of 200 ohms since the modified network sees this through the balun.

Target values at 7.15 MHz, design $Q = \mathbf{5},\ R, = \mathbf{500},$ and the same stray capacitance as before are

$R_5 = 300$	$X_c = -110.227$
$R_4 = 1552.42$	$L = 5.0697\ \mu H$
$R_6 = -947.583$	$C_d = 132.62\ pF$
$X_L = 227.753$	$C_c = 201.94\ pF$
$X_d = -167.847$	

Holding the inductor fixed and calculating the minimum R_a gives

$R_a(min) = \mathbf{103.743}$ (25.9357 at the line)
$Q_{min} = \mathbf{2.65087}$
$C_d = \mathbf{89.735}\ pF$
$C_c = \mathbf{94.735}\ pF$

Since minimum R_a is very close to a 2:1 vswr, n is set to 0.8 for an $R_a(high)$ of 400 ohms or 100 ohms on the other side of the balun. High resistance values are

$$Q_{high} = 7.33141$$
$$C_d = 164.97\ pF$$
$$C_c = 188.77\ pF$$

Calculation at equivalent line resistances of 30, 40, and 75 ohms shows that C_c and C_d will be maximum at 25.9 ohms. The C_d minimum occurs at 100 ohms but C_c minimum is close to the 50-ohm design target (it increases slightly going towards 100 ohms).

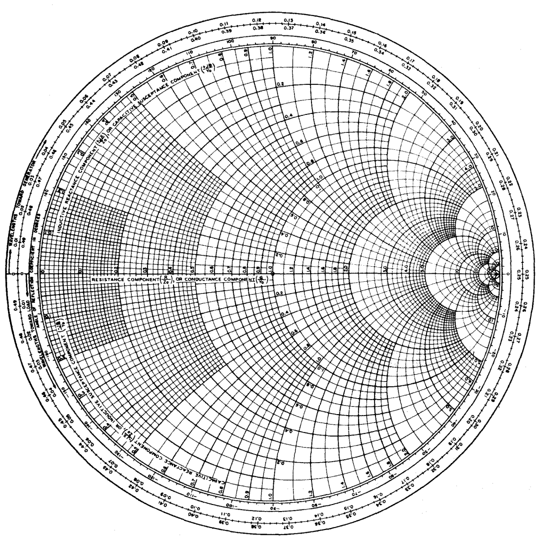


fig. 3. Admittance matching range of the pi network discussed the first example in the text (dashed lines). The solid lines show the modification of matching range available by using a balun.

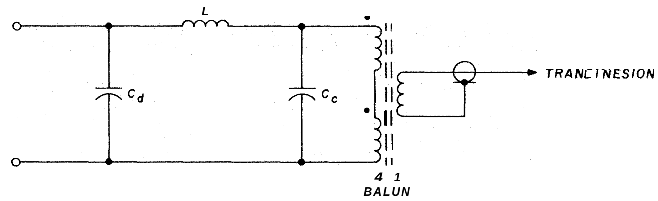


fig. 4. Basic arrangement of 4:1 balun between the transmission line and the output of the pi network.

Using a 50 per cent variation of C_c to compensate for variations in line susceptance requires a practical range of 55 to 342 pF, easily obtainable. C_d must cover 135 to 248 pF if used in a single network. The single network matching area is shown in **fig.** 3. Cascading networks allows a slight increase in area. If this is done, the virtual resistance must change in calculations but another penalty occurs: three variables are required for tuning.

Qand harmonic suppression

Harmonic suppression varies with design Q; and the design Qvaries with tuning when a fixed inductor is used. As a general idea of Q change with tuning, the first example's voltage response is shown in **fig. 5.** This set of curves must be considered on the basis that the data assumes *both* load and source resistances remain constant with frequency. In actual fact, both load and source admittance is frequency sensitive. Sensitivity varies with the final amplifier circuit and antenna when used in a transmitter. A general rule of thumb is that antenna admittance variation at harmonics will make little difference in suppression. An exception might occur if the feedpoint of the antenna is at a voltage node at a harmonic. Effect is still slight.

The major effect of the pi network at harmonics occurs at the source end. Regardless of circuit, the network appears as a single, slightly lossy capacitor to the amplifier. The impedance of the first example at 14.3 MHz (the second harmonic) is $1.14 - j79.5$ *ohms* looking into C and assuming a perfect load. Impedance at 21.45 MHz (third harmonic) is $\mathbf{0.08} - j45.1$ *ohms* under the same conditions. The amplifier must be able to work at the fundamental frequency yet be stable with a capacitive load at harmonics.

Source impedance at harmonics is generally low and depends on the particular circuit. This is a separate analysis problem with many variations. The network still looks capacitive at harmonics.

It is a good idea, regardless of the amplifier circuit, to use two sections in cascade. The source-end section can use a lower design Q but retain the virtual resistance for cascading. C_o' is tuned just once, leaving C_m and C_a variable for matching. This increases harmonic suppression capability without adding more variables.

The normalized resistance and reactance of the first circuit example is graphed in **fig. 6** and assumed to be looking into C_o. Curves are for design Qs of 3.86, 5, and 10. Other circuits will show a slight deviation but the general effect is the same.

A general rule is that the resistance and reactance change slowly near band center for stability. Design Qs lower than 5 are preferred. This is achievable by using the two-section configuration with a low-Q source end. Variations to the source due to tuning are reduced.

Individual component Q should be high to avoid losses; component Q of 100 or greater is preferred. This is easy with capacitors but inductors should be toroids or airwound. A 0.3 dB loss is insignificant at the other end whether running 10 watts or a kilowatt, but that loss with a kilowatt means local heating of about 67 watts.

receiver applications

All that was said about transmitter applications applies here. Many expensive receivers are mismatched — the so-called communications receivers with 3:1 tuning bands show wide input impedance variations.

An RX noise bridge can be used to check a receiver input. The noise source itself should be padded to avoid overload and saturation, particularly with solid-state receivers, because overloading the front end can change the impedance reflected back to the antenna terminals.

Old receivers with 300-ohm input options should have the pi network placed there. Variations will still be found so the impedance must be checked.

An advantage gained in the high-frequency bands is a reduction of interference from nearby CB rigs when the receiver is used at 15 meters and below. There is some advantage gained in attenuation below

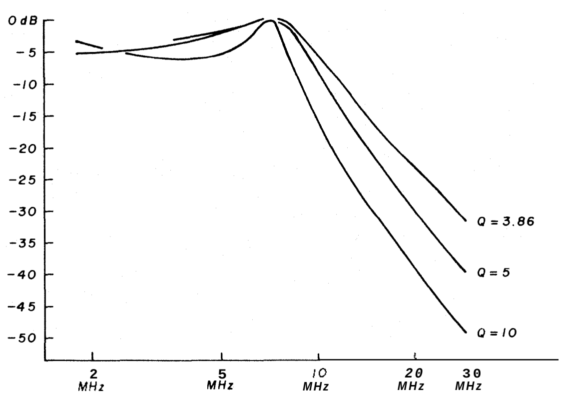


fig. 5. Voltage response of a pi network vs Q. Center frequency is at 7.15 MHz.

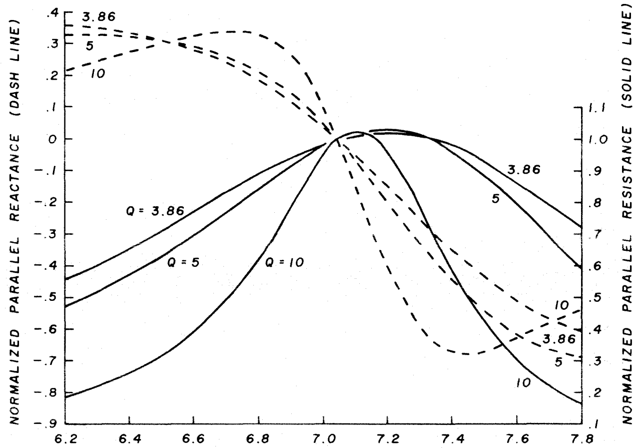


fig. 6. Normalized resistance and reactance vs Q around the design center frequency of the pi network (7.15 MHz in this case).

the center frequency due to mismatch loss. The input impedance to a pi network is reactive on either side of the center frequency, actual loss will depend on the antenna.

Mobile antennas in the high-frequency range can use the pi network in place of the common L-network. The same is true of short verticals. It is wise to use a wideband 4:1 balun with a bridge in measuring such antennas because of low resistances. Don't forget to divide the impedance by 4 (or multiply with admittances) when using the balun.

Interstage matching in any amplifier chain can also be done. Higher design Q can be used, but with caution: the reactance variation is greater and may result in instability. There are a dozen other 3-component matching networks available which can be found in the references. About half of these will be better for interstage applications.

Proper design and operation with any matching network depends on knowing the source and load characteristics. With amplifiers at either end, you should check to see what happens with reactive conditions out of band.

bibliography

1. Carson, Ralph S., High Frequency Amplifiers, John Wiley & Sons, New York, 1975.
2. Davis, Frank, "Matching Network Designs With Computer Solutions," Motorola Semiconductor Products Application Note AN-267, Motorola Semiconductor Corp., Phoenix, Arizona.
3. Hubbs, Robert A. and Doting, Frank A., "Improvements to the RX Noise Bridge." *ham* radio, February, 1977, page 10.
4. Richwell, Gunnar, "Design and Match RF Amplifiers," Electronic Design 10, May 10, 1969, page 106.
5. Smith, Phillip H., Electronic *Applications* of the Smith Chart, McGraw-Hill, New York, 1969.
6. ARRL Electronics Data Book, American Radio Relay League, Newington, Connecticut, 1976, Chapters 3, 4, and 5.

ham radio

transmitter matching networks

Complete data
for building networks
to match the
output impedance
of your transmitter
to the
input impedance
of a linear
power amplifier

Most transmitters are designed for 50-ohm output loads and the use of 50-ohm coax cable has become the standard for antenna systems used by amateurs. As the typical transmitter these days has 100 to 175 watts output, it is often used as an exciter to drive a linear amplifier to higher output power. These units normally are cathode-driven and are characterized by an input impedance that falls in the region of 20 to 200 ohms. Although in many cases the exciter can drive such an amplifier directly with satisfactory results, the use of a properly-terminated matching network can be beneficial in a variety of ways: It allows maximum energy transfer (most output), presents the best load to the exciter, minimizes harmonic radiation (TVI, etc.), and allows barefoot operation without retuning.

Perhaps other advantages will come to mind. Some exciters have only a 50-ohm output, and cannot be retuned for other impedances.

input impedance

The input impedance of linear amplifiers is rarely the same from one band to another. Some amplifiers are not operated at zero-bias and actually drive the grid through a passive resistor. These systems usually present about the same impedance from one band to another, of course, but are rarely 50 ohms to start with.

Formulas have been given to enable the calculation of the input impedance of a grounded-grid, cathode-driven amplifier. However, such formulas are all but worthless since they do not take the frequency into consideration. Measurements taken at the input of such amplifiers usually show a rather impressive variation from 10 to 80 meters, indicating that a formula would be quite misleading. These variations are caused by the manner in which the rf is isolated from the filament transformer (and hence the house wiring). Two methods are used to ac-

By Irvin M. **Hoff, W6FFC** (reprinted from the January, 1973, issue of *ham radio)* complish this: filament chokes, such as bifilar-wound coils, or low-capacitance filament transformers.

The best uniformity is normally obtained with the low-capacitance filament transformer, but such a transformer is not always available, and in any event would need to be mounted within a few inches of the tube base. This is not always convenient, so filament chokes are more commonly used. These chokes range from commercially-available units to home-made — the latter usually being two number-12 double-enameled wires wound simultaneously around a round ferrite rod until 11 turns (you would count 22 with the two wires) are on the rod. With proper bypassing these chokes allow the 60-Hz filament current to pass, but to not allow the high-frequency rf signal into the filament transformer.

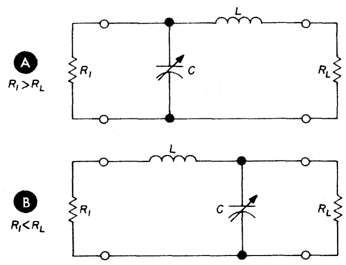


fig. 1. L-networks. The circuit in A is a step-down L-network; B shows a step-up L-network.

Factors which seem to contribute to variations in input impedance from band to band include the voltage on the final amplifier, the type of tube or tubes being used, the frequency involved, and the type of rf chokes used.

matching

I once had a Johnson Pacemaker 90-watt ssb transmitter. This unit could tune as high as 300 ohms on the output. I did not think any type of matching network to my linear was needed, but one day, while operating on 10 meters, I got a bad rf burn on my mouth when I came too close to the microphone. This led to an investigation of the input impedance, and I found on that particular transmitter it was only 15 ohms on 28 MHz; the Pacemaker could not handle this low impedance at all. A simple pi-network was used, and when incorporated for other bands, I found I not only increased the output power, but I could also then switch immediately from high power to barefoot, a distinct advantage over the previous system.

Various articles have been written regarding the use of networks between the exciter and the linear, and this is now standard practice for most commercial units. These usually have input networks incorporated into the design, and are often adjustable if you wish to optimize them for a specific part of the band. They are usually switched automatically as you change the band selector.

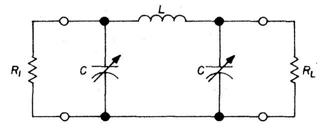


fig. 2. A typical pi-network. R_i is the input load, R_L the output load.

Such networks are usually made up of pi networks although a few use the more simple L-network. The pi network is usually preferred as greater control and uniformity are possible from band-to-band since the Q can be predetermined for consistent performance over a wider variety of impedances. The L-network is more simple, but it is also somewhat more difficult to adjust for optimum swr.

L-networks have been covered adequately in other texts,[1] so only an example will be shown here (see fig. 1). Although this is a very simple circuit, it has several minor disadvantages.

For one thing, in the L-network *Q* cannot be controlled, and is usually very low. Also, if the network is used for all hf amateur bands, the capacitor often has to be switched from one end of the coil to the other. Further, the L-network has very little exciter loading due to the low Q and it offers very little harmonic suppression.

A typical pi network is shown in **fig.** 2. It offers predictable performance because the Q may be preselected. It also offers a good load for exciter stability, and can easily be used for all hf amateur bands.

input impedance

The input impedance of the network may be determined by testing; use of formulas should be avoided

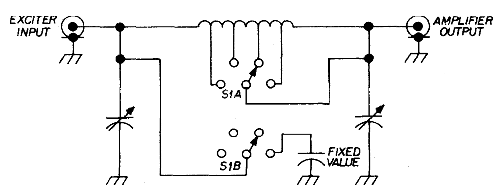
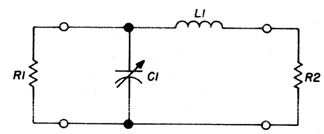
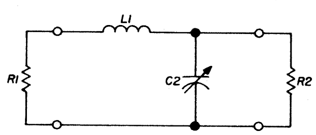


fig. 3. A typical pi-network for transmitter matching. The switch selects the proper tap for the various bands; the second switch section may be used to switch in parallel fixed values on the lower frequency bands.

table 1. L-network component values. Data is for matching a 50-ohm transmitter to a cathode-driven amplifier. The *Q* is set by the ratio of the input and output impedances and is shown for approximately the middle of each amateur radiotelephone band. The *Q* at the top of the band would be slightly less, at the bottom of the band it would be slightly greater.





R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
50	1.9	4188	2.09	--	10	2.0
50	3.8	2094	1.05	--	10	2.0
50	7.2	1105	0.55	--	10	2.0
50	14.2	560	0.28	--	10	2.0
50	21.2	375	0.19	--	10	2.0
50	28.5	279	0.14	--	10	2.0
50	1.9	3420	3.42	--	20	1.2
50	3.8	1710	1.71	--	20	1.2
50	7.2	902	0.90	--	20	1.2
50	14.2	458	0.46	--	20	1.2
50	21.2	306	0.31	--	20	1.2
50	28.5	228	0.23	--	20	1.2
50	1.9	3420	5.13	--	30	0.8
50	3.8	1710	2.56	--	30	0.8
50	7.2	902	1.35	--	30	0.8
50	14.2	458	0.69	--	30	0.8
50	21.2	306	0.46	--	30	0.8
50	28.5	228	0.34	--	30	0.8
50	1.9	4188	8.38	--	40	0.5
50	3.8	2094	4.19	--	40	0.5
50	7.2	1105	2.21	--	40	0.5
50	14.2	560	1.12	--	40	0.5
50	21.2	375	0.75	--	40	0.5
50	28.5	279	0.56	--	40	0.5
50	1.9	--	11.24	3746	60	0.4
50	3.8	--	5.62	1873	60	0.4
50	7.2	--	2.97	989	60	0.4
50	14.2	--	1.50	501	60	0.4
50	21.2	--	1.01	336	60	0.4
50	28.5	--	0.75	250	60	0.4
50	1.9	--	9.27	2649	70	0.6
50	3.8	--	4.64	1324	70	0.6
50	7.2	--	2.45	699	70	0.6
50	14.2	--	1.24	354	70	0.6
50	21.2	--	0.83	237	70	0.6
50	28.5	--	0.62	177	70	0.6
50	1.9	--	8.65	2163	80	0.8
50	3.8	--	4.33	1081	80	0.8
50	7.2	--	2.28	571	80	0.8
50	14.2	--	1.16	289	80	0.8
50	21.2	--	0.78	194	80	0.8
50	28.5	--	0.58	144	80	0.8
50	1.9	--	8.43	1873	90	0.9
50	3.8	--	4.21	937	90	0.9
50	7.2	--	2.22	494	90	0.9
50	14.2	--	1.13	251	90	0.9
50	21.2	--	0.76	168	90	0.9
50	28.5	--	0.56	125	90	0.9
50	1.9	--	8.38	1675	100	1.0
50	3.8	--	4.19	838	100	1.0
50	7.2	--	2.21	442	100	1.0
50	14.2	--	1.12	224	100	1.0
50	21.2	--	0.75	150	100	1.0
50	28.5	--	0.56	112	100	1.0
50	1.9	--	8.41	1529	110	1.1
50	3.8	--	4.21	765	110	1.1
50	7.2	--	2.22	404	110	1.1
50	14.2	--	1.13	205	110	1.1
50	21.2	--	0.75	137	110	1.1
50	28.5	--	0.56	102	110	1.1
50	1.9	--	8.50	1416	120	1.2
50	3.8	--	4.25	708	120	1.2
50	7.2	--	2.24	374	120	1.2
50	14.2	--	1.14	189	120	1.2
50	21.2	--	0.76	127	120	1.2
50	28.5	--	0.57	94	120	1.2
50	1.9	--	8.61	1324	130	1.3
50	3.8	--	4.30	662	130	1.3
50	7.2	--	2.27	350	130	1.3
50	14.2	--	1.15	177	130	1.3
50	21.2	--	0.77	119	130	1.3
50	28.5	--	0.57	88	130	1.3
50	1.9	--	8.74	1249	140	1.3
50	3.8	--	4.37	624	140	1.3
50	7.2	--	2.31	330	140	1.3
50	14.2	--	1.17	167	140	1.3
50	21.2	--	0.78	112	140	1.3
50	28.5	--	0.58	83	140	1.3
50	1.9	--	8.88	1185	150	1.4
50	3.8	--	4.44	592	150	1.4
50	7.2	--	2.34	313	150	1.4
50	14.2	--	1.19	159	150	1.4
50	21.2	--	0.80	106	150	1.4
50	28.5	--	0.59	79	150	1.4
50	1.9	--	9.04	1129	160	1.5
50	3.8	--	4.52	565	160	1.5
50	7.2	--	2.38	298	160	1.5
50	14.2	--	1.21	151	160	1.5
50	21.2	--	0.81	101	160	1.5
50	28.5	--	0.60	75	160	1.5
50	1.9	--	9.19	1081	170	1.5
50	3.8	--	4.60	541	170	1.5
50	7.2	--	2.43	285	170	1.5
50	14.2	--	1.23	145	170	1.5
50	21.2	--	0.82	97	170	1.5
50	28.5	--	0.61	72	170	1.5
50	1.9	--	9.35	1039	180	1.6
50	3.8	--	4.68	519	180	1.6
50	7.2	--	2.47	274	180	1.6
50	14.2	--	1.25	139	180	1.6
50	21.2	--	0.84	93	180	1.6
50	28.5	--	0.62	69	180	1.6
50	1.9	--	9.51	1001	190	1.7
50	3.8	--	4.76	501	190	1.7
50	7.2	--	2.51	264	190	1.7
50	14.2	--	1.27	134	190	1.7
50	21.2	--	0.85	90	190	1.7
50	28.5	--	0.63	67	190	1.7
50	1.9	--	9.67	967	200	1.7
50	3.8	--	4.84	484	200	1.7
50	7.2	--	2.55	255	200	1.7
50	14.2	--	1.29	129	200	1.7
50	21.2	--	0.87	87	200	1.7
50	28.5	--	0.64	64	200	1.7
50	1.9	--	9.83	937	210	1.8
50	3.8	--	4.92	468	210	1.8
50	7.2	--	2.59	247	210	1.8
50	14.2	--	1.32	125	210	1.8
50	21.2	--	0.88	84	210	1.8
50	28.5	--	0.66	62	210	1.8
50	1.9	--	9.99	909	220	1.8
50	3.8	--	5.00	454	220	1.8
50	7.2	--	2.64	240	220	1.8
50	14.2	--	1.34	122	220	1.8
50	21.2	--	0.90	81	220	1.8
50	28.5	--	0.67	61	220	1.8
50	1.9	--	10.15	883	230	1.9
50	3.8	--	5.08	441	230	1.9
50	7.2	--	2.68	233	230	1.9
50	14.2	--	1.36	118	230	1.9
50	21.2	--	0.91	79	230	1.9
50	28.5	--	0.68	59	230	1.9
50	1.9	--	10.31	859	240	1.9
50	3.8	--	5.16	430	240	1.9
50	7.2	--	2.72	227	240	1.9
50	14.2	--	1.38	115	240	1.9
50	21.2	--	0.92	77	240	1.9
50	28.5	--	0.69	57	240	1.9
50	1.9	--	10.47	838	250	2.0
50	3.8	--	5.24	419	250	2.0
50	7.2	--	2.76	221	250	2.0
50	14.2	--	1.40	112	250	2.0
50	21.2	--	0.94	75	250	2.0
50	28.5	--	0.70	56	250	2.0

because the calculations rarely approximate the observed results.

The easiest and quickest method of measuring input impedance would be to use a variable impedance bridge, such as the RX noise bridge.[2] The *ARRL Handbook* also contains an excellent rf impedance bridge that may be easily built. These rf impedance bridges are basically a small swr bridge with a variable leg in the bridge so you can match the load impedance.

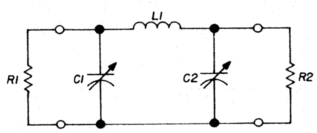
When making the impedance measurement the high voltage must be on the amplifier, and the meter hooked as close as possible to the place the network will be added.

There are probably no typical impedances, but as a general rule I have found that most amplifiers I tested fell in the neighborhood of 150 to 200 ohms on 80 meters, and around 15 to 30 ohms on 10 meters. The rest of the bands came somewhere in between. In many cases 20 meters offers a fairly decent match to 50 ohms with no network at all.

If the input impedance is measured directly at the filament of the power tube it will be considerably less than 50 ohms on ten meters, and considerably more than 50 ohms on 80. The data shown below is for my own 4-1000A linear with 6000 volts on the plate.

	impedance at tube base (ohms)	impedance at network (ohms)
80 meters	180	100
40 meters	155	60
20 meters	75	22
15 meters	50	40
10 meters	40	65

table 2. Pi-network component values. Data is for matching a 50-ohm transmitter to a cathode-driven amplifier. The *Q* has been chosen quite low to obtain broadband characteristics. The *Q* figure in the last column shows the worst-case condition at the bottom of the band using the inductance value shown.



R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
50	1.8	4909	1.85	7612	10	3.6
50	3.5	2600	0.94	4153	10	3.8
50	7.0	1179	0.49	1678	10	3.3
50	14.0	579	0.25	801	10	3.2
50	21.0	384	0.16	528	10	3.2
50	28.0	300	0.12	439	10	3.4
50	1.8	3828	2.56	4992	20	3.3
50	3.5	2028	1.30	2679	20	3.4
50	7.0	920	0.68	1157	20	3.0
50	14.0	451	0.34	562	20	3.0
50	21.0	300	0.23	372	20	3.0
50	28.0	234	0.17	298	20	3.1
50	1.8	3403	3.08	3978	30	3.3
50	3.5	1802	1.56	2120	30	3.4
50	7.0	817	0.82	940	30	3.0
50	14.0	401	0.41	459	30	3.0
50	21.0	266	0.27	305	30	3.0
50	28.0	208	0.20	241	30	3.1
50	1.8	3090	3.54	3310	40	3.2
50	3.5	1636	1.79	1757	40	3.3
50	7.0	742	0.94	790	40	3.0
50	14.0	364	0.47	387	40	3.0
50	21.0	242	0.32	257	40	3.0
50	28.0	189	0.23	202	40	3.1
50	1.8	2872	3.95	2872	50	3.2
50	3.5	1521	2.00	1521	50	3.3
50	7.0	690	1.04	690	50	3.0
50	14.0	339	0.53	339	50	3.0
50	21.0	225	0.35	225	50	3.0
50	28.0	176	0.26	176	50	3.1
50	1.8	2689	4.33	2542	60	3.2
50	3.5	1427	2.20	1346	60	3.3
50	7.0	643	1.15	611	60	3.0
50	14.0	315	0.58	300	60	3.0
50	21.0	209	0.39	199	60	3.0
50	28.0	157	0.29	149	60	3.0
50	1.8	2559	4.68	2306	70	3.3
50	3.5	1361	2.37	1221	70	3.4
50	7.0	609	1.24	554	70	3.0
50	14.0	298	0.62	272	70	3.0
50	21.0	198	0.42	181	70	3.0
50	28.0	148	0.31	135	70	3.0
50	1.8	2432	5.01	2104	80	3.3
50	3.5	1295	2.54	1114	80	3.4
50	7.0	576	1.33	506	80	3.0
50	14.0	281	0.67	248	80	3.0
50	21.0	187	0.45	165	80	3.0
50	28.0	140	0.34	124	80	3.0
50	1.8	2319	5.34	1939	90	3.3
50	3.5	1237	2.71	1027	90	3.4
50	7.0	546	1.42	466	90	3.0
50	14.0	267	0.71	229	90	3.0
50	21.0	177	0.48	152	90	3.0
50	28.0	133	0.36	114	90	3.0
50	1.8	2216	5.66	1799	100	3.3
50	3.5	1184	2.87	953	100	3.4
50	7.0	520	1.50	432	100	3.0
50	14.0	253	0.76	212	100	3.0
50	21.0	168	0.50	141	100	3.0
50	28.0	126	0.38	106	100	3.0
50	1.8	2120	5.96	1679	110	3.3
50	3.5	1135	3.02	889	110	3.4
50	7.0	495	1.58	403	110	3.0
50	14.0	241	0.80	198	110	3.0
50	21.0	160	0.53	131	110	3.0
50	28.0	120	0.40	99	110	3.0
50	1.8	2036	6.26	1577	120	3.3
50	3.5	1092	3.17	835	120	3.4
50	7.0	473	1.66	379	120	3.0
50	14.0	230	0.84	186	120	3.0
50	21.0	152	0.56	123	120	3.0
50	28.0	114	0.42	93	120	3.0
50	1.8	1959	6.54	1488	130	3.3
50	3.5	1052	3.32	788	130	3.4
50	7.0	453	1.74	358	130	3.0
50	14.0	220	0.87	175	130	3.0
50	21.0	146	0.58	117	130	3.0
50	28.0	109	0.44	87	130	3.0
50	1.8	1887	6.82	1410	140	3.3
50	3.5	1015	3.46	747	140	3.4
50	7.0	434	1.81	339	140	3.0
50	14.0	210	0.91	166	140	3.0
50	21.0	139	0.61	110	140	3.0
50	28.0	104	0.46	83	140	3.0
50	1.8	1821	7.09	1342	150	3.3
50	3.5	981	3.60	710	150	3.4
50	7.0	417	1.88	322	150	3.0
50	14.0	201	0.95	158	150	3.0
50	21.0	133	0.63	105	150	3.0
50	28.0	100	0.47	79	150	3.0
50	1.8	1757	7.36	1279	160	3.3
50	3.5	949	3.73	678	160	3.4
50	7.0	400	1.95	307	160	3.0
50	14.0	193	0.98	151	160	3.0
50	21.0	128	0.66	100	160	3.0
50	28.0	96	0.49	75	160	3.0
50	1.8	1697	7.62	1223	170	3.3
50	3.5	918	3.86	648	170	3.4
50	7.0	384	2.02	294	170	3.0
50	14.0	185	1.02	144	170	3.0
50	21.0	122	0.68	96	170	3.0
50	28.0	92	0.51	72	170	3.0
50	1.8	1641	7.88	1173	180	3.3
50	3.5	890	3.99	621	180	3.4
50	7.0	369	2.08	282	180	3.0
50	14.0	178	1.05	138	180	3.0
50	21.0	117	0.70	92	180	3.0
50	28.0	88	0.53	69	180	3.0
50	1.8	1589	8.12	1127	190	3.3
50	3.5	863	4.12	597	190	3.4
50	7.0	355	2.15	271	190	3.0
50	14.0	170	1.08	133	190	3.0
50	21.0	113	0.72	88	190	3.0
50	28.0	84	0.54	66	190	3.0
50	1.8	1538	8.37	1085	200	3.3
50	3.5	837	4.24	575	200	3.4
50	7.0	342	2.21	261	200	3.0
50	14.0	163	1.11	128	200	3.0
50	21.0	108	0.74	85	200	3.0
50	28.0	81	0.56	64	200	3.0
50	1.8	1512	8.58	1054	210	3.4
50	3.5	824	4.35	558	210	3.5
50	7.0	335	2.27	253	210	3.1
50	14.0	160	1.14	124	210	3.0
50	21.0	105	0.76	83	210	3.0
50	28.0	79	0.57	62	210	3.0
50	1.8	1515	8.75	1035	220	3.4
50	3.5	826	4.43	548	220	3.6
50	7.0	335	2.32	249	220	3.1
50	14.0	160	1.17	122	220	3.1
50	21.0	106	0.78	81	220	3.0
50	28.0	79	0.58	61	220	3.0
50	1.8	1518	8.91	1016	230	3.5
50	3.5	828	4.52	538	230	3.6
50	7.0	335	2.36	244	230	3.2
50	14.0	160	1.19	120	230	3.1
50	21.0	106	0.79	80	230	3.1
50	28.0	79	0.60	60	230	3.1
50	1.8	1520	9.08	999	240	3.6
50	3.5	829	4.60	529	240	3.7
50	7.0	335	2.41	240	240	3.3
50	14.0	160	1.21	118	240	3.2
50	21.0	106	0.81	78	240	3.2
50	28.0	79	0.61	59	240	3.2

The amplifier uses a low-capacitance filament transformer. The first column of figures is the impedance measured right at the tube base; the second column shows the impedance at the end of a 6-foot (1.8m) length of RG-58A/U where my matching network is placed.

You can instantly see the futility in trying to cut a piece of coax to just the right length to provide proper matching on a number of different bands. This table also illustrates how unpredictable it would be to try to use a formula to find the impedance!

In one rig I built, using a pair of 813s and a commercial FC-30 filament choke, the impedance varied widely: from 12 ohms on ten meters to over 200 ohms on 80 meters. Replacing the commercial filament choke with a homemade bifilar-wound unit gave results that varied much less, from about 30 ohms minimum on one band to 130 ohms on 80 meters. These figures are given only to illustrate the wide impedance variations possible from 3.5 through 29 MHz, and are unlikely to be typical of what you may experience with your own particular amplifier.

wattmeter method

Many amateurs don't have access to an rf impedance bridge. You can still match the exciter to the amplifier, but it will take longer. The name of the game is low swr between the two units, so a wattmeter makes a good trial-and-error method of initially tuning the network. Once the settings have been found, you can mark them on the box and paste on tabs or use the sheet of paper I use.

In this case you observe, from the computer charts, the approximate inductance and capacitance, and start out by setting the inductance somewhere near what you think would be appropriate. With about half-power on the transmitter, rotate the variable capacitors while observing the reflected power.

If it does not go to zero, tap up or down on the inductor and try again (the tap on the coil should be temporary until properly selected). This same technique is used on each different band.

using a swr bridge

This is the least desirable of the various methods. It will usually work, but is the most time-consuming of all and can be misleading. If you think you have gotten it just right, switch to the exciter barefoot and see if the antenna presents approximately the same load, plate current, output power, etc. without returning the exciter. This will provide a check on your accuracy, and is, of course, the desired end result anyway — the ability to switch from antenna to amplitier with similar results.

network placement

In commercial rf power amplifiers the matching network is usually quite near the tubes in the amplifier, and normally there is a separate network for each band. The appropriate network is switched in automatically with the band-selector knob.

It is not at all necessary to have the networks in the same cabinet with the rest of the transmitter. You may find it considerably more convenient to install the network a few feet away from the amplifier where it can be changed quickly whenever you bandswitch. This is the arrangement I have used successfully for a number of years. I have a short piece of coax connecting the network to the input of the amplifier. The length of the coax is in no way critical, but once the network is adjusted, of course, the coax length should remain the same.

A piece of paper was temporarily placed on the front panel of the enclosure, the correct settings for the various bands found, and the paper marked. Then a nicer looking paper was drawn up with markings for those settings, typewritten with the bandmarkings, and attached to the front panel. This allows very rapid setting of the network whenever I bandswitch, yet only one coil and two variable capacitors are used.

Other methods may come to mind that will work adequately for your purpose. Trying to put the networks into the amplifier usually makes additional problems with regard to space, synchronizing with the bandswitch, etc. Thus, the remote installation may appeal to some of you who do not have space in the amplifier or the technical capability of providing mechanical selection when the bandswitch is rotated.

components

Even with 100-watts output, there is only about 1.4 rf amps flowing. Consequently, rather small inductors, such as B&W stock can be used successfully. B&W type 3018 comes in 4-inch (10cm) lengths, 8 turns per inch (2.5cm); the full 4 inches (10cm) is 9.4 microhenries. B&W type 3014 is also 8 turns per inch (2.5cm), 3-inches (7.5cm) long, and 4.8 microhenries. These should give you ideas, and a wide variety of similar inductances are available.

Even with 100-watts output, the voltage across 50 ohms is only about 70 rms. Almost any type of variable capacitor, including the common 365 pF broadcast type, will be more than adequate. You can easily find these for free from junker a-m radios of another era, and usually in gangs of two or three on the same shaft.

You will probably want a bandswitch for the network. Any type of switch capable of handling small amounts of rf will be adequate, and the additional pole/poles may be used to switch in fixed values for the lower frequencies, if desired. Ceramic or steatite switches are recommended.

Fixed capacitors should be rated for at least 150 or 200 volts, and capable of handling rf currents. Mica transmitting types are excellent. Low-cost door-knob capacitors are also good and are usually capable of handling kilowatt outputs.

Some commercial amplifiers use fixed capacitors and a slug-tuned variable inductor. Unless you have some means of determining the actual impedance to be matched, tuneup could be very time consuming, and fairly costly unless a large supply of capacitors suitable for rf is available. Also, many of the available slug-tuned inductors will not handle the amp or two of rf current without damage.

summary

Some method of matching the exciter's 50-ohm output impedance to the input of a linear amplifier should be offered. A good, simple but effective method is to build a single, variable pi-network and place it in a convenient place a few feet from the amplifier. A rf wattmeter may be used for initial tuneup, and simple markings placed on the box containing the network so rapid band changes can be made. Tables are included for both pi networks and L-networks. These were computer-derived and include values for 1.9 through 29.7 MHz.

references

1. Robert Leo, W7LR, "How to Design L-Networks," *ham radio,* February. 1974, page 26.

2. Frank Doting, W6KNU, and Robert Hubbs, W6BXI, "RX Noise Bridge for Accurate Impedance Measurements," *ham radio,* February, 1977, page 10.

ham radio

introduction to operational amplifiers

Operational amplifiers have invaded amateur radio — here are the basic facts on their theory, selection, and application

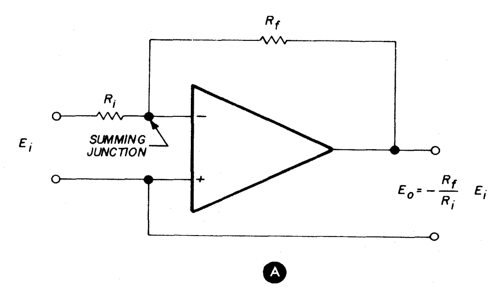
If you have been keeping up with the current electronics literature, you've surely seen articles on the integrated circuit operational amplifier. Perhaps you've wondered just what it is and what it can do for you. The fact is, it can do just about anything you wish, and do it better than conventional circuits.

Tube versions of the op amp have been around for a long time. They were originally used in analog computers to perform mathematical operations such as addition, subtraction, and averaging. The main objection to these circuits was their huge physical size. Recent advances in solid-state technology have produced op amps at very reasonable cost with active elements formed on a single chip of silicon. A complete amplifier now occupies less space than many of the discrete components used in the original vacuum-tube operational amplifiers.

The IC op amp is so useful in amateur radio applications that I've prepared this article to acquaint you with it. The first part of the article discusses some of the more popular circuits and gives the equations describing the relationship between input and output. Then comes a description of the op amp's gain characteristics. The last part of the article is devoted to some applications you'll find useful around your amateur station.

typical circuit

The input stage of the op amp is a high-impedance differential voltage amplifier. This is followed by



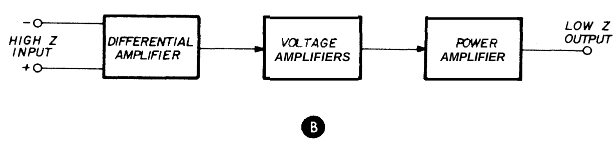


fig. 1. Typical operational amplifier circuit, (A), and block diagram of its three stages (B). Many arrangements of feedback elements are possible.

other voltage amplifiers. The output stage is a low-impedance power amplifier.

Fig. 1 shows a typical circuit. Resistor R_f feeds the output to the negative input, which is sometimes called the summing junction. The negative input is isolated from the driving signal, E_i, by resistor R_i, which represents the circuit's input resistance. The negative input is 180 degrees out of phase with the output and is at ground potential. Under these conditions, no current flows into the amplifier, because

By Donald W. Nelson, WB2EGZ, 9 Green Ridge Road, Voorhees, New Jersey 08043
(reprinted from the November, 1969, issue of *ham radio*).

current in R_f and R_i is equal and opposite, Ohm's law says that the output voltage, E_o, is related to the input voltage, E_i, in the same proportion as the values of R_f and R_i. The negative sign in the equation of **fig. 1** means that the phase has shifted 180 degrees.

definitive examples

The following op amp circuits are ideal representations. Nothing is perfect, of course, but I've used examples of a perfect amplifier to provide definitive examples. A perfect amplifier would have these characteristics:

1. infinite open-loop gain.
2. Infinite input impedance and zero output impedance.
3. Zero response time (the output changes simultaneously with changes in input).
4. Zero offset. With no voltage between the input terminals, the output voltage will be zero.

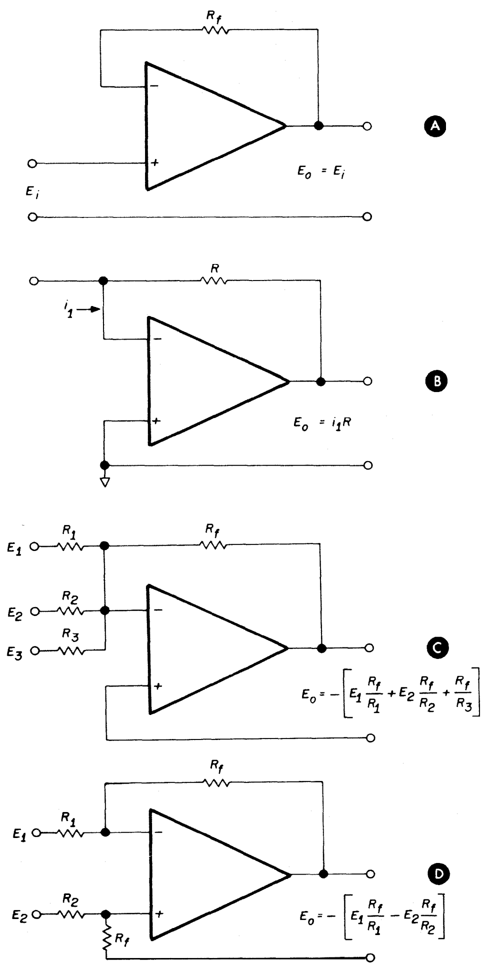


fig. 3. Other examples of op amp voltage amplifiers. A voltage follower is shown in (A), current-to-voltage transducer, summing amplifier, and difference amplifier are shown in (B), (C), and (D).

The important things to remember about these characteristics, which are called summing junction restraints, are:

1. No current flows at either positive or negative input.

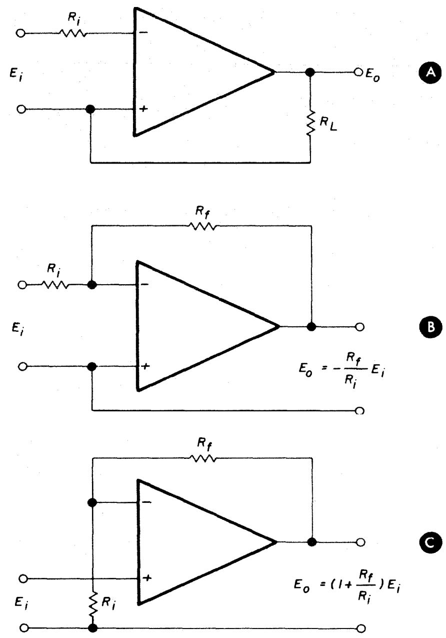


fig. 2. Three common op amp circuits. An open-loop circuit is shown in (A), useful as a voltage comparator. The circuits in (B) and (C) are inverting and non-inverting amplifiers.

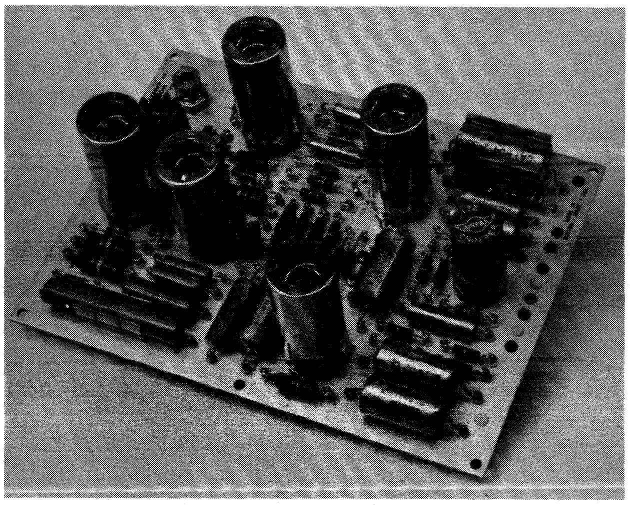
2. Both inputs are at the same potential.

open-loop operation

No feedback is used in the circuit of **fig. 2A.** The amplifier is running wide open. If the input is other than zero, the amplifier will be driven into saturation. An op amp isn't often used in the open-loop mode because of practical considerations. One use, however, is in a voltage comparator circuit. If two ac voltages are applied to the input, the open-loop amplifier will follow their potential difference. As the voltage on the positive terminal changes from that on the negative terminal and vice versa, the amplifier will swing as far as its supply will allow.

inverting and noninverting amplifiers

Two widely used arrangements of the op amp are illustrated in **figs. 2B** and **2C.** In the circuit of ***fig.* 2B** (which is the same as that of **fig. 1**), the output signal is inverted with respect to the input. If a square-wave, for example, with positive-going pulses is applied to the input, the output pulses will be negative. Gain will be proportional to R_f and R_i. The sign in the right-hand member of the transfer



The Philbrick-Nexus **USA4JT** "grandpappy of op amps." Very few amplifiers can match its performance. However, this fine unit has been retired to the back shelf because of its large size, aging characteristics, and high power consumption.

function (relationship of input to output) will be negative because of the 180-degree phase reversal.

In the noninverting circuit of fig. 2C input and output are in phase, which accounts for the plus sign in the equation.

voltage follower

A variation of the noninverting amplifier, the voltage follower, is shown in fig. 3A. Note that the output is connected to the negative input. The positive input is driven directly by the input signal, E_i. Output is equal to input: a unity-gain amplifier. This circuit is used for following voltage references. The limitations of the cathode follower (or emitter follower in transistor circuits) are minimized.

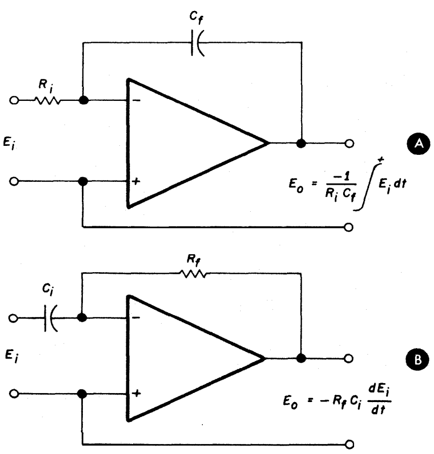


fig. 4. Voltage waveform integration and differentiation may be performed by (A) and (B). These circuits are used for precise filtering.

transducer, adder, subtractor

The circuit of fig. 3B is a current-to-voltage transducer. It can be used to drive a meter, recorder, or other voltage-operated indicating instrument from limited current sources.

Voltage inputs are added directly in the summing

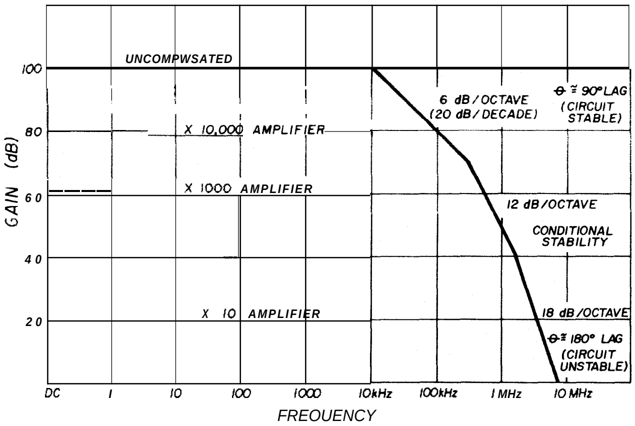


fig. 5. Gain-bandwidth characteristics of uncompensated amplifier. Instability occurs as gain attenuation exceeds **18 dB/octave** because of **180°** phase lag.

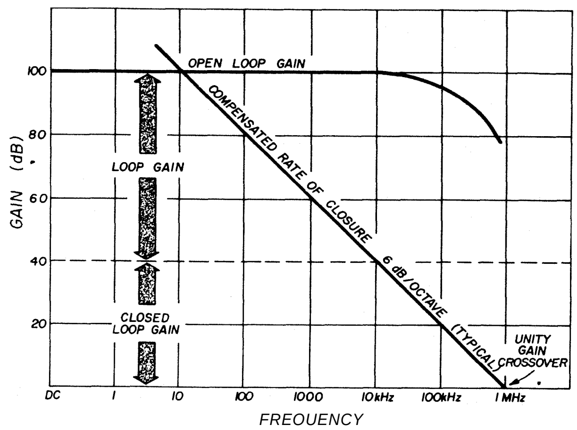


fig. 6. Bode plot of compensated amplifier. Response is limited so that **180°** phase shift occurs before unity gain is reached.

amplifier of fig. 3C. The op amp is shown here in one of the operations for which it was originally used. Each input may be weighted by using different resistor values. Input weighting is proportional to the gain of the particular input: E_1 will have a weight of 2 if $R_f = 2k$ and $R_1 = 1k$. If $R_2 = 500$ **ohms,** E_2 will have a weight of 4.

The circuit of fig. 3D is sometimes called a balanced input amplifier or symmetrical subtractor (difference amplifier). It's used when neither side of the signal being amplified is at ground potential, as across a current-sensing resistor. Other inputs may be added where inputs to the negative terminal are additive, and those to the positive terminal are subtractive.

integrator and differentiator

By using a capacitor in the feedback loop **(fig. 4A)**, the op amp may be used to integrate voltage waveforms. When the capacitor is in the input, **fig. 4B**, the signal is differentiated. Both differentiator and integrator, as shown, are purely theoretical.

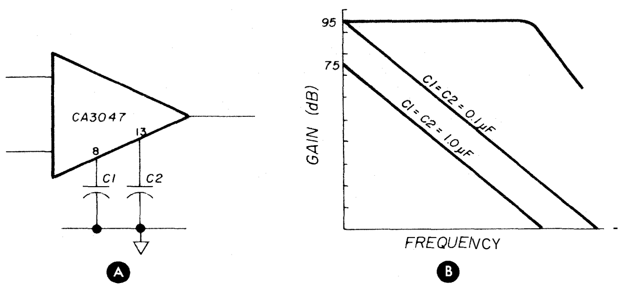


fig. 7. Compensating by bypassing. As capacitor values increase, (A), amplifier open-loop response moves to the left, (B).

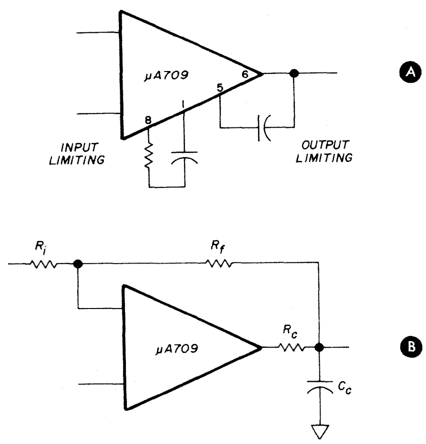


fig. 8. Response limiting at amplifier output. Capacitive compensation, (A), or the RC network in (B) are frequently used to supplement input compensation.

practical limitations

Most errors in a practical operational amplifier with known characteristics can be calculated. If the amplifier is properly chosen for a particular application, these errors may be negligible or can be compensated. With an understanding of amplifier gain, frequency response, and phase shift, you'll be able to apply compensation methods to tame the op amp of your choosing.

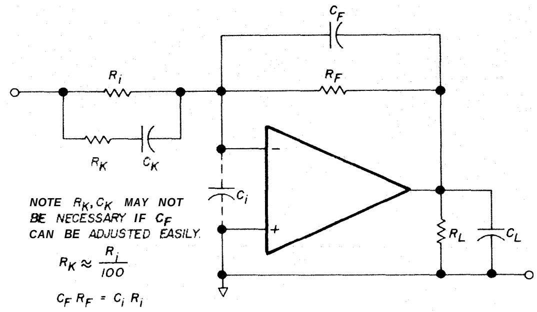


fig. 9. Compensation for op amps with discrete components. Input loading, stray wiring capacitance, and output loading must be compensated.

Several definitions of gain must be understood:

1. ***Open-loop* gain.** This is the ratio of output-to-input voltage at any frequency. No feedback is used. Typical open-loop gains are from 10^4 to 10^9 in commercially available amplifiers.
2. ***Closed-loop* gain.** When feedback is used, the amplification is called closed-loop gain. For reasons to be discussed, closed-loop gain is rarely less than unity.
3. ***Loop* gain.** This is the difference between open and closed-loop gain. Usually, errors are minimized with greater loop gain.

Characteristics such as gain attenuation with frequency (also called roll-off) and phase shift, which are common to all amplifiers, are especially important when considering operational amplifiers. As men-

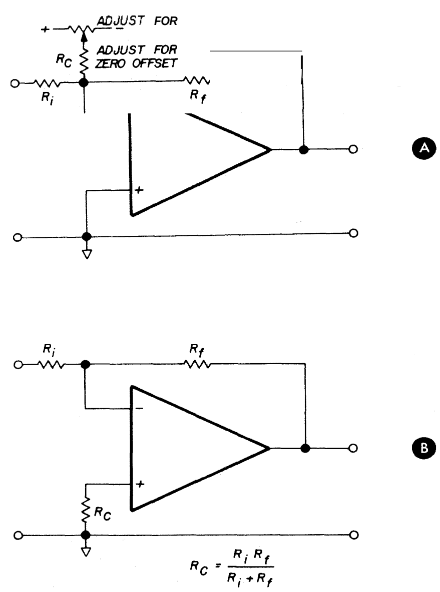


fig. 10. Compensation for dc offset error in an inverting amplifier. Adjustable current offset compensation is shown in (A); drift-compensating resistor in (B).

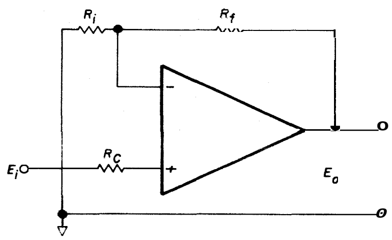


fig. 11. Drift compensation in a non-inverting amplifier. Input voltage swing is a limitation; R_c is now the input resistor.

tioned previously, phase shift through the amplifier must be less than 180 degrees when feedback is employed. Any additional phase shift must be compensated, or the circuit will oscillate.

In fig. 5, gain-bandwidth characteristics are shown for an uncompensated amplifier (not necessarily typical). The phase shift (lag) increases as the gain is affected by feedback. The amplifier becomes unstable when the roll-off exceeds 18 dB/octave because of the 180-degree phase lag. In well-designed amplifiers, this limit occurs below unity gain. Even with compensation, the amplifier can't be controlled when 18 dB/octave is reached; therefore, operating below unity gain is usually impractical. Some amplifiers may be difficult to control at gains slightly above unity.

compensation

Amplifier compensation will limit frequency response, but roll-off and phase shift will be controlled. A plot of a compensateed amplifier's response is shown in fig. ***6.*** This type of presentation is called a Bode plot; it illustrates the limited gain roll-off (rate of closure with the unity-gain point in the frequency response of the amplifier).

Most op amp data sheets give enough information to make a Bode plot. This will allow you to analyze the results of intended compensation. The Bode plot is the easiest way of showing the characteristics of compensation.

The simplest way to stabilize an amplifier in which a large amount of feedback is used is to bypass some signal point in the circuit. IC op amps such as the RCA CA3047 and Fairchild μA702 have terminals specifically provided for this. Fig. ***7*** shows an example. If the bypass capacitors (fig. 7A) are increased in value, the amplifier open-loop response will shift to the left (fig. **7B**).

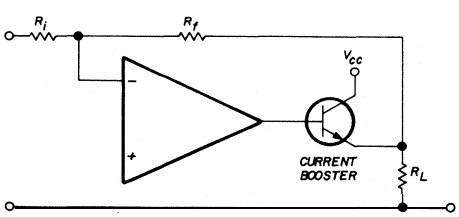


fig. 12. Increasing current output. If output current capability is too low. a booster amplifier can be used to reduce output impedance.

As the Bode plot shows, the high-gain, high-frequency characteristics are very limited with this configuration. The simple addition of a series resistor

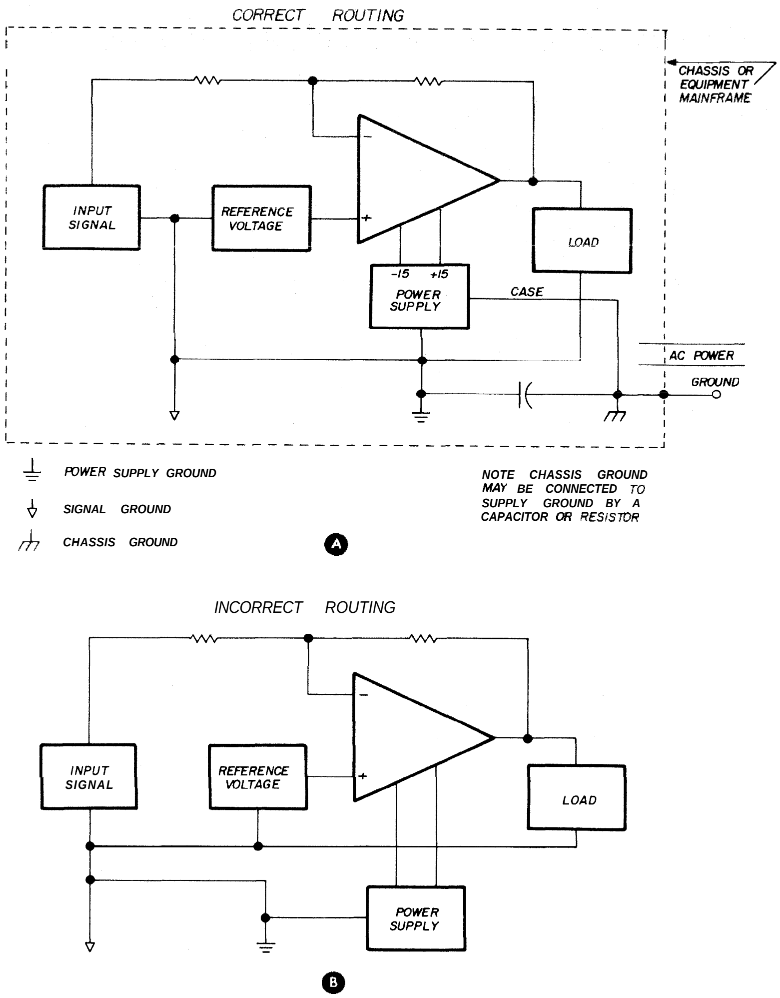


fig. 13. Correct routing (A), and incorrect routing (B), of load return in an op amp layout. Incorrect routing will cause an error in reference voltage.

with the bypass capacitor will yield greater bandwidth.

Output limiting is another popular form of compensation (fig. **8**). Amplifiers such as the Fairchild μA709 have a special terminal just for this purpose (fig. **8A**). The technique of fig. **8B** is also useful. Output compensation is frequently used to supplement some form of input compensation such as that suggested in fig. **8A.** While every compensation problem is unique, we may generalize and say that the compensations shown above are required by the peculiarities of integrated circuits.

Amplifiers using discrete components are internally compensated to a degree. With discretes, the main compensation is for output loading, C_L, stray wiring capacitance, and input loading, C_i. Compensation techniques are shown in **fig. 9.** This is by no means the last word in compensation; it's only in-

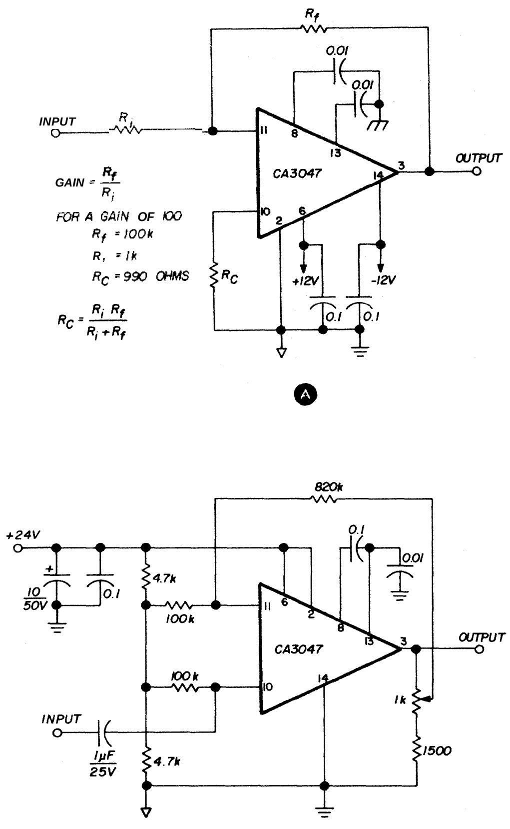


fig. 14. Complete circuits for the basic inverting amplifier, (A), and non-inverting amplifier, (B). Gain of non-inverting ac circuit is approximately 10 and may be trimmed for precise gain.

tended to help you when some published circuit won't work.

offset error

Among the imperfections of a practical op amp is the mismatch of components that prevents the amplifier from having exactly zero output with zero input. This may well be the most serious problem you'll encounter in dc operation of a high-gain amplifier. The basic compensation methods are shown in **fig. 10.**

input/output limitations

Input impedance and voltage swing generally may be neglected in the conventional inverting amplifier

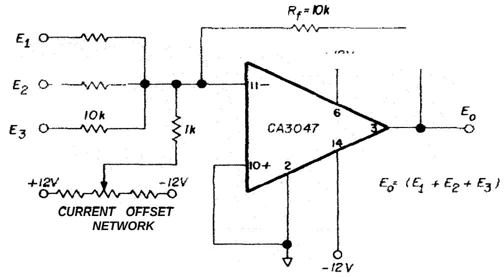


fig. 15. Typical adder. Offset compensation is similar for all computing circuits.

shown in **fig. 10.** The input impedance will be equal to the input resistor, R_i, because the input is a virtual ground. The amount of drift may be considered a limitation of input. The simplest compensation for this is shown in **fig. 10B**, where a resistor is used in the positive input.

Drift may be compensated similarly in a noninverting amplifier as shown in **fig. 11.** The difference here is that R_c is now the input resistor. In this circuit, the input voltage swing is a limitation. This is called "common-mode voltage swing" on the data sheet.

Output impedance, being some value greater than zero, will introduce small circuit errors. It is desirable to keep Z_{out} low. This may be done by using the greatest loop gain possible. A booster amplifier also reduces Z_{out} although such an addition probably wouldn't be considered unless the output current capability is too small. The most common current booster is the pass transistor in a precision power supply as suggested in **fig. 12.**

Beware of the limitation of output voltage swing, especially in IC amplifiers. This is a luxury that closely relates to the price of the op amp. The 30-volt p-p capability of RCA's CA3047A is high for an IC.

One input error that's elusive — at least to me — is the common-mode error. Variously defined, this error arises from the effect of a change in one input on the signal fed to the other input. Common-mode

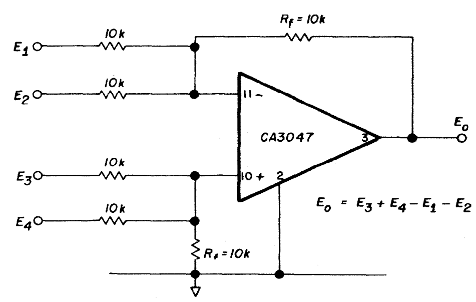
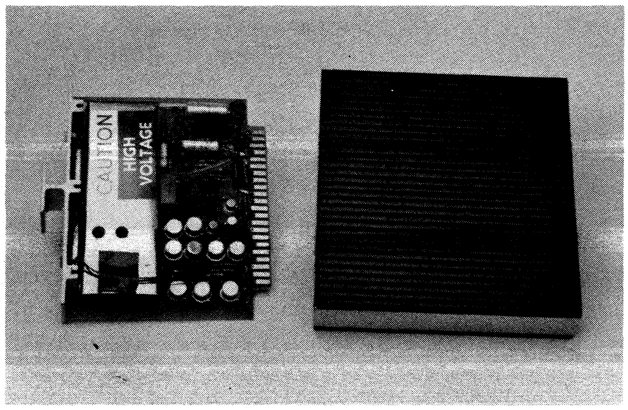


fig. 16. Adder-subtractor. Large values of feedback resistors will result in a gain greater than unity.



A noteworthy successor to the best tube op amps is the Fairchild A00-7. It features discrete components, built-in compensation, and chopper stabilization.

error is smallest when the common-mode rejection ratio is high. This error is important when differential inputs are used, or when the amplifier is operated in the noninverting mode.*

Always look for the least-expensive amplifier that will satisfy your requirements. Some suggestions to guide the newcomer are outlined below.

1. High loop gain is desirable. Usually this implies the need for high open-loop gain.
2. Sufficient output voltage swing and output current to the load must be considered.
3. Offset voltage and drift must be checked for compatibility with your circuit.
4. Offset current is particularly important in circuits such as the current-to-voltage converter.

*Slewing rate is another limitation of practical op amps. Briefly, it is the maximum rate of change of output voltage with time. It must be considered when pulses of fast rise time are employed or high frequency, high level sine wave signals must be processed. It also is a limitation to using operational amplifiers at high frequencies. A thorough treatment of this parameter would be quite lengthy; however, a careful examination of the Bode plot will show slewing rate will change with compensation. Interested readers will find a discussion related to an integrated-circuit op amp (MC-1530) in the Motorola *Integrated Circuit Handbook,* 1968 edition, p. 10-74. **Editor**

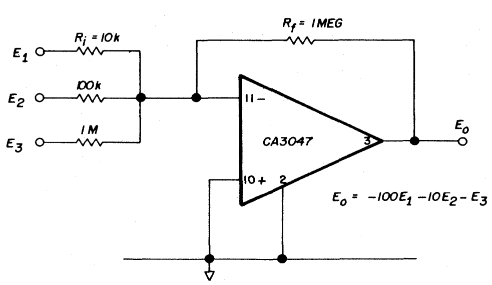


fig. 17. Weighted adder. The sum of the inputs is a function of the feedback resistor; and any reasonable combination of R_f and R, is permissible.

5. Common-mode voltage is important for noninverting and dual-input circuits.
6. Power-supply ripple, drift, and regulation are most important when the supply is used as a reference. However, all op amps work better with a high-quality supply.

The best over-all performance in op amps is obtained from those using discrete components – in fact, tube types. The least expensive and most interesting to experimenters are the integrated-circuit op amps. Despite their low cost. performance is excellent

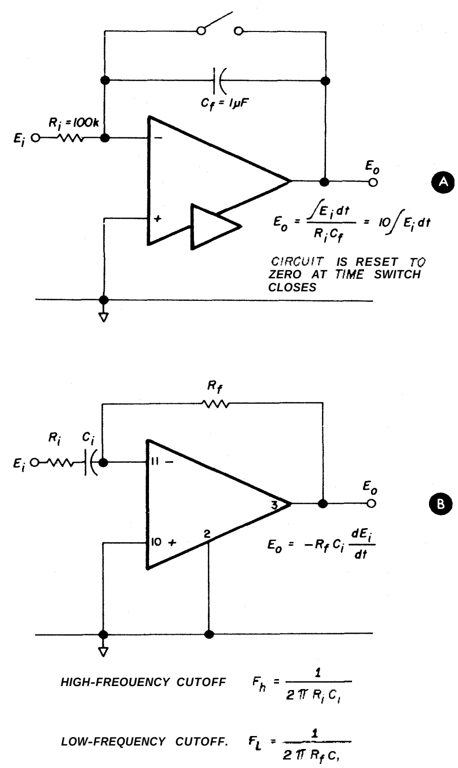


fig. 18. Integrator (A), and differentiator circuit (B). Double amplifier symbol in (A) denotes chopper stabilization required because of offset error due to C_f.

If you understand the parts of this article dealing with the ideal operational amplifier and the limitations of practical circuits, you're almost ready to warm up your soldering iron. First, however, I'd like to give a few precautions on layout and choice of components.

capacitors and bypassing

Poor layout in an op amp circuit may cause its response to peak at the higher frequencies. Under certain conditions, oscillation will result. The problem can exist even with a neat layout. In stubborn cases, peaking may be cured with a mica bypass capacitor (try 100 pF) directly at the noninverting input. This is appropriate only for an inverting amplifier. The problem is rare when the amplifier is used in the noninverting mode.

More frequently, oscillation results from improper bypassing in the power supplies. A 0.1-μF capacitor

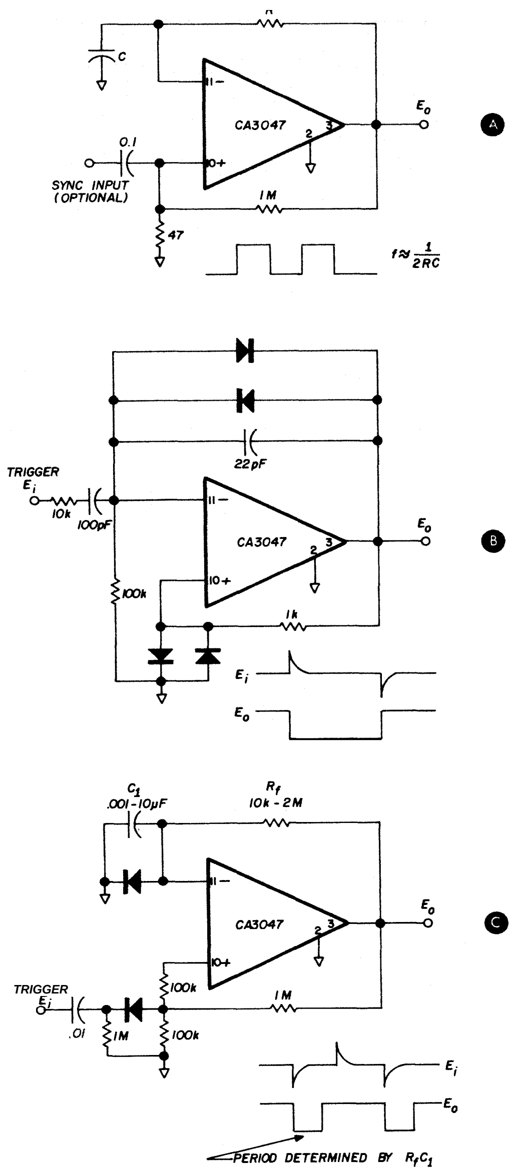


fig. 20. Examples of the multivibrator. Circuit at (A) is free-running. or astable; (B) is a bistable multivibrator, or flip-flop. Monostable, or one-shot, is shown at (C).

on each power-supply lead at the amplifier socket is good practice. Low-inductance, laminated ceramic capacitors are perfect for this.

Capacitors can be critical in some circuits where low leakage is important. Dura-mica types are excellent for compensation purposes. High values and higher precision, such as would be required for tim-

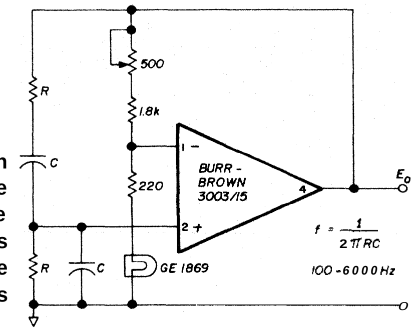


fig. 19. The Wien bridge oscillator. The circuit is temperature sensitive; stability is improved by the lamp, which operates at low-current level.

ing circuits, will call for Mylar or Polystyrene capacitors.

resistors and diodes

The giant called *loop gain,* which is restrained by an operational system, will create problems when noise and unwanted reactances exist. Therefore, certain precautions must be observed with respect to other circuit components.

Resistors must be chosen with care in systems where accuracy depends on the resistor. Wirewound resistors have low noise and excellent stability. However, they have the largest shunt capacitance and series inductance of all types. Also they're not usually available in values above one megohm, and they're expensive.

Carbon composition resistors shouldn't be used where high stability is required, such as in the input and feedback circuits. Although they produce noise, these resistors are inexpensive and are satisfactory in less critical parts of the circuit.

Metal film resistors have excellent characteristics and provide a good compromise between the wirewound and composition types. Their upper range is ten megohms. Higher resistance values are available from Victoreen and Pyrofilm in the form of glass-enclosed, deposited-carbon construction. While there's little choice in precision resistors above ten megohms, you should be aware that some high-

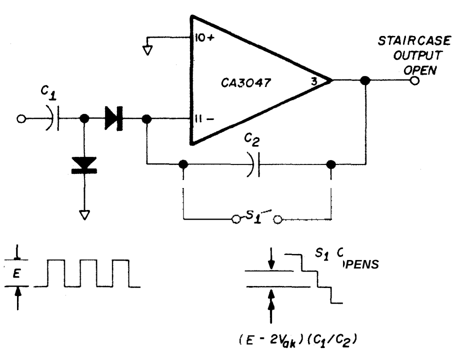


fig. 21. Staircase generator. Ramp output results if a dc signal is applied to pin 11 through a resistor.

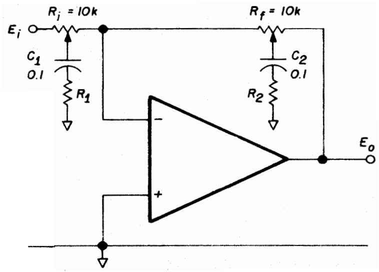


fig. 22. Phase-shift network for system stabilization. Lead and lag compensation are shown.

resistance types are voltage sensitive. They're not precise at voltages other than their test voltage – usually 10 volts. Be careful not to get dirt or perspiration on these, as it may reduce their resistance.

Clamping diodes are frequently used in op amp circuits. Low-leakage, low-capacitance diodes such as the 1N914 or 1N457 types should be used. *Never* use germanium diodes unless the leakage allowances are high enough.

triple grounds

Three grounds should be used: *signal ground, power-supply ground,* and *chassis ground.* This triple grounding technique is essential to minimize voltage drops that would create system errors. At some point, all grounds may be connected, but not necessarily. Consider each system with respect to the voltage drops that will develop. For example, with high output current (load current), the load return to power supply ground must be direct. The reference signal, using signal ground, must not be transmitted through the same wire. **Fig. 13** illustrates some basic grounding techniques; however, the subtleties of the ground loop aren't always easily controlled. A little experimentation with the preceding concepts in mind could lead to a better solution.

compendium of op amp circuits

I've devoted the remainder of this article to a description of some of the more common applications for the operational amplifier. These circuits are just a starting point. I'm sure that ham ingenuity will result in many more interesting variations. Who knows? Perhaps someone will adapt one of these circuits to a communications problem and revolutionize the industry. In any event, I hope these ideas will inspire more experimentation. If you come up with a new use for the op amp, the market is wide open for your ideas.

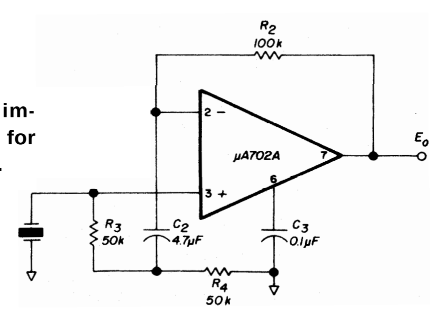


fig. 23. High input impedance amplifier for crystal transducer.

basic computer circuits

While basic computing circuits may not be your idea of a construction project, such applications of the op amp serve to identify what follows. As a matter of fact, with a little thought and planning, these circuits might be just the thing for a science fair presentation.

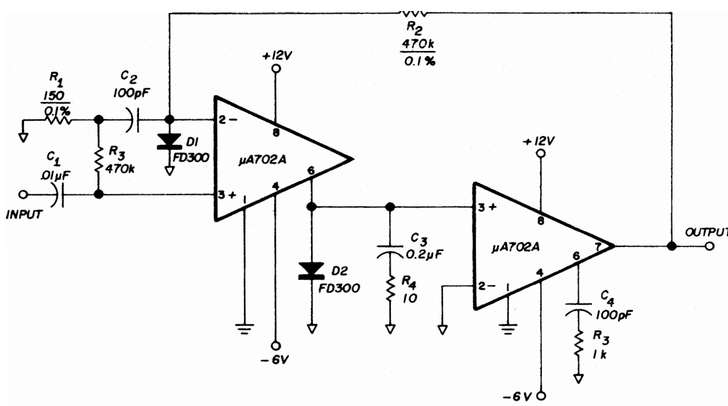


fig. 24. Precision ac amplifier. Gain is 70 dB; input impedance is 200 megohms.

To recap, the basic inverting and noninverting op amps are shown in **fig. 14** with all the component values. You'll recall that the inverting amplifier shifts the phase of the input signal 180 degrees; that is, a positive-going input produces a negative-going output. The output signal will be in phase with the input in the noninverting amplifier.

Typical compensation is shown in the circuits of **fig. 14.** The following circuits are simplified. Compensation and proper bypassing are essential, of course. The RCA CA3047 is inexpensive and altogether adequate for the applications shown.

An adder is shown in **fig. 15.** The offset network is typical for all computing circuits. An alternate would be a voltage offset circuit, which is usually connected to the positive input. The currents from these three inputs are summed, and the negative of this sum appears at the output. Feedback at the negative input means that the input is a virtual ground, so the three inputs are effectively isolated, and no interaction exists among them.

An adder-subtractor circuit is shown in **fig. 16.** Note the equation of the circuit: the output voltage equals the sum of the noninverting inputs minus those on the inverting inputs. Thus, we have a subtracting circuit. By making the two resistors in the feedback circuit larger, greater-than-unity gain may be obtained.

If we change resistor values, a *weighted* adder results, as shown in **fig. 17.** The feedback resistor value affects the sum of the inputs. The weight of the adder is proportional to input gain, which is determined by the feedback resistor.

Other mathematical operations in computers are integration and differentiation. The former is used to find the area under a curve; the latter determines the slope of a curve at any point. In the integrator of **fig. 18A**, an fet input amplifier should be used because of the error caused by bias current. Also the capacitor leakage must be very low — 1 nanoamp or preferably, less.

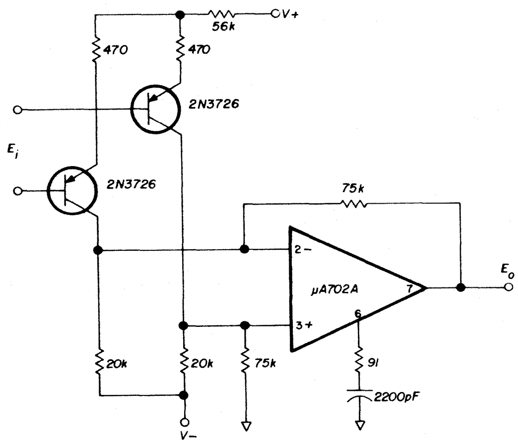


fig. 25. Low-noise tape head amplifier. Matched transistor pair reduces noise and increases input impedance.

Gain response of the integrator is maximum at the low frequencies and decreases linearly with increasing frequency. Amateur application of such a circuit would be in a lowpass filter following a speech clipper to attenuate harmonics.

In the idealized differentiator, gain increases indefinitely with frequency. To eliminate high-frequency noise problems, gain limiting is provided by R_i in the circuit of **fig. 18B**. This circuit is also useful for filter applications; frequency response is determined by the RC constants according to the equations shown.

oscillators and waveform generators

Of the many operational amplifier circuits used in computers, probably the most popular amateur adaptations are oscillators and their close relatives, the multivibrators.

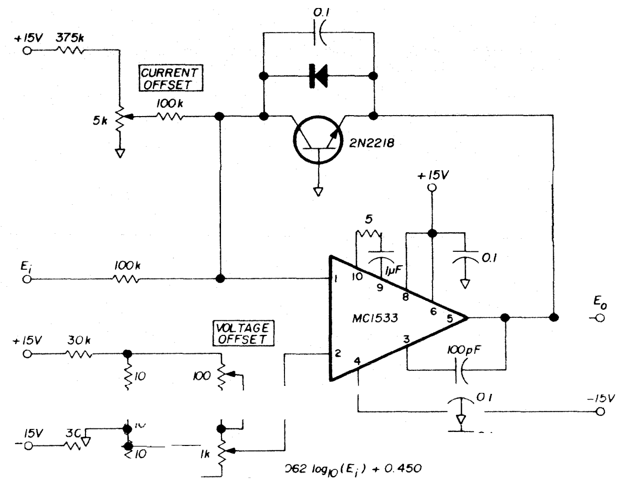


fig. 26. Logarithmic amplifier. Circuit operates over a frequency range of six decades.

If you need an oscillator with an unusually pure sine wave output, the Wien bridge[1] circuit in **fig. 19** is a good candidate. It is inherently temperature dependent, however. In the circuit shown, stability is improved with a lamp operating at very low current.

The multivibrator circuits in **fig. 20** have appeared in various forms in many amateur publications. They're used in electronic keyers, frequency counters, square-wave generators, and a host of other circuits where a controlled signal source is required.

The circuit of **fig. 20A** is an astable, or free-running multivibrator. Its uses include a timing-pulse

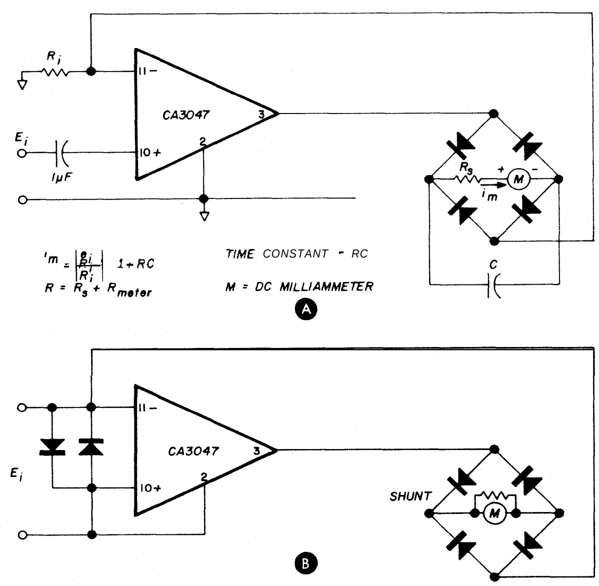


fig. 27. Lossless ac meter circuits. A high-impedance dc meter is preferred for the millivoltmeter circuit, (A); a low-impedance meter should be used in the milliammeter circuit, (B).

generator, or clock, in counters. Feedback to the positive input is called "bootstrapping." This effectively increases circuit gain until it approaches infinity.

The bistable multivibrator (fig. 20B) has two stable states, each of which changes only when triggered by a pulse of opposite polarity. This circuit is used as a memory storage, counter, or shift register in computers. Its principles are often used in amateur circuits with little or no modification.

The monostable multivibrator, fig. 20C, is also called a one-shot. It has one stable state, which can be changed by an external pulse, It will then return to its originai state after a time period determined by its RC constants. The one-shot is used for a time delay or to produce a pulse of specific width when triggered.

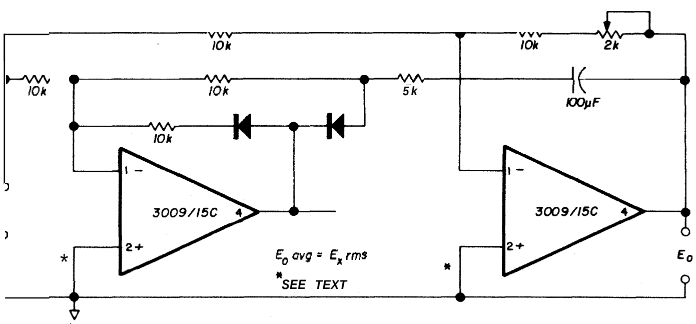


fig. 28. Ac-to-dc converter. Circuit has an input range from 6 mV to 6 volts rms at 10 to 1000 Hz. Amplifier is a Burr-Brown IC.

An application where the integrator feedback capacitor is allowed to charge is shown in fig. 21.[2] During a finite period, the input pulses will add algebraically until the amplifier saturates. When the switch is closed, the output returns to zero. The circuit shown generates a staircase waveform; it can be used as a ramp generator if a dc signal is applied to pin 11 through a resistor. Successively opening and closing the switch would give a sawtooth output. Systems frequently require phase compensation for stability. Precise adjustment may be made with the technique shown in fig. 22. Adjustable lag is obtained by changing the input bypass capacitor; lead adjustment is provided by varying the feedback resistor. Resistors R_1 and R_2 may be necessary to stabilize the system.

amplifiers

In addition to the basic amplifier circuits previously shown, I've included some useful variations.

The circuit of fig. 23 is often used in dynamic instrumentation such as vibration measurements. It's a high input impedance amplifier using a crystal as a

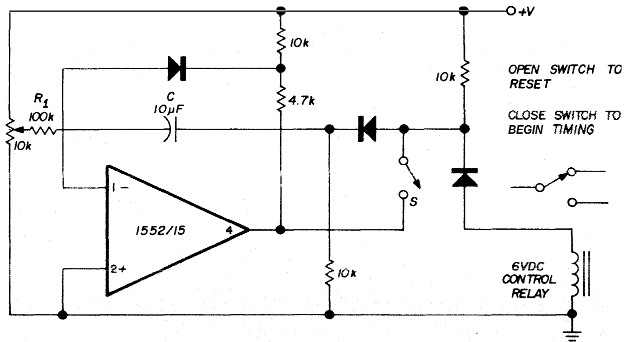


fig. 29. Timing control circuit. Minimum delay is determined by $R1 \bullet C$, maximum delay is infinite.

transducer.[3] A possible adaptation for amateur use would be a crystal microphone preamp.

The amplifier in fig. 24 has a gain of 70 dB and an input impedance of 100 megohms.3 Diodes are used to prevent latch-up. Because of the high-frequency characteristics (100 kHz) with the compensation shown, special attention should be given to layout and power-supply decoupling.

The tape-head amplifiers of fig. 25 uses a matched pair of 2N3726s to reduce noise and increase input impedance. Despite the fact that it uses no input resistor (purists may object to classifying this circuit with op amps), the circuit does suggest a technique for improving common-mode rejection and increasing the common-mode range for any op amp.

A widely used instrument is the log amplifier (or log converter). It has the capability of compressing input voltage ranges of several decades into a useful linear range. Some uses for this circuit (fig. 26) are in filter measurements, leakage measurements, and as a computer power-function generator. The amplifier shown uses a diode-transistor combination in the feedback circuit to achieve the conversion function.4 Both current and voltage offsets are required for operation over a 6-decade input range. With an input of 0.13 mV to 100 volts, the output is from 220 to 580 mV.

Fig. 27 shows two lossless ac meter circuits. The millivoltmeter circuit, **A,** uses an op amp to compensate for diode, resistor, and meter losses. The

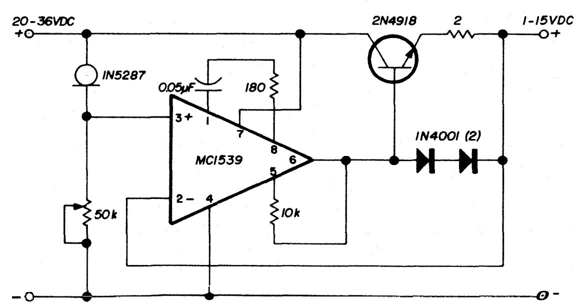


fig. 30. A power supply that provides up to 200 mA between 1 and 15 volts. Regulation is better than 0.01%!

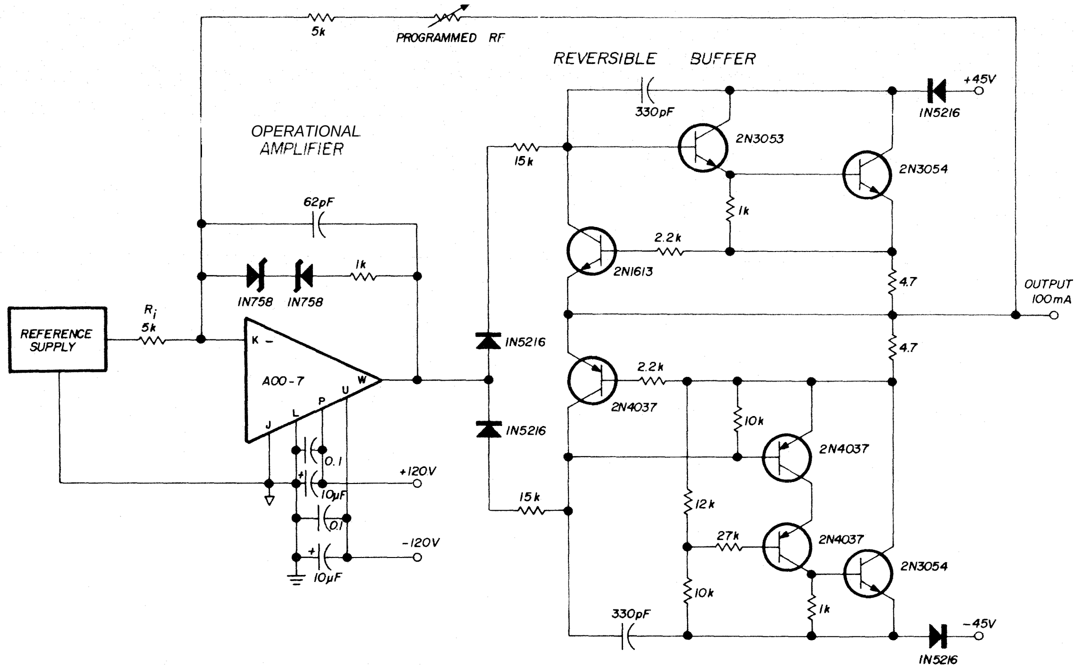


fig. 32. The op amp in a filter circuit. Filter in the feedback circuit yields an output with response characteristic of the filter element.

response time, which is usually low, can be increased by increasing either the meter series resistor, ***R,*,** or the averaging capacitor, *C*.

The current-sensitive counterpart of the millivoltmeter, shown in **fig. 27B**, has zero drop across its terminals. Limiting diodes at the input should have very low leakage. No charging capacitor is necessary, because the current is averaged by the meter. Low-impedance dc meters are practical in this circuit, whereas the millivoltmeter performs more efficiently with a high-impedance meter.

In measurement and control circuits, it's frequently necessary to convert ac to dc. The circuit of **fig. 28** using Burr-Brown amplifiers' consists of a full-wave rectifier and a filter.

The time delay circuit in **fig. 29** requires an amplifier with a high-impedance input such as that provided by an fet. The Burr-Brown 3521H is such an amplifier. Of the many uses for this circuit in amateur applications, an example would be to control timing of voltage turn-on in a power supply. Circuit response time would be limited by relay action.

The Motorola MC1539 op amp is the center of precision in the circuit of **fig. 30**[5]. This supply provides up to 200 mA at any voltage between 1 and 15 volts. Regulation is better than 0.01%. The unusual reference supply consists of a constant-current diode (type 1N5287) and a 50 kilohm potentiometer.

The amplifier has a gain of 120,000, so it won't load the reference. Note that output compensation is between pins 5 and 6, and input compensation is between pins 1 and 8. The circuit is protected against burnout from short circuits.

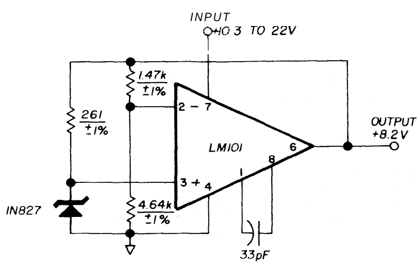


fig. 31. Voltage reference source. Regulation is 0.01 mV/V; temperature stability is ±0.5% from −55° to 125°C.

voltage reference

A more sophisticated reference supply uses its own op amp, a National Semiconductor LM101 **(fig. 31)**.[6] The 1N827 reference diode is temperature compensated. Regulation is 0.01 mV/V, and temperature stability is ±0.05% from −55°C to 125°C. Short-circuit protection for the reference is provided internally. The LM101 needs only one compensating component, the 33-pF capacitor (the commercial version is the LM301A). If you'd like an alternate circuit, see National Semiconductor LM343 op amp data sheet, ±65 V output!

active filters

A nice thing about active filters is that you don't need inductors to achieve near-ideal mathematical

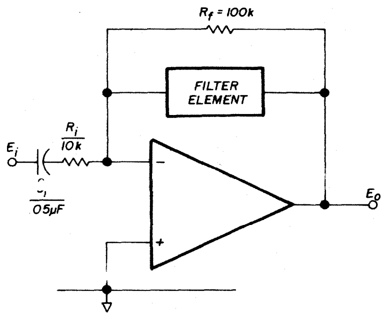


fig. 33. A Twin T filter for use in an op amp circuit. Bandpass is 1000 Hz; with input and feedback resistors of 10k and 100 the gain would be 10. The product C_iR_i should be greater than twice CR.

response characteristics. Another good feature is high input impedance, which means that matching is not a consideration." While compensation for the operational amplifier is necessary, filter reactance trimming is not. Once you've calculated component values for a specific response, you're done.

A circuit that's easy to understand is shown in fig. 32. A filter in the feedback circuit of a conventional inverting amplifier yields an output with the response characteristic of the filter.

A possible filter is the Twin T shown in fig. 33. A 1000 Hz bandpass filter is in the basic circuit. If $R_i = 10k$ and $R_f = 100k$, the gain will be 10 at 1000 kHz. The Twin T is one of the simplest (first order) filter elements; however, it has relatively low Q, so don't expect miracles from it.

The circuit of fig. **34** may be used with an active high pass, or low pass, or rejection-notch filter by inserting the appropriate filter element. Reference 7 provides more information for active filter designs.

Another practical approach toward building an improved filter is to precede a conventional filter with an op amp follower (fig. **35).** This circuit eliminates filter input loading problems. Although resistor R is chosen to equal the filter input impedance, the resistor is really used to match the input of the preceding stage. The input impedance of the op amp is arbitrary.

A follower on the filter output would be useful if a varying load is used. The purist will argue that this isn't a true active filter. I'm willing to concede the point, but I hasten to add that it's a *handy technique.* I encourage the amateur to take it from here.

'True for this circuit, but not for controlled source and negative immitance converter techniques. **Editor**

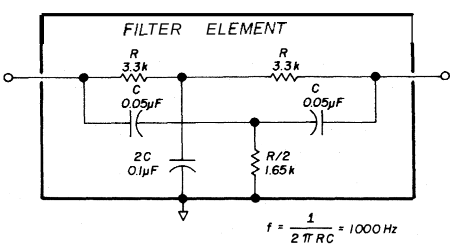


fig. 34. Preceding a conventional filter with an op amp follower to eliminate input loading. Input impedance of amplifier is arbitrary, as explained in the text. Resistor *R* should equal the input impedance to the filter.

a parting thought

I've presented some basic data on one of the most interesting and challenging products of modern solid-state technology. The circuits shown are the most commonly used, but by no means do they cover the entire field of possible applications.

If you wish to adapt these circuits to your needs, a good grasp of op amp theory is essential; the material listed in the references will supplement that

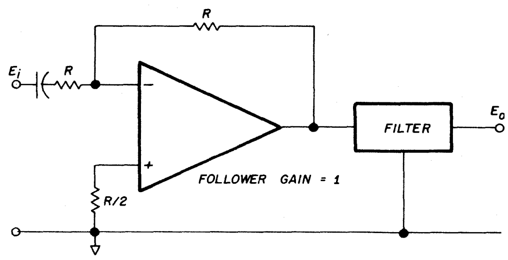


fig. 35. Preceding a conventional filter with an op amp follower to eliminate input loading. Input impedance of amplifier is arbitrary, as explained in the text. Resistor *R* should equal the input impedance of the filter.

in the first part of this article. Some possible projects that come to mind are:

1. An ultra-stable oscillator (for system synthesis).
2. A precision filter for selective calling.
3. A high-impedance meter for measurement of h_{fe} or g_{fs}.
4. A precision digital power supply.
5. Science Fair computer projects.

I'm sure you've thought of a few projects, too.

references

1. "Handbook of Operational Amplifier Applications," Burr-Brown Research Corp., Tucson, Arizona.
2. H. A. Wittlinger, "Application of RCA CA3033 and CA3033A, etc.," RCA, Electronic Components, Harrison, New Jersey.
3. "Linear Applications Manual," Fairchild Semiconductor, Mountain View, California.
4. "Integrated Circuit Data," Motorola Semiconductor Products, Inc., Phoenix, Arizona.
5. "Motorola Monitor," Vol. 7 No. 1, Motorola Semiconductor Products, Inc., Phoenix, Arizona.
6. "Ideasfor Design," *Electronic Design,* January *18. 1968,* p. 128.
7. "Handbook of Operational Amplifier Active RC Networks." Burr-Brown Research Corp., Tucson. Arizona.
8. "Application Manual for Computing Amplifiers,'' Philbrick-Nexus, Dedharn, Massachusetts.

ham radio

an IC op amp update

Following the work of Nelson, Author Jung brings you up to date with state-of-the-art IC operational amplifiers

The first operational amplifier article in this issue discussed the fundamentals of op amp design theory, some specs, and also earlier representative examples. Although there has been a great deal of new device activity since that article was originally published, take heart; it can all be sorted out. In this update article I'll discuss a number of different communication related applications for op amps, with attention directed to which device to choose for optimum performance, for a given application. The emphasis will be on the simple and straightforward devices, their uses, and limitations.

As a starting point, what is the standard in IC op amps? It has to be said that for a general purpose device, it is the 741. General purpose has come to mean a unit which can be used with ±5V to ±18V (dual) or +10V to +36V (single) supplies. In addition, it has a small-signal bandwidth of 1 MHz, a slew rate of 0.5 V/μS, input bias currents of 100 nA or less, and an offset voltage of 2 mV, all of these specs

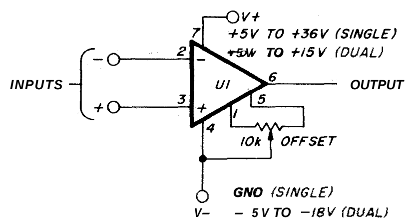


fig. 1. Standard pinout diagram for 8-lead IC op amps. When shown in a schematic diagram, the power connections are always mandatory. though sometimes they may just be implied. The offset null connections are not 100 per cent universal, and therefore the individual data sheets should be consulted.

being typical. There are probably a hundred or more IC op amps which can meet this definition. However, I'll concentrate on the more common and readily available types **(table 1)**.

Fig. 1 shows what has now become the standard pinout for single IC op amps. You'll note that the pin numbers marked on the leads correspond to the 8-pin (round can or mini-dip) configuration. The 5

By Walter Jung, 1946 Pleasantville Road, Forest Hill, Maryland 21050

pins, exclusive of the offset pins, are the *absolute minimum* required for op amp connections, and you cannot make a circuit work without all of them. Power pins ($V+$ and $V-$) are always there, although in some drawings they may only be implied for simplicity.

A 741 can be made to function in about 90 out of 100 op amp circuits, so it's a handy device to have around. For basic experiments and study, it is easily the best device since it's essentially foolproof. You don't have to bother with compensation components either, because it is internally compensated. Other op amps (such as the 709,748, and 301A), use additional external compensation components. If you have some of these on hand, they also can be used in many of the following circuits (if the devices are compensated for unity gain as shown in **fig. 2**). Generally though, other devices mentioned will all be internally compensated, unless otherwise noted.

Standard op amp power supply voltages are ±15 volts, but this is not a very rigid requirement. If you have a balanced dc power supply providing between ±5 and ±15 volts, it will suffice for most of the circuits we'll discuss; you could also use a pair of 9 volt batteries. In general, you'll want to include a pair of

table 1. Common op amps and their manufacturers.

devices	manufacturer
μA709, μA740, μA741, μA759, μA791, μA798, μA799	Fairchild Semiconductor 464 Ellis Street Mountain View, California 94040
ICL8007	Intersil 10710 N. Tantau Avenue Cupertino, California 95014
MC1456, MC1436, MC1741S, MC3403, MC1458, MC3471	Motorola Semiconductor Box 20924 Phoenix, Arizona 85036
LM301A, LM307, LM318, LM324, LM348, LM349, LF356, LF357, LM358	National Semiconductor 2900 Semiconductor Drive Santa Clara, California 95051
RC4558, RC4136	Raytheon Semiconductor 350 Ellis Street Mountain View, California 94040
CA3140, CA3130, CA3160	RCA Solid State Division Route 202 Somerville, New Jersey 08876
NE532, NE534, NE535, NE536, NE5534	Signetics 811 E. Arques Avenue Sunnyvale, California 94086
TL080 Series TL071 Series TL061 Series	Texas Instruments Dallas, Texas 75222

table 2. Standard pinout single op amp devices.

bipolar		fet	
device	remarks	device	remarks
741 (748)	general purpose	3140	general purpose fet
307 (301A)	general purpose	356 (357)	high performance fet
1456	high slew rate	TL081 (TL080)	general purpose fet
1436	high voltage	TLC71	low noise TL081
343 (344)	high voltage	3160 (3130)	CMOS output
759	high current		
799	single supply	740	first
		536	generation
		8007	fets
318	very high speed		
5534	low noise		
1741S	high slew rate		
535	high slew rate		

good rf bypasses, such as 0.1 μF ceramic capacitors across the supply lines, preferably near the IC. Or, you may want to construct a simple dc supply, (2, 3, 4) using IC regulators. For most IC op amp circuits, regulation is not at all critical and IC voltage regulators are more than adequate.

With regard to **fig. 1,** the only other circuit detail not yet mentioned is the offset pot. The method shown is common to many units, such as the 741, but is not *completely* universal. If your circuit requires offset nulling, double check the data sheet to be sure of the method. Often, the need for an offset adjustment can be eliminated by careful circuit design.

The recent arrival of the fet input IC op amp, with wide availability and low cost, is one of the happier developments in recent years. A number of bipolar and fet devices are listed in **table** 2. All ICs listed in this chart can be substituted in the pinout diagram shown (excluding offset null). In some cases a dual listing is shown; the second is the uncompensated version of the basic device.

A fact which further serves to demonstrate the maturity of IC op amps is the preponderance of multiple units, both duals and quads. A few of the more popular ones are listed in **table** 3. As can be seen, many of them are dual or quad versions of single op amps. There are no wholly universal types, but the 1458 and 4558 are probably the most popular duals **(fig. 3A)**, likewise the 324 and 4136 are the most popular quads. In the quads, there are two generally accepted pinouts **(fig.** 3B) corresponding to the 324 and 4136.

applications

Having very briefly looked at popular standard ICs, it's time to examine some specific op amp applications. This treatment is somewhat unique, because it shows only a few device part numbers. It's intended that almost any device listed in tables **2** and 3 can be used in these circuits with the appropriate connections. The circuit discussion, however, will emphasize which device is the best choice, and why.

gain blocks

Probably one of the most common uses of IC op amps is as a gain block, to raise signal levels or buffer a source. The basics of inverting and non-inverting gain stages are quite straightforward, but a lot can be said about tailoring a gain stage to specific uses, while getting around device limitations.

In **fig. 4A,** an op amp inverter is connected for split supply use. Gain is simply R2/R1, while the input impedance is equal to R1. The practical problem with this circuit is that high gains tend to result in a low input impedance. This can be alleviated by fixing R1, and raising R2. If a fet input amplifier is used, R1 and R2 can be made as large as convenient, many megohms being entirely practical. The practical limit is that at very high values of R2, stray capacitance will begin to limit bandwidth.

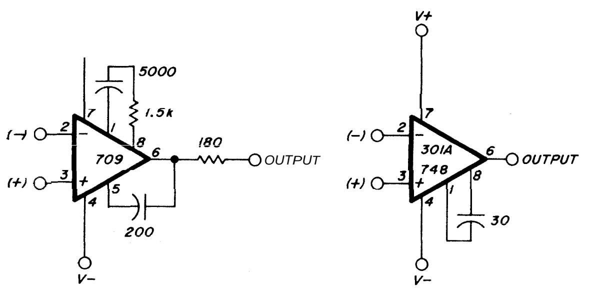


fig. 2. Some op amps are not internally compensated and may require external capacitors for compensation. The examples shown here have been compensated for unity gain.

A dc response is obtained with the input directly connected to RI. At high gains (100 or more), offset at the output may be prohibitive, requiring offset nulling. For ac use, the blocking action of C1 causes the dc gain to be unity (for any ac gain) and thus offset is not amplified. This is the preferred connection when only ac gain is required. Note that C1 and R1 have a low-frequency rolloff which sets the lowest usable frequency. The high-frequency bandwidth can be controlled by either of two means. If gain is very high (⩾ 40 dB), the amplifier's gain bandwidth will cause a rolloff at the op amp's unity-gain frequency divided by the stage gain. For example, a 741's unity response occurs at 1 MHz, thereby limiting bandwidth to 10 kHz with 40 dB of gain. Wider

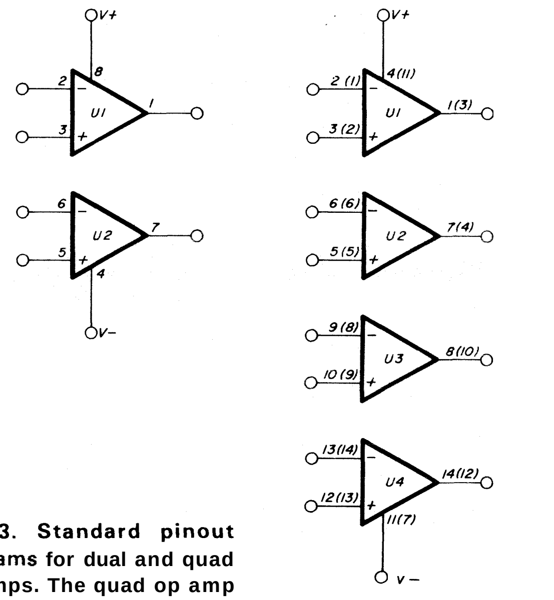


fig. 3. Standard pinout diagrams for dual and quad op amps. The quad op amp can be in either o. two different patterns, with the second set in parenthesis.

bandwidth units, such as the 4558, will increase the total bandwidth.

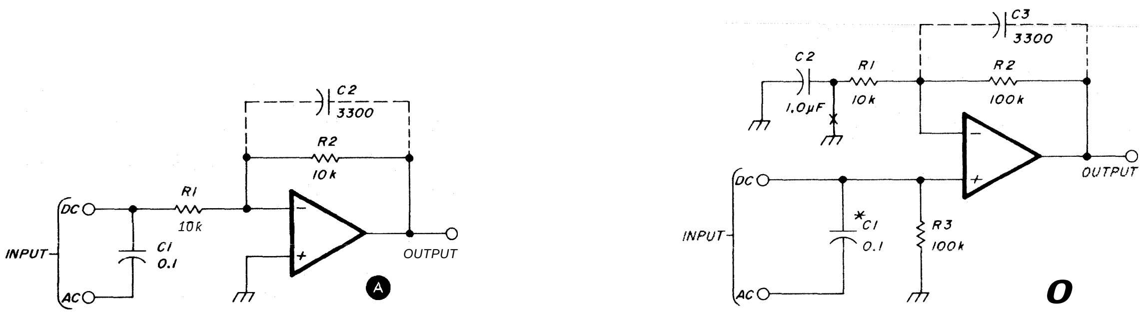
For working gains not limited by amplifier rolloff, the bandwidth can be set by a capacitor (C2) connected across R2. In the example shown, C2 sets the 3 dB bandwidth to 5 kHz. An interesting basic characteristic of the inverter is that it can be used for gains *less* than unity, when R2 R1. This can be useful when it is required to reduce the gain of a stage to zero, or some low value less than unity. Compression agc or amplifiers make good use of this factor.

A non-inverting gain block is shown in **fig. 4B;** the

table 3. Standard pinout dual and quad op amp devices.

bipolar		fet duals	
device	**remarks**	**device**	**remarks**
1458	**dual** 741	TL082	**dual** TL081
358	**low power, single supply**	TL072	**low noise**
2904	**low power, single supply**	TL062	**low power**
532	**low power, single supply**		
798	**low power, single supply, class AB output**		
4558	**"faster"** 741, **low noise**		
bipolar quads		**fet quads**	
device	**remarks**	**device**	**remarks**
324	**low power, single supply, class** B **op**	TL084	**quad** TL081
3403	low **Power, single supply, class AB** op	TL074	**low noise** TL084
4136	**"faster"** 741, **low noise** (pnp **input)**	TL075	**low noise,** 4136 **pinout**
		TL064	**low power**
348 (349)	**quad** 741 (npn **input)**	3471	**high speed**

fig. 4. A standard op amp inverting gain block is shown in A. The gain of the IC is $R2/R1$, with the input impedance being $R1$'s value. For ac use, the low frequency rolloff occurs at $f_L = \frac{1}{2\pi R1C1}$, while the high frequency rolloff is $f_H = \frac{1}{2\pi R2C2}$. B illustrates a non-inverting configuration. In this case the gain is $\frac{R1+R2}{R1}$. The low frequency rolloff will occur at two frequencies, $f_{L1} = \frac{1}{2\pi R3C1}$ and $f_{L2} = \frac{1}{2\pi R1C2}$. There will only be a single high frequency rolloff, $f_H = \frac{1}{2\pi R2C3}$



general gain of this stage is $\frac{R1 + R2}{R1}$. The intrinsic input impedance is very high (assuming R3 is not connected), as it looks directly into the input to the amplifier. If the amplifier is a fet type, the bias current will be only a few picoamps, while in bipolar units it is typically on the order of 50-100 nA. The non-inverting stage is therefore inherently best when minimal loading of the source is required, such as timing capacitors or high impedance transducers.

Gain can be manipulated by either R1 or R2, as convenient, with no effect on input impedance. The minimum gain of this circuit is unity, with R1 open and R2 shorted. Breaking the dc path of R1, and inserting C2 causes amplification of only the ac component. Input ac coupling is provided by C1, with R3 as a bias return. R3 is shown as nominally 100k ohms but a fet amplifier can allow 10 megohms or more here, without compromise.

The same bandwidth limitations apply to the non-inverting amplifier as the inverting amplifier. C3 can be used to reduce bandwidth at a specific point. Note that in this circuit, C3 can reduce the gain to a minimum of unity.

For both of these circuits, large-signal bandwidth is limited by the slew rate of the op amp used, and can be quite independent of the external components. If you require high output voltages (10 volts peak) at frequencies of more than 10 kHz, a high-slew rate op amp is in order. Most fet amplifiers have slew rates of 5 V/μS or more, allowing full power to 100 kHz or more. Reference 5 will provide some further insight into optimizing general purpose gain blocks. The absolute limit on voltage swing depends either on the op amp or its power supplies. Standard units can swing about ±10V with ±15 V supplies. High voltage devices, like the 1436 or 344, can swing ±20 V or more, with +28 V supplies. For rated output, loading should be 1k ohm or more. If lower impedances are used, they will not necessarily damage the device, but may result in reduced output due to current limiting.

Since a great many op amp audio amplifiers use a single power supply, it is appropriate to configure the previous gain blocks for a single voltage. **Fig. 5A** shows the inverting mode. Again R1 and R2 set the gain, with C1 and C_O providing input and output coupling. Assuming a zero dc level at the input and

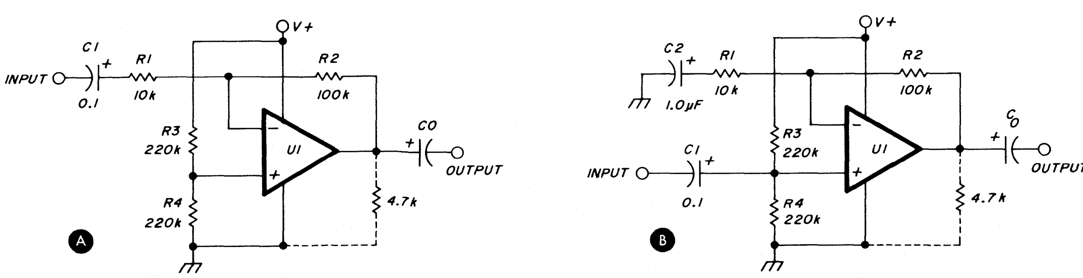


fig. 5. The inverting and non-inverting configurations can be connected for a single supply voltage. The voltage divider provides one half of the supply voltage as a reference to the op amp. In the case of the class-B output stages, the resistor can be added to reduce cross-over distortion.

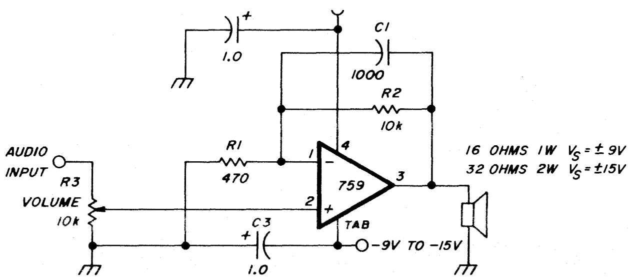


fig. 6. The Fairchild 759 can be connected as a 1 or 2 watt monitor amplifier. For 1 watt operation, the supply must be ±9 volts with a 16 ohm speaker: far 2 watts, ±15 volts with a 32 ohm speaker. The IC package must be connected to a heat dissipating device.

output, C1 and C_0 must be polarized as shown. R3 and R4 form a $\frac{V+}{2}$ divider, which biases the amplifier output for the maximum signal swing. For most general-purpose op amps, this is all that is required.

Some op amps have class-€3 output stages for reduced power drain. Examples are the 358 and 324 units which have 0.7 and 1.5 mA quiescent drains, making them highly suited for battery or other power uses. However, the class-€3output stage does generate cross-over distortion, which may be objectionable. An optional pull-down resistor (4.7k) can be used to minimize this affect. It should be adjusted to suit the particular application.

monitor amplifier

Recently there have appeared on the market several op amps which can furnish substantial output power, with the convenience and simplicity of general op amps. One of these is the Fairchild 759 illustrated as a 1W/2W monitor amplifier circuit in **fig. 6.** The 759 has a peak output current of 350 mA, and a supply range similar to other op amps (36 volts). Thus, the power it can deliver to a given load is related to the impedance and supply voltages used. As shown here it can furnish 1 or 2 watts to a 16 or 32 ohm speaker, with supplies of ±9 or ±15V, respectively.

The device is furnished in a heatsink type package, which should be attached (but insulated) to a chassis or other heat radiator. This circuit is attractive because of its simplicity, and can be adapted to suit other gain requirements. Gain is 20 for the values shown, with the response rolled off by C1 at 15 kHz.

A higher power "op amp with muscle" is the 791 (Fairchild) which has a 1 A output. There are also several high voltage devices which can be used to drive external transistors providing many watts of output. Examples are the Signetics NE540 and NE541, the Motorola 1436, and the National 343 and 344.

push-pull driver

An interesting technique which will produce 6 dB more voltage and power output, for a given supply voltage, is shown in **fig.** 7. This circuit is a push-pull driver, which effectively doubles the output voltage swing across a floating load. The circuit is quite cost effective when used with one of the dual op amps in **table** 3. Gain is adjusted by R1 or R2, and it may he adapted for single supply bias by lifting the grounded end of R5 and applying a potential of $\frac{V+}{2}$ to the input of U2 and the bottom of R5.

parallel driver

A handy idea, when a single op amp just won't supply enough output power, is the parallel driven circuit shown in **fig. 8.** Here U1 and U2 are two op amps of a similar type, with their inputs driven in parallel. The outputs are combined through low value resistors, with output current being approximately doubled. Additional similarly connected stages can also be used, such as 3 or 4 sections of a quad unit. The output resistors force the current to be shared equally between the op amp outputs. While their values are not critical, they should be at least 50 ohms.

Some single op amps have noteworthy power or voltage characteristics and are very attractive for use in this type of circuit. Examples are the Signetics 5534 which has a 30 mA output stage, and high voltage types such as the 1436, 343, or 344 which can swing up to 80 V p-p in this circuit.

variable voltage reference

The circuit in **fig. 9** is useful for the generation of a

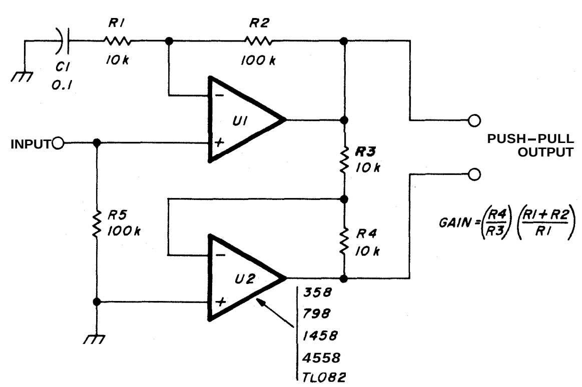


fig. 7. If the load is floating, two op amps can be connected for push-pull operation. This method will provide 6 dB more voltage and current output, for a given supply voltage.

buffered and stable reference voltage source, while being readily adaptable to a wide range of output voltage and current requirements. This method takes advantage of the ability of a number of op amps to operate from a single supply voltage, with their inputs ***at or at near ground*** potential. The basic reference voltage is developed by the LM336, a stable, 2.5 V, monolithic zener diode with a low (20 ppm/°C) temperature coefficient. Because of the low-dynamic impedance of this diode (less than 1 ohm), the 2.5 V is extremely stable when the diode is biased for a current of 1 mA. R4 applies some fraction of the 2.5 V to the op amp, which amplifies it by a factor of 4 to yield a +2.5 to 10 V output.

Output current rating is dependent upon op amp, of course, and will be about 10 mA for general purpose types. The 759 can supply up to 350 mA if desired, or other devices can be buffered by an npn

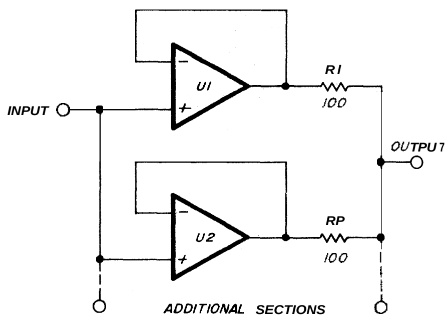


fig. 8. Similar op amps can be connected in parallel, providing additional output current. The outputs must be connected through low-value resistors.

emitter-follower stage. If greater output range is desired, the circuit can be operated from a higher supply voltage with R2 adjusted accordingly. R3 should be selected to maintain about 1 mA of current in the LM336.

digitally-programmable voltage source

A buffered, digitally programmed voltage source is shown in **fig. 10.** This circuit is quite useful as a repeatable, programmable lab source with a basic range of 0 to 2.55 V (10 mV per step), or 0 to 25.5 V (100 mV per step). The output is adjustable, in binary fashion, with an 8 bit TTL compatible input control.

This circuit uses an MC1408 8-bit D/A converter, which provides 1.99 mA full scale. The current in turn is converted to a buffered voltage by U2. R3 determines the basic voltage range, being 1280 ohms for a 2.55 V scale, or 12.8k for 25.5 V (the op amp used must be capable of handling these output voltages). A 741 or other general purpose op amp is adequate for a 2.55 V range, but a high-voltage single-

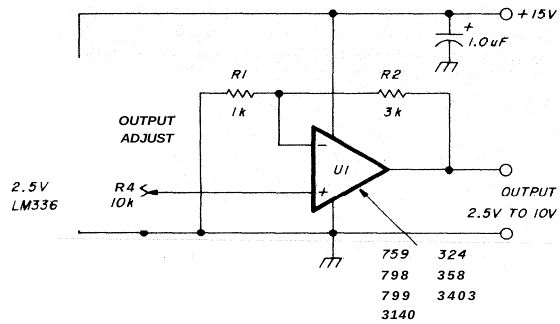


fig. 9. Combining a high-quality zener diode with an op amp will produce a variable voltage reference. In this case, the op amp has a dc gain of 4, giving a voltage range of 2.5 to 10 volts.

supply type is more appropriate for the 25.5 V range. For this, the 759 is suggested. It can also be used for the lower scale, of course, and is attractive because of its high output current.

Like the voltage-reference source, regulated current sources are also useful as basic circuit elements, especially for control and measurements applications. **Fig. 11A** illustrates a simple current source, which uses only a reference diode and a single resistor to set the output current. The reference diode is driven in bootstrap fashion by the op amp, causing the reference voltage V_z to appear across R5. The regulated output current is I_o, which may be any value less than I_z, but must be substantially greater than the op amp's bias current. If a fet op amp is used, this circuit can be used from 1μA up to a level approaching I_z. In this case, it is 10 μA.

The current in the zener is set by R1 to provide a minimum zener current (1 mA), taking into consideration the supply voltage and R_L. With $R_L = 1$ *meg,* the circuit has a compliance voltage of 10 volts. The weak point of this circuit is the fact that the zener

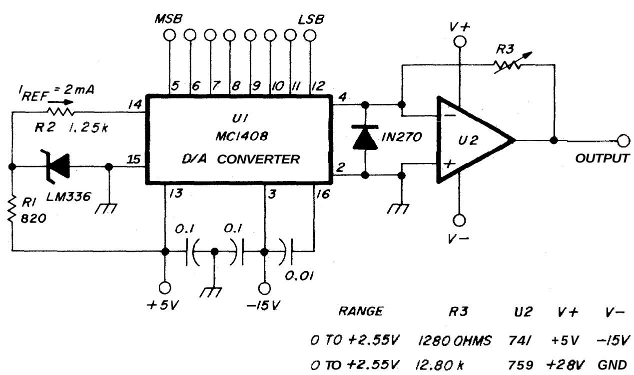


fig. 10. Instead of using a straight analog control, this circuit uses a D/A converter to form a digitally-programmable voltage source. The binary coded, TTL information will produce either of two voltage ranges, 0 to 2.55, or 0 to 25.5 volts. Full scale calibration is achieved by trimming R3.

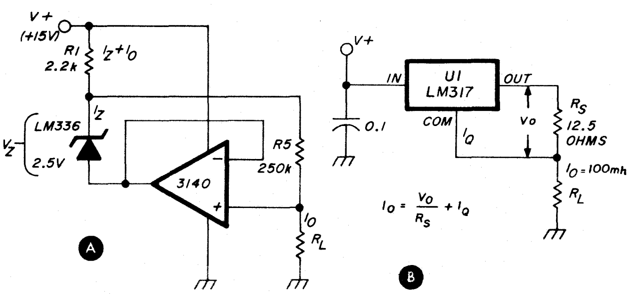


fig. 11. A simple current source combines a zener diode with an op amp in a bootstrap configuration (*A*). For higher currents, any of the three-terminal regulators can be used as constant-current sources (*B*). The current setting resistor is connected between the output and ground terminals of the device.

current changes with supply voltage and R_L. Although the LM336's low impedance mitigates this, it is still the ultimate limit to precision. For a higher performance version of this circuit, the zener can be replaced by a 2.5 V three-terminal voltage reference, such as the Analog Devices AD580 or the Motorola MC1403. Reference 6 includes a discussion of this type of circuit

For currents higher than a few mA three-terminal references and regulators can be used very effectively, with the addition of only one resistor, to set the output current, as shown in **fig. 11B**. This schematic is quite general as shown, and can use any of the three-terminal devices. The AD580 and MC1403 are usable up to 10 mA, while the LM317 can handle one ampere or more (it must, however, have a minimum load current of 10 mA).

low-voltage ohmmeter

A circuit which employs several of the previously described principles, as a low-voltage ohmmeter, is shown in **fig. 12.** Two desirable factors incorporated in this ohmmeter are a low applied voltage, (0.1 V), and an output which is linearly proportional to the unknown resistance. It can accurately measure resistances to below one ohm.

The heart of the circuit is a current source, composed of U1, CR1, and R This circuit, with R_Y selected for the appropriate range, furnishes a constant current to the unknown, R_X. For example, with the values shown, I_x is 100 μA, and 1k ohm R_X will drop 0.1 V. The voltage dropped across R_X is indirectly read from U1's summing point by amplifier U2. This node, being of a much lower impedance, allows U2 to be a relatively high bias current device such as a 741. U1 is a fet input unit (356) giving best accuracy at high R_X levels (low levels of I_X). Alternately, a 3140 can be used with somewhat less precision.

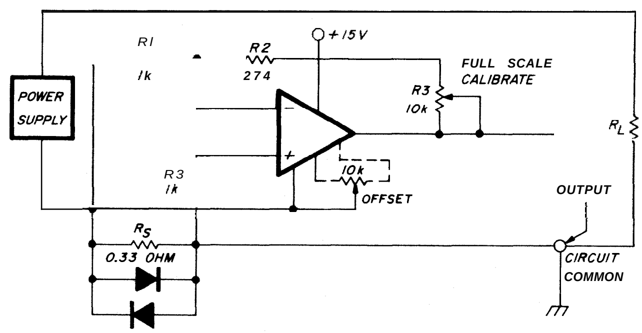
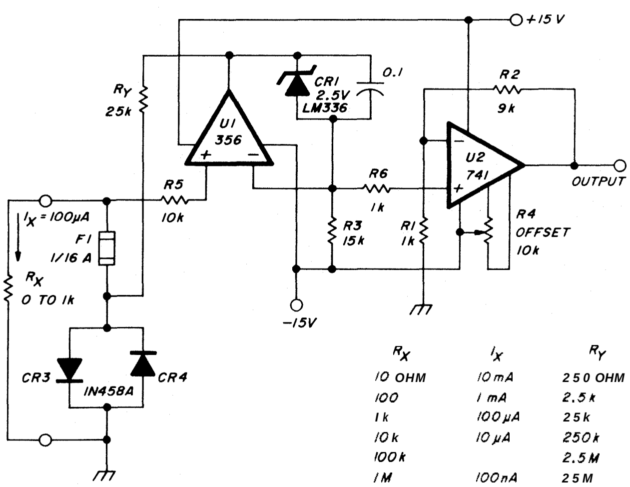


fig. 13. The supply-current monitor uses the voltage drop across a low-value resistor to indicate the current being drawn in the circuit. One side of the resistor and op amp are referenced to ground. The voltage difference across the resistor is amplified by the op amp, producing a 10-volt output that corresponds to one ampere of current.

U2 operates as a straight dc amplifier, with a gain of 10, scaling the 0 to 100 mV unknown voltage to 0 to 1 V at the output. Thus, a 1k R_X resistor can be read as 1.000 (k) on a DVM scale. The circuit can overrange at least 100 per cent, therefore, 2 volt scaled instruments can read up to 2k ohms full scale (or 200 mV at the input). The dynamic range of the circuit is over five decades.

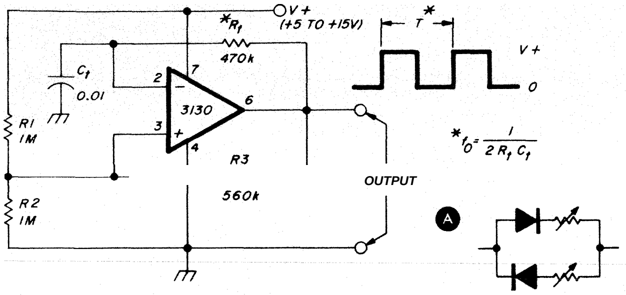
Since the maximum voltage handled by U2 is only 100 mV, it should be offset nulled to eliminate zero error for best low-scale accuracy. This is done by shorting the input and adjusting for 0.000 V out of U2. For full-scale calibration, the individual range values of R_Y should be trimmed for correct output, with a reference value for R_X.

If the circuit is to be used to probe equipment, overvoltage/current protection is warranted. A 1/16 A fuse and clamp diodes CR2-CR3 protect the range resistors; with R5 protecting the op amp. The diodes used for CR2 and CR3 must be low leakage types, such as those specified in **fig. 12.**



R_X	I_X	R_Y
10 OHM	10 mA	250 OHM
100	1 mA	2.5k
1k	100 μA	25k
10k	10 μA	250k
100k		2.5M
1M	100nA	25M

fig. 12. This low-voltage ohmmeter combines a stable constant current source with a dc amplifier. In this diagram the ohmmeter will read 1000 ohms full scale.



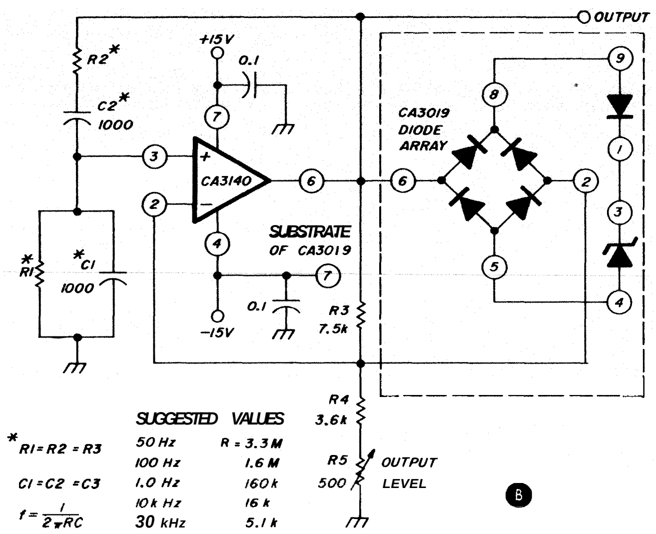


fig. 14. This square-wave generator (*A*) can cover the range of 1 Hz to over 1 MHz by proper component selection. Symmetry of the output waveform is controlled by connecting diodes in series with R_t (see text). The wide-range Wein bridge oscillator shown in B at right uses a diode array to provide amplitude control.

Fig. 13 illustrates how the differencing input voltage feature of the op amp can be used to monitor the current in a supply line. A sampling resistor R_s is inserted in the line to develop a voltage proportional to the supply current. This voltage is then amplified and referenced to the circuit common point by the op amp, which can be a 799 for medium to high currents ($\geqslant 10\mu A$), or a 3140 for very low currents. R3 is trimmed to calibrate the output, in this case, 10V = 1 ampere.

Techniques such as this, which functionally do nothing more than replace a series ammeter, will become more important as forms of automated control pervade the amateur station. The output as shown, could be directly processed by an A/D converter, for instance.

op-amp signal sources

The test bench can always use simple, inexpensive, and high performance signal sources. **Fig. 14** illustrates two examples of oscillators which capitalize on some features displayed by modern IC op amps.

A simple (and probably familar) op amp based astable multivibrator is shown in **fig. 14A.** This circuit generates square waves over an extremely wide range, from well below 1 Hz to over 1 MHz (with suitable values, of course). The RCA 3130 used, a +5 to +15 V device, has a CMOS output stage. Thus, it can drive either 5 V TTL or 10-15 V CMOS logic stages directly, since its output swings from rail to rail (V+ to ground). Rise and fall times of the circuit are quite fast, on the order of 100 nS. Although shown here as a 100 Hz source, R_t and C_t can be readily scaled for different ranges. For control of symmetry, R_t can be replaced by two resistors in series with a reverse connected diode. If higher output swings are desired, other (uncompensated) op amps can be used. Examples which are capable of high speed are the 301A, 748, TL080, and 357 units.

The classic Wien bridge oscillator is often seen in the literature[7] and is a true stalwart for the generation of low-distortion sine waves. The circuit of **fig. 14B** shows how the 3140 can be used in conjunction with diode array, providing amplitude control.

Two problems which beset this type of oscillator are high distortion at high frequencies, (due to limited slew rate in the amplifier) and amplitude "bounce." The 3140 has a high inherent slew rate of 9 V/μS, which allows full output (20 V p-p) to over 100 kHz. The use of a zener diode clamp for amplitude control allows fast agc, without the bounce or overshoot of thermistors. A range of suggested values is given in this figure, and RCA's data sheet[8] for the 3140 discusses this circuit in further detail.

references

1. Donald Nelson, WB2EGZ, "What's This We Hear About Op Amps?" ***ham radio.*** November. 1969.
2. Douglas Schmieskors, WB9KEY, "Adjustable Voltage Regulator IC's," ***ham radio,*** August, 1975, page 36.
3 Henry Olson, W6GXN, "Second Generation IC Voltage Regulators," ***ham radio,*** March, 1977, page 31.
4. Chris Cogburn, K5VKQ, "How To Design Regulated Power Supplies," ***ham radio,*** September. 1977, page 58.
5. Walter G. Jung, "Improve Op Amp Audio Circuits," ***Electronic Design,*** September 27, 1973, page 6.
6. Walter G. Jung, "Applications of the AD580," ***Analog Dialogue,*** 9-2, page 20, Analog Devices, Norwood, Massachusetts, 1975.
7. Walter G. Jung, "Sine Wave Oscillators," ***db,*** July, 1976, page 38.
8. RCA CA3140 Operational Amplifier Data Sheet, number 957, RCA/Solid State Division, Somerville, New Jersey 08876.

bibliography

1. Hall, Courtney, WA5SNZ, "Circuit Design With the 741 Op Amp," ***ham*** *radio,* April, 1976. page 26.
2. Jung. Walter G., "IC Op Amp Cookbook," Howard W. Sams, Indianapolis. Indiana.

ham radio

high-frequency hybrids and couplers for amateur applications

Hybrid and coupler circuits borrowed from the microwave domain have many uses at the lower frequencies — here are some applications for amateur use

Microwave hybrids are extremely versatile devices. They have many applications not necessarily restricted to the microwave region. This article explains how these circuits may be put to use at the higher amateur frequencies where communication may be enjoyed without the interference and noise created by thousands of commercial kilowatt transmitters.

When hybrids are mentioned, many hams think of bifilar-wound coils on toroidal forms. However, the circuits described here may be constructed from coaxial line for vhf or uhf use. For the higher frequencies, they may be constructed using stripline, microstrip, or waveguide techniques. Three devices are considered:

1. The half hybrid **(fig. 1)**. This is a degenerate form of a 4-port device that may be used as a power combiner or divider.

2. The branch directional coupler, **fig.** 2, which is a quadrature hybrid with some interesting applications for moonbounce work and ssb.

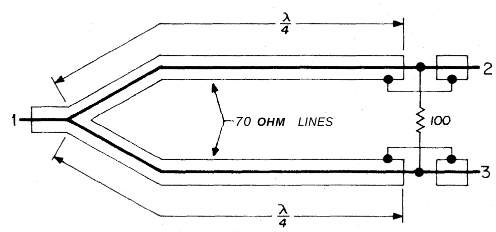


fig. 1. The half hybrid, useful as an isolator between two power sources, Impedance at each port is 50 ohms.

3. The coaxial rat race **(fig. 3)**. Sometimes called a 180° hybrid, this circuit may be used as part of a balanced modulator or to match or balance two equal loads (as in combining equal sections of an antenna array).

the half hybrid

This is the simplest of the devices described. It consists of a Y or T junction, two quarter-wavelength matching transformers, and a bridging resistor.

By Henry S. Keen, W2CTK, (reprinted from the July, 1970, issue of *ham radio*)

If the half hybrid is fed at port 1 **(fig. 1)**, the signal will divide equally between ports 2 and 3. Because no phase difference exists at ports 2 and 3 when properly terminated, no voltage appears across the resistor; therefore, no power is absorbed. If an imbalance exists due to a mismatch, however, part of the signal will be absorbed by the resistor and part will be reflected to the generator. If the generator impedance is 50 ohms, it will absorb the reflected portion. The isolation between output ports is independent of the match provided by the loads.

If you look at the circuit quickly, the source of this isolation may seem vague; but if the circuit is redrawn as in **fig. 4,** the path from port 2 to port 3 resembles the familiar bridged-? network. In this circuit, a signal at port 2 will be nulled at port 3. Therefore, the load impedance at each port is not a factor in the isolation between ports.

The half hybrid may be used to provide isolation between two power sources, such as a pair of power

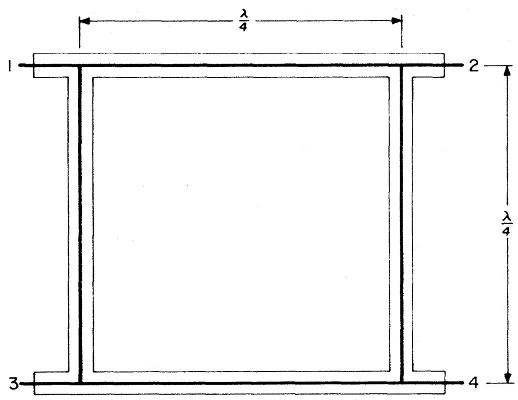


fig. 2. The branch directional coupler. This device divides power between two matched loads.

transistors. A fail-safe arrangement is thus obtained, whereby failure of either component will not affect the load presented to the other unit. Power output will decrease by 6 dB because input power will be dissipated in the bridging resistor, but loading conditions presented to the source will remain unchanged.

In applications requiring high reliability during prolonged unattended operation (as in fm repeaters), half hybrids as combining networks offer a passive means of ensuring uninterrupted service without resorting to complex switching mechanisms.

branch coupler

The branch coupler is a 4-port device. It divides input power between two matched loads. The isolation between two input ports is a measure of the match provided by the loads. A 90° phase difference exists between the signals at the two output ports, because one signal travels one-quarter wavelength farther than the other. This device can be used for sampling a portion of the signal for reference or comparison purposes.

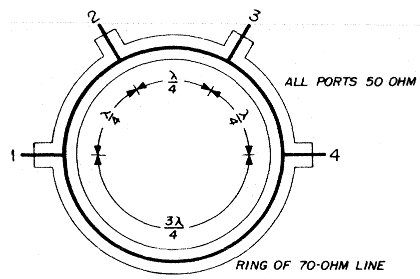


fig. 3. Coaxial rat race, or ring coupler. It can be used to balance similar sections of an antenna or as a balun.

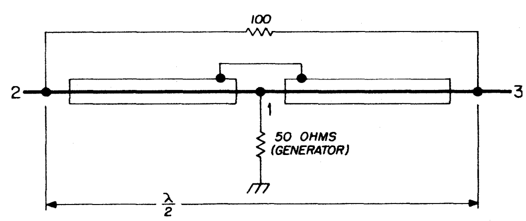
analysis

To understand the design principles of the branch coupler, consider the case of the 3-dB version in which the input power is divided equally between the two load ports.

In a perfectly matched coupler no signal exists at port 3, so this port can be short-circuited without affecting power distribution **(fig. 5)**. This would make branches 1-3 and 3-4 shorted quarter-wavelength stubs, shunted across ports 1-4. Thus they may be removed, leaving only branches 1-2 and 2-4.

If power is to divide equally between ports 2 and 4, port 4 must present a 50-ohm load at port 2. The characteristic impedance of branch 2-4 must therefore be 50 ohms, thus establishing an impedance of 25 ohms at port 2. To match this to a 50-ohm input, branch 1-2 must have a characteristic impedance of 35 ohms, which can be obtained with two 70-ohm coaxial line sections in parallel.

When the network is "reassembled," branch 1-3 will be the same as branch 2-4; while branch 3-4 will be the same as branch 1-2. For a general solution of



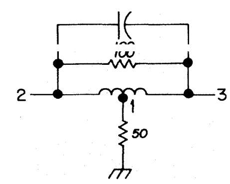


fig. 4. Equivalent circuit of the half hybrid. Note the resemblance to the bridged-T network.

the design, the branch impedances will be:

$$1\text{-}3 = 2\text{-}4 = \sqrt{\frac{Power\ (2)}{Power\ (4)}}$$

$$1\text{-}2 = 3\text{-}4 = \sqrt{\frac{Power\ (2)}{Total\ power}}$$

Several applications of the branch coupler are of interest for amateur work. For example, a 3-dB coupler can be used as a phasing power divider to feed a circularly polarized antenna. Another use would be as a 90° phase shifter for phasing-type ssb generators.

Let's first consider the power divider. if a signal fed to port 1 produces clockwise phase rotation, feeding port 3 will produce counter-clockwise rotation. If both ports are fed simultaneously, linear polarization

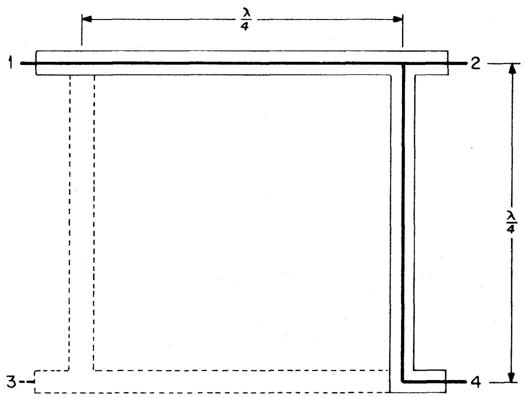


fig. 5. The branch coupler. If port 3 is short-circuited, branches 1-3 and 3-4 may be removed.

will result. A line stretcher in one of the inputs would permit adjustments to any desired phase angle.

A received signal of the same polarization as that transmitted would appear at the same port from which it was transmitted. A signal of opposite phase rotation, such as a reflected signal, would appear at the other port.

In microwave applications, isolation from 40 to 50 dB has been obtained under ideal conditions. The thought occurs that this idea might be useful for moonbounce work; however, I have no information as to how much the circular polarization would be degraded.

single sideband

Single-sideband phasing techniques have been used in microwave receiver design to phase out the image signal. This method also offers a theoretical 3-dB reduction in front-end noise.

A block diagram of such a system is shown in **fig.** 6.1 The second 3-dB hybrid operates as a combining network designed for the intermediate frequency. Balanced mixers could be used to cancel the noise contributed by the local oscillator.

At lower frequencies, the branch coupler may be synthesized with appropriate values of L and C. An equivalent quarter-wavelength line may be constructed for any desired characteristic impedance **(fig. 7)**. The absolute values of each reactance at the design

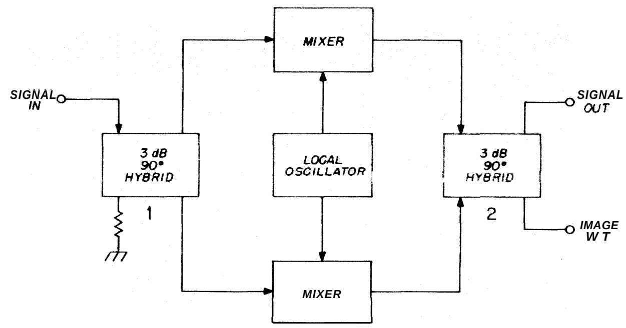


fig. 6. The branch coupler used to phase out frontend images in ssb receiver applications.

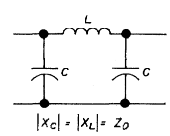
frequency should equal that of the line being synthesized. The capacitors in the final version, **fig. 8,** are identical in value.

coaxial rat race

The standard form of the 50-ohm rat race, or ring coupler, is shown in **fig.** 3. It consists of a closed loop of 70-ohm line with a circumference of three *1/2* wavelengths. The four ports are located 1/4 wavelength apart, with first and fourth ports connected by a 314-wavelength line.

A signal fed to port 2 divides in two; each half travels around the loop in opposite directions. The path to port 4 is a half-wavelength longer than that to port *2,* so the two signals arrive at their respective

fig. 7. Synthesized quarter-wavelength line of impedance Z_0.



loads in phase opposition. Port 3, located midway between the two loads, will therefore receive no signal. The loads must be identical for this cancellation to occur.

As a matter of interest, both loads can be removed, leaving only the loop with ports 1 and 3. Cancellation will occur at the center frequency. This dual-path structure is known as a re-entrant filter.

If the signal is fed to port 3, the two loads will be fed in phase. Any in-phase reflections of equal magnitude from ports *2* and 4 will arrive at port *1* out of phase and will therefore cancel. If the loads are unequal, and the reflected signals differ in amplitude or phase, or both, then cancellation will be incomplete, causing a signal to appear at port 1. In some applications, a matched load may be placed at the odd port to absorb the imbalance.

The rat race offers an excellent means of adjusting signal balance between similar sections of an antenna array. A detector-indicator, such as a receiver with an S-meter connected at port 3, would show imbalance between array sections. Identical lengths of transmission line must be used between ports 2 and 4 and their respective loads to avoid complications due to phase differences.

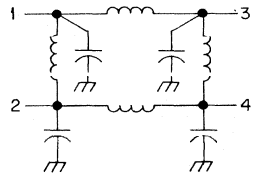


fig. 8. Lumped-constant equivalent circuit of the branch coupler.

The rat race also functions well as a balun. When used for this application, the balanced load impedance is twice that of the coax input line, and port 3 is usually grounded.

capacitively-coupled hybrid

The capacitively-coupled hybrid shown in **fig. 9** is another form of the 90° or quadrature hybrid. Coaxial line of any convenient characteristic impedance can be used in its construction as long as the correct line length and proper coupling capacitances are used. The electrical length of the line elements is computed from:

$$Coupling\ (dB) = -20 \log_{10} \cos \theta$$

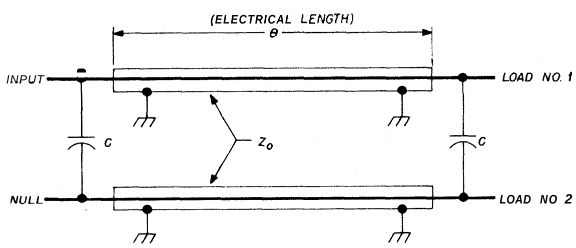


fig. 9. Capacitively-coupled hybrid is another form of the 90° or quadrature hybrid.

The reactance of the coupling capacitors is:

$$X_c = Z_o \tan \theta$$

Thus, for a 3 dB hybrid using common 50-ohm coaxial cable, the lines would have an electrical length of 45°, and the coupling capacitors would each exhibit 50 ohms reactance.

50-ohm rat race

If you don't happen to have any 70-ohm line handy, you can make a rat race with 50-ohm line, which is suitable for spot frequency or narrowband work. In this version (fig. 10) ports 1 through 4 are separated by 0.153 wavelength of 50-ohm line. The long side is 0.653 wavelength, taking into account the cable's velocity factor.

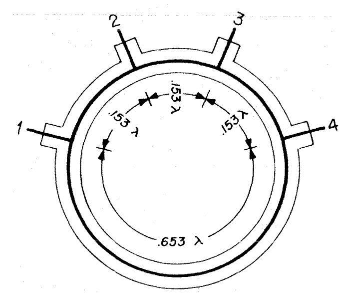


fig. 10. Coax rat race constructed with 50-ohm line.

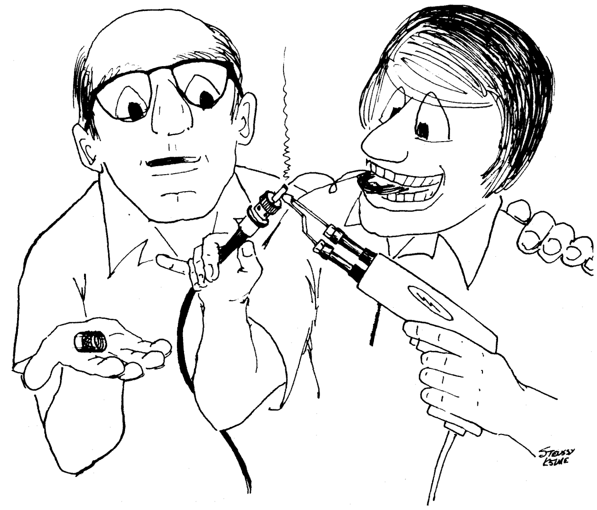
At lower frequencies, the rat race is replaced by a center-tapped transformer. In the higher-frequency regions, where waveguide is used, a device known as the "magic Tee" performs the same function.2

In all regions of the radio spectrum, hybrid devices exist in one form or another, which can contribute much to the versatility of equipment design.

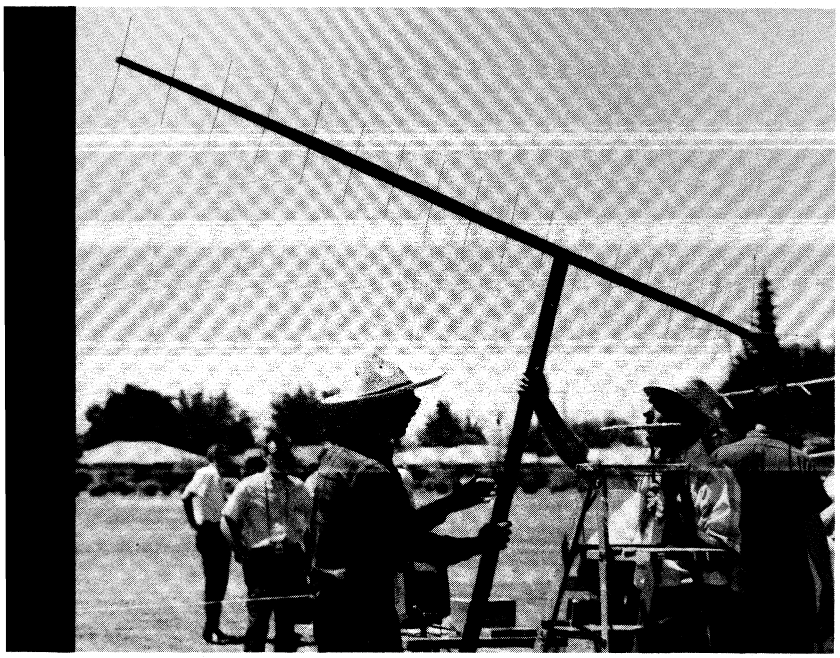
references

1. M Loss, "Phasing Unwanted Images Out of Microwave Recetvers," *Electronics*, July 12, 1965.
2. Keith Henney, "Radio Engineering Handbook." Fifth Edition, McGraw-Hill, New York, pages 6-31.

ham radio



"You forgot something.. ."



direct methods for measuring antenna gain

How to obtain meaningful vhf antenna gain data using simple equipment

For the amateur interested in top station performance on any band, antenna refinement definitely produces the most rewarding return per unit of effort and expense.

Only in the antenna system, which includes the feedline and supporting structure as well as the radiator, can improvements increase performance for both transmitting and receiving. Unfortunately, however, the antenna system is usually the most neglected part of an amateur station. Performance tuning, if done at all, is usually limited to adjusting the driven element length, sliding the clamps on the T match, or adjusting the gamma capacitor for the lowest standing wave ratio. Except for using the swr bridge, *antenna* scope impedance bridge, or field-strength meter, most amateurs seem content to leave antenna tuning to the manufacturers.

The manufacturers can't build antennas to meet all performance demands. Commercially built antennas are designed for "average" installation conditions. All too often these just don't exist in many amateur installations. Most amateurs are plagued by poor soil conductivity, height restrictions, nearby objects, and a host of other adverse conditions that affect antenna performance. These adverse effects can be reduced by tuning the antenna system once you have some dependable quantitative data as a baseline for optimization.

The degree of improvement by tuning is limited with simple antennas. With the more elaborate arrays used above 14 MHz, it's possible to obtain performance increases up to 3 dB with small antennas. Improvements of 7 to 8 dB are possible with larger arrays.

By Bruce Clark, K6JYO (reprinted from the July, 1969, issue of *ham radio)*

The following paragraphs present simple methods for measuring vhf antenna gain directly, with good accuracy. Once you *know* what your antenna is doing, you can make the right adjustments to optimize performance. A few examples are also given of some rather startling results obtained by amateurs who were introduced to these methods.

direct measurements

The average amateur can measure antenna gain with adequate precision using simple equipment. The measurement results are much more meaningful than, say, a measured standing-wave ratio of 1.02-to-l on the transmission line. All this indicates is that the feedline is taking power. The antenna may or may not be radiating in the desired direction or with the desired efficiency.

Of the many methods of measuring antenna gain, two are within the capability of the amateur. These are the attenuatorlreceiver method, and the matched-detector method. Both are comparison tests using a reference antenna and the antenna to be tested. Received signals provide the measurement data.

These methods are more reliable and provide more repeatable data under varying site conditions than those using transmitted signallfield-strength meter or measured-pattern methods.

attenuatorlreceiver method

The attenuatorlreceiver method is block diagrammed in **fig. 1.** Basically, the system uses an accurate attenuator combined with the station receiving system. The signal-source output should be as low as possible and still provide a usable signal at the receiver S-meter when the reference antenna is connected. For most situations 100 mW is adequate. The source should be stable and free of spurious outputs.

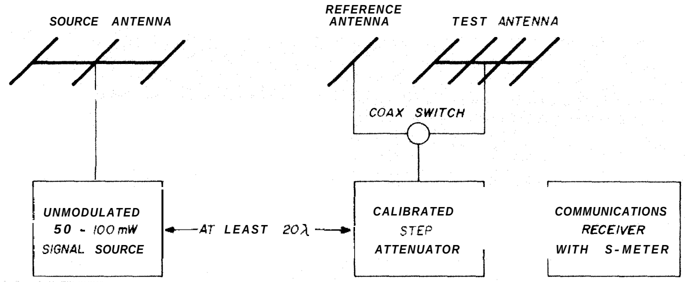


fig. 1. Test equipment for the attenuatorlreceiver method. The source antenna should be as high as possible, in the clear, and at least 20 wavelengths from the antenna under test.

WA6KKK and WB6MGZ aim 18.6-dB 1296-MHz dish.

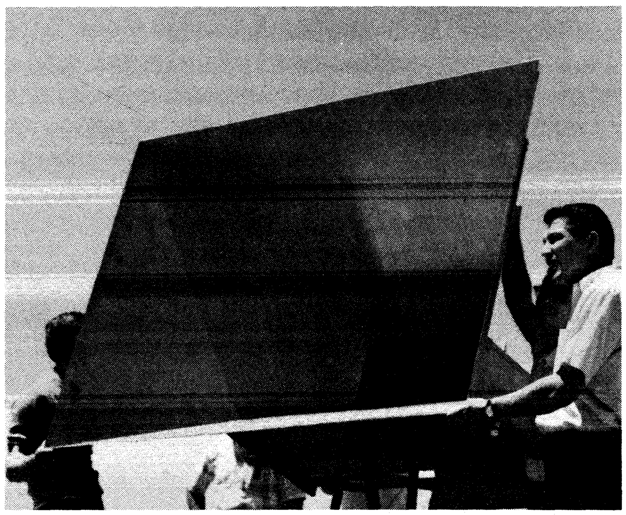
procedure

Set up the source antenna in the clear at least 20 wavelengths from the test antenna. A nearby amateur's tower, flag pole, or TV mast is a good support. Turn on the source, and adjust the attenuator for a reference level on the receiver (anywhere between S-6 and S-9 will do). Record the number of dB used on the attenuator to obtain the reference value on the S meter. Switch to the test antenna, and peak the antenna for maximum signal. Adjust the attenuator for the same S-meter reading obtained with the reference antenna. Record the new attenuator reading. The difference between attenuator readings is the amount of gain (or loss) between the two antennas.

Repeat the process several times, moving the reference antenna for an average level. Note that some variation is introduced by moving the reference antenna. This can be reduced by using a directional source antenna to reduce ground reflection contributions to the received signal (discussed later). In addition, the source antenna should be moved between several different sites at varying distances. Several measurements should be made at each site. The resultant gain figure should be the average of at least six readings.

Note also that feedline losses are included in these measurements. If known, they can be added to the measured antenna gain to get the actual gain of the antenna. Although less impressive, the measured figure is a more practical value, especially above 50 MHz where feedline loss contributions are significant.

The attenuatorlreceiver method will give accuracies on the order of ±1 dB. It's limited by the accuracy and resolution of the attenuators, but is probably the most applicable method for amateur work.



W6MMU's big horn for 432 — 11.7 dB.

The matched detector method is very popular with the West-Coast vhf crew. It requires more sophisticated equipment, but gives greater resolution and quicker readout. The absolute accuracy is still limited by the reference antenna performance due to reflections. The averaging procedure should be used here also if absolute gain figures are desired.

Either a high (1 watt) or low (10 mW) source signal, modulated with a 1-kHz audio tone is used (fig. 2). The type of source determines the detector type. A crystal diode detector similar to a Telonic XD-series, or a home-built equivalent,' will give a square-law output at low input levels. This is ideal for the vswr meter readout.

procedure

The vswr meter is a 1-kHz, sharply tuned, gain-stable, low-noise audio amplifier driving an rms ac vtvm. The 1-kHz modulation is detected and amplified. This signal drives the meter, which is calibrated directly in dB. By adjusting the vswr meter gain range (0-60 dB in 10-dB steps), a reference level can be obtained with the reference antenna. The test antenna is then connected to the detector, and the gain increase (or decrease) noted.

Although the initial cost is high ($200), the vswr meter is available in surplus outlets for approximately $40 to $60 for the earlier Hewlett-Packard HP-415 series. Others by PRD and General Radio are also available.[2]

If the source power is too high, the crystal diode detector will be driven out of the non-linear, square-law portion of its curve. The resultant output will deviate, and the vswr meter reading will be high. This can be prevented by:

1. Keeping the source power output very low.
2. Inserting a calibrated attenuator (3 to 6 dB) ahead of the detector mount (fig. 2).
3. Using the vswr meter with a wider range detector called a bolometer (thermistor) mount. Although not as sensitive, the bolometer mount provides good results when used with sources of 1 to 2 watts output.

reference antennas

No study of antenna gain measurement would be complete without a word on reference antennas. Classically, the isotropic radiator is a point source that illuminates all points equally on the inside surface of a sphere. It is used as the reference antenna in nearly all theoretical work. However, it's not possible to produce such an antenna, so the matched 112-wave dipole is often used as a reference antenna.

The dipole has a disadvantage. Because of its broad pattern, it's extremely sensitive to ground effects and to near-field reflections from the signal source. These reflections add or detract from the desired free-space signal and produce an output that varies from the ideal (average) value.

NBS standard reference antenna

Absolute accuracy of measurements depends on the accuracy of the reference dipole, so it is im-

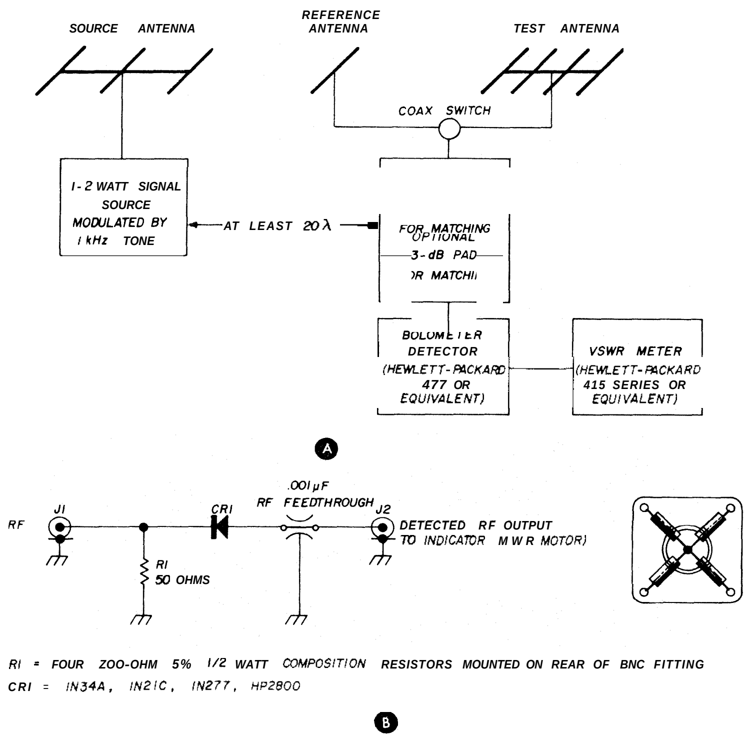


fig. 2. Setup for the matched detector method. A bolometer is used in (A). An easily constructed diode detector substitute for the bolometer is shown in (B). The 3-dB pad will improve the match between detector and antenna, especially at uhf.

portant to average the reference dipole readings under different site conditions. Recently, highly accurate standard reference antennas have been designed and employed by the National Bureau of Standards (NBS) and some amateurs, among them W6VSV and W6HPH. Basically a simple directional array designed for low side-lobe content and high front-to-back ratio, the NBS standard antenna has a gain of 7.7 dB over a reference dipole, measured under laboratory conditions in an anechoic chamber (see fig. 3).

The measurement repeatability is on the order of ±0.1 dB or better. The NBS standard antenna is used in a manner identical to that of the reference dipole, but there is less variation due to reflections. Also, one must remember to add the 7.7 dB reference-antenna gain figure to those from the vswr meter with the test antenna in the line. For example, if the test antenna measures 2.3 dB when the reference antenna measures 0 dB, the antenna gain is 10 dB.

results

These techniques are regularly employed by top vhf-uhf amateurs to obtain the most from homebrew and commercial arrays. In the past few years, antenna contests at hamfests have become popular proving grounds where new winning combinations have been discovered. A case in point is the reawakened popularity of the Yagi antenna at 432 MHz. It has resulted from careful optimization of several scaled-down designs that didn't work at all (or poorly at best). Another case is the 1 to 2 dB gain increase from adding directors to collinear arrays – a method now adopted by at least one manufacturer.

The accuracy of the results is amazing. My own 32-element, 432-MHz array measured 15 dB at the West-Coast Uhf Conference in Fresno and 16.2 dB at the Hughes Radio Club contest in Fullerton (after some matching deficiencies were discovered).

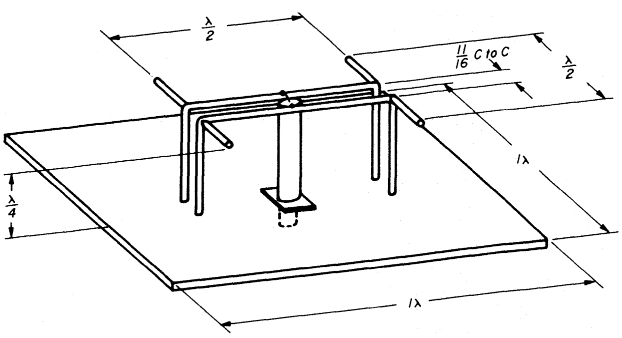
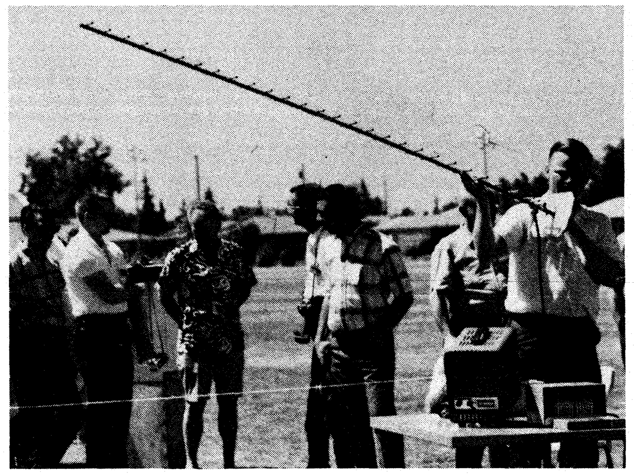


fig. 3. Layout of the standard gain antenna used by National Bureau of Standards. When accurately built. this reference antenna will provide 7.7 dB over an isotropic radiator 10.25 dB. Element diameter is about 0.01λ (3/8" or 10mm at 432 MHz).



K7ICW's 30-element Yagi for 1296 MHz yielded – 2.5 dB!

As for the repeatability of results from site-to-site, tests of the popular 6-foot boom Tilton Yagi at 432 MHz resulted in consistent measurements yielding 12 to 13 dB in contests from Missouri to California. W5ORH's twin bi-square beam measured 8.0 dB at three different sites using three different test methods. These examples are exceptions. Typically, however, results haven't varied more than ±2 dB when good equipment and normal care were used in making the measurements.

some surprises

At one contest several owners of supposedly high-gain commercial arrays really had their eyes opened. One 432-MHz Yagi, with a manufacturer's claim of "over 17 dB forward gain," measured *negative* 2 dB off the front and +6 dB off the back. Cutting the antenna in half got about +8 dB forward gain.

Another homebrew 13-element Yagi from a popular vhf handbook measured +1.9 dB gain over a dipole. (The owner had substituted a wooden boom for the original metal boom and hadn't reduced the element lengths to compensate. Trimming the elements and matching the feed brought the gain up to 12.3 dB – not a bad increase.)

It should be obvious that antenna gain measurement is worthwhile for the amateur. From my experience, it gets results we all desire: better reports and more consistent contacts.

references

1. Fred W. Brown, W6HPH, "The Matched Detector," VHF'er, June, 1965.

2. Robert Stein, W6NBI, "Using the SWR Indicator," *ham radio*, January. 1977. page 66.

3. Loren Parks, K7AAD, "The Reference Dipole," *VHF'er*, April, 1965.

ham radio

graphical solution
of impedance-matching problems

Using simple geometry to design and analyze a variety of impedance-matching networks

One of the most common problems in radio circuits is matching one impedance to another. The problem might be that of matching a transmitter output stage to a resistive load, or the load may have a reactive component, as is usually the case when attempting to transfer power to an antenna.

Many articles have been written covering the mathematics of this problem and also the application of the Smith chart.[1] Impedance-matching problems can be solved readily with sufficient accuracy for practical purposes with no more equipment than a straightedge, compass, and graph paper. The graphical method lends itself to multiple-component networks involving complex impedances, without resorting to trigonometry or complex algebra. It allows a visual choice of constants and shows forbidden approaches in choosing impedance paths.

The method presented in this article will allow you to solve most impedance problems encountered in amateur work. The geometric principles are easy to follow, and you'll need to make only a few simple computations. Rules are given for constructing the diagrams. Typical examples and solutions are shown. The examples are presented without mathematical proof, however. For those who wish to pursue the classical approach, some excellent material will be found in references 2, 3, and 4.

a starting point

First consider the familiar methods known as the "leaning ladder" diagram for determining the resultant of two resistors or reactances in parallel (**fig. 1**).

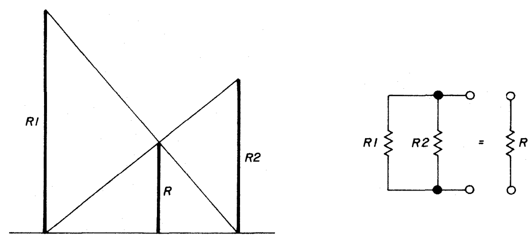


fig. 1. The "leaning ladder" diagram for finding the resultant of two resistances or reactances in parallel.

Two perpendicular lines with lengths proportional to the two resistors or reactances are erected with arbitrary separation from a common baseline. Lines are then drawn from the top of each perpendicular to the base of the other. A third perpendicular is now drawn from the intersection of these lines to the baseline. The length of this new perpendicular is proportional to the combined resistance or reactance of the two parallel elements.

What happens, however, when two reactances of opposite sign are to be evaluated? The same pro-

By **I. L. McNally, W1NCK**, and **Henry S. Keen, W2CTK** (reprinted from the December, 1969, issue of *ham radio)*

cedure is followed as before, except that the perpendicular lines representing the reactances will be located on opposite sides of the baseline **(fig. 2)**. Again connect the end of each perpendicular to the base of the other, extending the lines until they intersect. The length of a perpendicular from this point of intersection to the baseline represents the combined reactance of the two paralleled elements. The side of the baseline where the intersection takes place determines whether the resultant, X_R, is inductive or capacitive.

Now suppose a reactance is to be paralleled with a resistance. How do you determine the impedance of such a combination? Semicircles are constructed upon rectangular coordinates, with diameters proportional to the paralleled resistance and reactance, intersecting at point **A (fig. 3)**. A line, **0-A,** from the

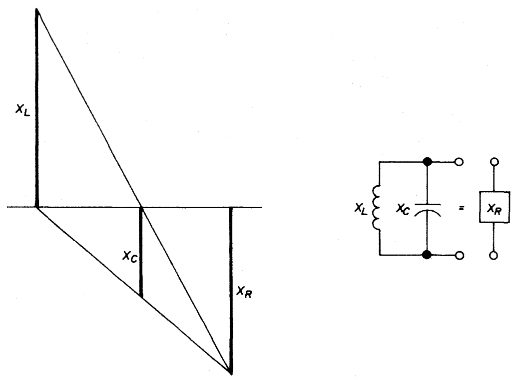


fig. 2. Diagram for finding resultant impedance of two reactances of opposite sign.

origin to the point of intersection, will be proportional to the impedance of the combination. The projections of this point of intersection upon the resistive and reactive axes will then be proportional to the resistance, R_S, and reactance, X_S, respectively, which make up the series equivalent of the parallel combination.

Because an angle inscribed in a semicircle is always a right angle, it is easily shown that the point of intersection, **A,** lies on a straight line connecting the ends of the two diameters. This construction leads to a well-known diagram frequently used to solve L networks, **(fig. 4)**. An L network is merely a transformation from a parallel resonant circuit, seen looking in at $Z1$, to a series resonant circuit, seen looking in at 22.

rules for construction

By combining these diagrams, it's possible to solve a variety of matching-network problems. The geometry of **fig.** 5 is the basis of solving all problems us-

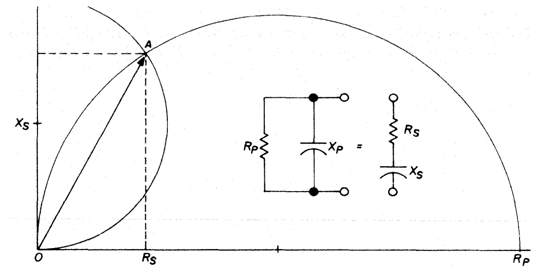


fig. 3. Geometry for solving parallel-to-series transformation.

ing this method. General rules for using the method are:

1. Adding a series of reactances moves the impedance on a vertical line — up for inductive reactance and down for capacitive reactance.

2. Adding a parallel reactance moves the impedance along a circle with its center on the horizontal axis. It rotates clockwise for capacitive reactance and counter clockwise for inductive reactance.

3. When choosing impedance paths, it is not permissible to use a path passing through the origin of coordinates.

The method permits rapid comparison of different network designs without a knowledge of complex algebra, and a clear picture is given of what happens when parameters are modified.

We'll begin with the pi network since this is one of the most-used circuits in amateur work. Other circuits will then be described which will provide a foundation for solving most impedance-matching problems. Some numerical examples will then be given to show step-by-step procedures.

pi networks

The pi network can be considered as two cascaded L networks, designed to transform both input and output impedances to a common internal transfer im-

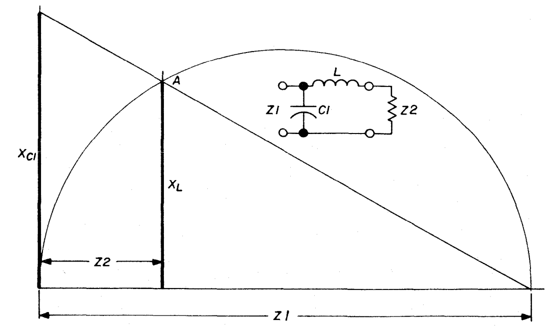


fig. 3. Geometry for solving parallel-to-series transformation.

pedance, which must be lower than either terminal impedance. This internal transfer impedance determines the network Q, a fact that becomes apparent from a consideration of the design diagram (**fig. 6**).

To design a pi network, begin at the origin of a set of rectangular coordinates, and construct a semicircle above the horizontal axis, with diameter proportional to $Z1$. ($Z1$ is the greater of the two terminal impedances.) Similarly, from the origin construct a second semicircle below the axis. Its diameter is proportional to $Z2$, the lesser of the two terminal impedances.

Because an infinite number of solutions exist to a pi-network probiem when terminai impedances are specified, an assumption must be made for one of the three reactances. This is necessary to establish the internal transfer impedance. There are certain advantages if the reactance of the output capacitor, C2, is made equal to the load resistance, $Z2$. However, network Q requirements frequently dictate a lower value as discussed later.

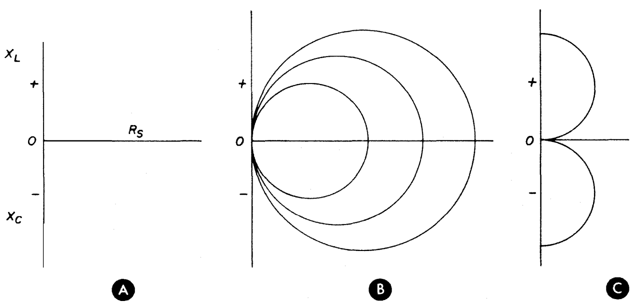


fig. 5. Basic geometry for graphical solution of impedance problems. Series resistance is plotted along the horizontal axis and series reactance on the vertical axis, as at (*A*). Parallel resistance and parallel reactance circles are constructed as in (*B*) and (*C*) respectively.

The assumed reactance X_{C2}, of the output capacitor becomes the diameter of a third semicircle, beginning at the origin and constructed downward below the horizontal axis. The point of intersection between this and the $Z2$ semicircle is point **A.** From this point a vertical line is drawn to intersect the original $Z1$ semicircle at point **B.** The length of the line segment, AB, represents the required reactance of X_L.

A straight line is now drawn from the extreme end of the $Z1$ diameter through point **B,** intersecting the vertical axis at point C. Line OC will then be proportional to X_{C1}, the reactance of the required input capacitor.

The intersection of inductive reactance line **AB** with the horizontal axis is point D. The significance of this point is that line segment OD represents the internal transfer impedance of the network. The Q of

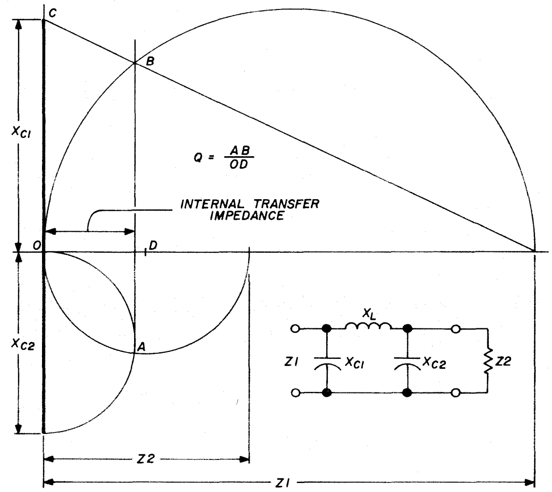


fig. 6. The pi network diagram. Circuit Q is determined by the internal transfer impedance.

the network, when driven by a current generator such as a screen-grid tube or a transistor, will be equal to the inductive reactance, **AB,** divided by the internal transfer impedance, OD. The Q will also be equal to $Z1/X_{C1}$ plus $Z2/X_{C2}$, which can be proven identical.

When driven by a resistive source, such as a triode, the network is loaded from both ends, and the effective Q may be cut in half.

tee networks

Although the T network is not as well known as the pi network, it is a very usefui circuit and is quickly solved graphically. With the T network, we may assume the internal transfer impedance as equal to or greater than the sum of the terminal impedances $Z1$ and $Z2$, usually by a factor of two or more. The graphical design procedure, with reference to **fig.** 7 is as follows:

Construct a semicircle with horizontal diameter *greater* than the sum of the terminal impedances.

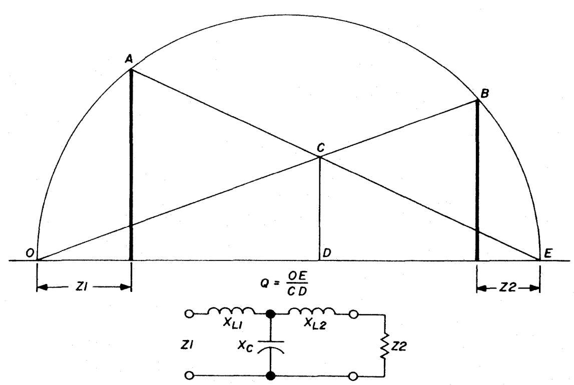


fig. 7. Solving the T network. The internal transfer impedance is equal to or greater than $Z1 + Z2$ by a factor of two.

Mark off, from opposite ends of the diameter, line segments proportional to the two terminal impedances. From these two points erect perpendiculars to intersect the semicircle at points **A** and B respectively. Connect points **A** and B to the remote ends of the diameter, intersecting each other at **C.** A perpendicular, **CD,** to the diameter will be proportional to the reactance of the capacitor, **C.** The Q of this network, as driven by a current generator, wiii be equal to *the* diameter of the semicircle divided by the line segment **CD.** The sum of $X_{L1}/Z1$ plus $X_{L2}/Z2$ will give an identical result.

Although the derivation of this diagram may seem obscure, if perpendiculars are erected at the ends of the diameter, and the slant lines extended to intersect these perpendiculars, we will have the two super-imposed L-network diagrams. The line segments of these end perpendiculars will each represent a capacitive reactance corresponding to one of the two cascaded L networks making up the complete T network. The extended slant lines can then be seen to represent the leaning-ladder diagram, with line **CD** being the result of both capacitive components in parallel. All construction exterior to the semicircle, therefore, will be redundant and can be omitted.

If the semicircle is constructed so that its diameter is equal to the sum of the two terminal impedances, all reactances will be of the same magnitude, differing only in sign, and will be equal to the geometric mean of the terminal impedances. The Q of such a network would be quite low, being equal to the sum of the two terminal impedances divided by the square root of their product.

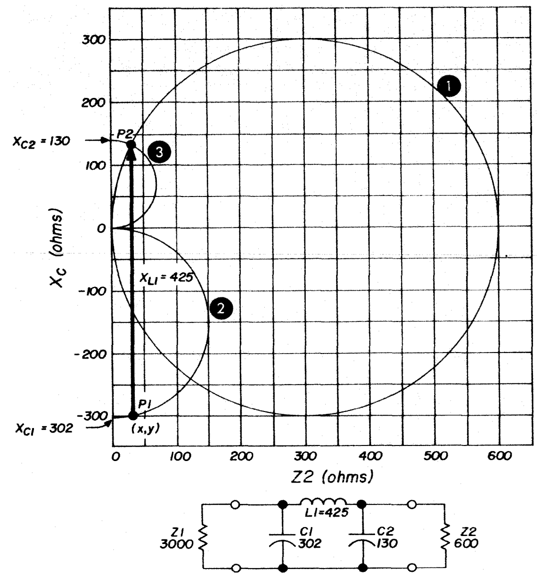


fig. 8. Solution to problem I: matching 3000 + j0 ohms to 600 + j0 ohms with a pi network. Arrows indicate impedance path.

When the pi network is designed so that $22 = X_{C2}$, excursions of 22 will have minimum effect on $Z1$. Resonance will be maintained by retuning X_L. A network is possible whereby a two-to-one range of 22 (assumed purely resistive) will, in turn, cause $Z1$ to vary from the target impedance by less than five per cent.

Similarly, design of the T networks so that $Z1 = X_{L1}$ will permit $Z1$ variations of the same magnitude, with the network output still presenting a match to the load (within the same limits). Resonance is maintained by retuning C1.

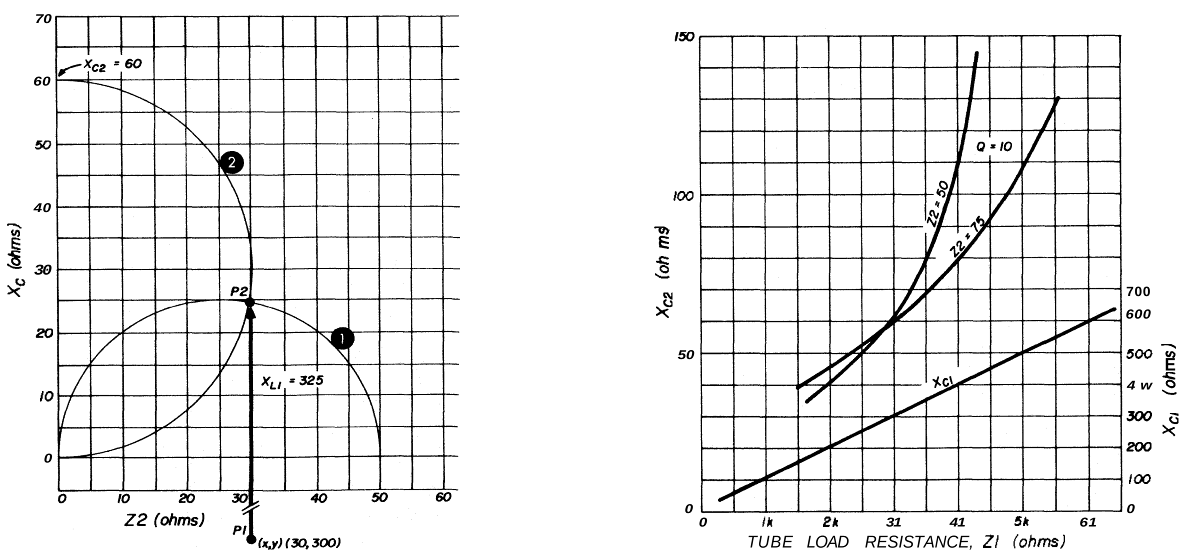


fig. 9. Solution to problem 2: matching 3000 + j0 ohms to 50 + j0 ohms with a pi network – a common problem in transferring tube output impedance to an antenna transmission line. Expanded scales for R2 = 50 ohms are shown at (*A*). At (*B*) the curves are limited to R2 = 50 or 75 ohms and Q = 10.

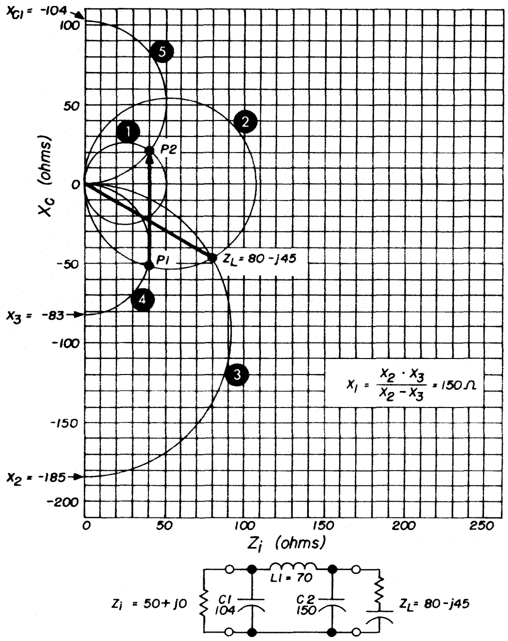


fig. 10. Solution to problem 3: matching an antenna load of 80 − j45 ohms to 50 + j0 ohms with a pi network. This has a three step impedance path: Z_l to P1 to P2 to Zi.

In either network the terminal impedance, $Z1$, is assumed the higher of the two. Although $Z1$ has been treated at the input end, either network is completely reciprocal.

examples using pi networks

Problem 1. Match 3000 + j0 ohms to 600 + j0 ohms with a pi network. In this case, $Q = 10$.

1. Draw a 600-ohm circle (1), fig. 8.

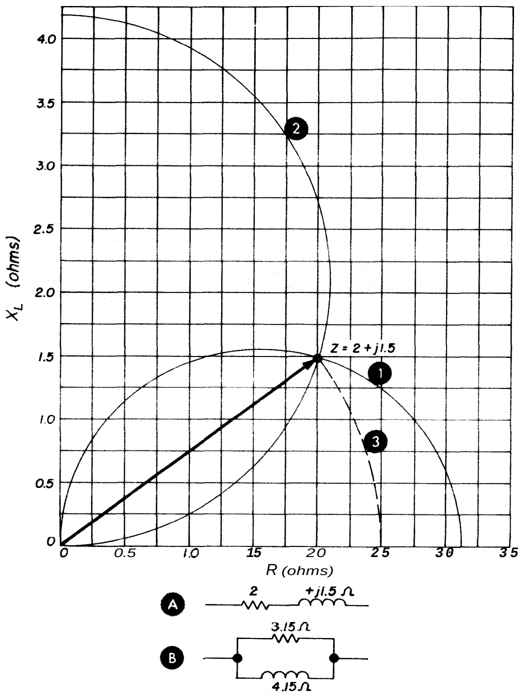


fig. 12. Solution of series-to-parallel transformation. The resultant impedance is 2.5 ohms.

2. Calculate point (x,y) and plot:

$$y = \frac{R1}{Q} = \frac{3000}{10} = 300$$

$$x = \frac{R1}{Q^2} = \frac{3000}{100} = 30$$

3. Erect a vertical line from point P1 (x,y) to intersect the 600-ohm circle at point P2. This is X_L to scale (425 ohms).

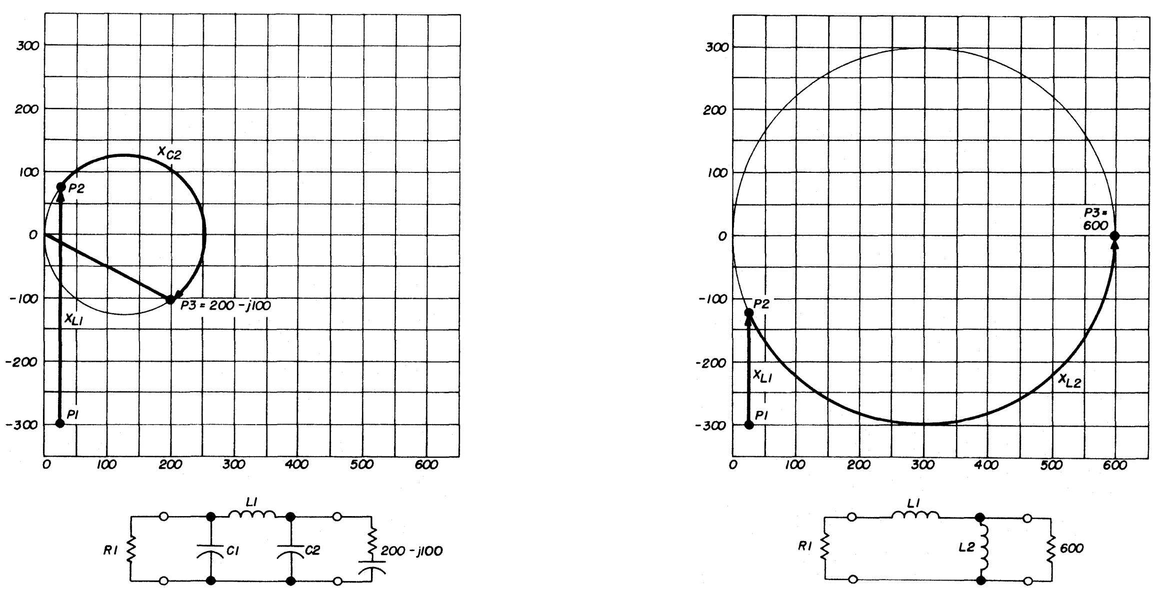


fig. 11. Examples of correct impedance paths. Both show networks for matching typical power amplifier tube impedances to various loads. Note that the impedance paths must not pass through the origin of coordinates.

4. Draw circle (2). Its intercept on the vertical axis at 302 ohms is X_{C1}.

5. Draw circle (3). Its intercept on the vertical axis at 130 ohms is X_{C2}.

6. Solution: L1 = 425 ohms
   C1 = 302 ohms
   C2 = 130 ohms

Problem 2. This is the same as problem 1, except $Z2 = 50\ ohms$ (fig. **9**).

1. Draw a 50-ohm circle (1), fig. **9A**.

2. Erect X_{L1} through X = 30 to intersect the 50-ohm circle at point **P2.** This scales to y + 25, or X_{L1} = 325 ohms.

3. Draw circle (2) through point **P2.** It will intersect the vertical axis at X_{C2} = 60 ohms.

4. Solution: L1 = 325 ohms
   C1 = 302 ohms
   C2 = 60 ohms

Some pi-network curves for common tube load resistance are shown in fig. **9B**.

Problem 3. Match an antenna load of 80 – j45 ohms to 50 ohms using a pi network (fig. **10).**

1. Construct a 50-ohm circle (1).

2. Plot Z_L = 80 – j45 ohms.

3. Construct circle (2) through Z_L.

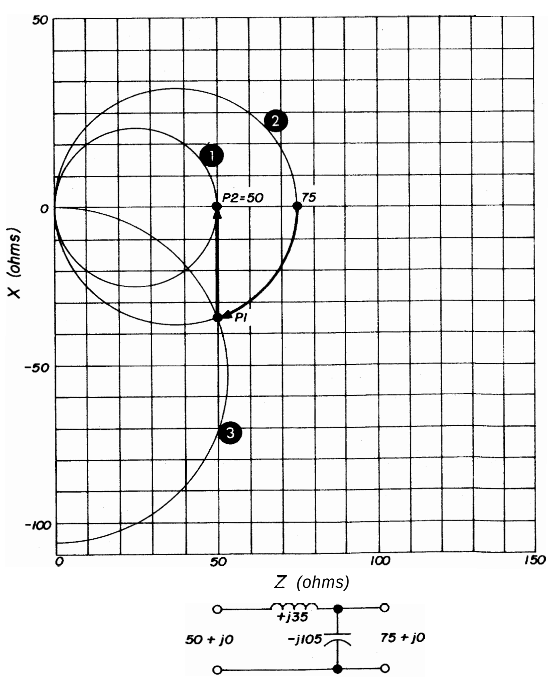


fig. 13. Examples of 2-step impedance path from 75 ohms to 50 ohms in an L network.

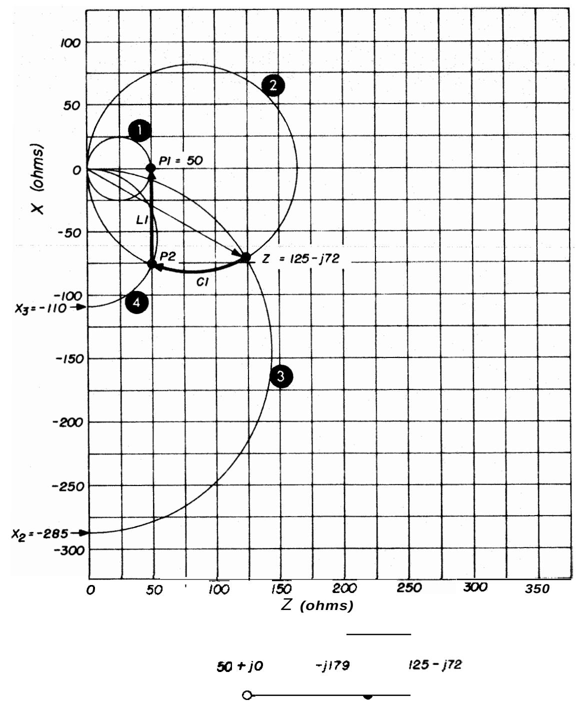


fig. 14. Matching a 50-ohm resistive load to a complex load. The impedance path is from Z_L to P2 to P1.

4. By inspection the maximum value of X_L is about 75 ohms. Select a value of 70 ohms and fit it vertically as **P1**, **P2** between circles (1) and (2).

5. Construct circle (3) through Z_L. It will intersect the vertical axis at – 185 ohms ($X2$).

6. Construct circle (4) through **P1**. It will intersect the vertical axis at – 83 ohms ($X3$).

$$X1 = \frac{X2 \bullet X3}{X2 - X3}$$

$$= \frac{(-185)\ (-83)}{(-185) - (-83)} = \frac{185 \bullet 83}{-102}$$

$$= -150.5 \text{ ohms}$$

This is a capacitive reactance added by moving from Z_L to **P1**, along circle (2).

choice of impedance paths

Recall that the internal transfer impedance must be lower than either terminal impedance. The internal transfer impedance determines the Q of the pi network. From fig. **6,** line segment OD determines this parameter. Therefore, from the rules of construction for graphical solution to these problems, it is not permissible to choose an impedance path through the origin of coordinates. Examples of correct impedance paths are shown in fig. **11;** hence, these show transformation between typical tube output impedances and various load impedances. Note that

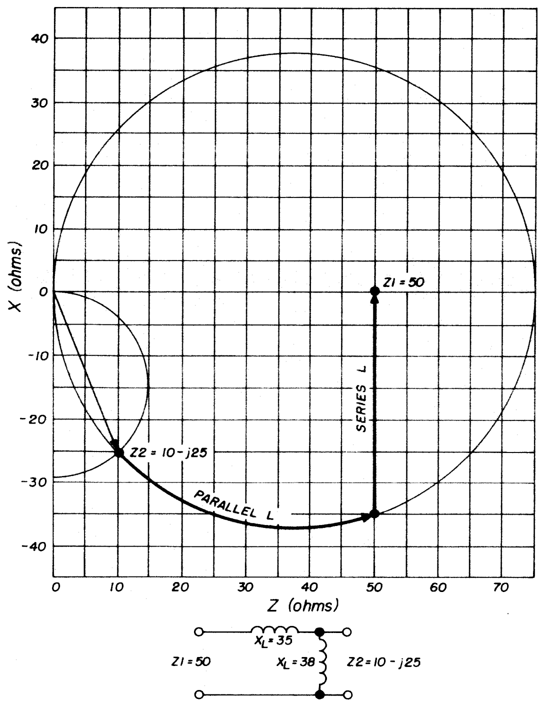


fig. 15. Another example of impedance paths in network design. Three different 2-step paths are shown from $Z2$ to $Z1$ as well as the "forbidden" path through the origin.

the paths do not pass through the origin. While a 600-ohm terminal impedance is not too common in most rf circuits these days, the example does indicate the principles to be followed when designing these networks.

solving series and L networks

Problem 1. Given the series circuit of fig. **12A**, find the equivalent parallel circuit.

1. Plot the impedance vector $Z = 2 + j1.5$.
2. Construct circle (1) with its center on the horizontal axis, which passes through the origin and Z as shown. It will intersect the horizontal axis at 3.15 ohms resistance.
3. Construct circle (2) with its center on the vertical axis, which passes through the origin and Z. It will intersect the vertical axis at 4.15 ohms inductive reactance. The equivalent parallel circuit is shown in fig. **12B.**
4. Solution. Scaling the Z vector gives an impedance of 2.5 ohms (3).

Problem 2. In the network of fig. **13,** it is desired to find X_L, X_C, and C for a frequency of 3.9 MHz.

1. Construct circles (1) and (2) through 50 ohms and 75 ohms as shown.
2. Construct line **P1**-**P2**, which is X_L series and scales 35 ohms.
3. Construct circle (3) through **P1**. It will intersect the vertical axis at – 105 ohms. This is X_C parallel capacitive reactance obtained in moving clockwise from 75 ohms along circle (2) to **P1,** which is directly below **P2,** the 50-ohm point.
4. Solution.

$$L = \frac{X_L}{2\pi f} = \frac{35}{2\pi \times 3.9} = 1.43\ \mu H$$

$$C = \frac{10^6}{2\pi f X_C} = \frac{10^6}{2\pi \times 3.9 \times 105} = 386\ pF$$

$X_L = 35$ ohms

$X_C = 105$ ohms

Problem 3. Match 50 +j0 ohm to $Z_L = 125 - j72$ ohms, a complex load (fig. **14**).

1. Construct a 50-ohm circle (1).
2. Plot $Z_L = 125 - j71$
3. Construct circle (2) through Z_L.
4. Construct a vertical line from **P2** to **P1**. This is the series X_L and scales 77 ohms.
5. Construct circle (3) through the origin and Z_L. It will intersect the vertical axis at – 285 ohms ($X2$).

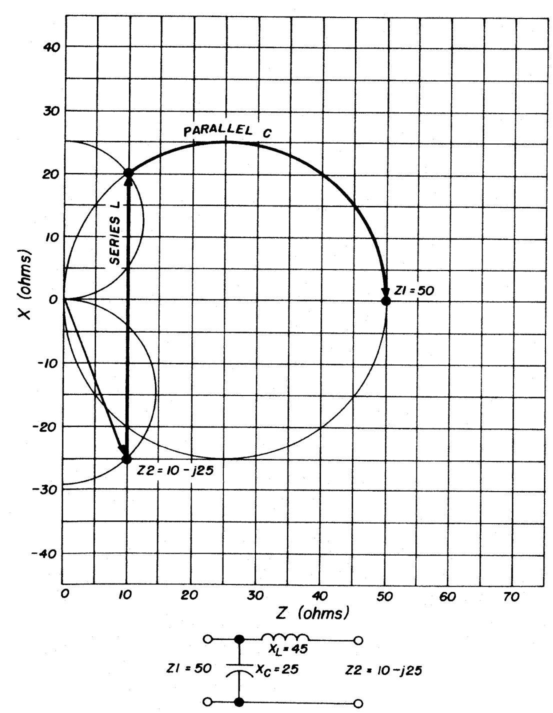


fig. 15A. Impedance path through matching network consisting of series L and parallel C.

6. Construct circle (4) through the origin and **P2.** It will intersect the vertical axis at – 110 ohms ($X3$).

7. When neither of the terminal points is on the horizontal axis, as in this case with Z_L and **P2,** it is necessary to compute the value of reactance involved in moving from $Z2$ to **P2.**

$$X1 = \frac{X2 \bullet X3}{X2 - X3}$$

$X1$ = *added reactance*

$X2$ = *initial reactance*

$X3$ = *final reactance*

$$X1 = \frac{(-285)\ (-110)}{(-285)-(-110)}$$

$$= -179 \text{ capacitive}$$

8. Solution $L1 = j77$ *ohms*, and $C1 = j170$ *ohms*.

In **fig. 15** are examples of three different choices of impedance paths in going from $Z2$ to $Z1$ using three network combinations to match a resistive 50-ohm load to a complex load of 10 – j25 ohms. Again, the "forbidden" path is not to be used because it passes through the origin.

summary

We have shown examples of solving the most common impedance-matching problems using simple geometric methods. The following notes are offered in adopting these methods for solving a wide variety of problems.

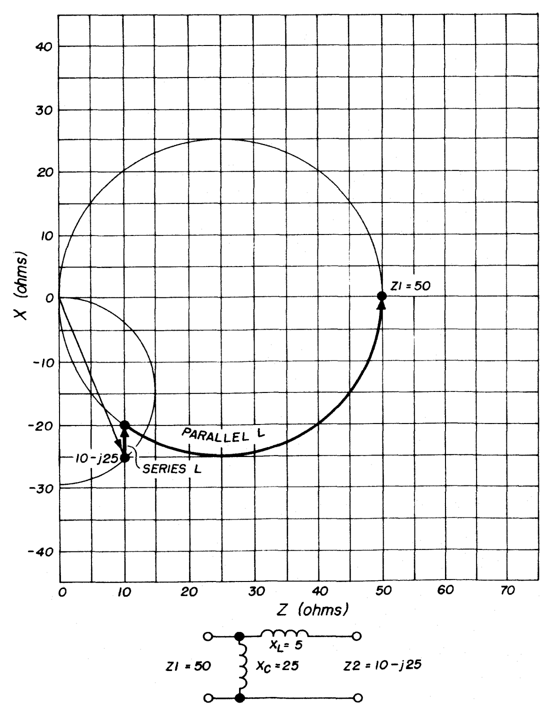


fig. 158. Impedance path through matching network consisting of series and parallel *L*, respectively.

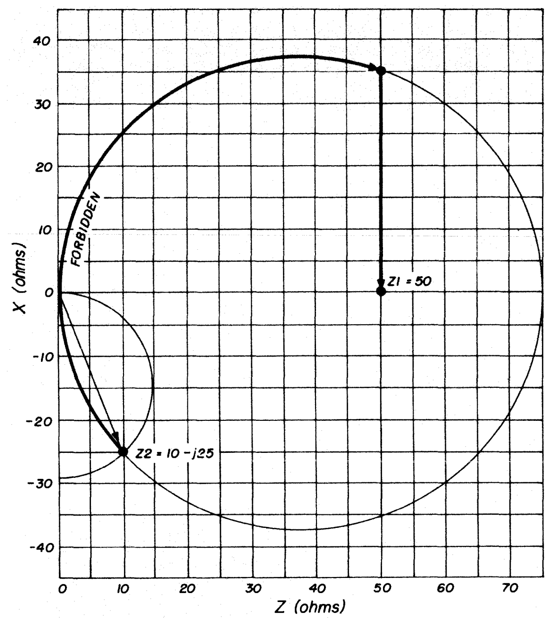


fig. 15C. Impedance path through parallel and series *L*, respectively.

1. The L-network is one of the most useful circuits known for matching nearly all direct-coupled tank systems. The examples show how to match a high-resistance to a low-impedance reactive load. If the converse is desired, it is only necessary to convert the reactive load to its equivalent parallel components.

2. T networks are useful as harmonic attenuators in low-impedance transmission lines. These can be readily solved by treating them as two cascaded L sections and combining the capacitances.

3. The pi network is used to match a wide range of load impedances with reasonable tank-circuit *Q*. Contrary to some popular notions, the pi network will not match a tube to any length of wire. The circuit is load-limited by the ratio of tube load and circuit *Q* if it is to perform as an efficient transformer.3

references

1. J. R. Fisk, W1DTY, "How to Use the Smith Chart," *ham radio*, November, 1970, page 16 (reprinted this issue).
2. F. E. Terman, "Electronic and Radio Engineering," McGraw-Hill, New York, 1955.
3. Keith Henney, "Radio Engineering Handbook," McGraw-Hill, New York, 1959.
4. C. E. Magnusson, "Alternating Currents," McGraw-Hill, New York, 1931.

ham radio

how to use the Smith Chart

A discussion of the Smith chart with examples of its use in transmission-line problems

By James R. Fisk, W1HR, (reprinted from the November, 1970, issue of *ham radio)*

Although articles on the Smith chart have appeared in the amateur magazines from time to time, amateurs have made little use of this handy transmission-line calculator – probably because it has been difficult to measure complex impedances with simple homebuilt equipment. However, this problem has been solved with the simple impedance bridge described by W2CTK – at least for the high-frequency range.[1] With careful attention to lead dress and component layout this instrument should be usable on six and two meters as well.

A quick glance at the Smith chart suggests a formidable array of curved lines and circles that would cause the most hardened technician to go into fits of despair. On the other hand, if you spend a little time with the chart and look at each of its component parts, it's not really very complicated. Perhaps the one thing that scares many prospective users is its unfamiliar circular shape; it's not at all like the straight-line graphs you're accustomed to. However, when you understand the chart and have mastered its use you'll be able to solve complex impedance and transmission-line problems much easier and faster than ever before.

layout of the chart

The Smith chart is basically a circle which contains various circular scales. The horizontal line through the center marked *resistance component* is the only straight line on the chart and is called the "axis of reals" (see fig. 1). Constant resistance circles are centered on the axis of reals, tangent to the rim of the chart at the infinite resistance point. All the points along a constant-resistance circle have the same resistive value as the point where it crosses the axis of reals.

Superimposed over the resistance-circle pattern are portions of other circles tangent to the axis of reals at the infinite resistance point, but centered off the edge of the chart (fig. 2). The large outer rim of the chart is calibrated in relative reactance and is called the "reactance axis." Any point along the same constant-reactance circle has the same reactive value as the point where it intersects the reactance axis on the rim of the chart. All points on the Smith chart above the axis of reals contain an inductive-reactive component and those below the axis of reals contain a capacitive-reactive component. Since the calibration points go from zero to infinity, any complex impedance can be plotted on the chart.

The impedance coordinates on the Smith chart would be of little use without the accompanying peripheral scales (fig. 3). These scales relate to quantities which change with position along a transmission line. Two scales are calibrated in terms of wavelength along the transmission line: one, in a clockwise direction, is "wavelengths toward generator," and the other, counter-clockwise, is "wavelengths toward load." The entire length of the circumference of the chart represents one-half wavelength.

normalized numbers

Normalized values must be used when plotting impedances on the Smith chart.* Normalized impedance is defined as the actual impedance divided by the characteristic impedance of the transmission line.

Normalizing is done to make the chart applicable to transmission lines of any and all possible values of characteristic impedance. For example, a 50-ohm coaxial transmission has a normalized value of 50/50 or 1. On this basis an impedance of 120 ohms would have a normalized value of $120/50 = 2.4$ ohms. Similarly, $z = 0.8$ ohms (the lower case indicates a normalized value) would correspond to a value of 0.8 times the characteristic impedance of the line or $0.8 \times 50 = 40$ ohms.

What has been said about coaxial cable with regard to normalized impedance applies equally to waveguide, where a characteristic impedance of 400 ohms at a specific frequency would be considered unity in normalized form. All other values would be related to this value, so that a 560-ohm component would have the value $560/400 = 1.4$ ohms in normalized terminology, while $z = 0.9$ in normalized form would actually be $0.9 \times 400 = 360$ ohms.

plotting values on the chart

Any complex impedance, regardless of value, may be plotted on the Smith chart. For example, assume the load on a 50-ohm transmission line is 42.5 – j31.5 ohms. This is equal to 0.85 – j0.63 when normalized. To plot this point on the chart, locate 0.85 on the axis of reals and note the corresponding constant-resistance circle **(fig. 4)**. Next locate 0.63 on the periphery of the chart. The quantity $(-j)$ indicates a capacitive-reactive component so the value 0.63 is on the lower half of the chart. Note the constant-reactance circle representing – j0.63. The complex impedance 0.85 – j0.63 is at the intersection of the constant-resistance and constant-reactance circles.

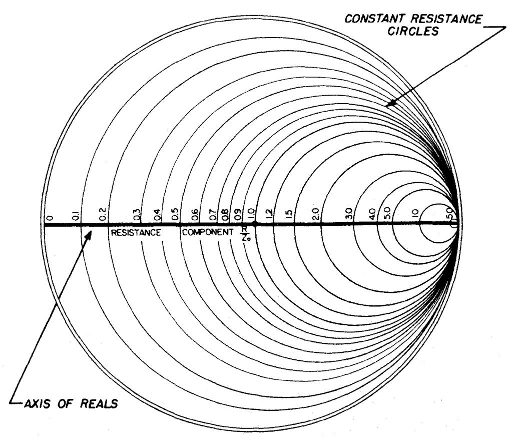


fig. 1. Smith chart resistance scales.

Draw a line from the center of the chart through this point to the outer rim. With the point 1.0 on the axis of reals as the center, scribe a circle that intersects the impedance point. This circle is known as the "constant-gamma circle," and its radius is equal to the coefficient of reflection. The constant-gamma circle crosses the axis of reals at two points; the point

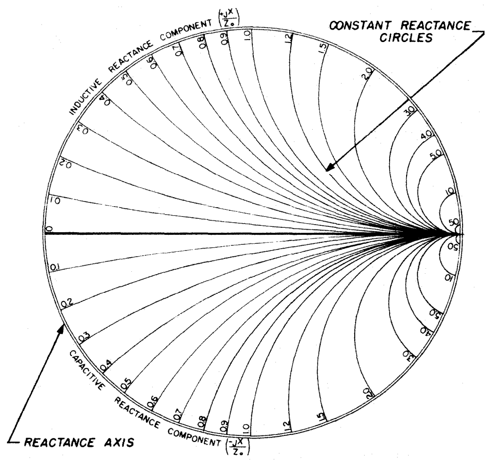


fig. 2. Smith chart reactance scales.

of intersection to the right of center is the standing wave ratio (2.0 in this case).

If the voltage were measured at this point on the transmission line, it would be found to be a maximum. Conversely, the point of intersection one-quarter wavelength away on the left-hand axis of reals is a point of voltage minimum (this point is also equal mathematically to the reciprocal of the swr).

The point at the intersection of the radial line and the angle of reflection coefficient scale represents the phase of the coefficient of reflection. This is the angle by which the reflected wave leads or lags the incident wave. When these two waves add in phase to give maximum voltage, the impedance is resistive and greater than the characteristic impedance of the line and the angle of the coefficient of reflection is zero.

As you move away from the zero-phase-angle point in a clockwise direction toward the generator, the reflected voltage lags the incident voltage, and the phase angle is negative for the first quarter wavelength. The reactive component of the impedance in this region is negative or capacitive.

At the quarter-wavelength (90°) point the incident and reflected waves are out of phase and the angle of the coefficient of reflection is $\pm 180°$. As you continue in a clockwise direction the two waves become

*Since 50-ohm systems are standard for military and industrial use, 50-ohm Smith charts are available. On a 50-ohm Smith chart the center point has a value of 50 ohms.

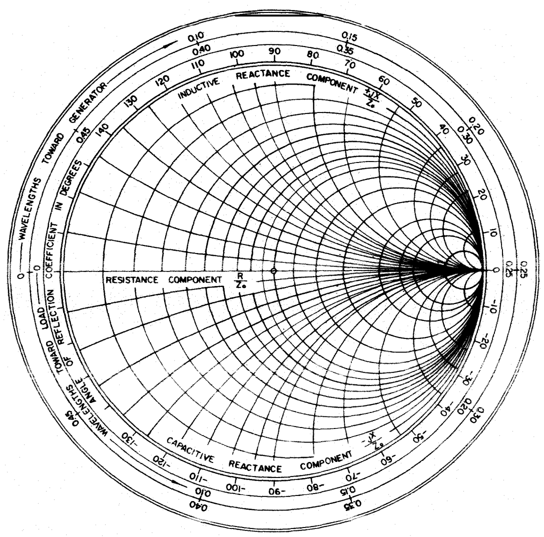


fig. 3. Smith chart peripheral scales.

increasingly more in phase and between one-quarter and one-half wavelength from the voltage maximum the reactive component is inductive, the reflected wave leads the incident wave, and the reflection coefficient has a positive angle.

A number of parameters are uniquely related to one another as well as to the magnitude of reflections from the load and are conveniently plotted as scales at the bottom of the Smith chart. These parameters are vswr, coefficient of reflection, vswr in dB, reflection loss in dB, and attenuation in 1-dB steps.

using the smith chart

The general utility of the Smith chart is best illustrated by showing examples of its more common uses. Use of the radially-scaled parameters will be shown in the same way.

Example 1. Finding standing-wave ratio. A 75-ohm transmission line is terminated with a load impedance $Z_L = 30 - j90$ ohms. What is the swr? (See **fig. 5.**)

1. Normalize the load impedance by dividing by 75

$$\frac{30-j90}{75} = 0.4 - j1.2$$

2. Locate this point on the chart.

3. Construct a constant-gamma circle so its circumference passes through this point.

4. The swr is defined by the point where the constant-gamma circle crosses the axis of reals on the right-hand side. In this case swr = 6.4.

5. The swr may also be determined with the radial nomograph. This is simply accomplished by marking a distance equal to the radius of the constant-gamma circle on the radial scale labeled "standing wave voltage ratio." The value of swr in dB may also be determined from this scale.

$$swr_{dB} = 16.1\ dB$$

Example 2. Finding the reflection coefficient (ρ) and angle of the reflection coefficient (θ) for voltage and current. A 50-ohm transmission line is terminated with a load impedance $65 - j75$ **ohms.** What is the reflection coefficient and angle of reflection coefficient? (See **fig. 6**).

1. Normalize the load impedance

$$\frac{65-j75}{50} = 1.3 - j1.5$$

2. Locate this point on the chart and draw a line from the center of the chart through it to the outer scale.

3. Construct a constant-gamma circle.

4. The reflection coefficient may be calculated by measuring the radii of the constant-gamma circle and the Smith chart to its first periphery and by computing their ratio. Smith-chart radius = 57/16 inch; constant-gamma radius = 32/16 inch.

$$\rho = \frac{32}{16} \div \frac{57}{16} = 0.56$$

5. The coefficient of reflection may also be found on the radial nomograph. Simply mark the radius of the constant-gamma circle on the scale labeled "reflection coefficient of voltage." The constant-gamma radius intersects the radial scale at 0.56. The "reflection coefficient of power" may also be determined from this same scale at 0.314.

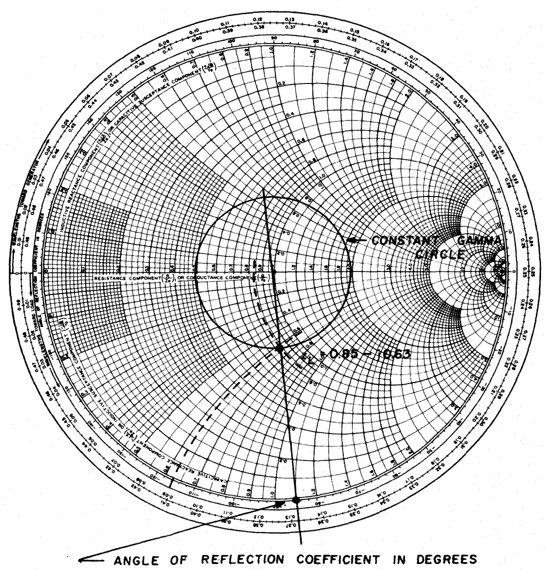


fig. 4. Plotting impedance coordinates on the Smith chart.

6. The angle of the reflection coefficient is defined by the intersection of the radial line plotted in **step 2** and the "angle of reflection coefficient in degrees" scale on the rim of the chart.

$$\rho = -46^{\circ}$$

Example 3. Finding input impedance. A 50-ohm transmission line 20 feet long is terminated with $Z_L = 50 - j50$ *ohms.* What is the input impedance at the sending end of the line at 14.1 MHz? (See **fig. 7.**)

1. Normalize the load impedance

$$\frac{50-j50}{50} = 1 - j1$$

2. Find the length of the transmission line in meters by multiplying by 0.3048."

$$20 \text{ feet} \times .3048 = 6.096 \text{ meters}$$

3. Find the electrical length of the transmission line at 14.1 MHz. First, determine the wavelength at 14.1 MHz. Free-space wavelength is found by dividing the speed of light by frequency

$$\lambda = \frac{3 \times 10^8 \text{ meters per second}}{14.1 \times 10^6 \text{ cycles per second}} = 21.276\ m$$

Calculate the electrical length of the transmission line

$$\theta = 360^{\circ}\left(\frac{6.096\ m}{21.276\ m}\right) = 102^{\circ} = 0.28 \text{ wavelength}$$

4. Plot the impedance coordinates from **step 1** on the chart and draw a line from the center of the chart through this point to the outer scale.

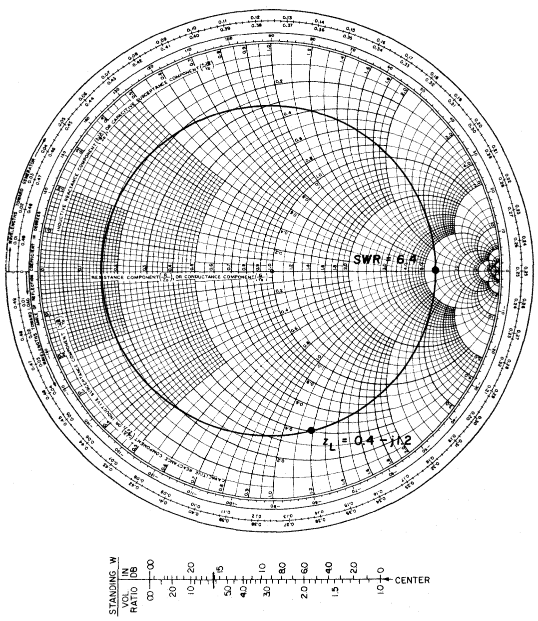


fig. 5. Using the Smith chart to find swr (example 1).

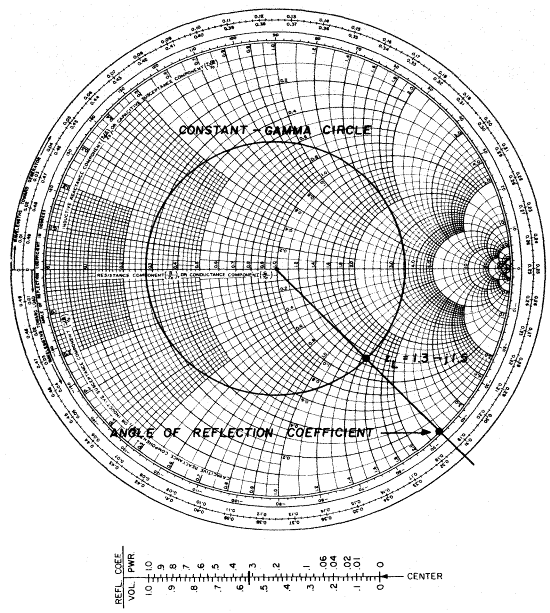


fig. 6. Finding reflection coefficient with the Smith chart (example 2).

5. Draw another line from the chart center to the outer scale at a point 0.28 wavelength clockwise (toward the generator) from the line drawn in **step** 3. Swing an arc from the center of the chart through z_L to this line. The intersection is at $z_L = 0.62 + j0.7$, the normalized input impedance. To find the actual impedance this value must be multiplied by the line's characteristic impedance

$$Z_i = 50(0.62 + j0.7) = 31 + j35$$

Example 4. Calculating load admittance. The impedance of a load terminating a 50-ohm transmission line is 75+ j82 ohms. What is the admittance of the load? (See **fig. 8.**)

1. Normalize the load impedance

$$z_L = (75 + j82)/50 = 1.5 + j1.64$$

2. Plot this point and draw a line through the center to the outer scale on the opposite side of the chart.

3. Swing an arc through z_L to the line on the opposite side of the chart. The point of intersection denotes the *normalized* admittance

$$y_L = 0.305 - j0.33$$

4. Calculate the actual admittance by multiplying the characteristic admittance of the system times the normalized admittance. The characteristic admittance (Y_o) is equal to the reciprocal of the character-

"Although all the computations may be made in feet (or inches) the metric equivalents are easier to work with. To convert from inches to centimeters, multiply by 2.54.

istic impedance

$$Y_o = \frac{1}{Z_o} = \frac{1}{50} = 0.02\ mho$$

Therefore, the admittance is

$$Y_L = 0.02\,(0.305 - j0.33)$$
$$= .0061 - .0066\ mho$$

Example 5. Determining the effect of a characteristic impedance change. A 50-ohm transmission line, 0.15 wavelength long, is terminated with 100 − j0 ohms. The 50-ohm line is fed from a 72-ohm line. What is the vswr in the 72-ohm line? (See **fig. 9**.)

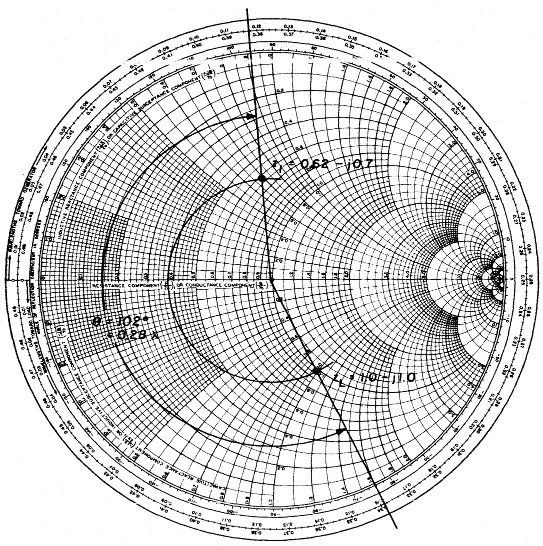


fig. 7. Using the Smith chart to find input impedance (example 3).

1. Normalize the load impedance

$$z_L = (100 - j0)/50 = 2 - j0$$

2. Determine the input impedance at the point where the two transmission lines are connected, 0.15 wavelength from the load. Plot the normalized load impedance on the chart and draw a line from the center of the chart through this point. Note that the line crosses the "wavelengths toward generator" scale at the 0.25 wavelength mark **(fig. 9A)**.

3. Move 0.15 wavelength in a clockwise direction along the "wavelengths toward generator" scale to the 0.40 wavelength mark. Draw a line from this mark through the center of the chart. Swing an arc through z_L. The intersection of the arc and the radial line denote the input impedance to the 50-ohm transmission line 0.15 wavelength from the load

$$z_A = 0.68 - j0.48$$

4. Find the impedance at point X **(fig. 9C)** and normalize to the 72-ohm line. The impedance at point X

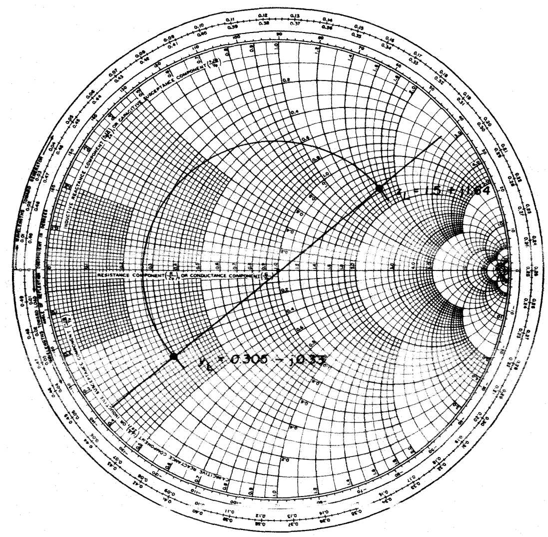


fig. 8. Calculating load admittance (example 4).

is $50(0.68 - j0.48) = 34 - j24$ ohms. Normalize this value to the 72-ohm line

$$(34 - j24)/72 = 0.47 - j0.33$$

5. Plot this point on the chart **(fig. 9B)** and draw a circle through z_A to the "axis of reals." The vswr in the 72-ohm line is 2.5:1. The vswr can also be found with the radial nomograph as outlined in **example 1.**

In the upper vhf region ordinary capacitors and inductors cannot be relied upon to act as pure reactances, and sections of transmission line are often used in their place since any input reactance may be obtained with the proper length of open- or short-circuited line.

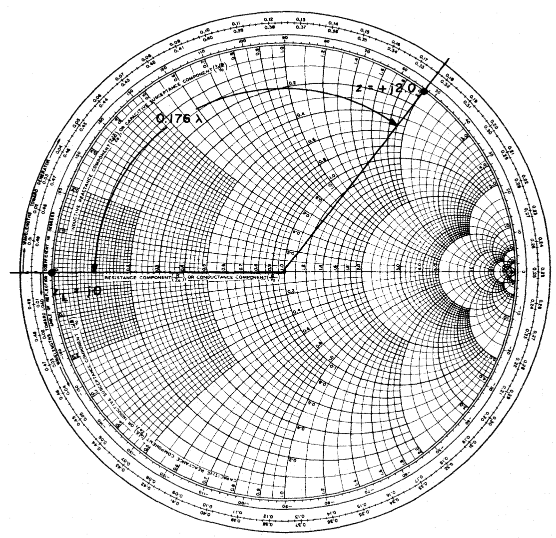


fig. 10. Using a transmission line as a circuit element (example 6).

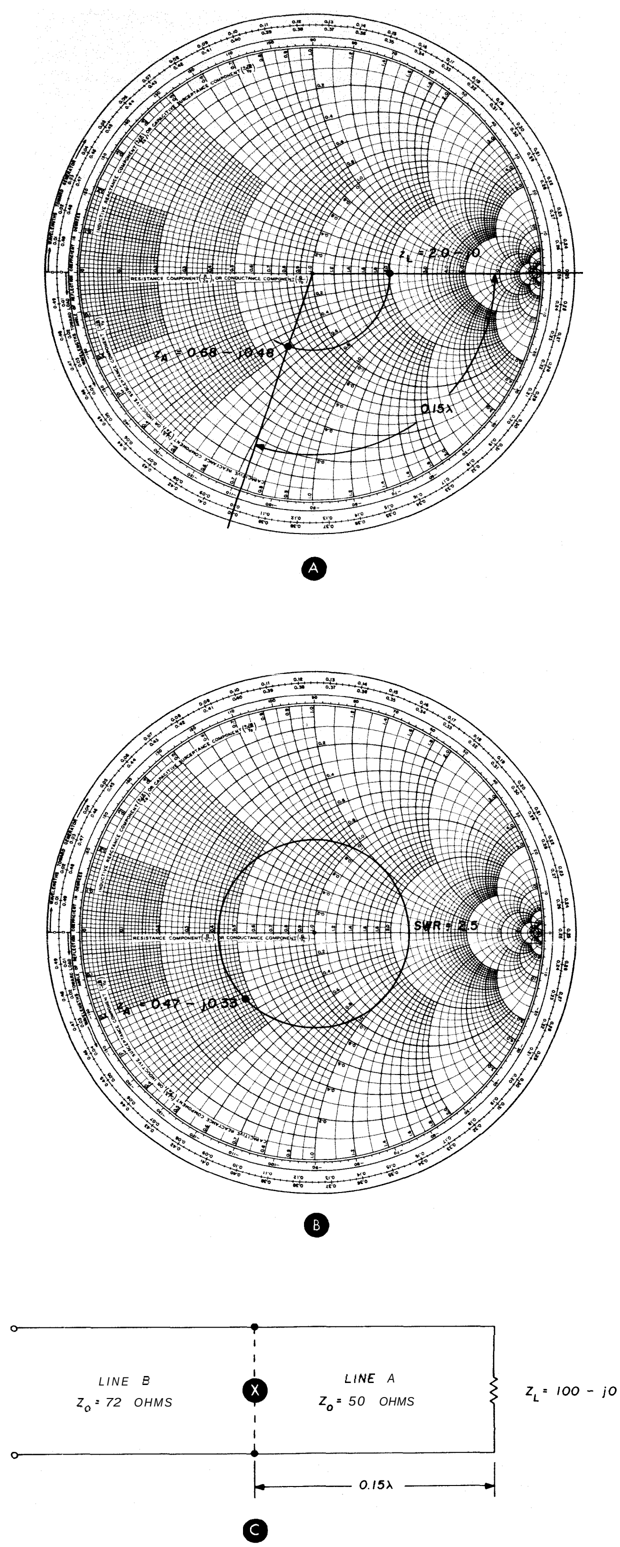


fig. 9. Determining the effect of a characteristic impedance change (example 5).

Example 6. Transmission lines as circuit elements. It is desired to obtain +j100 ohms reactance with a 50-ohm short-circuited transmission line as the circuit element. What length is required? (See **fig. 10.**)

1. Normalize the desired reactance

$$z = (+j100)/50 = +j2$$

2. Since the line is short-circuited,

$$Z_L = 0 + j0, \text{ and } z_L = 0 \text{ ohms.}$$

3. Plot these two points on the chart and draw lines from the center of the chart through each of them. On the "wavelengths toward generator" scale there is a distance of **0.176** wavelength between the two lines. Therefore, a transmission line **0.176** wavelength long is required for a reactance of +j100. (At **144** MHz, +j100 represents an inductance of **0.11** μH.)

Example 7. Finding matching stub length and location. A 50-ohm transmission line is terminated with a load impedance of **32 + j20** ohms. A matching stub is to be used to provide a match to the line. Both the length of the stub (l_s) and its distance from the load (l_d) are variable; find l_s and l_d. (See **fig. 11.**)

1. Normalize the load impedance

$$z_L = (32 + j20)/50 = 0.64 + j0.4$$

2. Locate this point on the chart and draw a line through it and the chart center, extending the line through the peripheral scales in the negative, or bottom, portion at $0.336\lambda (\theta = -62^\circ)$

3. Construct a constant-gamma circle through z_L, on through the admittance point y_L, and intersecting the unity *conductance* circle ($G = 1$) at point **A.**

4. Draw a line from the chart center through point **A** to the outer scale at $0.348\lambda (\text{or } \theta = -71^\circ)$. l_d, the distance from the load to stub, is the distance from **0.336** to **0.348.**

$$l_d = (0.348 - 0.336) = 0.012\lambda$$

$$8 = 71^\circ - 62^\circ = 9^\circ \text{ (4.5 electrical degrees)}$$

5. To find the length of the stub, determine the amount of susceptance necessary to match out the load. The required susceptance is the difference between the susceptance at point A and the susceptance at the center of the chart. The susceptance at point **A** is **−j0.67.** The required stub susceptance is

$$B = +j0.67$$

6. Determine the equivalent stub reactance by taking the reciprocal of the susceptance (as described in **example 4**).

$$X = -j1.49$$

7. Locate the reactance **−j1.49** on the rim of the chart (point **B**). Determine the distance between the short-circuit point and the required reactance (point

B) along the "wavelengths toward generator" scale. $l_s = 0.344\lambda$. ($\theta = 248°$, 124 electrical degrees).

For practical reasons it may not be possible to place a shunt stub only 4.5° from the load. It may be necessary to increase the distance l_D to the next point where $G = 1$ (not R = 1), represented by point C, **fig. 11A**. in this case l_D wouid be measured, clockwise from 0.336 through 0.50 to 0.151. Using the reflection coefficient scale, from $\theta = -62°$ to 180" plus 180" to $+71°$, which totals 227°R, or 113.5 electrical degrees. This represents 0.316λ. This will require a +jX stub, length shown as l_s (C), of the same numerical reactance value as before.

lossy lines

All the examples shown so far have assumed no attenuation in the transmission line. Since all lines have some loss, this must be considered to find the actual case. However, at many amateur frequencies loss is low enough to be neglected. Nevertheless, at 144 MHz and above, line loss should be considered when using the Smith chart.

Attenuation along a uniform transmission line causes the impedance point to spiral inward toward the center of the chart when moving toward the generator; when moving toward the load the impedance point spirals outward toward the rim of the chart. The rate at which the spiral approaches the center (or the rim) depends upon the attenuation as well as the starting point. Impedance points near the rim are affected more per dB of attenuation than points near the center.

The attenuation effect is easily determined with the scale at the bottom of the Smith chart labeled "transmission loss, l-dB steps." Since the intial point on this scale must apply to any point on the chart, it is laid out without numerical calibration. The opposite attenuation effects of moving toward the load as opposed to moving toward the generator are indicated by arrows on the scale which show the proper direction to move the corrected impedance point. Thus, to determine the effect of 2-dB attenuation, simply mark off two l-dB intervals in the proper direction along the scale from the initial starting point before reading the actual impedance coordinates.

Example 8. Impedance transformation through a lossy line. A 50-ohm transmission line 24 centimeters long is terminated with 10 − j10 ohms. What is the input impedance to the line at 250 MHz if the attenuation of the line is 2 dB? (See **fig. 12.**)

1. Normalize the load impedance

$$z_L = (10 - j10)/50 = 0.2 - j0.2$$

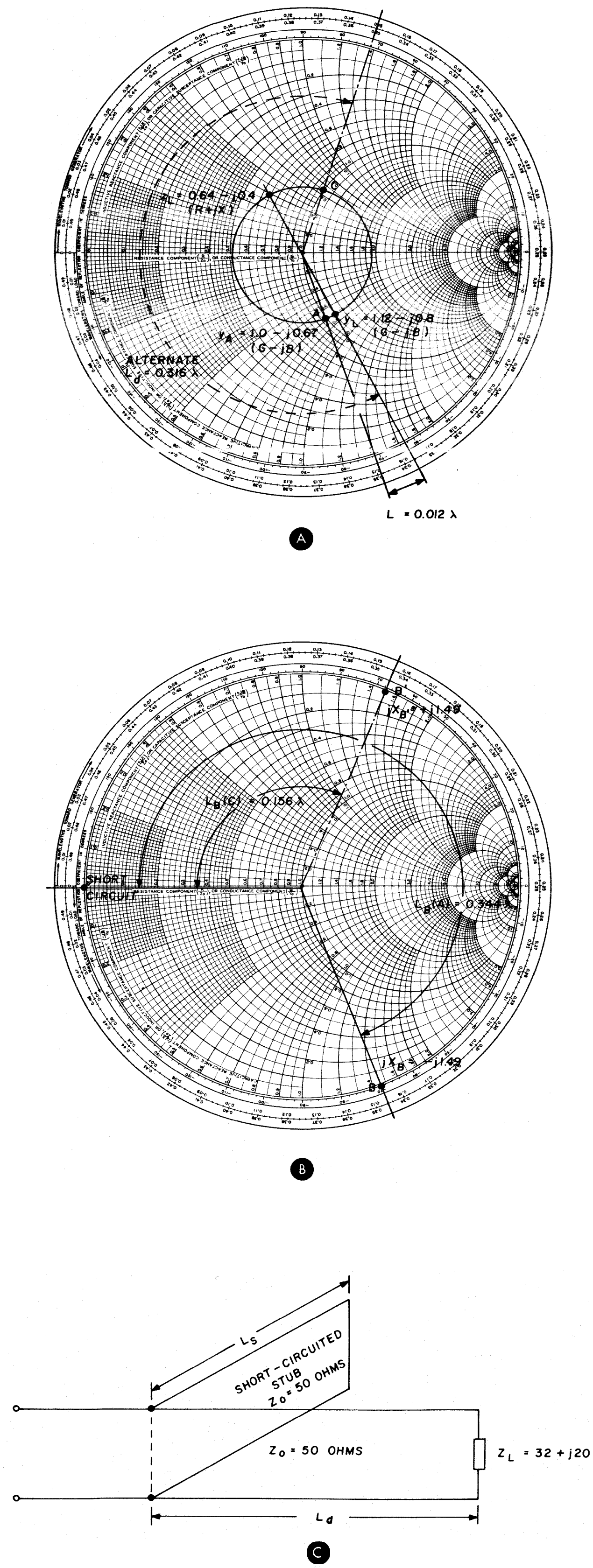


fig. 11. Finding matching stub length and location (example 7).

2. Find the electrical length of the line at 250 MHz.

$$\lambda = \frac{300 \times 10^8}{250 \times 10^6} = 120\ cm$$

The electrical length of the line is

$$\theta = 360°(\frac{24cm}{120cm}) = 72° = 0.2\ wavelength$$

3. Plot the impedance from **step 1** on the chart and draw a line from the center of the chart through this point to the outer scale.

4. Draw another line from the chart center to the outer scale at a point 0.2 wavelength clock-wise (toward the generator) from the line passing through z_L. Swing an arc through z_L to this line. The intersection point denotes $z_i = 0.71 + j1.52\ ohms$. This is the normalized solution for the lossless case. The rf energy from the generator is attenuated 2.0 dB on reaching z_L, and the voltage reflection coefficient is lower than the lossless case. Since the voltage reflection coefficent varies directly with the *power ratio* of one-way line attenuation, the reflection coefficient is reduced to

$$antilog\ \frac{2.0\ (dB)}{10} = 0.631$$

5. The reflection coefficient (ρ_0) for the lossless case is 0.68 (found on the scale at the bottom of the chart). The actual coefficient of reflection may be calculated by multiplying the lossless coefficient of reflection by the power ratio from **step 4.**

$$0.631\rho_o = 0.631\ (0.68) = 0.429$$

6. Swing an arc equal to the ratio $\rho_1 = 0.429$ so it intersects the line drawn through z_i; the radius of this arc can be found on the "voltage reflection coefficient" scale on the bottom of the chart. The normalized impedance for the lossy case is $1.08 + j1.05$. The actual input impedance is

$$Z_i = 50(1.08 + j1.05) = 54 + 52.5\ ohms$$

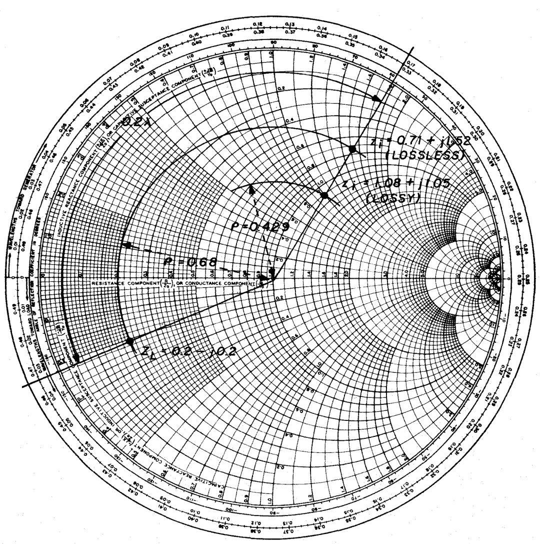


fig. 12. Impedance transformation through a lossy transmission line (example 8).

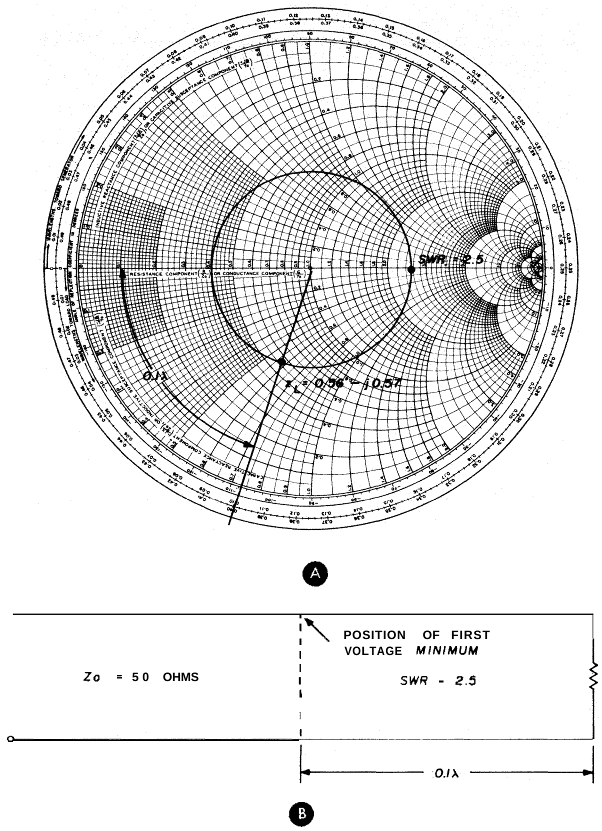


fig. 13. Using the Smith chart to find load impedance from vswr and position of the first voltage minimum on a slotted line (example 9).

slotted lines

At frequencies above 300 MHz conventional impedance-measuring instruments give way to the slotted line. A slotted line is essentially a section of transmission line with a small opening so you can use a probe to measure the voltage along the line. Vswr is easy to determine with the slotted line since it's the ratio of the maximum voltage along the line to the minimum. With the known vswr and position of the first voltage minimum, the impedance of the load can be quickly found with the Smith chart.

Example 9. Calculate the load impedance from the vswr and position of the first voltage minimum. A 50-ohm transmission line has a vswr of 2.5; the first voltage minimum is 0.1 wavelength from the load. What is the impedance of the load? (See **fig.** 13.)

1. Draw a radial line from the center of the chart

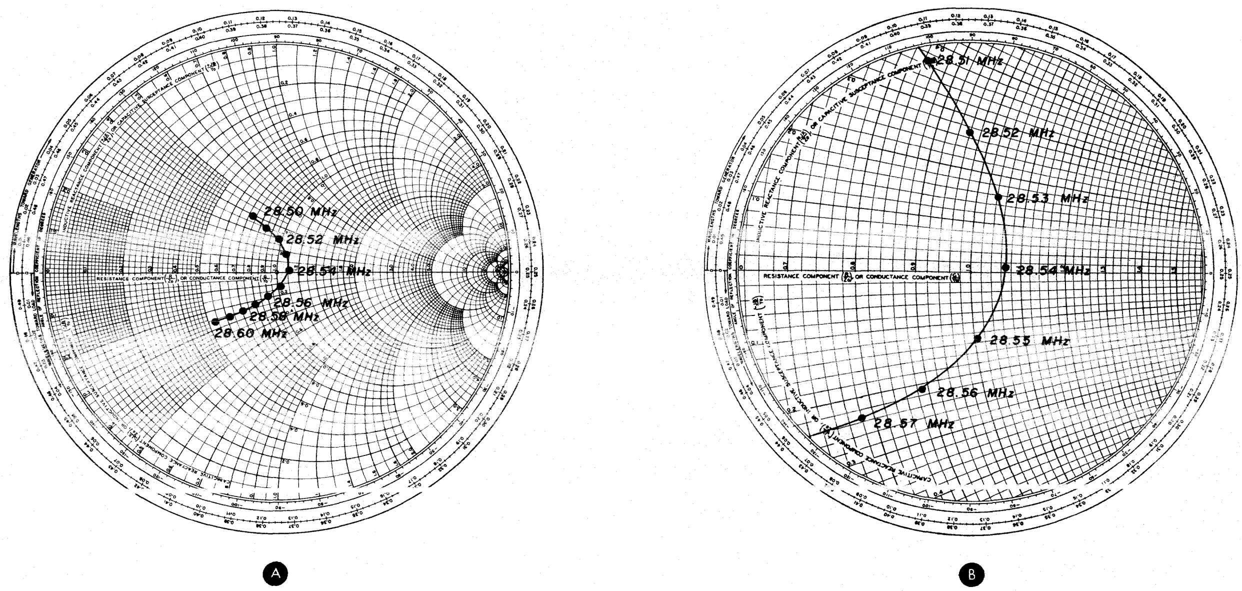


fig. 14. Use of the expanded Smith chart. Impedance points in A are too close together; expanded chart in B is easier to work with.

through the 0.1 wavelength mark on the "wavelengths toward load" scale.

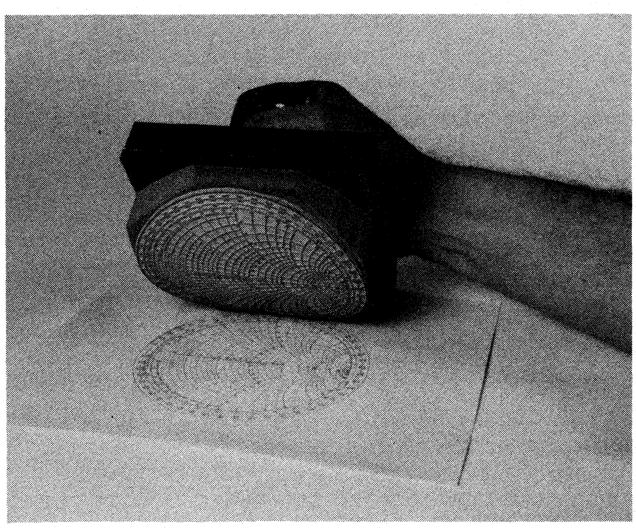
2. Find the 2.5 point on the axis of reals and draw a constant-gamma circuit through this point to intersect with the 0.l-wavelength line.

3. Read the coordinates of this intersection to obtain the normalized impedance of the load

$$z_L = 0.56 - j0.57$$

$$Z_L = 50(0.56 - j0.57) = 28 - j28.5 \text{ ohms}$$

If you use twin-lead or open-wire feedline this technique could be used to determine the impedance of your antenna. However, the voltage probe must be held a uniform distance away from the line for all measurements, and must not be so close that it disturbs the electric field around the conductors.



The Smith chart rubber stamp is 10cm (4 in) in diameter

expanded smith charts

The more closely an antenna is matched to a transmission line, the closer the impedance points are to the center of the Smith chart. In a well-designed system the impedance points may be so close to the center of the chart that it's difficult to work with them. When this happens it's best to use an expanded Smith chart. Two versions are commonly available: one with a maximum swr of 1.59, the other with a maximum swr of 1.12.

The use of the expanded Smith chart is shown in **fig. 14.** In **fig. 14A** the impedance plot of a well-matched 10-meter beam over the low end of the phone band falls very close to the center of the chart. When these same impedance points are plotted on the expanded Smith chart in **fig. 14B** they are much easier to read and work with.

where to buy them

Smith charts can usually be purchased at college bookstores in small quantities, or in larger quantities from Analog Instruments Company or General

"Smith charts from Analog Instruments come in packages of 100 sheets, $4.75 the package. For standard charts order 82-BSPR; expanded charts (maximum swr = 1.59), order 82-SPR; highly expanded (maximum swr = 1.12), order 82-ASPR. Analog Instruments Company, Post Office Box 808, New Providence, New Jersey 07974.

Smith charts from General Radio are available in pads of 50 sheets. $2.00 per pad. For standard charts, normalized coordinates, order 5301-7560; 50-ohm coordinates, order 5301-7569; normalized, expanded coordinates, order 5301-7561. General Radio, West Concord, Massachusetts 01781.

Radio." If you buy directly from the manufacturer, there's a minimum order quantity, so it might be a good idea to get your radio club to sponsor the purchase.

Another solution is the Smith-chart rubber stamp shown in the photo. This stamp is 10 cm (about 4 inches) in diameter and presents an adequately detailed grid structure for most engineering problems. The rubber surface of these stamps is cast from metal dies, and is dimensionally compensated for rocker-mount ellipticity and shrinkage. The capacity is well over a million impressions so you should never be able to wear it out. The stamps are available in standard (vswr $= \infty$) or expanded form (vswr = 1.59

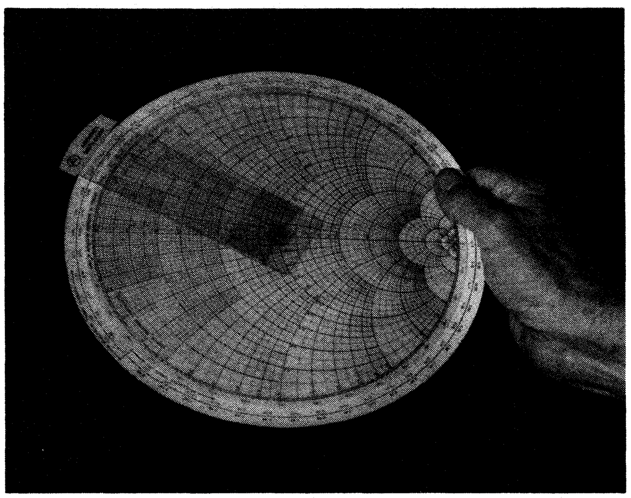


fig. 15. Smith chart calculator provides rapid answers to complex impedance problems.

or 1.12) from the Analog Instruments Company. Cost is $14.75 each.

If you don't need a permanent record of your Smith chart calculations, the calculator shown in **fig. 15** provides rapid answers to complex impedance problems. This calculator is constructed from two laminated plastic discs and a radial arm pivoted at the center with a sliding cursor. This calculator, which is 9-1/2 inches (24 cm) in diameter, is priced at $9.95 and is available from the Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048.

references

1. Henry S. Keen, W2CTK, "A Simple Bridge for Antenna Measurements," *ham* radio, September, 1970, page 34.
2. Philip H. Smith, "Transmission Line Calculator," Electronics, January, 1939, pages 29-31; and "An Improved Transmission Line Calculator," Electronics, January, 1944, pages 130-133 and 318-325.
3. Philip H. Smith, Electronic *Applications* of the Smith Chart in Waveguide, Circuitand *Component Analysis,* McGraw-Hill, New York, 1969.

ham radio

numerical smith chart

How to use the hand-held programmable calculator to execute Smith-chart problems

In this day of handheld programmable calculators, the graphical solution of transmission-line problems with the Smith chart seems unnecessarily cumbersome. Presented here is a derivation of the formulas upon which the Smith chart is based along with a program I have written for the HP-25 programmable calculator which numerically does Smith Chart transmission-line calculations.

The derivation and program are presented in such a way that it is possible to go directly to the program (page 6) and the explanation of its use.

formula derivation

Suppose you have a *lossless* transmission line of characteristic Z_o (ohms), of electrical length θ (in degrees), at frequency **f** (in Hz), terminated with the complex impedance $R+jX$ (ohms) at the far end (fig. 1). You wish to determine the impedance Z_i of this line as seen from the near (input) end.

A constant voltage source of frequency **f** is connected to the input and time is allowed for the system to relax into a steady state. The voltage of the source is now the sum of two distinct and measurable voltages: *incident* (outgoing) and *reflected.* Suppose at the input end the outgoing voltage at time t (seconds) is

$$V_o(t) = ae^{j360ft} \qquad (1)$$

referenced to the bottom conductor.* Let t_o be the time required for this voltage to reach the far end. Then the outgoing voltage which appears across the termination at time t is exactly $V_o(t-t_o)$, the outgoing voltage that appeared at the input terminals t_o seconds in the past (fig. 2)!

Let the reflected voltage appearing at the termination at time t be $V_r(t)$. Then the reflected voltage appearing at the near end is $V_r(t-t_o)$.

In effect there are two generators: one at the input end generating an outgoing voltage of value $V_o(t)$, and a generator at the far end simultaneously generating a reflected voltage of value $V_r(t)$.

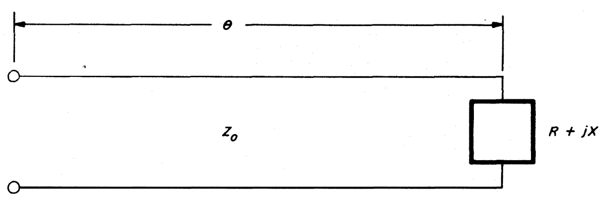


fig. 1. Lossless transmission line with characteristic impedance Z_o, electrical length θ, and terminated with $R+jX$. Formulas are given in the text for calculating the input impedance of the line Z_i.

Let's first examine what is happening at the far end. At the termination the outgoing current, together with the reflected current, must total the current through the termination.

$$\frac{V_o(t-t_o)}{Z_o} - \frac{V_r(t)}{Z_o} = \frac{V_o(t-t_o)+V_r(t)}{R+jX} \qquad (2)$$

The minus sign before the second term corrects for the fact that reflected current is moving right to left rather than left to right (this is essentially why the standard reflectometer circuit can separate incident

*To generalize Ohm's law to ac circuits it's necessary to write voltages and current in the complex form $ae^{j\omega t}$ rather than in the more familiar form $a \cdot \sin\omega t$.

By C. R. MacCluer, W8MQW, 1105 Orchard, Lansing, Michigan 48912

voltage from reflected voltage). Dividing through by $V_o(t-t_o)$ and setting

$$\rho = \frac{V_R(t)}{V_o(t-t_o)}$$ (the complex coefficient of reflection)

we obtain

$$\frac{1-\rho}{Z_o} = \frac{1+\rho}{R+jX}$$

Solving for ρ

$$\rho = \frac{R+jX-Z_o}{R+jX+Z_o} = \frac{r+jx-1}{r+jx+1} \tag{3}$$

Therefore

$$\rho = -\frac{1-z}{1+z} = \frac{z-1}{z+1} \tag{4}$$

where

$$z = \frac{R+jX}{Z_o} = r + jx$$ (normalized termination)

Meanwhile, back at the input end of the transmission line, the source voltage V is the sum of the outgoing and reflected voltages

$$V = V_o(t) + V_r(t-t_o) \tag{5}$$

The current I drawn from the source is the total of the outgoing and reflected currents

$$I = \frac{V_o(t)}{Z_o} - \frac{V_r(t-t_o)}{Z_o} \tag{6}$$

(The minus sign is used for reverse current, as before.) Therefore, the impedance Z_i is found by dividing **equations 5** and **6.**

$$Z_i = \frac{V}{I} = \frac{V_o(t) + V_r(t-t_o)}{\frac{V_o(t)}{Z_o} - \frac{V_r(t-t_o)}{Z_o}}$$

$$= Z_o \, \frac{1 + V_r(t-t_o)/V_o(t)}{1 - V_r(t-t_o)/V_o(t)} \tag{7}$$

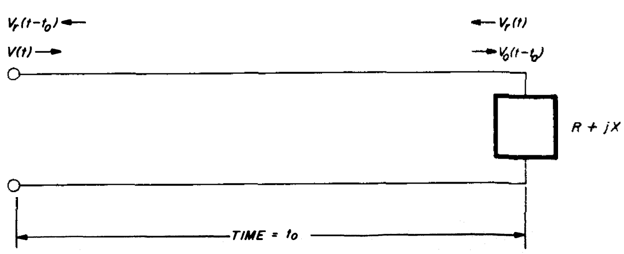


fig. 2. If a constant voltage source, $V(t)$, at frequency f is connected at the input, and t_o is the time required fo: the voltage to reach the far end, then the outgoing voltage which appears across the termination at time t is exactly $V_o(t-t_o)$, the outgoing voltage that appeared at the input terminals t_o seconds in the past.

The quantities $V_o(t)$ and $V_r(t-t_o)$ differ possibly in amplitude and phase, but not in frequency, so their quotient is constant! Therefore

$$\frac{V_r(t-t_o)}{V_o(t)} = \frac{V_r(t-2t_o)}{V_o(t-t_o)} =$$

$$\left(\frac{V_r(t)}{V_o(t-t_o)}\right)\left(\frac{V_r(t-2t_o)}{V_r(t)}\right) \tag{8}$$

But

$$V_r(t-2t_o)/V_r(t) = e^{-j2\theta}$$

Hence

$$\frac{V_r(t-t_o)}{V_o(t)} = \rho \, e^{-j2\theta} \tag{9}$$

Combining **equations 7** and **9** yields the Smith chart formula for lossless lines:

$$Z_i = Z_o \left(\frac{1+\rho\, e^{-j2\theta}}{1-\rho\, e^{-j2\theta}}\right) \tag{10}$$

where

$$\rho = \frac{z-1}{z+1}$$

$$z = \frac{R+jX}{Z_o} = r + jx$$

lines with loss

Let's now assume that the transmission line is

$V_r(t) = ae^{\,j(\omega t+\phi)}$ and so

$$\frac{V_r(t-2t_o)}{V_r(t)} = \frac{ae^{\,j[\omega(t-2t_o)+\phi]}}{ae^{\,j(t+\phi)}} = e^{-j2\omega t_o}$$

But

$$2\omega t_o = 2 \cdot 360 f t_o = 2\theta$$

lossy, say A dB per degree of length. Power P_1 introduced at one end of the line will be attenuated to power P_2 at the other end where

$$-A\theta = 10 \log \frac{P_2}{P_1} \tag{11}$$

Dividing through by 10 and exponentiating yields

$$P_2 = P_1(10^{-A\theta/10})$$

Thus outgoing voltage will be decreased (attenuated) by a factor of $\alpha = 10^{-A\theta/20}$ at the termination and reflected voltage has also decreased by the factor a at the near end **(fig. 3)**.

table 1. Example of HP-25 Smith chart calculations to find the complex impedance at the generator (input) end of a transmission line, reflection coefficient, and vswr, given the complex impedance of the termination.

	R_L (ohms)	jX_L	L (ft)	f (MHz)	v	θ (deg)	Z_o (ohms)	R_i (ohms)	jX_i	ρ	vswr
1	73	+j16	85	3.6	0.66	169.62	50	63.29	+j22.45	0.226	1.58:1
2	32	−j5	100	7.1	0.66	393.57	50	34.98	+j12.49	0.227	1.59:1
3	50	j0	250	3.8	0.66	526.61	73	51.47	−j 8.99	0.187	1.46:1
4	100	+j100	100	14.2	0.66	787.14	50	18.14	−j35.40	0.620	4.27:1
5	18.14	−j35.40	100	14.2	0.66	787.14	50	15.18	+j26.18	0.620	4.27:1
6	75.00	−j30.00	—	—	—	90	50	28.74	+j11.49	0.304	1.87:1
7	28.74	−j11.49	—	—	—	90	50	75.00	−j30.00	0.304	1.87:1

With these losses in mind let's again derive the Smith chart forrnuias. in our calculations at the termination replace $V_o(t-t_o)$ with $\alpha V_o(t-t_o)$. In the calculations at the input end replace $V_r(t-t_o)$ with $\alpha V_r(t-t_o)$. The result will be the Smith chart formulas for transmission lines with loss

$$Z_i = Z_o \frac{1+\rho\ e^{-j2\theta}10^{-A\theta/10}}{1-\rho\ e^{-j2\theta}10^{-A\theta/10}} \tag{12}$$

where

$$\rho = \frac{z-1}{z+1}$$

$$z = \frac{R+jX}{Z_o}$$

In practice suppose a transmission line has length L (feet) and a loss of d dB per 100 feet. Since $A9 = Ld/100$, the attenuation factor

$$\alpha^2 = 10^{-A\theta/10} = 10^{-Ld/1000} \tag{13}$$

HP-25 program

Suppose you have a *lossless* transmission line of characteristic impedance Z_o (ohms) of electrical length 9 (degrees) at the frequency f (MHz) terminated in a complex impedance $R_L + jX_L$ (ohms). You wish to compute the impedance Z_i of this line as seen from the input end (**fig. 1**). The HP-25 program in **fig. 4** can be used to calculate Z_i. To use this program, follow the following steps:

1. Key in program.

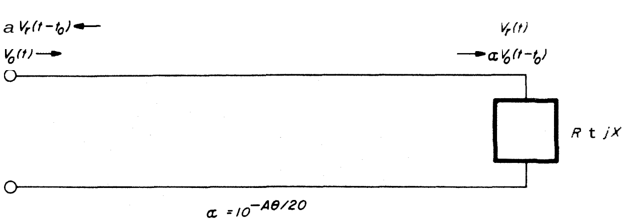


fig. 3. When a transmission line is lossy, the outgoing voltage is attenuated by the factor $\alpha = 10^{-A\theta/20}$ at the termination, and reflected voltage is also decreased by the factor a at the input end.

2. Store load resistance R_L in Register 0 (R_L STO 0).

3. Store load reactance jX_L in Register 1 (X_L STO 1).

4. Calculate line length θ in degrees using the following formula, and store in Register 2 (θ STO 2)

$$\frac{360Lf}{984v} = 0.3659 \frac{Lf}{v} \quad (feet)$$

$$\frac{360Lf}{299v} = 1.204 \frac{Lf}{v} \quad \text{(meters)}$$

where L is the length of the line in feet or meters, f is the frequency in MHz, and v is the transmission line's velocity factor.

5. Store transmission line impedance Z_o in Register 3 (Z_o STO 3).

6. Press R/S key to run program.

7. Calculator displays real part R_i of the input impedance (reactive part jX_i is in they register).

8. Press $x \rightleftarrows y$ key to display reactive part jX_i of the input impedance.

9. Coefficient of reflection ρ is available in Register 5.

10. Manually calculate vswr from

$$vswr = \frac{1+\rho}{1-\rho}$$

Several sample runs are shown in **table 1**.

It is a commonly held misconception that transmission lines transform symmetrically; many amateurs believe that if a transmission line terminated with Z1 yields Z2 at the input (generator) end of the line, the same line terminated with Z2 will yield Z1 at the generator. Except for lines which are one-quarter wavelength long ($\theta = 90$ degrees), or multiples thereof, this is not true. For example, a shorted 45° (λ/8) length line of characteristic impedance Z_o has impedance

$Z = jZ_o \tan 45^\circ = jZ_o$, yet this same line, when terminated by jZ_o, is equivalent to a shorted quarter-wavelength line and hence has infinite impedance.

The special case of the quarter-wavelength transmission line is illustrated in lines 6 and 7 of **table 1.** Here a 50-ohm transmission-line termination of 75 – j30 ohms is transformed to 28.74 + j11.49; the same line terminated with 28.74 + j11.49 yield 75 – j30 at the generator end.

To use the HP-25 program in **fig. 1** to calculate the impedance of the termination Z_L given the impedance Z_i at the generator, store the real part R_i in register 0 (R_i STO 0, Step 2), and the reactive part jX_i in register 1 (X_i STO 1, Step 3). In Step 4 store $-\theta$ in Register 2 ($-\theta$ STO 2).

program modification for lines with loss

Suppose that the transmission line is lossy. Modify the previous HP-25 program by inserting an R/S between Steps 34 and 35, renumber succeeding steps, and at Step 28 replace the GTO 43 with GTO 44. To use the modified program, Steps 1 through 5 are the same as before.

6. Compute the factor $\alpha^2 = 10^{-Ld/1000}$ where d is the loss of the line in dB per 100 feet. Store the result in Register 6 (α^2 STO 6).

7. Press the R/S key; the routine will stop at Step 35 at the inserted R/S.

8. Recall the factor α^2 from Register 6 and press X.

9. Press R/S key.

10. Calculator displays the resistive part of Z_i (reactive part of Z_i is contained in the y register).

11. Compute the swr at the input as before.

To compute the swr at the termination, calculate

$$\rho' = \rho/\alpha^2 = \rho 10^{Ld/1000}$$

and proceed as before with ρ replaced by ρ'.

To use this program "backwards" (to find the termination impedance Z_L given the input impedance Z_i), store Z_i in Steps 2 and 3, store $-\theta$ in Step **4,** and in Step 6 store $1/\alpha^2 = 10^{Ld/1000}$. The swr calculated at Step 11 will now be the swr at the termination. To calculate the swr at the input end of the line, the value ρ' found in Register 5 must be replaced by $\rho = \rho'\alpha^2$ before proceeding as in Step 11.

As an example, consider the amateur who uses his 75-meter dipole for two-meter fm. Assuming 120 feet (36.6m) of RG-58/U coaxial cable ($v = 0.66$, $d = 5.7$ *dB loss/100 feet* at 144 MHz), what is the swr at the input end of the line? (The antenna feedpoint impedance is assumed to be 2000 – j800 ohms.)

"Copies are available from *ham radio* upon receipt of a self-addressed, stamped envelope.

HP-25 Program Form

Title Transmission line impedance transformation (complex Z_i given Z_L)

Switch to PRGM mode, press [f] [PRGM] then key in the program. LOSSLESS LINE

DISPLAY LINE	CODE	KEY ENTRY	X	Y	Z	T	COMMENTS
00							
01	00	0					
02	236104	STO X4					
03	2403	RCL 3					
04	237100	STO ÷0					
05	237101	STO ÷1					
06	2401	RCL 1					
07	32	CHS					
08	01	1					
09	2400	RCL 0					
10	41	–					
11	1509	→P					
12	01	1					
13	2400	RCL 0					
14	51	+					
15	2401	RCL 1					
16	21	X↔Y					
17	1509	→P					
18	21	X↔Y					
19	22	R↓					
20	71	÷					
21	22	R↓					
22	41	–					
23	32	CHS					
24	21	X↔Y					
25	22	R↓					
26	2404	RCL 4					
27	1561	X≠0					
28	1343	GTO 43					
29	22	R↓					
30	2402	RCL 2					
31	02	2					
32	61	X					
33	41	–					
34	21	X↔Y					
35	2305	STO 5					
36	1409	→R					
37	2300	STO 0					
38	21	X↔Y					
39	2301	STO 1					
40	01	1					
41	235104	STO +4					
42	1306	GTO 06					
43	22	R↓					
44	21	X↔Y					
45	2403	RCL 3					
46	61	X					
47	1409	→R					
48	1300	GTO 00					
49							

REGISTERS	
R_0	R
R_1	X
R_2	Θ
R_3	Z_O
R_4	
R_5	
R_6	
R_7	

fig. 4. HP-25 program for calculating the impedance as seen at the input end of a lossless transmission line while terminated in a complex impedance.

$$\theta = 0.3659 \frac{Lf}{v} = 0.3659 \frac{120 \cdot 144}{0.66} = 9579$$

$$\alpha^2 = 10^{-Ld/1000} = 0.207$$

$$Z_i = 52.83 - j21.35 \text{ ohms}$$

$$swr = 1.49{:}1$$

Because of high line loss, the swr at the input end of the line is only 1.5:1, even with the outrageous mismatch at the antenna! In contrast, consider the same case with a lossless transmission line:

$$Z_i = 2.99 - j64.16$$

$$swr = 44{:}1$$

The two HP-25 programs presented here are invaluable time savers when it comes to making transmission line calculations. I have also written a facile program for the HP-67 programmable calculator which includes all the above features."

ham radio

comments

10 GHz Gunnplexers

Dear HR:

I was very happy to see the article on 10 GHz in April, 1977, *ham radio.* This band has been of interest to me for more than 20 years. In 1955, I built a system using reflex klystrons and obtained good results. Two years ago, I built a system similar to the one described and have obtained results much the same as discussed in the article.

In the new system, two Microwave Associates 86656D transceivers are used. As can be seen in the photograph, the horns are homemade. In the enclosure, the microphone amplifier is on the left, an inexpensive fm broadcast receiver is on the right, and the voltage regulator is at the rear. These transceivers do not have a varactor diode; the Gunn diode bias voltage controls all frequency changes, both modulation and basic frequency. This system seems to work quite well.

I've experienced the same frequency stability problems described, even with very good voltage regulation. Temperature variations are mostly to blame.

So far I have not developed any form of afc system. The crystal control method looks like it would keep the Gunn transmitter close to a given frequency but it also appears quite complicated. I like the idea of using the afc or discriminator output to control the voltage regulator. It appears to be quite simple yet, will provide a perfect lock.

Allen C. Webb, W5RLG
Richardson, Texas

receiver recovery in the SB102

Dear HR:

The *ham notebook* article (March, 1977) by W2CNQ, regarding excessive recovery delay in the CW mode of Heathkit's SB102, is not unique as the same problem occurs in other Heath transceivers. The root of the problem may well lie in the fact that all SB100 series, HW100s, and HW101s are not the same, some having been modified, while others are later models.

Opening up the screen voltage supply line to cut off pentodes and tetrodes is not without its problems, especially since small amounts of current can flow from plate to screen. There's an additional problem if other tubes are connected to the same screen line. This was originally described in the article by Peterson and Williams (*QST*, January, 1969) which points out that the screen voltage from 6146s (with continuous plate voltage) had been observed to rise as much as 20 volts when the screen supply was opened.

Accordingly, the placement of very high back resistance silicon diodes in the screen circuit of the 6146s and 6CL6 driver, will in most cases prevent the screen voltage, at V2, from rising during the receive condition (in the CW mode).

Changing tubes will help in many instances, but the root of the problem will remain if it is due to small amounts of current circulating back through the supposedly "open" screen circuit.

W. H. Fishback, W1JE
Chatham, Massachusetts

regulated power supplies

Dear HR:

I must congratulate K5VKQ on his article concerning the design of regulated power supplies. I feel that this type of article is needed by most amateurs.

I would like to point out that in the design example, with an average input voltage to the regulator of 13.5 Vdc, and a current drain of 1 ampere, the regulator must dissipate 8.5 watts, with an output voltage of 5 volts. However, the LM309K will only dissipate 3.5 watts, without a heatsink at an ambient temperature of 25° centigrade. Therefore, potential builders/designers should consider power dissipation in the regulator, $P_{DISS} = (V_{IN} - V_{OUT})(I_{LOAD})$. Allowable power dissipation values *vs* ambient temperature curves are readily available in most data books, along with recommended heatsink types.

Wayne Whitman, W9HFR
Oconomowoc, Wisconsin

mospower fet

Dear HR:

As pointed out by Johnson in the *RCA Review* (March, 1973) and in the United States Patent 3,174,462, all types of fet devices have a diffusion mode of operation which ideally provides a device transconductance — per-unit-current of approximately

39000 micromhos of transconductance per milliampere of drain current. The junction fet can provide this transconductance over as much as five orders of magnitude variation in drain current. Over this range, it behaves essentially like a bipolar transistor.

Transconductance efficiency, the percentage of transconductance per unit current which the user can obtain from a practical device in his circuit, is of major importance to amateurs since it controls both the efficiency of the device as a low noise (front end) amplifier, and also the power handling ability of mospower devices. In the low noise application, maximum efficiency is required, while in power applications, quite low values of efficiency may be desirable.

All solid-state amplifiers are voltage-gain limited, not current-gain limited. For this reason, the output supply voltage which may be used with an rf amplifier can be expressed in terms of a simple equation:

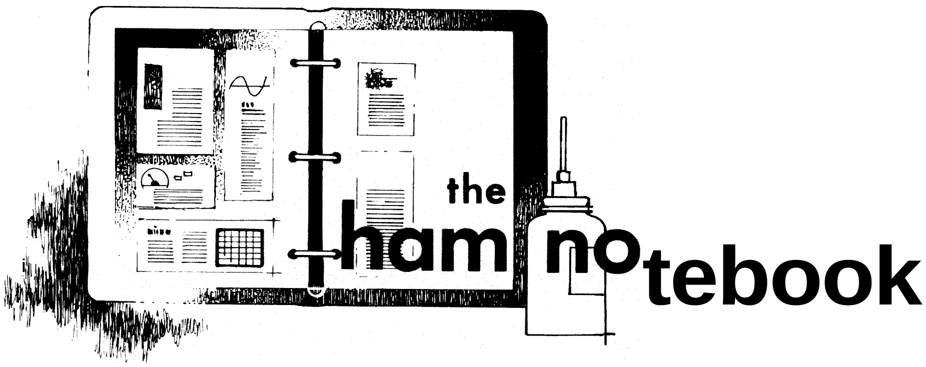
$$V_{dd} = A/x$$

where V_{dd} is the drain supply voltage, x is the (decimal) value of transconductance efficiency, and A is a small number, typically near unity, but almost never over ten. (This equation may easily be confirmed from data sheets on power electron tubes as well as transistors of all kinds.)

If freedom from oscillation and spurious transients is necessary, overall stage voltage gain for any solid-state rf amplifier must be limited to approximately ten. (To prevent TVI, for example.) Use of voltages higher than indicated by the equation lead to a variety of problems, excess phase shift, improper bandpass responses, excessive dissipation, or birdies.

The field-effect transistor, and particularly the mospower unit, may lead to a breakthrough in the application of semiconductors to rf power amplifiers, possibly breaking the present practical limit of 50 watts per device for bipolar devices in the common-base configuration. The fet uses an insulated-gate structure similar to that described in the patent for diode fet devices, with corresponding power and frequency response characteristics. It looks like a bipolar transistor but forms a channel against the insulated gate. Since the device is forward biased, it behaves as a bipolar transistor so is limited by the high value of x, near unity, which is typical of bipolar devices.

Keats A. Pullen, W3QOM
Kingsville, Maryland



remote-switching circuit

A reliable and inexpensive circuit for remote frequency control of a vhf transceiver can be made from three NAND gate ICs. The circuit could also be used for other applications requiring remote selection of one of a number of mutually exclusive functions. The use of CMOS gates permits operation from an automobile electrical system or other 12-volt source, providing a high degree of noise immunity, and freedom from rf interference. Suitable ICs are the RCA CD4011A and 4012A, or the corresponding Motorola MC4011CP and 14012CP.

The circuit shown in **fig. 1** consists of four flip-flops, each made up of one 4-input and one 2-input NAND gate. Momentarily grounding any input will drive the corresponding output high, and all other outputs low. Unless power is interrupted, additional pulses on the same input have no effect; the circuit remains in a stable state until another input is momentarily grounded. The outputs may be used to drive other logic devices directly, but an external buffer will be needed if the current approaches the safe limit of 10 mA.

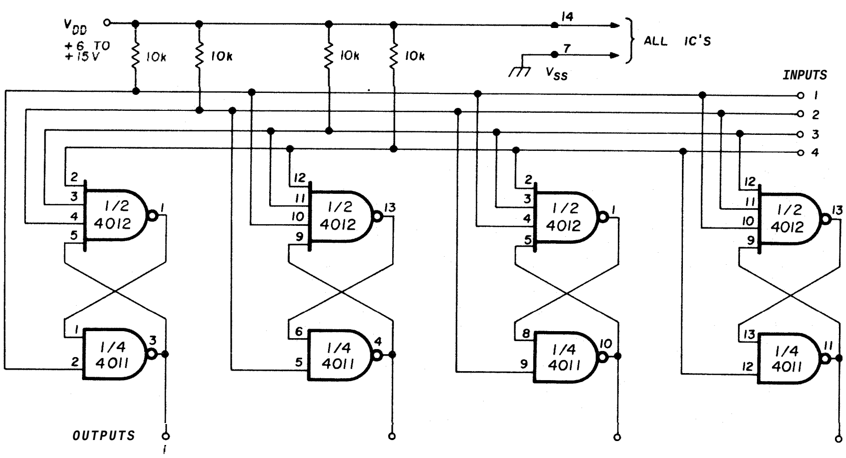


fig. 1. Remote switching circuit employing CMOS NAND gates.

Pat Shreve, W8GRG

a TTL and CMOS logic probe

The circuit shown in **fig. 2** is designed to indicate the logic states in TTL and CMOS circuits. In addition, it will indicate the presence of positive- and negative-going pulses, substituting for a high-speed triggered oscilloscope. In the schematic, Q1 is used as a high-impedance buffer preceding one of the NAND gates. CR1 acts as the buffer before a second gate. With a logic 0 input, the **A** section of the 4011 will turn LED2 off, and the **B** and **C** sections will turn LED1 on. A logic 1 on the input reverses the levels causing LED2 to be on and LED1 to be off.

A 555-type timer IC acts as a one-shot multivibrator, triggered by either positive- or negative-going pulses. With S1 open, LED3 will come on for approximately 200 ms regardless of the input pulse width. After the input pulse, with S1 closed, LED3 will remain permanently on.

In normal operation, a logic 0 input will cause both LED1 and LED3 to light, with LED3 remaining on for only 200 ms. For a logic 1 input, only LED2 will light. With S1 closed, the circuit will indicate whether a negative- or positive-going pulse has occurred. If positive-going, LED1 and LED3 will remain on; LED2 and LED3 will remain on after a negative-going pulse has occurred. The input and correct displays are summarized in **table 1.**

table 1. **LED** displays for different ir conditions.

input	LED1	LED2	LED3	S1 clos
logig 0	on	off	on200 ms	no
logic 1	off	on	off	no
positive-going pulse	on	off	on	yes
negative-going pulse	off	on	on	ves

Howard M. Berlin, W3HB

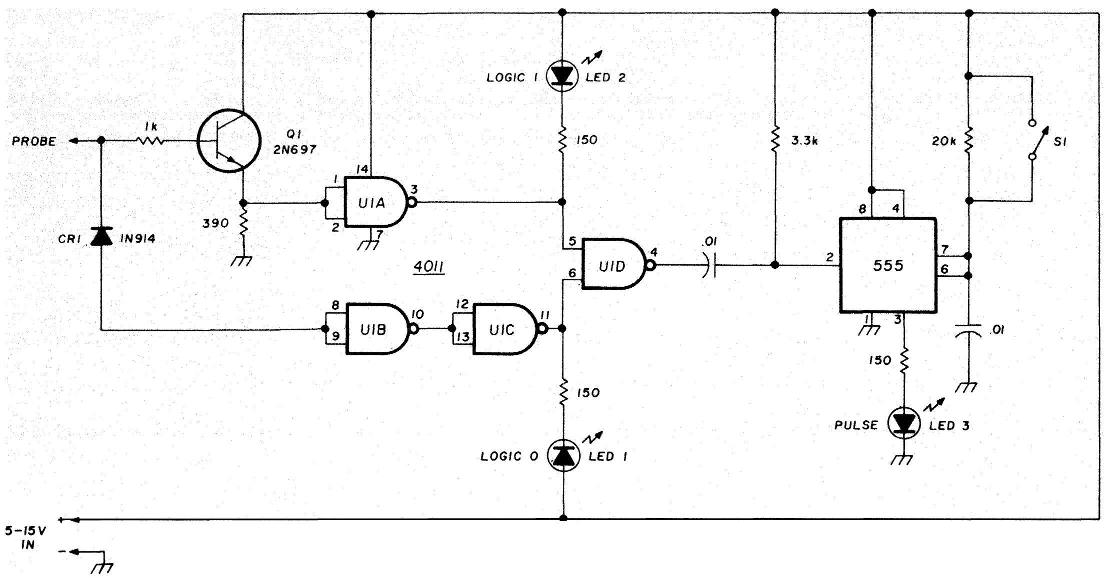


fig. 2. Schematic diagram of the TTL and CMOS logic probe. Since a CMOS gate is used in the probe, the supply voltage can be between 5 and 15 volts. The 555 timer will produce a 200 ms pulse after being triggered.

voltage adapter for MSI/LSI circuits

For experimenting with integrated circuits, it is necessary to have a bench power supply with at least two voltage levels. With the scheme shown in **fig.** 3, it is possible to supply +12, −12, and +5 volts from a single regulated 24-volt source for use with many LSI circuits. The +12V and −12V supplies can be adjusted either in the same direction by varying the 24V source, or in opposite directions by adjusting the potentiometer.

Resistor R1 is used to decrease the power dissipated in the LM309K voltage regulator. A 2.2 ohm value is correct, but it is possible to change it depending upon the current necessary on the +5 volt line (the input voltage to the LM309K must always be greater than 7 volts for a good regulation). All the small components are mounted on a 2 x 1-314-inch (5x4.5cm) pc board, and the two TO-3 cases are mounted on a commercial heatsink.

J. A. Piat, F2ES

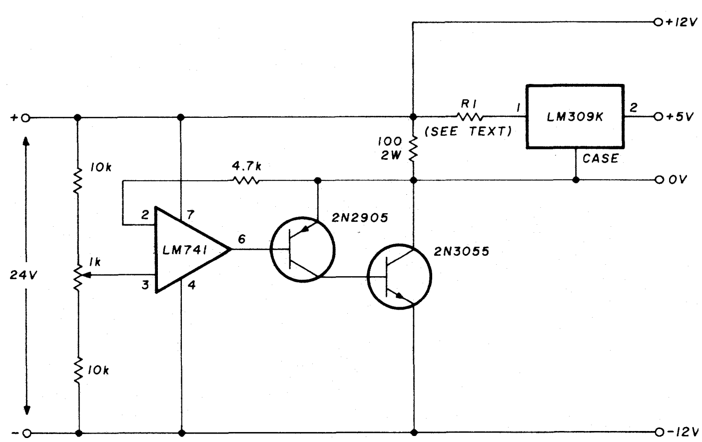


fig. 3. Schematic of the voltage adapter for **LSI/MSI** circuits. Value of resistor **R1** is discussed in the text. Printed-circuit for this simple circuit is shown in fig. 2.

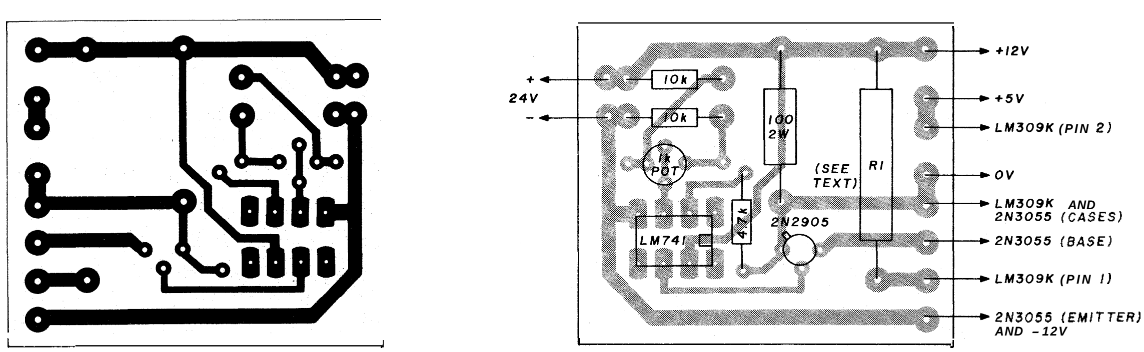


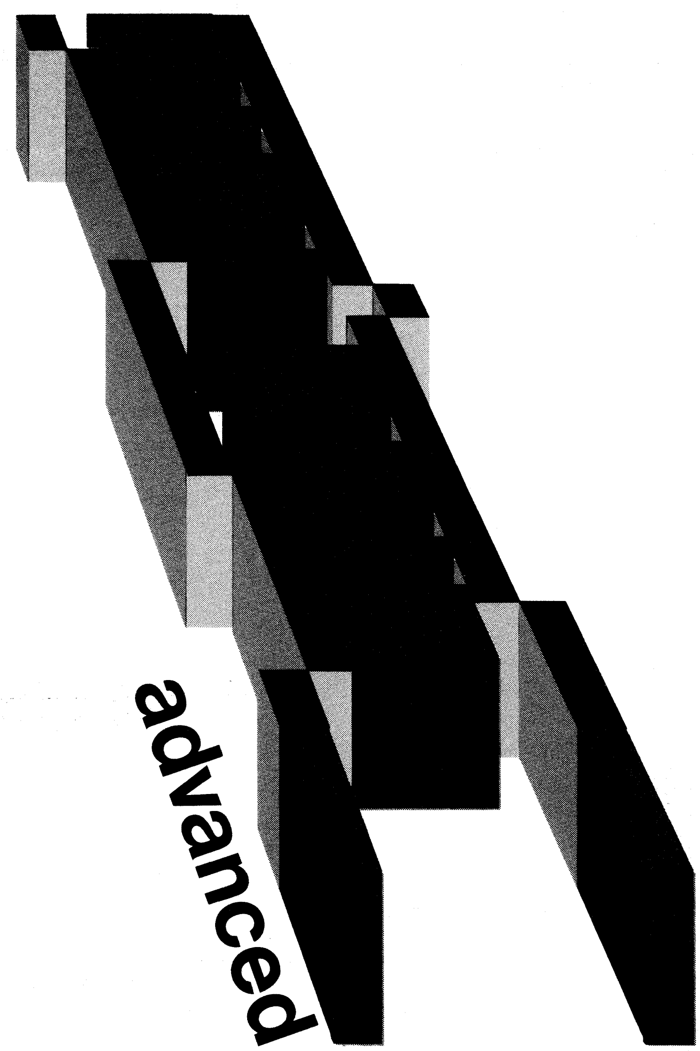
fig. 4. Full-size printed-circuit board layout and component placement diagram for the **MSI/LSI** voltage adapter.

$1.50

ham radio magazine

hr

APRIL 1978



advanced electronic keyer

ham radio magazine

APRIL 1978

volume 11, number 4

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

U.S. and Canada: one year, $12.00
two years, $22.00
three years, $30.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

Foreign subscription agents are
listed on page 101

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents

a second look

by Jim Fisk

If you look at technological advances with the eye of a historian, you'll see periodic peaks in new technology, followed by nulls where everything seems to come to a standstill. This phenomenon is most noticeable in the semiconductor industry during the past few years, but if you look back to the 19th century, you'll see the same thing happened then, with George Stephenson's steam-powered locomotives and Eli Whitney's cotton gin early in the century, and later with the development of mass production techniques, the electric telegraph, and Marconi's wireless. Every few years, it seems, a major breakthrough in technology occurs which produces an avalanche of new products on the market place. If you analyze it carefully, in most cases you'll find that the "major breakthrough" didn't happen all at once as it first appears, but was the culmination of years of research by many different workers in many diverse fields.

In many ways the cyclic rise and fall of technological achievement is an extension of the well-known domino theory – the apparent lulls in activity occur when the dominoes are being lined up; the breakthrough comes when the last domino is set in place and the whole line is knocked down, one domino after the other. It takes but one missing or misplaced domino to prevent the whole line from going down.

In the field of gallium-arsenide field-effect transistors (GaAsfets or "gas" fets, for short), the last of the dominoes has been set in place, and in the near future there will be devices on the market which will provide noise figures of less than 1 dB at 4000 MHz, and power outputs of 6 watts or more at 10 GHz. At a conference last summer at Cornell University, researchers from Bell Labs reported on a GaAs fet amplifier which yielded 10 dB gain at 4 GHz with a noise figure of 0.7 dB; at 6 GHz the noise figure increased to 1.25 dB and gain dropped to 9 dB. At the same conference, scientists from Rockwell International were talking about a device which gave a 2.2 dB noise figure at 10 GHz with 8.5 dB gain. These devices and similar ones from other manufacturers will be available on the commercial market this year, but be prepared to pay a pretty healthy price for the privilege of building a circuit around them. As manufacturing processes are improved and yields go up, however, I expect device cost will drop within the amateur price range. If you can't wait, in this issue JH1BRY describes a high-performance GaAs fet preamp for 432 MHz that has a noise figure below 1 dB.

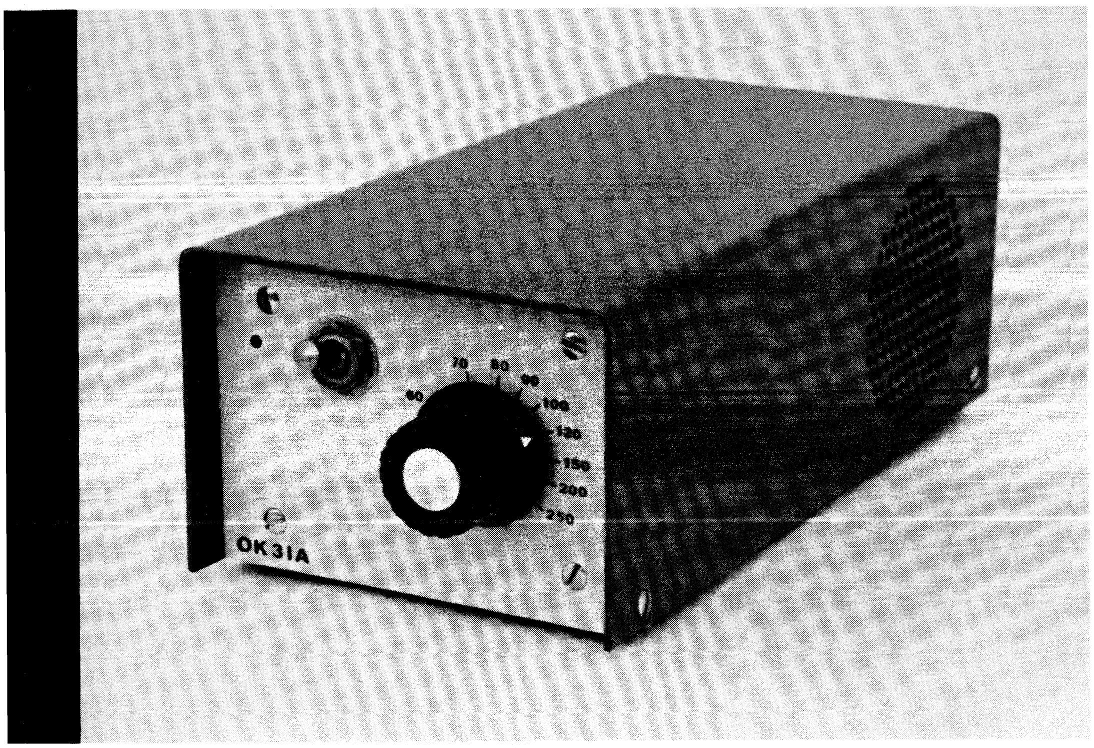
At Texas Instruments the accent is on power GaAs fets with claimed power output of 4.2 watts at 10 GHz; and RCA has announced 140 mW output at 22 GHz. The most impressive of the power GaAs fets, however, is a device from Fujitsu which provides 10 watts output at 4000 MHz. In theory 14 or 15 watts should be possible at 4 GHz with GaAs fet devices, and 6-7 watts at 10 GHz, but there are still a number of problems to be solved, so this capability may be five years in the future.

Some of the same technology that has produced high-performance microwave GaAs fets is also being used in other areas. At the International Solid State Conference in February, an engineer from Hewlett-Packard described a monolithic 4-GHz integrated amplifier which provides 28 dB gain from dc to 2000 MHz (gain is 7 dB from dc to 4 GHz). The three-stage amplifier is built on a single GaAs chip and uses reverse biased Schottky diodes as capacitors and MESFETs as resistors. At the same conference, Bell Labs reported on a uhf operational amplifier which has a unity gain frequency in excess of 1000 MHz. Silicon NPN transistors are used in the design, which provides 20 dB gain at 300 MHz; the operational frequency range is 10-500 MHz.

While GaAs fets threaten to displace bipolars above 3000 MHz, silicon MOSFETs are quietly moving in on microwave bipolars from the lower end of the frequency spectrum. Attention has been focused recently on advances in VMOS devices for vhf and uhf applications. Since the first commercial VMOS device was announced by Siliconix two years ago (*ham radio,* September, 1976), a number of semiconductor manufacturers have gotten on the VMOS bandwagon including Intersil, Motorola, and Westinghouse. One firm has reportedly obtained 10 watts at 1000 MHz and more than 5 watts at 1500 MHz with a VMOS transistor. One of the big advantages of VMOS is its negative temperature coefficient which eliminates thermal runaway and secondary breakdown. This means that emitter resistors, temperature sensing diodes, and other protection circuitry required for bipolars can be eliminated, resulting in a substantial cost savings. VMOS also offers better linearity with third-order IMD typically 3 to 5 dB better than bipolar power amplifiers.

As I said earlier, technological breakthroughs seem to come in spurts. During the past year or so there have been some remarkable achievements in the world of digital electronics, but it has been pretty quiet on the analog front. With the latest rf devices just now emerging from the laboratory, how long will we have to wait before all the dominoes are gathered for the next breakthrough?

Jim Fisk, W1HR
editor-in-chief



advanced electronic keyer

By defining keying intervals the author has developed a practical keyer that permits maximum time for character keying — enabling you to send letter-perfect code.

About thirty years ago, the bug was the most popular keying device, being widely used by hams all over the world. Only a small minority were using a new device, the electronic keyer, which had just begun to appear on CW bands, to produce fast and perfect machine keying. In the late 1940s, W6OWP described in *QST*[1] a new and simple principle for the electronic keyer. In Europe, this keyer was modified and popularized on the amateur bands by OZ7BO.

The W6OWP electronic keyer was simple, yet reliable. It did not use any clock-pulse generator, with the timing of dots, dashes and spaces performed by RC circuits, two triode vacuum tubes, and two relays. The dot, dash, and space time ratio was practically constant over the whole speed range. The most important features, though, were the self completing dots and dashes (including the following space), instant response when the paddle was closed, and sufficient time to release the paddle when the character generation was to be stopped.

electronic keyer requirements

Are there new requirements for the electronic keyer today? The optimum timing of characters, (dots, dashes, and spaces) over the entire speed

By Pavel Horvath, OK3IA, Institute of Physics, Slovak Academy of Sciences, 899 30 Bratislava, Czechoslovakia

range, and simple but reliable control of the keyer with the paddles are, of course, still expected. This can be easily understood by examining the timing of characters as shown in **fig. 1.** The time durations of the dash, dot, space within a character, space between characters, and space between words are defined as 3, 1, 1, 3, and 7 elementary time intervals. This timing of characters seems to be optimal for fast speeds. A different timing could be required at very low speeds; this problem, however, is not discussed here. Even though the keyer must accept imperfect keying, the output characters should be generated without errors.

Control of the keyer with the paddles is very important. Two types of paddles are generally used; the single paddle for standard keying and the dual paddle

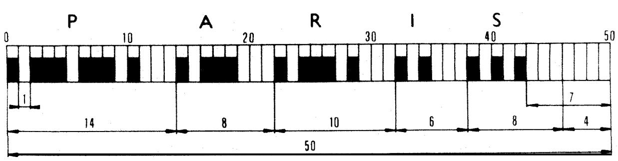


fig. 1. Timing of characters and spaces.

for squeeze or iambic keying. The single paddle, when released, slips into the neutral position and can be closed for a dot or for a dash. The dual paddle can be closed simultaneously for a dot and for a dash. Keying with a dual paddle is somewhat different than that with a single one.

Using the single paddle, there are two distinct groups of characters. First, those characters having an alternate sequence of dots and dashes (C for example) are generated by alternatively closing the

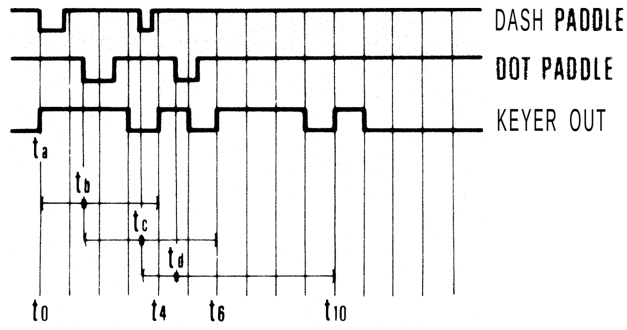
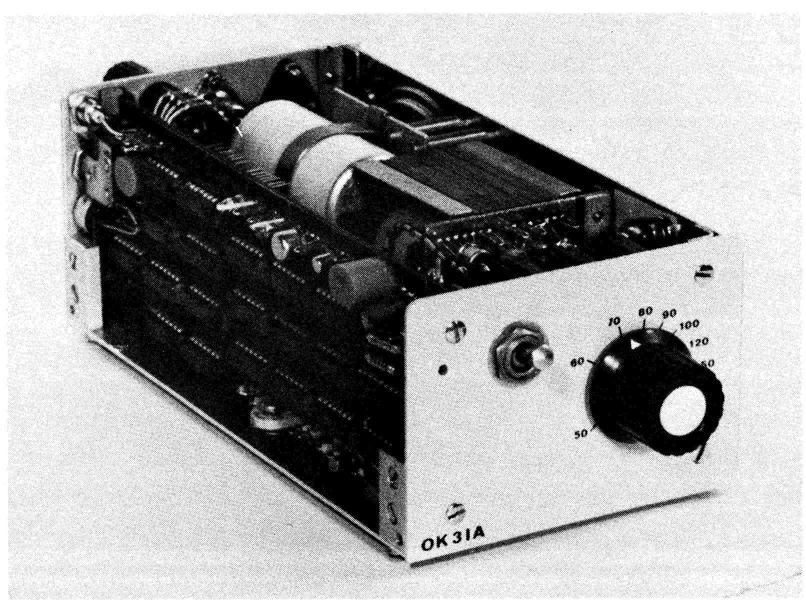


fig. 2. Permitted time intervals for closing the paddles to generate the character C.

paddle for a dash and then a dot. The first dash starts at the instant $t_a = t_\emptyset$, when the paddle is closed. No delay time should occur. By using dot and dash memories, the paddle can be moved to the dot side after the first dash begins. **Fig.** 2 shows the permitted time intervals for closing the paddle to generate the character C without errors.

The second group of characters are those which have a series of dots or dashes. **Fig.** 3 shows the keying required for the character = **(BT).** The paddle is first closed to generate a dash at the instant $t_a = t_\emptyset$, but yet, must be released and moved to the dot side before t_4. During the interval t_c to t_{10}, the paddle must be repositioned to form the last dash. For example, the paddle could be closed during t_c to t_{10} and released during t_d to t_{14}.

A somewhat different situation arises if you use dual paddles for iambic keying. The output with both paddles closed is called an iambic sequence. **Fig. 4** shows the keying for CQ. You start by closing the dash paddle, followed by closing the dot paddle. The



Front view showing the construction and placement of the printed circuit boards, power supply, speaker. and other components.

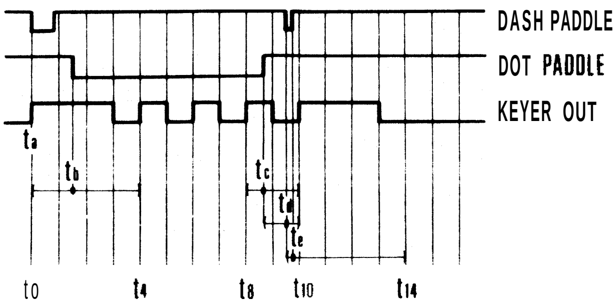


fig. 3. Permitted time intervals for closing and releasing paddles to generate the character = (BT).

dot paddle can be closed any time during the t_0 to t_4 interval. Both paddles are held until you recognize the last dot in the character C. Then you release both paddles. The sequence for releasing the paddles is not important, but both must be released during t_{10} to t_{12}. If the keyer also completes the spaces between characters, you can re-close the dash paddle anytime during the t_{12} to t_{14} interval. After the second dash in the Q has started (t_{18}), you must close the paddle for a dot during t_{18} to t_{22}, keeping both paddles closed to produce the character in the iambic mode. If both paddles are released anytime during t_{24} to t_{28}, the Q will be completed. When both paddles are released, the iambic sequence must stop after one bit (dash or dot) is completed. No second bit should be generated.

The permitted time intervals previously discussed are the maximum intervals in which the paddles can be closed or released, allowing the characters to be generated without errors. Any shorter time intervals make the keying worse. A summation of the previous points produces a list of features which an electronic keyer should incorporate:

1. Keying by both single and dual paddles.

2. Maximum possible permitted time intervals to make the keying easy.

3. At the beginning of the message, the character should start the instant the paddle is closed.

4. Optimum timing of dots, dashes, and spaces, throughout the entire speed range.

5. Stable clock generator, without clock pulse variations after triggering.

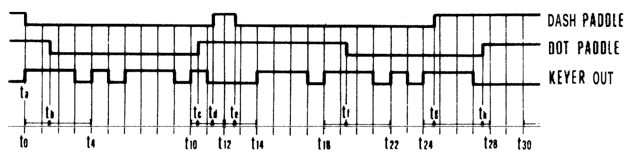


fig. 4. Keying of the word CQ in the iambic mode.

I began my design with the circuit described by WB2DFA.[2] In my version, I use very similar timing circuits, modified for a triggered clock and complete timing of spaces. My efforts were concentrated on improving the control of the keyer with paddles, making keying easy and convenient.

A block diagram of this keyer, with the significant signal names, is shown in **fig. 5.**

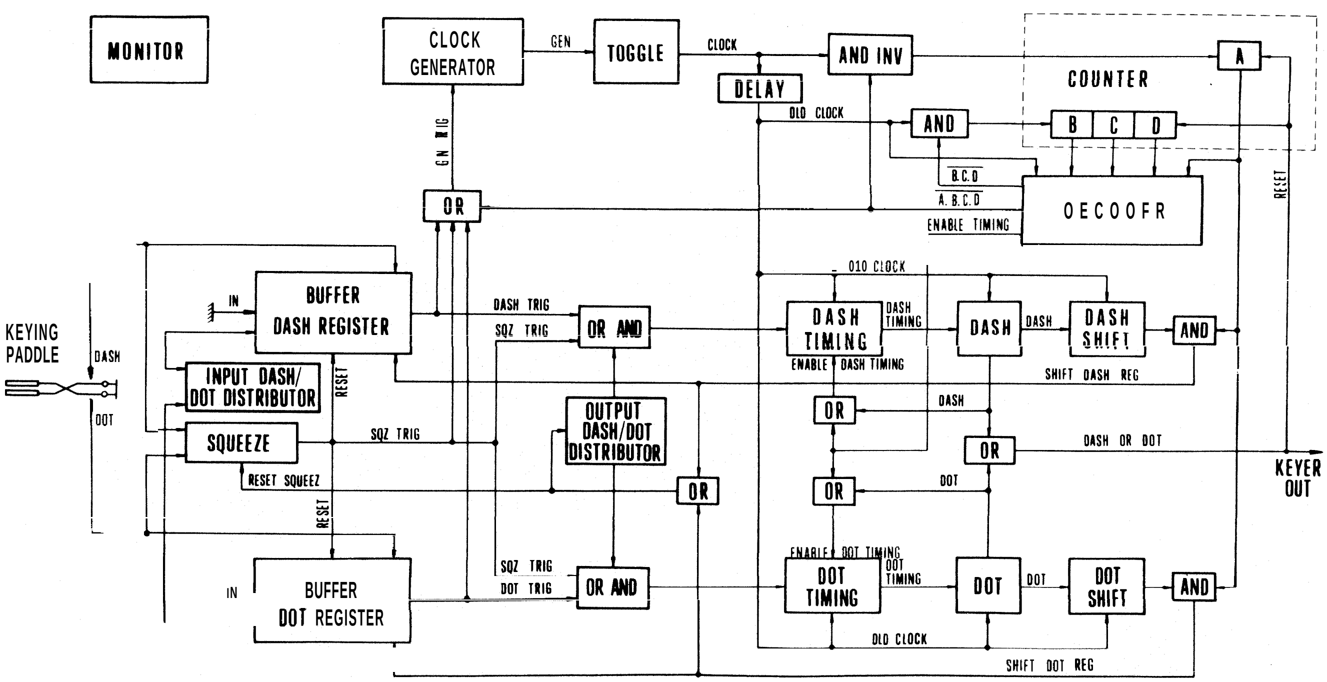
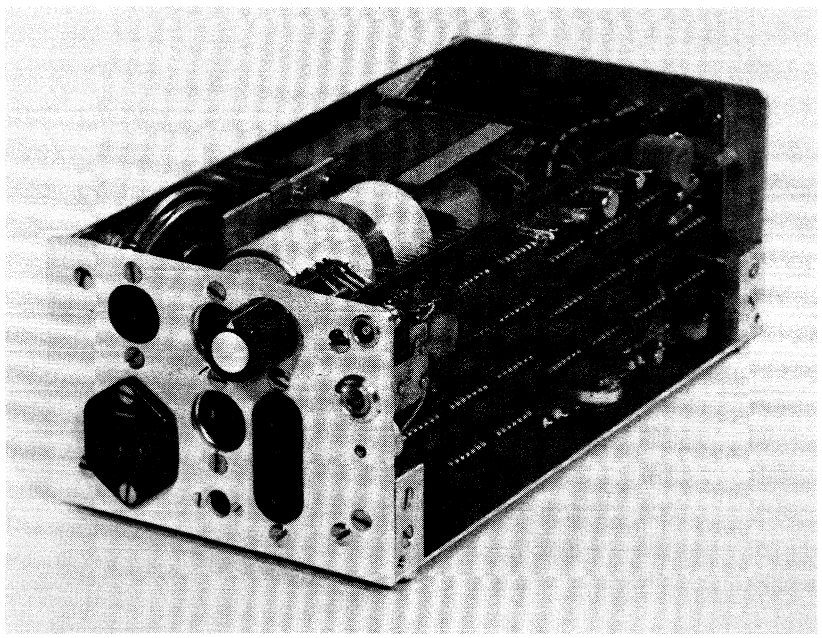


fig. 5. Block diagram of the advanced electronic keyer showing the most significant signals.

Clock generator. As shown in fig. **6,** the schematic diagram, the clock pulses are generated by transistors T_4, T_5, and the associated logic. In the initial state, the clock generator is off since GEN TRIG is low (0). The signal GEN TRIG is the logical sum of DASH TRIG, DOT TRIG, SQZ TRIG, MESSAGE START, and A•B•C•D. Because of the low from GEN TRIG, T4 is turned on by the current flowing through D3 and subsequently T5 is also turned on. At the instant the paddle is closed, GEN TRIG goes high, enabling the clock generator. Simultaneously, a short, negative pulse is formed by the gates connected to D4. This negative pulse turns off T4 and starts the generation of clock pulses. This circuit philosophy allows the generator to start in synchronization with the paddle. The clock pulses (GEN) are perfectly formed, the first pulse being the same as the following ones. The clock frequency is continuously variable and corresponds to speeds of 10 to 50 words per minute.



Rear panel view showing connectors for the ac line, 12-volt battery, paddles, manual key, external speaker, message memory, keyer output, S1 and S2, and speaker volume control.

The toggle and delay circuits divide the GEN signal by two, producing CLOCK and DLD CLOCK signals, with a delay of about 200 nS.

Space generation timing circuits. The circuitry in this portion of the keyer is used to provide a constant-width space following each element of a character, whether a dot or a dash. This self-completing space corresponds to the length of one dot. In addition, the timing of the space between characters and words can also be made self-completing, corresponding to 3 and 7 elementary-time intervals.

In the initial state (completion of a space), the four outputs of the binary counter are all high, causing the ENABLE TIMING line to also be high. When this line is high, the circuits for generating a dot or dash are not inhibited, and can be activated by pressing either paddle. Now, if a dash is to be sent, for example, the paddle is pressed, causing the clock generator to be activated, and the DASH signal to be generated. This signal, DASH, is the actual eventual output from the keyer. And, in addition to being the output, it is used to reset the binary counter for the duration of the signal. When DASH resets the counter, through the RO(1) and RO(2) inputs, the B•C•D and A•B•C•D are high, thus feeding the clock pulses to the binary counter. But, the pulses will not be counted until the completion of the dash since the resets override the normal clock input.

When the dash is completed, the CLOCK and DLD CLOCK pulses will then be counted. The gates connected to the B and D outputs of the binary counter will give an ENABLE TIMING pulse at the completion of the first or third timing interval; if there is a dot or dash waiting in the buffers, it will be generated after the first interval and the process will start again. If a dot or dash is not waiting, another ENABLE TIMING pulse will occur after the third timing interval (character space). If a character is still not in the input buffers, the word space will be generated (seven time intervals) by the gates connected to the B, C, and D outputs of the counter. At the completion of the word space, the A•B•C•D line will go low, stopping the pulse generator. If a character had been keyed in as the beginning of a new word, however, the pulse generator will not stop. Thus, if you are keying properly, the clock generator runs without interruption through the whole message.

Buffers and registers. The input buffers (or dot and dash memories) are used to store the code elements as they are put into the keyer. The dash and dot memories are actually 2-bit shift registers set up as FIFO registers, composed of D-type flip-flops. During keying, the dots and dashes are not only stored, but are also simuitaneously shifted into the timing circuits Therefore, it is possible to buffer characters in which the sum of the dots and dashes is greater than four.

Dash and dot distributors. The input and output dashldot distributors remember which kind of a bit, dash or dot, was stored first. The input distributor stores the dots and dashes in a sequence depending on the first bit keyed in: the output distributor correctly loads the content of the buffers ınto the timing circuits.

The character C, if stored in the buffers, is processed to the keyer output in the following manner. The output dashldot distributor, being enabled for a dash, allows the DASH TRIG signal to pass through the gates into the dash timing circuit, triggering the first dash. When the dash and the following single width space are completed, the signal SHIFT DASH REG shifts the dash register and clocks the output distributor to the dot memory, thus enabling the DOT TRIG signal through the gates into the dot timing circuit, triggering the dot. When the dot and the single width space are completed, the signal SHIFT DOT REG shifts the dot register, turning the output distributor back to the dash buffers. The same procedure is followed until the character C is shifted out of the buffers,

Dash and dot timing. The timing of a dash will start if the DASH TRIG or SQZ TRIG signal is low, the output dash/dot distributor enabled for a dash, and the signal ENABLE DASH TIMING high. The pulses, DLD CLOCK, transfer the high level on the D input to the output of the flip-flop, causing DASH TIMING to go high. With this signal high, a four-state binary counter receives the DLD CLOCK pulses. The counter states are then decoded into the signal DASH, which is three elementary-time intervals long. While the DASH is executed, the DOT circuit is disabled.

Dash shift and dot shift. The signals SHIFT DASH REG and SHIFT DOT REG, shown in **fig. 7,** are short pulses positioned at the end of the single width space following the dash or the dot. By having these pulses at the end of the space, you have the maximum amount of time in which to key in another dot or dash.

The block diagram in **fig.** 5 illustrates how these signals are formed. The signal CLOCK is delayed 200

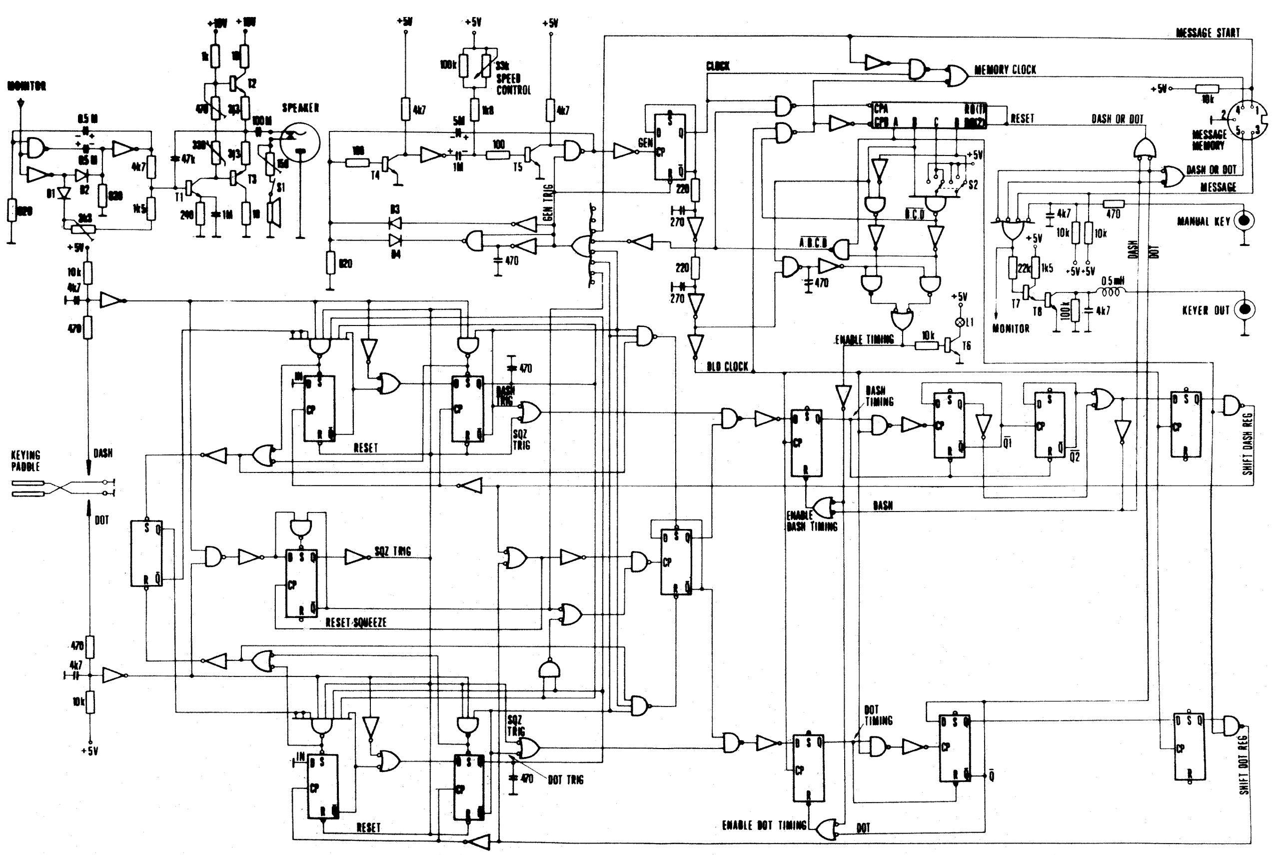


fig. 6. Schematic diagram of the advanced electronic keyer.

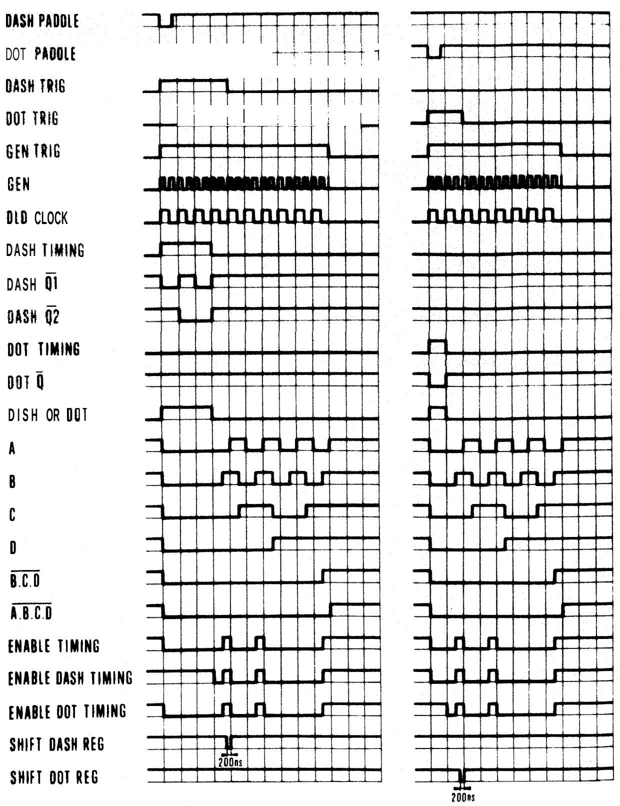


fig. 7. Timing diagrams for the signals shown in *figs.* 5 and 6.

nS, producing DLD CLOCK. Also, CLOCK is used to drive the A section of the binary counter. After the DASH and DOT signals are formed, they're shifted into another D-type flip-flop, which is clocked by DLD CLOCK. The dash or dot therefore initiates the shift register signal, while the DLD CLOCK stops it, 200 nS later.

Squeeze. During the initial phase of keying, the buffers, dash and dot registers, and input and output dash-dot distributors are involved. When both paddles are squeezed, SQZ TRIG goes low, clearing both buffer registers, enabling the clock generator, and executing the iambic sequence. When both paddles are released, RESET SQZ resets the squeeze circuit forcing SQZ TRIG to go high at the end of the first completed single-width space.

Monitor. The keyer contains an internal side-tone oscillator, which consists of a multivibrator, an integrating amplifier, and internal speaker. The multivibrator generates a l-kHz rectangular waveform, which is integrated in the amplifier. The triangularly shaped sidetone drives an internal 8-ohm permanent magnet speaker.

Message memory. Provisions have been made to connect the keyer to an external random-access memory. Four signals are available on a rear panel connector for operation of the memory. During the "message write operation" the MEMORY CLOCK and DASH or DOT are used for synchronous storing of the message. The MEMORY CLOCK automatically starts when the paddle is activated and continues to run during the manual keying. When the manual keying stops, the MEMORY CLOCK runs for seven clock pulses after message generation ceases. MEMORY CLOCK can be enabled to run through the remaining addresses by taking MESSAGE START low. During the "message read operation" MESSAGE represents the memory content previously stored and is transmitted through the OR gate to the keyer output.

construction

The electronic keyer is built in a 7 x 10 x 21 cm (2 3/4 x 4 x 8 1/4 inch) steel cabinet. The overall view is shown in the photographs. The ICs are mounted on a universal printed-circuit board with wire wrap used for the circuit connections. The parts list is shown in **table 1.**

concluding thoughts

This advanced electronic keyer could seem to some readers to be rather complicated. However, even an inexperienced homebrewer can expect perfect performance by exercising reasonable care during construction. Additionally, there are no special circuit adjustments necessary.

table 1. Semiconductors used in the electronic keyer

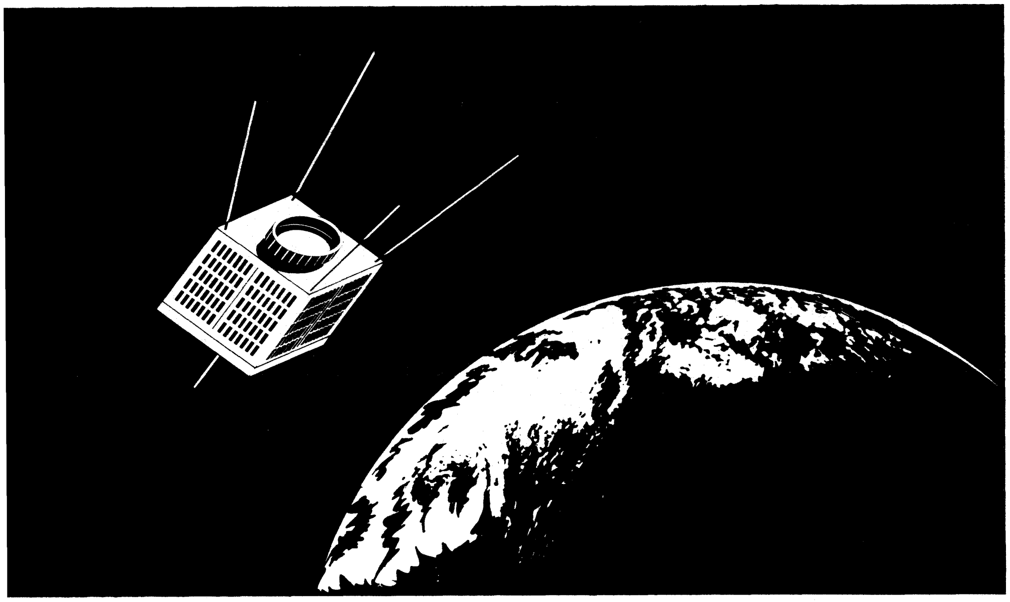
integrated circuits		**transistors**
7400	9 each	KC507 Silicon, 0.3 W T1, T4 - T7
7404	6 each	KF508 Silicon, 0.8 W T2, T8
7410	2 each	KF517 Silicon, 0.8 W T3
7420	1 each	
7430	3 each	
7474	8 each	Diodes
7493	1 each	KA221 Silicon, Switching D1 - D4

Only inexpensive TTL ICs were used in this project. The more experienced builder could change some circuit components or integrated circuits to ones more available in the United States. Or he could use NOR gates, monostable multivibrators, or MSI shift registers to minimize the number of parts and energy consumption. This should not be in contradiction with the aim of this article, however, which is to provide optimum keyer control and timing.

references

1. F. A. Bartlett, W6OWP, "Further Advances in Electronic Keyer Design," QST, October, 1948, page 27.
2. J. W. Pollock, WB2DFA, "Cosmos IC Electronic Keyer," ***ham radio,*** June, 1974, page 6.

ham radio



AMSAT-OSCAR D

The AMSAT-OSCAR D spacecraft is scheduled for launch sometime this month — here are the complete operating parameters for this new amateur satellite

AMSAT-OSCAR D, the next spacecraft in the OSCAR series, is a Phase II spacecraft which was built over the past two years by radio amateurs in the United States, Canada, Japan, and West Germany and is the first spacecraft in which AMSAT, Project OSCAR, and the ARRL have joined together to build flight hardware. The spacecraft makes extensive use of parts left over from the OSCAR 7 and Phase III programs, and was built primarily because the Phase III spacecraft will not be available until 1979. By stretching its resources almost to the limit, AMSAT has been able to work on both the Phase III spacecraft (with lots of publicity) and OSCAR D (with little publicity).

The new spacecraft carries transponders for two modes of operation. There is a conventional 145.9 MHz to 29.4 MHz Mode A transponder, and a new 145.9 MHz input, 435.1 MHz output, Mode J transponder – a similar frequency combination was used on the short-lived OSCAR IV spacecraft in 1966. In addition, six channels of telemetry are provided to monitor the onboard status of the satellite.

mission objectives

The principal objective of AMSAT-OSCAR D is its use as an educational tool in schools. Other objec-

By Perry Klein, W3PK, and Joe Kasser, G3ZCZ, AMSAT, Post Office Box 27, Washington, D.C. 20044.

tives include the continuation of communications demonstrations by means of stations in the amateur-satellite service, of the feasibility of using satellites with small amateur terminals for emergency communications, communications between medical centers and isolated areas, aeronautical, maritime, and land-mobile communications, direct satellite-to-home voice "broadcasting" to simple amateur receivers, and other similar applications. Further objectives are to demonstrate special operating techniques that enhance the usefulness of low orbits for these satellite applications, and to test the suitability of a new communications transponder frequency combination (Mode J) for small terminal users.

AMSAT-OSCAR D will permit the continuation of the education program, which began with OSCARs 5, 6, and 7, over the next several years, the anticipated lifetime of the satellite. OSCAR satellites have begun to play an important role in a new approach to science education. Used as remote laboratory tools, these satellites represent a pioneering utilization of an active space system in the classroom. Using inexpensive ground terminals for OSCAR satellites in schools, students can gain firsthand experience in space science. This type of direct, active involvement has relevance to the study of communications, astronomy, engineering, physics, mathematics, and meteorology. The low-cost OSCAR ground terminal puts an active satellite system at the disposal of the instructor and student.

spacecraft description

OSCAR D is a communications satellite in the Phase II (low-orbit) series which is designed to operate with small stations in the amateur-satellite service on a non-commercial basis. The spacecraft contains two communications transponders and command and telemetry systems; it is solar powered, weighs 27 kg (60 pounds), and is a 38 cm (15-inch) rectangular solid 33 cm (13 inches) high. Its anticipated useful operating lifetime is three years.

Two types of communications transponders are aboard the spacecraft. Normally, only one transponder will be operated at a time because of spacecraft battery constraints.

Two-to-ten meter transponder. The *Mode* A transponder is a two-to-ten meter unit similar to the one used on OSCAR 7 (input frequency passband 145.85-145.95 MHz, and output frequency passband between 29.40 and 29.50 MHz). A 250 mW telemetry beacon provides telemetry data in Morse code at a frequency of 29.402 MHz. Approximately −95 dBm is required at the transponder input terminals for an output of one watt. This corresponds to an effective radiated power from the ground of 80 watts for a distance to the satellite of 1930 km (1200 miles) and a polarization mismatch of 3 dB. The transponder translation frequency (input frequency minus output frequency) is 116.458 MHz. Thus, the relationship between the uplink (f_u) and downlink (f_d) is

$$f_d = f_u - 116.458 \pm \textit{Doppler}$$

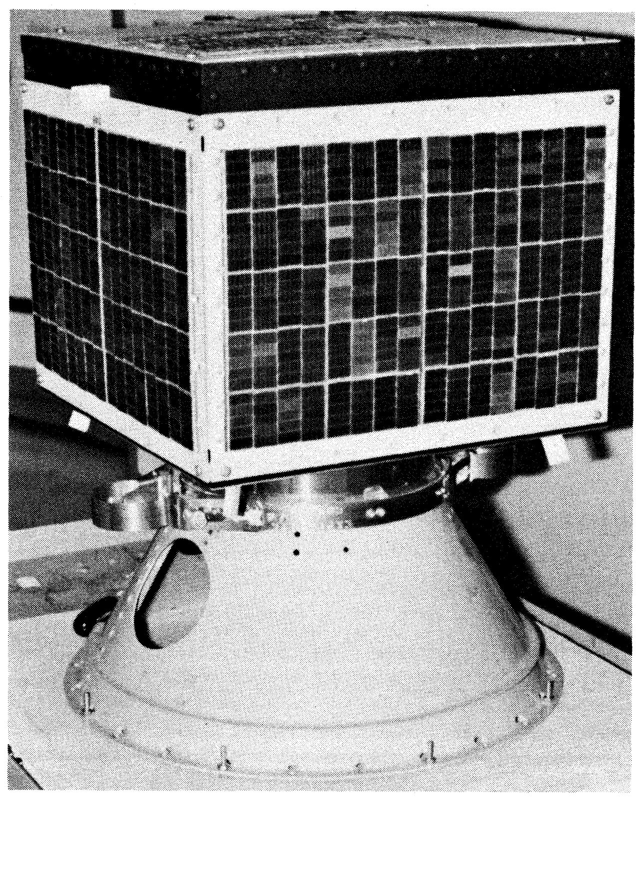
where both f_d and f_u are in MHz.

An uplink signal at 145.900 MHz, for example, will produce a down-link signal from the transponder on 29.442 MHz ± Doppler. As in the two-to-ten meter transponders in OSCAR 6 and 7, the passband is *not* inverted, and upper-sideband uplink signals become upper-sideband downlink signals. Output power is 1 to 2 watts.

Note that the downlink frequency will be slightly different (8 kHz) from that of the equivalent OSCAR 7 Mode A transponder which has an equivalent frequency relationship of

$$f_d = f_u - 116.450 \pm \textit{Doppler}$$

AMSAT OSCAR D spacecraft to be launched in March, 1978, by NASA. When it is in orbit it will be designated AMSAT OSCAR 8.



Two meter-to-70cm transponder. The Mode J transponder, constructed by members of the Japan AMSAT Association in Tokyo, uses a two-meter input, 70 centimeter output combination which has not yet been flown in the AMSAT Phase II series. This transponder operates with an input frequency passband of 145.90-146.00 MHz, and an output frequency passband of 435.10-435.20 MHz. Power output is about 1-2 watts PEP, and the output passband is *inverted, ie.*, upper-sideband uplink signals become lower-sideband downlink signals. The transponder translation frequency (input frequency plus output frequency) is 581.1 MHz ± Doppler. Uplink sensitivity for one-watt output is – 105 dBm, corresponding to an effective radiated power from the ground of 8 watts for a distance to the satellite of 1930 km (1200 miles).* Note the greatly improved sensitivity of this mode, and keep your power down. A 100-milliwatt beacon carries telemetry at a frequency of 435.095 MHz. The relationship between the uplink (f_u) and downlink (f_d) is

JA1CBL of Japan AMSAT Association (JAMSAT) building parts of the Mode J transponder.

$$f_d = 581.1 - f_u \pm \text{Doppler}$$

where both f_d and f_u are in MHz.

antenna system

Both the Mode A and Mode J transponders use the same receiving antenna, a canted turnstile comprised of four 48 cm (19-inch) lengths of 12.5 mm (112-inch) carpenter's rule fed by a hybrid and matching network so as to develop circular polarization. One port of the hybrid feeds the Mode A receiver; left-hand circular polarization is required by users in the Northern Hemisphere and right-hand circular polarization is required in the Southern Hemisphere. A second port of the hybrid is connected to the Mode J receiver; right-hand circular polarization is required in the Northern Hemisphere, and left-nand circular polarization in the Southern Hemisphere. The antenna gain should approach 5 dB in the $-Z$ direction (*i.e.*, toward the bottom of the satellite).

The Mode A ten-meter downlink antenna is a linearly-polarized dipole which is oriented perpendicular to the stabilization magnets in the spacecraft as in OSCAR 6 (but unlike OSCAR 7, which has the ten-meter antenna parallel to the axis of the magnets).

The 435-MHz Mode J downlink antenna is a simple monopole, linearly polarized, and located on the top of the spacecraft. Note that its location may result in some radiation shielding at high Southern Hemisphere latitudes.

"Sensitivity may decrease by a factor of ten (10 dB) under different conditions of battery voltage and satellite operating temperature, so that as much as80 warts may be required at certain times.

telecommand system

A five-function telecommand system of a new design is carried on OSCAR D. The system is based on the best features of the OSCAR 6 and 7 telecommand systems, and is designed to be virtually immune to noise and interference. The command functions are:

Mode A Select (Two-to-ten meter transponder ON)
Mode J Select (2 meter to 70 cm transponder ON)
Mode D Select (recharge mode; both transponders OFF)
Ten-meter Antenna Deployment
Ten-meter Antenna Reset

telemetry system

OSCAR D contains a six-channel Morse code telemetry system similar to the units flown in OSCARs 6 and 7. Telemetry is sent at 20 words per minute as three-digit numbers; *A1* emission is used in keying the Mode A or Mode J telemetry beacons, depending upon which transponder is in use. The six telemetry parameters are:

Channel 1. Total Solar Array Current –
$I_T = 7.15\,(101\text{-}N)$ ma.

Channel 2. Battery Charge-Discharge Current –
$I_{Bat} = 57\,(N\text{-}50)$ ma.

Channel 3. Battery Voltage –
$V_B = 0.1\,N + 8.25$ Volts

Channel 4. Baseplate Temperature –
$T_{bb} = 95\ 8 - 1.48N\ (^{\circ}C)$

Channel 5. Battery Temperature –
$T_{Bat} = 95.8 - 1.48\ N\ (^{\circ}C)$

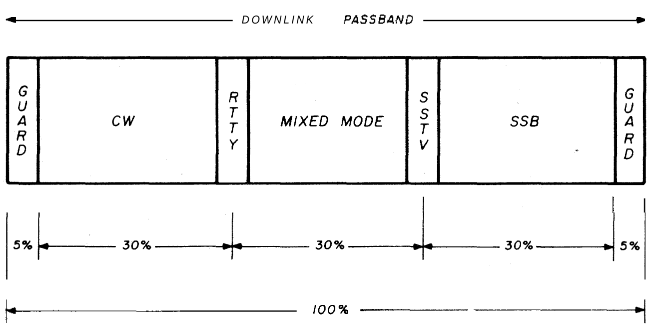
Channel 6. RF Power Output, Mode J –
$P_{JT} = 23N$ *milliwatts*

Note that, unlike OSCAR 6 and 7 telemetry, OSCAR D has only one parameter per line (OSCARs 6 and 7 had 4). As a result, a complete telemetry frame is sent in approximately 20 seconds.

power supply

The spacecraft contains solar panels on its four sides (along the +X, –X, +Y, and –Y axes), and on the top (the +Z axis). No panels are contained on the bottom (–Z axis) since this is where the spacecraft is attached to the launch vehicle. The solar cells, combined with a 12-cell, six-ampere-hour rechargeable nickel-cadmium battery should be adequate to power the spacecraft with a positive power budget in Mode A for several years, even considering solar cell degradation in the radiation environment. The power drain in Mode J, however, is somewhat larger, so the Mode J transponder probably cannot be operated continuously.

A battery charge regulator is also contained which converts from the 28-30 volt solar array voltage to the 14-16 volts required by the battery. It also tapers the charge rate so the battery trickle-charges as the battery approaches full charge (as indicated by the battery voltage).



1. Guard bands to avoid interference to beacons. These frequencies are available for emergency and bulletin stations.
2. RTTY and slow-scan television are placed at the edge of the CW and ssb passbands. This conforms to their high-frequency use where RTTY is present within the CW space, and SSTV is transmitted in the ssb sub-band.
3. Mixed-mode area is recommended for crystal-controlled stations, DXpedition stations, or anyone wishing to work both CW and ssb stations.

fig. 1. Basic satellite band plan proposed by G3ZCZ and adopted by the AMSAT Board of Directors in October, 1977. This band plan allocates a percentage of the available radio frequency spectrum as seen on the downlink to different modes of communication. The relative amount of spectrum for each mode is thus the same for any transponder in any satellite. Band plan frequencies for AMSAT OSCARs 7 and 8 are shown in *fig.* 2.

stabilization system

Four permanent magnets located inside the spacecraft and aligned along the Z axis provide stabilization, as in OSCARs 6 and 7. The polarity of the magnets is such that the top (+Z axis) of the spacecraft always points toward the magnetic north pole of the earth. Hysteresis permalloy damping rods mounted behind the +X, –X, +Y, and –Y solar panels are designed to reduce the spin of the spacecraft about the Z axis; they function in a manner similar to a shorted transformer turn as it cuts the lines of flux of the earth's magnetic field. The permalloy rods are left over from OSCAR 7, which successfully used the same type of stabilization system.

launch interface and orbit

The OSCAR D spacecraft is being launched from the NASA Western Test Range as a secondary payload with the NASA Landsat-C earth resources technology satellite and the NASA PIX (Plasma Interaction Experiment). The spacecraft will be ejected from the second stage of the two-stage Thor-Delta 2910 launch vehicle 5120.6 seconds after lift-off, at an approximate position of 78 degrees N. latitude and 15 degrees W. longitude. Programmed orbital parameters are:

Apogee	928 km (577 statute miles)
Perigee	884 km (549 statute miles)
Period	103 minutes
Inclination	99.0 degrees

Time of Descending Node 9:30 AM
(launch window from 1754-1824 UTC)

The orbit is planned to be sun-synchronous with passes repeating at the same time each day on a one-day cycle (as opposed to the two-day cycle of OSCARs 6 and 7).

spacecraft initialization

AMSAT-OSCAR D will automatically be powered up upon ejection from the Thor-Delta launch vehicle over northern Greenland at which time it will assume the next available number in the OSCAR series. It is designed to initialize itself in Mode J (two-meter-to-70-cm transponder ON). The two-to-ten meter (Mode A) transponder will be initialized OFF and should be kept off until the spacecraft is nearly completely stabilized, which may require as much as a

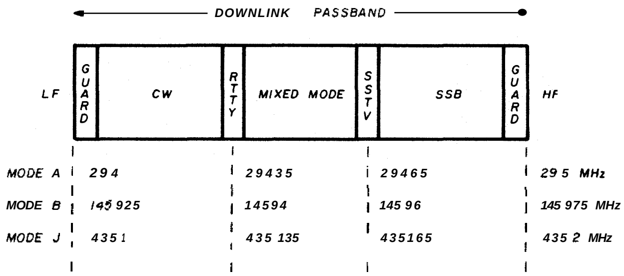


fig. 2. Satellite band plan for AMSAT OSCAR 7 and 8. As shown in *fig. 1*, this plan is based on percentages of the downlink passband, and applies to both inverting and non-inverting transponders. In Modes A and J the guard channels are 5 kHz wide; the guard channels in Mode B are 2.5 kHz wide.

week. Because of the non-rigidity of the deployable ten-meter dipole antenna, this antenna must not be deployed until the spacecraft spin rate is less than 1 rpm; otherwise, the antenna may be severely damaged or may not deploy properly. OSCAR D's ten-meter antenna is comprised of tubular extendable members which are slowly unreeled from the spacecraft by small motors. The deployment process takes approximately 15 seconds and is non-reversible (*i.e.*, the antenna elements cannot be retracted once they are deployed). During the time when the antenna is being deployed, the telemetry beacon switches from its normal Morse code format to a series of keying pulses, the rate of which is a function of the tip-to-tip length of the ten-meter dipole; the rate-of-change of the telemetry pulses will be carefully tape-recorded during deployment of the antenna to permit analysis later to verify success.

telecommand verification

OSCAR D's telecommand and telemetry systems have been designed to provide two means to verify whether the spacecraft is accepting commands. First, when the telecommand system has been enabled and is ready to accept a command, the Morse code telemetry will be interrupted and an unmodulated carrier will be heard on the beacon frequency. The beacon will revert back to Morse code telemetry when the telecommand system is no longer enabled.

The second method of telecommand verification is to use the "Ten-meter Antenna Deployment" command. This will cause a series of keying pulses to be heard on the telemetry beacon in place of the Morse code telemetry if the command has been accepted. The "Ten-meter Antenna Reset" command should be sent soon afterward to restore the beacon to the Morse code telemetry mode.

telemetry interpretation

The most important telemetry channel that will affect operations decisions is channel 3 (battery voltage). In Mode A the spacecraft should maintain a positive power budget so there should not be a net discharge of the battery over an orbit average. Mode J operation, however, requires somewhat more power, which may result in a net discharge of the battery, especially under conditions of high transponder loading; therefore it will be necessary for telemetry and telecommand stations to keep a close watch on the battery voltage so that action can be taken as necessary to command the spacecraft into Mode D (the recharge mode) before the battery discharges too far. Three cutoff levels are specified below:

Red Level **A** (1.2 volts/cell) Channel 3 = 61 counts
Red Level **B** (1.1 volts/cell) Channel 3 = 50 counts
Red Level C (1.0 volts/cell) Channel 3 = 38 counts

Red Level **A** should be used during the first year or so of the spacecraft's life as the cutoff point below which telecommand stations should command the satellite into Mode D for recharging. Later in the spacecraft's life, as the battery discharge characteristic curve changes, Red Level B should be used; Red Level C should be used if there is evidence of battery deterioration or if it is desired to recondition the battery.

Channel **1** (solar array current) provides an indication of whether the spacecraft is in the sun or eclipse (it should read in the nineties in counts when in eclipse). Fluctuation in channel 1 telemetry is the best indicator of the rate of spin of the spacecraft, along with observations of fading, particularly of the 435-MHz Mode J downlink signal from the quarter-wave 435-MHz monopole antenna.

Channel **2** (battery charge-discharge current) gives information on whether the battery is charging or discharging. A reading larger than 50 counts indicates that the battery is charging, while a reading of less than 50 counts means the battery is discharging. There is a two-second integration time associated with the current telemetered on this channel. The total power drain of the spacecraft can be determined by observing channel 2 while the spacecraft is in darkness (as indicated by channel 1, which should read in the 90s in darkness).

Channels **4** and **5** (baseplate temperature and battery temperature) should generally track within a few degrees (except perhaps in the first day or so after launch when the spacecraft has not yet stabilized at thermal equilibrium). Experience from OSCARs 6 and 7 indicate that the battery can overcharge and overheat during periods of the year when the spacecraft sees the most sunlight. If this is the case, channel 5 may exceed channel 4 in temperature by 10 degrees celsius or more, and action should be taken to reduce this overheating. This can be accomplished by keeping the spacecraft in Mode J to consume any extra charge current from the battery.

Channel 6 is a measure of the Mode J transponder 435-MHz rf power output. Associated with the telemetered readings is an integration time of 2.5 seconds, so that it is average power rather than peak power that is telemetered. There is no telemetry of the Mode A transponder. The Mode A transponder power consumption (largely determined by power amplifier current) can be measured by observing channel 2 telemetry as noted above.

operating schedule

Since the prime mission of the AMSAT-OSCAR D spacecraft is to use the Mode A transponder for the ARRL OSCAR educational program in schools, the spacecraft may be left in Mode A during weekdays (Monday through Fridays, United States time), and put in Mode J on weekends. Note that all communications should conform to the G3ZCZ band plan shown in **figs. 1** and **2.** Additionally, if not an excessive burden on the telecommand stations, evening orbits in the Western Hemisphere (morning orbits in the Eastern Hemisphere) can be switched to Mode J, battery permitting. In any case, all operation in Mode J will require careful monitoring of the battery charge level (as indicated from channel 3 telemetry). The power budget may not support the Mode J transponder for continuous fulltime operation in this mode over an entire weekend. In any event, details of the operational modes of the spacecraft will be announced by AMSAT in the *Newsletter,* with late updates on the AMSAT Nets.

OSCAR D will operate in a 900 km (560 statute mile) orbit, *ie.,* at just over half the altitude of the orbit of OSCAR 7. Thus, communication ranges will be different. The usable time on an overhead pass will be about 18 minutes instead of the 22 minutes provided by OSCAR 7, and the horizon range will be 3220 km (2000 miles) instead of the 3940 km (2450 miles) of OSCAR 7. This means that transatlantic communications will still be possible, but not as often as with OSCAR 7.

Keeping track of this satellite is going to be much simpler than for OSCAR 7. It will come into range at approximately the same time each day; the overhead descending node pass is planned for 9:30 AM local time.

credits

It is impossible to single out all those who contributed to the construction of the spacecraft, but a few calls can be listed.

JAMSAT – Mode J Transponder: JA1ANG, JA1CBL, JG1CDM, JA1VDV, JA1JHF, JR1SWB

AMSAT – Mode A Transponder: WA4DGU, W3PK

Morse Code Telemetry System: W5CAY, WA4DGU

Telecommand System: W3GEY, WA3LND, WA3ZCE, W3HUC, W3ITO, K1RT/WA1JZC

Antenna and Antenna Deployment Module: W3GEY, W3HUC, W3ITO, K1RT, WA3LND

Power System: DJ4ZC, JA1TUR, JF1DMQ, K1RT, W3HQ

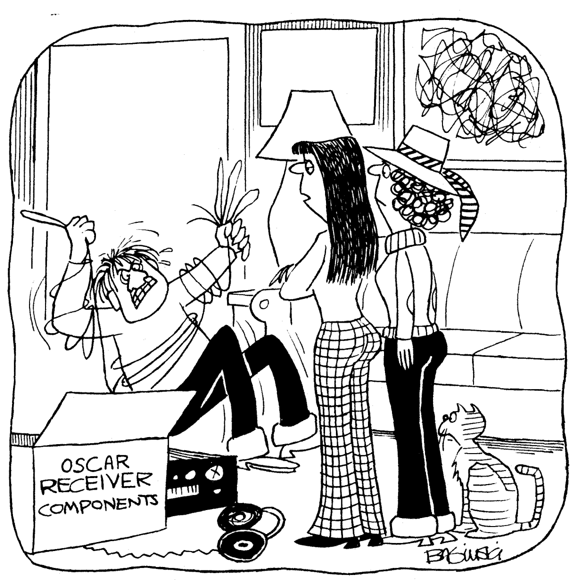
Structure and Module Containers: K6GSJ and Project OSCAR, K1JX/WA1JLD, K1RT, WA4DGU, VE3DPB, W3HSO, WBØGIM, Henry Smith, David Vanderbeke, W3ZKI

Cables and Wiring: Marie Marr, W3TMZ

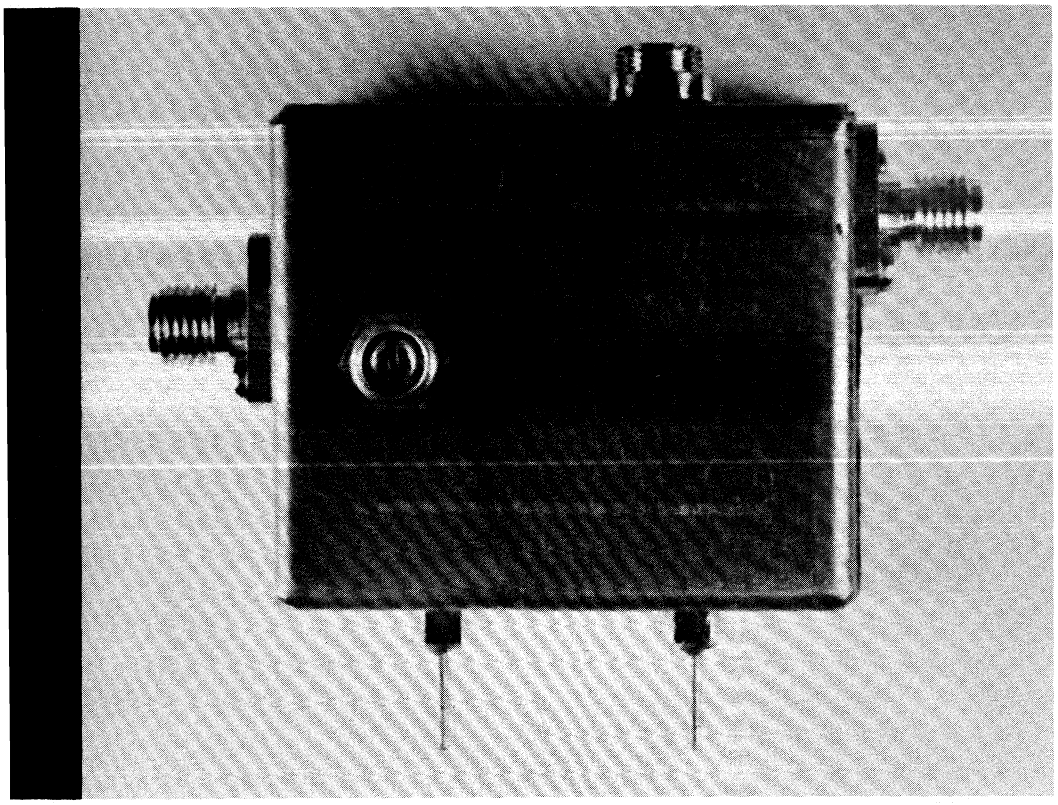
Engineering Drafting: WB4GIB

To others who contributed or assisted, our thanks and the thanks of thousands of radio amateurs, school children, and educators. Let's use the satellite wisely so that it will further help the educational program until the Phase III satellites are flying high.

ham radio



"Alfred was never any good in a crisis!"



very low-noise GaAs FET Preamp for 432 MHz

Construction details for a uhf GaAs fet preamp that will provide a 0.7 dB noise figure with 18 dB gain at 432 MHz

Every amateur who is active on the uhf bands is looking for ways to improve his system performance. One of these ways is to use a receiver preamplifier with lower noise figure and higher gain; some excellent preamplifier circuits have been published in *ham* radio. The low-noise uhf preamplifier described by W1JR in the March, 1975, issue is perhaps the best known of these circuits,[1] and is widely used by EME operators and others who are interested in long distance, weak-signal uhf communications.

This article describes a low-noise, state-of-the-art preamplifier for 432 MHz which uses an NEC NE24406 (2SK85) GaAs fet. This preamp is capable of a 50°K (0.7dB) noise figure at 423 MHz, so it can improve the performance of your uhf receiving system by about 100°K, as compared with a conventional bipolar transistor.

By Shigeru Sando, JH1BRY, 8-17-204 Sakonyama Asahi-ku, Yokohama 241, Japan. Mr. Sando is a microwave semiconductor development engineer at the Semiconductor Division of the Nippon Electric Company.

field-effect transistors

The history of the fet is older than most people realize. In fact, it's one of the oldest three-terminal solid-state devices. Dr. Shockley first proposed the fet in 1952, but because of a variety of technological and fabrication difficulties, fets did not become practical until the early 1960s.

Basically, there are three different types of fets. The junction fet or *jfet* is the simplest of the three types and became commercially available about the same time as the first bipolar microwave transistors.

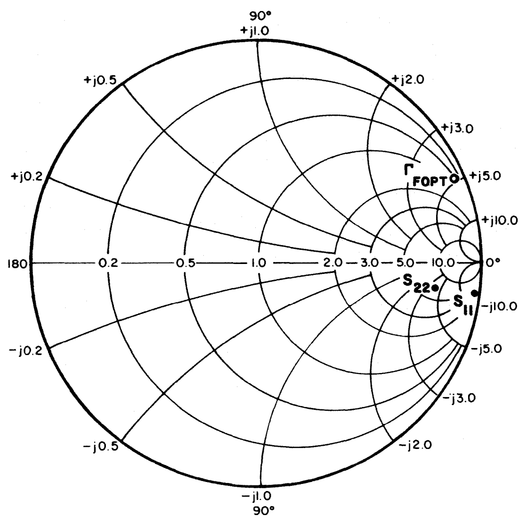


fig. 1. Smith chart plot of the NE24406 impedance characteristics at 432 MHz. Γ_{FOPT} is the optimum source impedance for noise figure at 423 MHz.

Advances in fabrication techniques and requirements for lower power fostered the development of the metal-oxide-semiconductor fet or *mosfet.* Both the jfet and mosfet are widely used in applications which require high input impedances, such as in the input stages of test instruments. The same fabrication processes developed for mosfets are also useful in the manufacture of integrated circuits.

In the microwave region the bipolar transistor has reigned supreme for a number of years — it wasn't threatened by either the jfet or the mosfet. However, a third type of fet has changed that. This new fet, which uses a Schottky barrier at the gate electrode, is called the metal-semiconductor fet or *mesfet.* Mesfets use Gallium Arsenide (GaAs) as the semiconductor material and are usually referred to as GaAs fets (pronounced gas fets).

Gallium arsenide has high electron mobility (5 to 7 times as high as silicon), and offers significant advan-

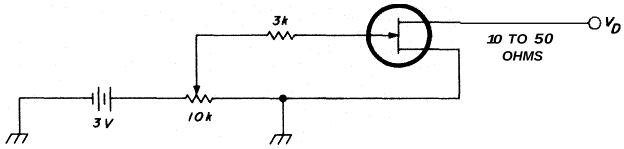


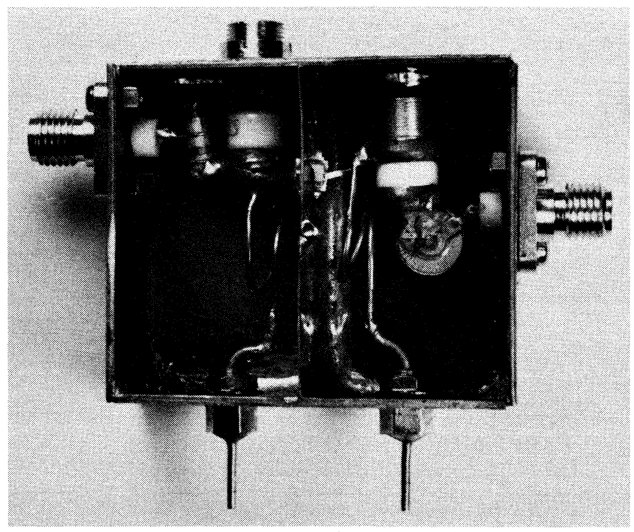
fig. 2. Recommended GaAs fet bias circuit uses a series resistor in the gate circuit to protect the device from transients.

tages over silicon at microwave frequencies. The intrinsic characteristics of GaAs result in shorter transit times and lower resistance, thereby providing higher gain, lower noise figures, and extremely high cut-off frequencies — all important characteristics for microwave transistors.

Gallium arsenide has been under development for several years, and practical microwave GaAs fets are now available off the shelf. Recently I had an opportunity to experiment with the NEC NE24406 (2SK85) GaAs fet. Although there have been many published reports which describe the performance of the NE24406 in amplifiers up to X band (about 12 GHz), there have been no published data on their use below 500 MHz. This article describes the first experimental results of the NE24406 GaAs fet on the amateur 70 cm band.

noise figure

Obviously a transistor with a low noise figure is required in a low noise preamplifier, but that's not the



Layout of the low-noise 432-MHz preamplifier. The input is to the left, output to the right. The GaAs fet is installed in the center shield partition. Capacitors C1 and C3 are supported by their own leads; C2 (left compartment) and C4 (right compartment) are mounted in the side walls of the chassis

only requirement. Also to be considered are the effects of the next stage's noise figure and the requirements for a low-loss matching circuit, low feedback, and good stability.

The overall noise figure of a receiving system is given by the following formula

$$NF_T = NF_1 + \frac{NF_2 - 1}{G_1} \qquad (1)$$

Where

NF_T = Total overall noise factor

NF_1 = Noise factor of the first stage

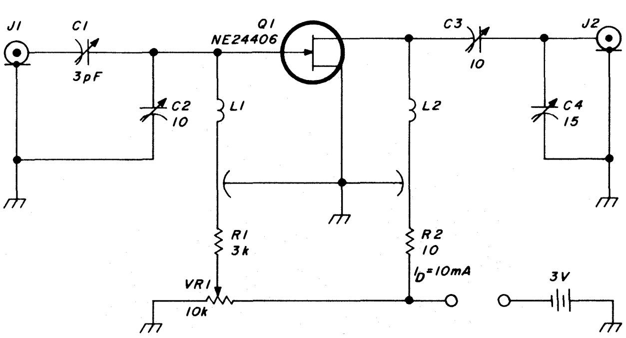
NF_2 = Noise factor of the second stage

G_1 = Available power gain of the first stage

Note that the noise factors and power gains must be in ratios, *not* in dB. For example, consider that you have a preamplifier with a 1.0 dB noise figure ($NF_1 = 1.259$) and 10 dB power gain ($G_1 = 10$); the noise figure of the second stage is 5 dB ($NF_2 = 3.162$). The calculated overall noise factor NF_T is 1.475 (or a noise figure of 1.688 dB).

Note that the system noise figure has increased about 17 per cent as compared to the preamplifier's noise figure. For best results, it's recommended that you use a lower noise second stage with a higher gain preamp.

One question that often arises is, "Can the noise figure of the preamplifier be as low as the noise figure of the transistor itself?" No, it cannot, even on the



C1 3 pF piston trimmer (Johanson 7274 or 7284)
C2, C3 10 pF piston trimmer (Johanson 5202)
C4 15 pF piston trimmer (Johanson 5402)
L1, L2 1 turn no. 20 (0.8mm), 1/4" (6mm) ID (see **fig. 5**)
J1, J2 SMA type coaxial connectors
Q1 NEC NE24406 GaAsfet

fig. 4. Schematic diagram of the low-noise preamp for 432 MHz which provides a 0.7 dB noise figure and 18 dB gain. Lowest noise figure occurs at drain current flow of 10 mA (see fig. 7).

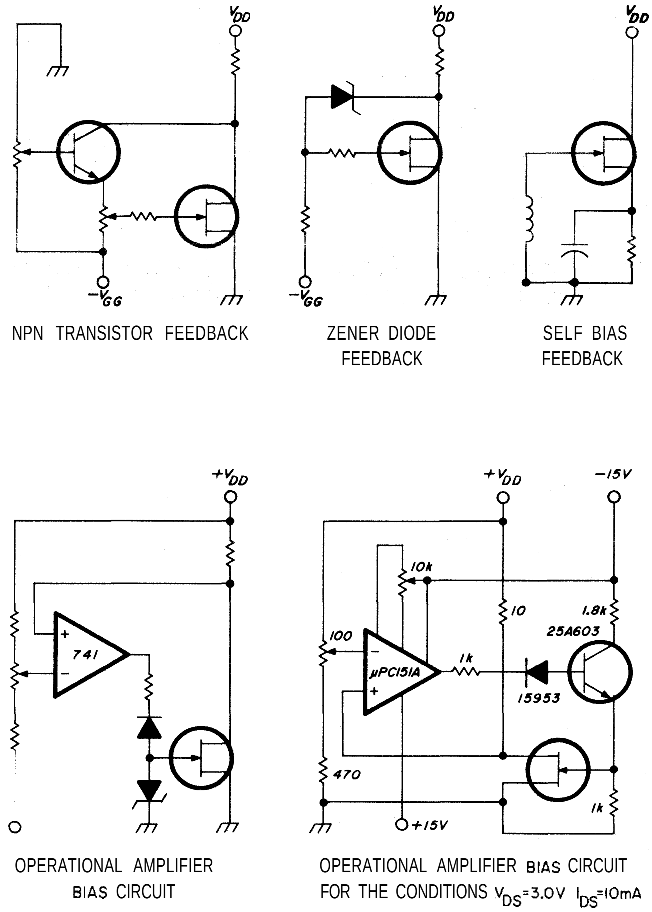


fig. 3. Several circuits that may be used to provide bias in fet circuits. Note that all *circuits must* be bypassed *a?*rf.

assumption that there are no matching or circuit losses. One way to describe the optimum noise figure of a transistor is *noise measure* which is expressed as

$$M + 1 = \frac{NF - 1}{1 - \frac{1}{G}} \qquad (2)$$

where

NF = Noise figure of the transistor

G = Associated gain of the transistor

The value of $M + 1$ shows the ideal total system noise figure, and indicates the same result where an infinite number of transistors with the same characteristics are used in cascade.

Assume you have a transistor which has a specified noise figure of 1.2 dB ($NF = 1.318$) and 14 dB gain ($G = 25.12$). From **eq.** 2

$$M + 1 = \frac{1.318}{1 - \frac{1}{25.12}} = 1.351 \text{ or } 1.243 \text{ dB}$$

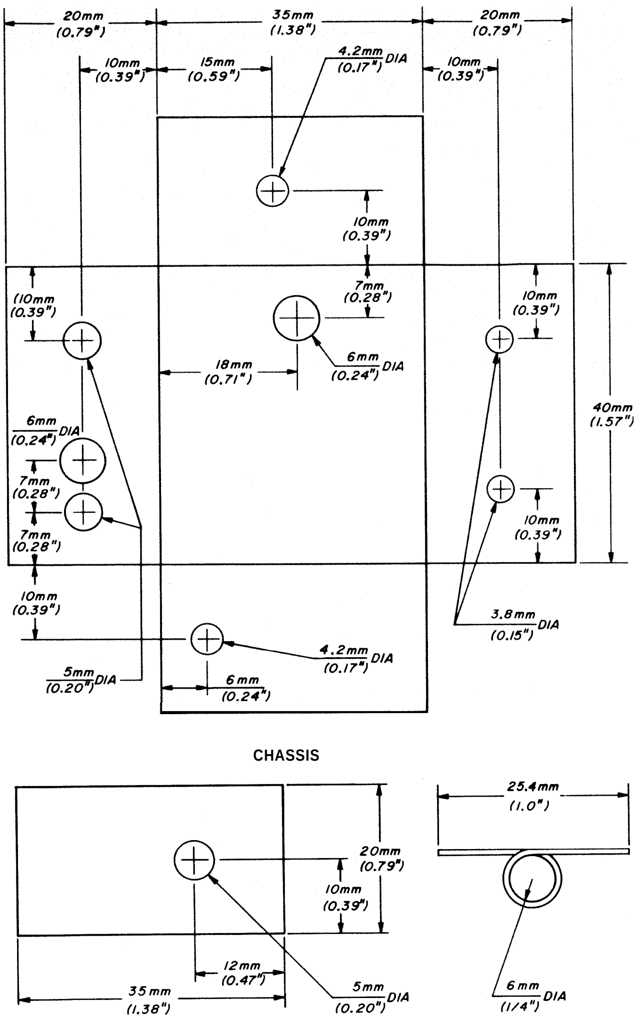


fig. 5. Layout of the low-noise GaAs fet preamp for 432 MHz. The enclosure is made from brass sheet, 0.028"–0.048" (0.7–1.2mm) thick.

This shows the minimum value of amplifier noise available from a transistor with a 1.2 dB noise figure and 14 dB gain (assuming no matching or connector losses).

circuit description

The GaAs fet has several disadvantages which make it somewhat more difficult to use than the bipolar transistor. For one thing, as shown in the Smith chart plot of **fig. 1,** the impedances are very high. This also indicates that the GaAs fet is a high Q device and not easy to match to low impedances such as 50 ohms. Also to be considered is the bias circuit. Once an operating point has been established for the GaAs fet, a bias circuit must be chosen which will provide stable operation over the required temperature and frequency range.

The importance of affording adequate protection against transients and change in I_{DSS} cannot be over emphasized. As every designer who uses GaAs fets will eventually discover, transients are the leading "fet killer." The most likely burn-out mode is a short from gate to drain or from gate to source which is caused by high field or high current transients. The highest field in common-source operation is between the drain and gate, and should never exceed 10 volts.

The best way of applying bias is to use a battery through a series resistor to the gate, as shown in **fig. 2.** Although two power supplies are normally required, a source resistor may be used to develop the necessary reverse bias, but it may reduce both gain and noise figure. Several bias circuits are shown in **fig.** 3.2 It should be mentioned that all bias circuits must be bypassed to rf; a series gate resistor from 1000 ohms to 10k will protect the gate from high-frequency transients.

For best performance a low-loss input matching circuit is very important; low-loss components should be used in the simplest possible low Q circuit. As can be seen in the Smith chart plot of the NE24406's S-parameters, $|S_{11}|$ and $|S_{12}|$ are very large; the optimum source impedance for noise figure, Γ_{FOPT}, is also very high. Fortunately, the optimum source impedances for gain and noise figure are not greatly different. Therefore, the input matching circuit can consist of a series capacitance or inductance. Since the series reactance must be carefully selected, the capacitor has the advantage of being easily tuned.

Theoretically the preamplifier will work fine with only a series capacitance in the input circuit, but the NE24406 can provide 12 dB gain in the 4-GHz band, so there's a possibility of oscillation outside the 430-MHz band. The parallel resonant circuit, L1-C2, in **fig. 4** suppresses these oscillations. L1 is also used to supply bias to the gate of the GaAs fet. The resonant impedance of L1-C2 is very high, and input matching is actually provided only by C1.

As was shown in **eq. 2,** the noise figure of a transistor amplifier depends greatly on the-available gain of the device being used. Since a gain increment of 4.0 to 4.5 dB ($|S_{22}| \cong 0.8$) can be anticipated at 432 MHz for the NE24406 when the output is matched,

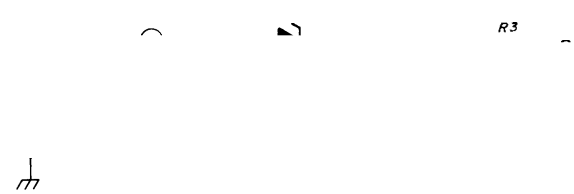


fig. 6. Recommended drain voltage supply circuit for the NE24406 GaAs fet preamplifier.

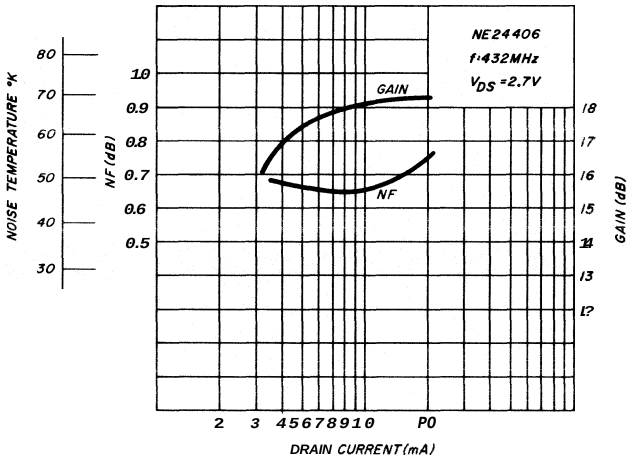


fig. 7. Preamp noise figure and power gain as a function of drain current. Note that lowest noise figure occurs at I_D = 10 mA.

improper output matching will degrade noise figure. Since the output impedance of the GaAs fet is extremely high, I installed a simple parallel resonant circuit at the output (L2, C2, C3), and matched to the 50-ohm output with the capacitive portion of the network. Both the input and output circuits have relatively high *Q*, so stable operatin is obtained over a narrow bandwidth.

construction

Basic construction of the low-noise 432-MHz preamplifier is shown in **fig. 5.** The enclosure is made from no. 18 to no. 22 (0.028-0.048 inch or 0.7-1.2mm thick) brass sheet. Do not omit the center shield – it's absolutely necessary to obtain stable operation.

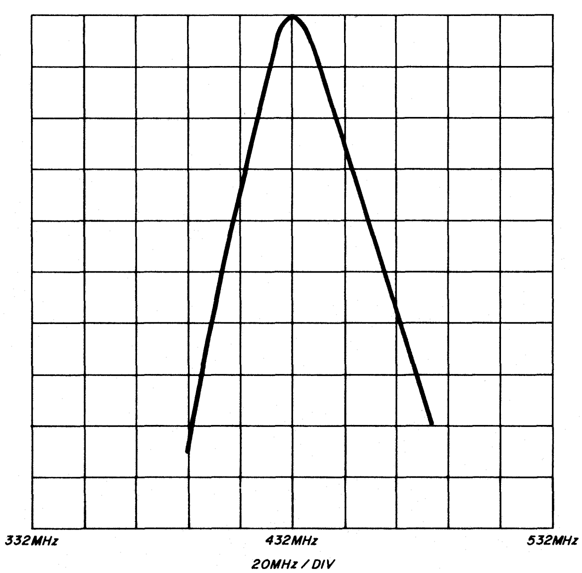


fig. 8. Passband response of the low-noise 432-MHz preamp; gain is 3 dB down at approximately 414 and 456 MHz.

Although no particular care is required when mounting the GaAs fet, the source leads should be separated and soldered directly to the holes in the shield plate. Also, when soldering the drain and gate leads to the input and output coils, take care not to pull on them too strongly.

Any type of coaxial connector may be used at the input and output, but for best noise performance it's important that the connector have good uhf characteristics. The losses of BNC connectors will degrade the noise figure. I used type SMA connectors. Type N or TNC connectors could also be used, bur SMA types are much smaller and therefore more suitable for a miniature low-noise preamplifier such as this one.

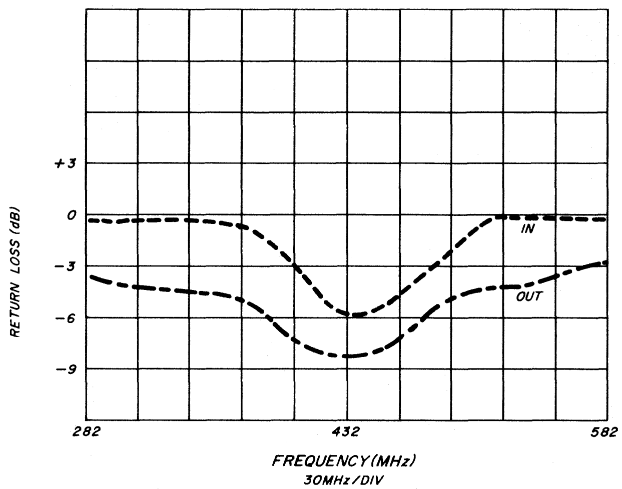


fig. 9. Return loss (vswr) of the GaAs fet preamp

operation and test

After construction is completed, inspect each part to make sure you haven't made any wiring errors. The components in the bias circuit should be checked and double checked. When you are satisfied that the circuit is correctly built, the bias and drain supply voltages should be applied to the GaAs fet in the following manner:

1. Voltage is initially applied to the gate circuit with a 3-volt battery (see **fig. 4**). This reverse biases the gate and prevents current flow in the drain circuit, which may reach the magnitude of I_{DSS}. There is no problem in allowing current flow up to I_{DSS}, but the intention here is to suppress any transient phenomenon due to this current (there are examples where I_{DSS} reaches 100 mA).

2. Next apply the drain voltage, but make sure that the voltage between drain and source, V_{DS}, does nor exceed 3 volts. When the drain voltage approaches 2.7 volts, the drain current should be set to 10 mA by adjusting the 10k pot in the bias circuit.

With the completion of these two steps, amplifier bias is established. A recommended drain bias circuit is shown in fig. ***6.***

Note that when a reverse bias is applied to the gate and drain, and current flows with only a slight application of drain voltage, either the gate circuit is open or a breakdown has occurred in the GaAs fet.

After completing the bias adjustments, adjust the input tuned circuit to resonance with a grid dipper. Apply a ***weak*** signal in the 430-MHz band and tune C1, C3, and C4 for maximum gain. When the preamp is adjusted for maximum gain, the noise figure will deteriorate slightly, but not seriously.

performance

The performance of the GaAs fet preamplifier is shown in the graphs of figs. 7, **8,** and **9.** The plot of noise figure ***vs*** drain current (fig. 7) shows that lowest noise figure occurs at $I_D = 10\ mA$. The bandpass characteristic of the preamp is shown in fig. **8;** when the preamplifier is peaked up to 432 MHz, gain is 3 dB down at approximately 414 and 456 MHz. In most applications no external bandpass filter should be required. Fig. **9** shows the vswr at the input and output of the preamplifier. The third-order IMD products are shown in fig. 10; 1 dB compression occurs at about 0 dBm — the third-order intercept point is at a very respectable +20 dBm (100 mW).

When this preamplifier is adjusted for maximum gain, the measured noise figure is about 0.75 dB. If an automatic noise figure meter is available, the input and output circuits can be adjusted for best noise figure — my measurements indicate that a noise figure improvement of about 0.1 dB is possible.

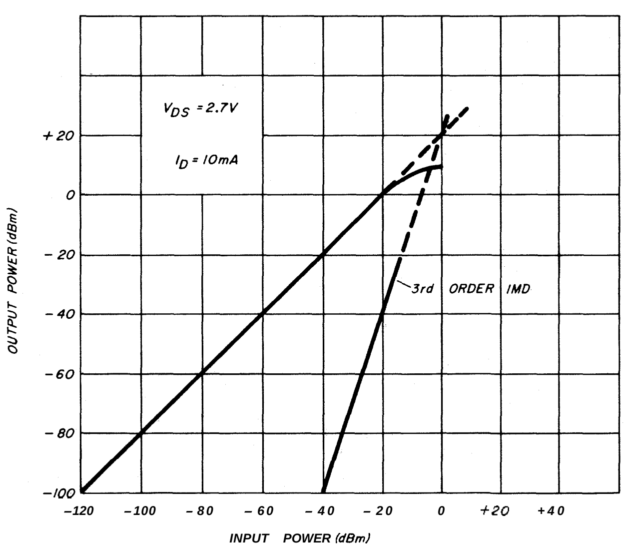


fig. 10. Input-output characteristics of the GaAs fet preamp. Third-order intercept point is at +20 dBm (100 mW); 1 dB compression is at approximately 0 dBm (1 mW).

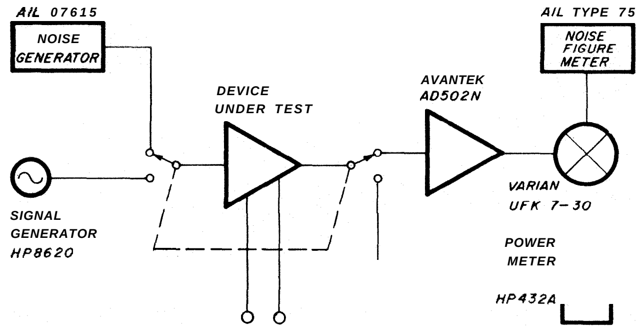


fig. 11. Block diagram of the noise-figure measurement setup used by JH1BRY when evaluating preamplifier performance.

When making noise-figure measurements, I use an AIL noise-figure meter with a ***solid-state*** noise source (fig. 11). Gaseous-discharge noise sources are also available for this frequency range, but they should ***never*** be used with fragile GaAs fet circuits which are susceptible to damage from surge transients.

summary

This GaAs fet preamplifier should bring you right up to the state-of-the-art in noise figure at 432 MHz. The NE24406 GaAs fets are available in the United States from California Eastern Laboratories* or one of their sales representatives.

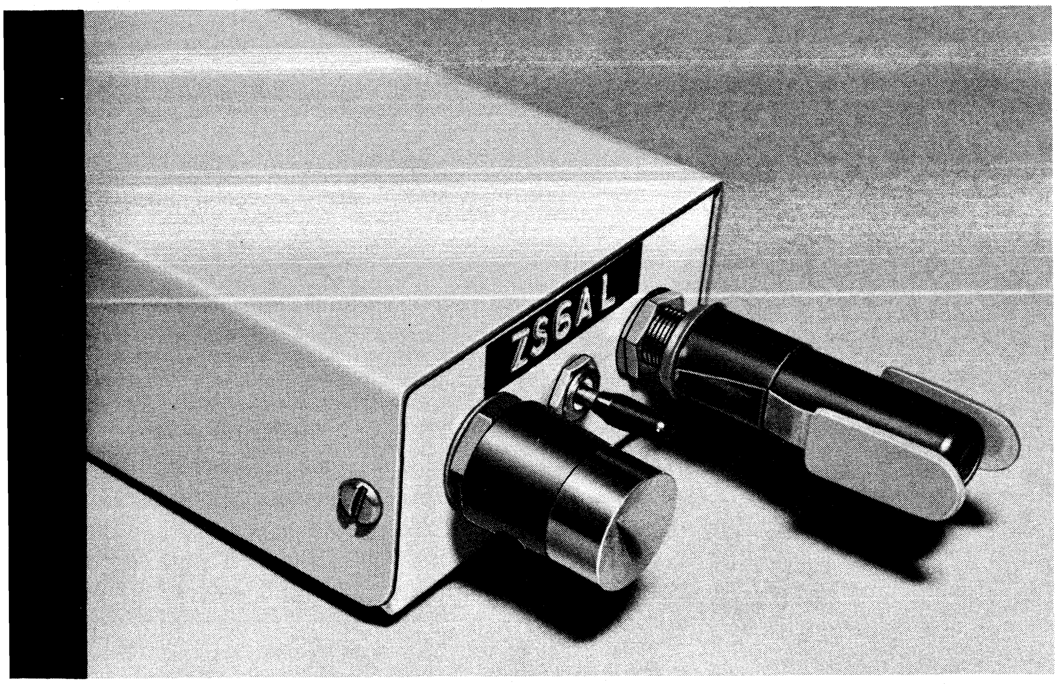
I would like to express my deep appreciation to Aki Munezuka, JA1VDV, for providing me with detail information on 432-MHz EME and kindling the fire of uhf ssb, and to Carl Peterson, K6VJN, who gave me an opportunity to publish this article. I also wish to thank the Nippon Electric Company for the use of test equipment and devices necessary to design this preamplifier. Special thanks go to Haruo Yoneda, JA1ANG, for all his helpful suggestions.

*California Eastern Laboratories, Inc., One Edwards Court. Post Office Box 915, Burlingame, California 94010; telephone (415) 342-7744. The NE24406 is priced at $190.00 in small quantities. After this article was written, NEC announced the basic GaAs fet in a smaller package at a lower price; this device is designated the NE24483 and is priced at $120.00.

references

1. Joseph Reisert, W1JR, "Ultra Low-Noise UHF Preamplifier," ***ham*** *radio,* March, 1975, page 8.
2. Jerry Arden, "The Design, Performance, and Application of the NEC V244 and V388 GaAs FET," California Eastern Labs, Burlingame, California, June, 1976.

ham radio



simple paddle

for electronic keyers

Construction details for the Ambidextrous Paddle for Electronic Keyers or APEK

The need may arise for the serious CW operator to take a battery operated electronic keyer on a mobile trip or to environments where a heavy fixed station dual-lever paddle simply has to stay home. I have improvised several keying devices in the past which provide comfortable keying of the mobile or portable rig under the varied conditions typical of such excursions. These devices were, however, all of the single lever variety that either had to be wedged, clamped, or clipped on to something, and moreover, had an extra lead going to the keyer – a nuisance at the best of times.

The keying paddle described here is essentially a dual-lever device which has proved to be so versatile that most of the problems typical of portable or mobile operation could be surmounted. As a matter of fact since inventing this Ambidextrous Paddle for Electronic Keyers, which I call APEK, the heavy dual-lever paddle at the fixed station has hardly been used at all.

The APEK is basically a three-contact jack plug with a few bits and pieces added, the total cost of which is hardly worth mentioning. Should a particular keyer, however, not be fitted with a jack type paddle socket, such a socket will have to be mounted on the front panel in a convenient position and connected in parallel with the existing socket. Since the

By Vidi la Grange, ZS6AL, 3 Elgar Street, Vanderbijlpark, 1900 Republic of South Africa

APEK plugs directly into the keyer, it can be rotated to have the dot lever at the right or left — a real ambidexter! As a matter of fact, it can be rotated into any convenient angle depending on the orientation of the keyer, which, during mobile or portable work can be just about anything imaginable.

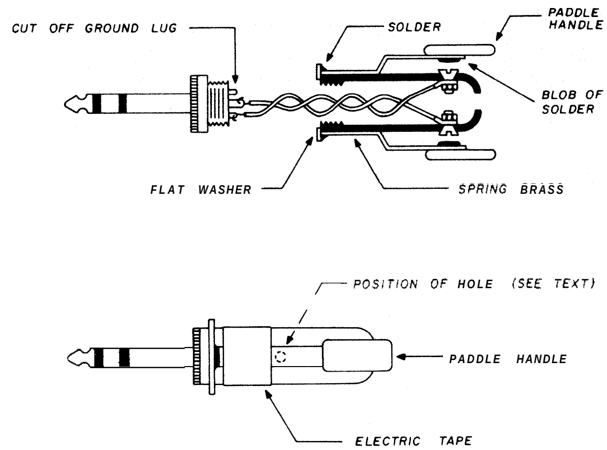


fig. 1. Construction of the Ambidextrous Paddle for Electronic Keyers or APEK. The paddle is built completely from junk-box parts, as described in the text.

The keyer shown in the photograph was built into a box which fits snugly into the car's door box. The APEK sticks out vertically so you can reach it without effort. Another attractive feature is the fact that the APEK is compatible with dual-lever keying devices. Squeeze keying features, if provided, don't have to be sacrificed when using the APEK, a feature impossible to achieve with single-lever keying devices.

construction

Obtain a standard three-connector jack plug and remove the shell. Cut off the ground connection close to the jack body. Drill and countersink two opposing holes in the plug shell about 8 mm (5/6 inch) from the unthreaded end. The holes should be suited to take two 2.3 mm by 4 mm (3/56 x 5/32 inch) countersink machine screws.

Take two small solder lugs and connect about 60 mm (2-1/4 inch) of thin flexible hook-up wire to each. Fit one of these lugs to each screw on the inside of the shell and secure with a nut. Find a suitable flat washer which will fit over the threaded part of the jack body and having an outside diameter slightly larger than that of the shell.

Prepare two strips of spring brass approximately 0.5 mm (1/64 inch) thick, 5 mm (3/16 inch) wide and 40 mm (1-1/2 inches) long. De-burr the edges and bend them into the shape shown in **fig. 1.** Solder these strips to the rim of the flat washer in opposing positions. This step can be simplified by temporarily assembling the plug. Mark off, on the underside of each strip, the positions of the screw heads fitted to the shell.

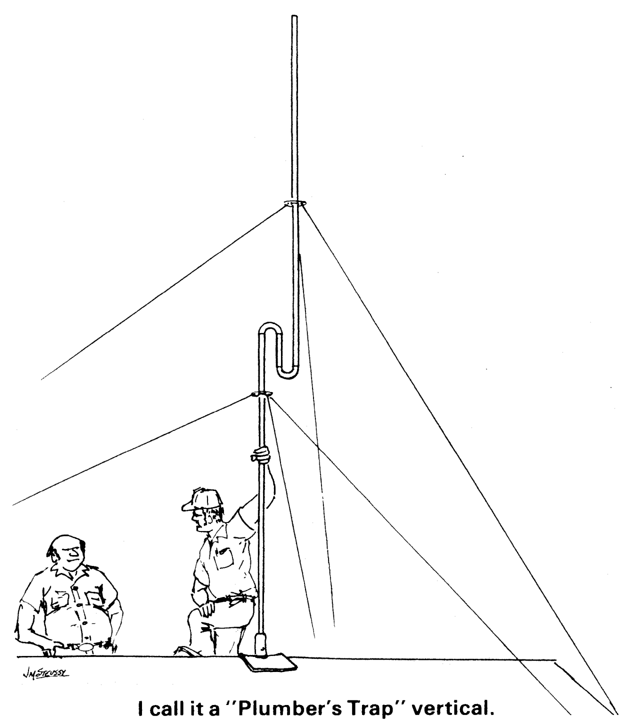
Disassemble the plug, clean and tin the positions marked off on the strips. Form a small blob of solder at these points and file them lightly to give a flat contact surface. Solder the two pieces of hookup wire, already connected to the screws on the shell, to the inner and outer lugs of the jack. Leave these wires long enough to survive the twist they will get when putting the jack together.

Shape two pieces of old PC board according to taste or as shown in **fig. 1** to form the paddle handles. Fix them to the brass strips with a small amount of epoxy. Finally, one or two wraps of electric tape cut to the correct width should be wound onto the jack shell. This will provide the necessary damping of the otherwise too springy brass strips.

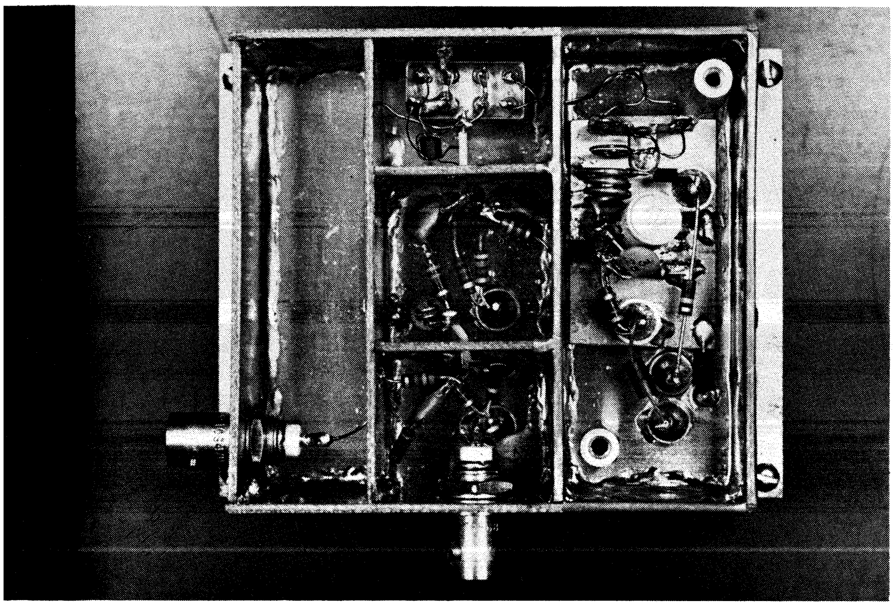
final adjustment

Contact spacing and spring tension can be adjusted by means of long nosed pliers. If you find that the spring tension is too high, a small hole drilled through the spring at the position indicated should do the trick. Then determine by experiment which side of the APEK activates dots or dashes and mark it for your convenience.

ham radio



I call it a "Plumber's Trap" vertical.



spectrum analyzer tracking generator

What is a tracking generator? When used with a spectrum analyzer it generates a CW signal corresponding to the frequency at which the spectrum analyzer is tuned. It's useful for looking at filter response. Filter blowby and undesired responses can be readily observed.

The generator described here was built for use with the spectrum analyzer described in reference 1. A tracking generator identical to this one is now being used with a Hewlett-Packard 85541141. The tracking generator can be used with almost any spectrum analyzer that provides first-local oscillator output and has a first i-f of 200 ± 20 MHz.

Because of the tracking generator's oscillator instability, narrow-bandwidth measurements can't be made, such as the bandpass response of crystal filters. Nevertheless, it's useful for measuring responses at other frequencies (parasitic resonances) often found in crystal filters.

circuit description

The schematic is shown in **fig. 2.** Q1 and Q2 provide some gain and isolation between the 200-MHz oscillator, Q3, and the first i-f amplifier of the spectrum analyzer. R1 provides fine tuning and should be located for easy access by the operator. MX1 mixes the 200-MHz output with the signal from the first local oscillator to provide the 100-kHz to 100-MHz tracking signal. An optional 130-MHz lowpass filter is shown. The lowpass filter attenuates the 400-500 MHz component generated by the mixer in the tracking generator.

The tracking generator is built in a box made from 1116-inch thick (1.5mm) copper-clad board. The same board is used as separators between stages. Paper-thin copper, available from hobby shops, is wrapped over the surface where the cover for the 200-MHz oscillator attaches. The other stages do not have shield covers. A blank compartment is available for the optional 130-MHz lowpass filter. **Fig. 1** shows the layout.

The oscillator is built on a separate piece of copper-clad board, 7/8 x 1-112 inch (22x38mm) and

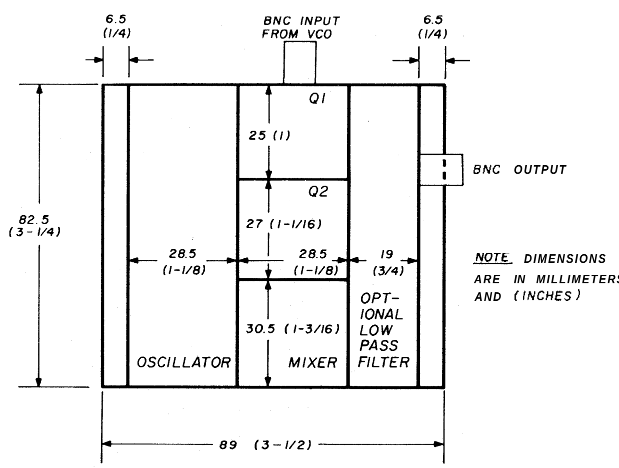


fig. 1. Circuit-board layout. Shielding is important for the 200-MHz oscillator compartment (see text).

By Wayne Ryder, W6URH, 115 Hedge Road, Menlo Park, California 94025

is held in place with double sticky-back tape. Two ferrite beads are strung on the ground wire that connects the oscillator ground to the compartment ground to minimize ground-loop currents, which can cause radiation from the oscillator. The mixer can be a standard mixer or the home-made mixer described in reference 1.

Dimensions shown inside are inside dimensions. The assembly is 1 inch (25.5mm) high. The 1/4 inch (6.5mm) overhang at each end is for mounting.

operation

The spectrum analyzer vfo output is connected to the tracking generator input of the spectrum analyzer.

1. Set spectrum-analyzer bandwidth to 250 or 300 kHz and scan width to 100 MHz.

2. Set R1 to the center of its range.

3. Adjust C1 so that the baseline on the spectrum analyzer shifts up.

4. Move L2 away from L1 until the baseline starts to move down.

5. Install the shield on the 200-MHz oscillator and adjust R1 so that the baseline is as high as possible, consistent with a flat response. As the tracking generator is being used, the 200-MHz oscillator will drift, and some readjustment of R1 will be required.

design considerations

One of the more challenging problems here is preventing the 200-MHz oscillator from radiating into the first i-f. If this happens, the baseline on the spectrum analyzer will shift up without a signal input. The signal from the oscillator will leak through MX1 backward through buffers in the tracking genera-

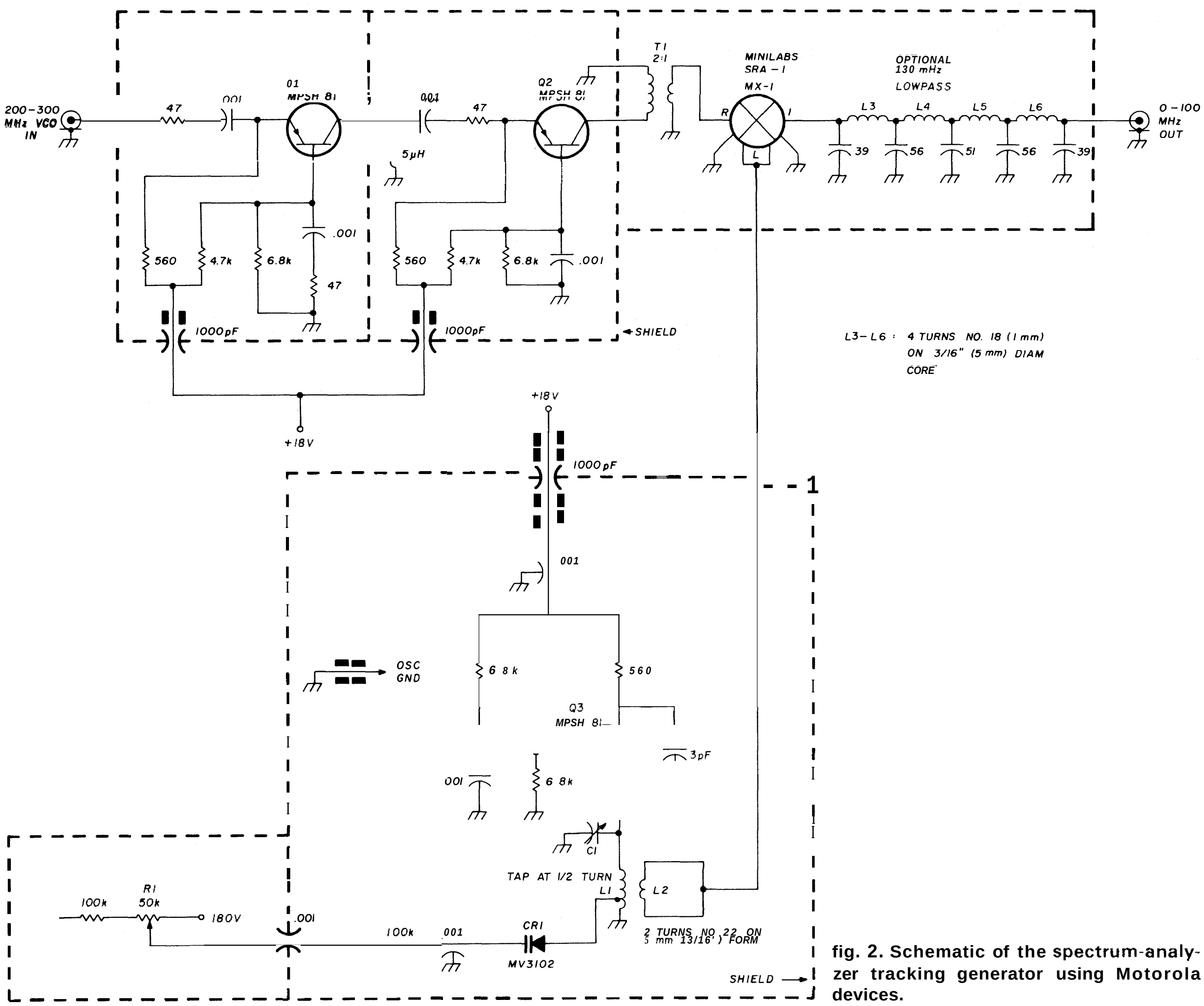


fig. 2. Schematic of the spectrum-analyzer tracking generator using Motorola devices.

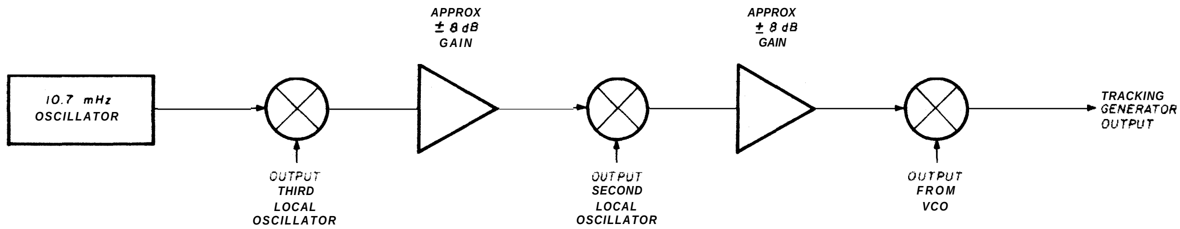


fig. 3. Suggestion for a tracking generator for measuring narrowband signals. Oscillators within the spectrum analyzer are used. A highly stable signal is provided.

tor, backward through any buffer in the vco and will follow the path of the vco to the first mixer in the spectrum analyzer and ieak through the first mixer into the first i-f. This is not surprising, since the output from the oscillator is about 0.5 volt and the spectrum analyzer sensitivity is only a few microvolts. Also, the 200-MHz oscillator can radiate directly into the spectrum analyzer if both are not carefully shielded.

The tracking generator described here was mounted on the outside back of the spectrum analyzer to minimize coupling. If your spectrum analyzer has a phase-locked vco, or if the vco is stabilized in some other way; you might consider a different form of tracking generator for measuring narrow bandwidth. This method uses the local oscillators within the spectrum analyzer and provides a more stable tracking-generator signal. The block diagram (**fig. 3**) assumes that the last i-f is 10.7 MHz.

reference

1. Wayne F. Ryder. W6URH, "High-Performance Spectrum Analyzer," *ham radio*, June. 1977, pages 16-30.

ham radio

zip-cord feedlines

Many years ago I watched as someone set up a rig in the desert. The power plant was set out, the rig was set up, and the antenna was strung out between two convenient cactus plants (cactus attair! a respectable height here). The thing that caught my eye was the feedline. It consisted of a long length of garden variety zip cord. My funny look at it gained a quick assurance that lamp cord was a perfectly good feedline.

Everything seemed to work just fine and the operators had no trouble working out on the band of the day, which was 75-meter phone. In those days it was common to run 100 to 200 watts input with high level a-m.

After a recent move I decided to put up a 75-meter antenna at home and use the most economical feedline. In the process of getting on the air, intermediate forms of antennas were used. That means some wire of about the resonant length was thrown up on the roof, and the near end was run into the rig. It was somewhat of a disappointment to discover that the intermediate antenna worked better than the well elevated final installation. A little checking showed that when the feedline itself was loaded up, it worked better than the antenna it was supposed to be feeding. In short, it appeared that something was not quite right.

A little further investigation revealed some rather interesting facts. RG-8/U, RG-58/U, and lamp cord were tested at 4 and 21 MHz. The rig was tuned up into a terminating type of wattmeter and the feedline under test was inserted between the rig and the meter. Power out with 60 cm (24 inches? of feedline *vs* power out with 9 to 18 meters (30 to 60 feet) of feedline was measured and the results were tabulated.

RG-58/U coax showed a 58 per cent loss at 21 MHz; it showed almost no attenuation at 4 MHz. RG-8/U gave about 12 per cent loss at 21 MHz.

Zip cord looked like it was best suited for use on the other end of a lamp or soldering iron. If you really want to know, it showed about a 60 per cent loss at 4 MHz. There was no need to measure it at 21 MHz!

About 20 meters (65 feet) of coax was used in the tests, and only 9 meters (30 feet) of zip cord (the rest of the zip cord was still attached to the antenna). Obviously, more lamp cord would have shown more loss. These tests were not conducted under laboratory conditions, but the variables were held to a reasonable level so that it was possible to reach a reasonable conclusion.

In any case, the zip cord came down and the RG-58/U went up. A crosstown telephone call got K7OXS on the air, again. Bill indicated that my 3-5 watt rig was back up to its normal signal strength at his location. He was almost as glad as I was that my antenna problems were finally resolved.

Evert Fruitman, W7RXV

constant-current battery charger for portable operation

Tired of pulling out your nicads for charging? Here's a neat method for keeping your batteries fully charged at all times in any location

The use of portable battery-operated equipment requires fully charged batteries to obtain maximum usefulness of the equipment. For handheld transceivers, keeping the batteriesfully charged would normally require frequent overnight charging. If the handheld is operated in a mobile environment it must be removed often from the car to the battery charger. However, it's possible to charge the equipment's batteries while in use, thus always assuring a full charge and maximum lifetime from the batteries between charges.

A charging system has been developed that permits constant-current charging from an automobile electrical system to a 10-cell nicad battery pack, which is commonly found in handheld and similar portable transceivers. Since a 12-volt nicad battery is not easily charged from an automobile 12-volt system, a special circuit was designed to furnish a charging current to the nicads.

nicad charging systems

The most obvious method of charging a nicad battery is shown in **fig. 1A**. A source potential, $V1$, delivers current into a battery pack whose potential is $V2$. Resistor R limits the charging current to a safe and maximum value. The charging current, I, is defined by

$$I = \frac{V1 - V2 - 0.6}{R}, \quad V1 > V2 \qquad (1)$$

This simple circuit works only if $V1$ is greater than $V2$, which may not be true in all instances. For example, if a nicad battery has a discharge potential of 10 volts, a 12-volt source can be used as a charging source. However, as the nicad battery becomes charged, its potential will increase to 13.5 volts, in which case the charging current would actually stop flowing during the charge cycle. In an automobile environment, source potential $V1$ is not constant but may vary from 11 to 14 volts depending on the condition of the automobile battery and engine rpm. Although the worst-case maximum voltage condition can be assumed, and the maximum charge current

By Gene Hinkle, K5PA, I/O Engineering, 9503 Gambels Quail, Austin, Texas 78758

defined and limited, this simple circuit will not charge a nicad battery with a constant current all the time.

If a large value of $V1$ is available, the circuit in **fig. 1B** works well. Since $V1$ is much larger than $V2$, the battery potential, the change in $V2$ versus charge time has little effect on the charging current. A lamp is used to decrease the voltage to the nicad at a specified current. This circuit is representative of many chargers on the market today. Its disadvantage is that the charging current is not exact and $V1$ must be much greater than $V2$, which is not the case in an automobile system.

Fig. 2 shows an excellent method of obtaining a constant-current from a common three-terminal voltage regulator. The three-terminal voltage regulator would normally have its common terminal grounded and would deliver a constant voltage between the output and common terminal. However, if the common line is not grounded, but left floating, and a fixed resistance is connected from the output to the common terminal, the regulator will try to furnish a fixed voltage across the resistance. The current through the resistor is given by Ohm's law:

$$I_{OUT} = V/R$$

where I_{OUT} is current through resistor R; V is the voltage regulated by the device.

Once the circuit is completed between the three-terminal regulator's common and power return (ground connection), current will flow through this connection, even if a resistance or voltage exists in the path to ground. The only requirement is that the input voltage must be equal to or greater than the full charge battery potential plus 5 volts (for a 5-volt regulator) plus 2 volts (overhead voltage). Thus, for a 13-volt full-charge nicad potential and a 5-volt-type voltage regulator, the input voltage to the current regulator circuit must be greater than 20 volts.

In the circuit shown, a transformer and rectifier were used to provide 30 volts dc. A 50 mA, 6-volt lamp was used in series with the regulator to indicate when the 50 mA charge current was flowing; it also dropped the power-supply voltage to 24 volts dc.

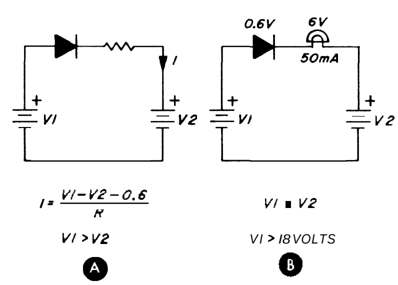


fig. 1. Simple charging systems for nicads. In A a resistor is used to limit the current source; the current is not constant and is nonlinear. Sketch B shows a lamp current-limited source. Voltage $V1$ must be much larger than $V2$ for good regulation. This is a very popular charging method.

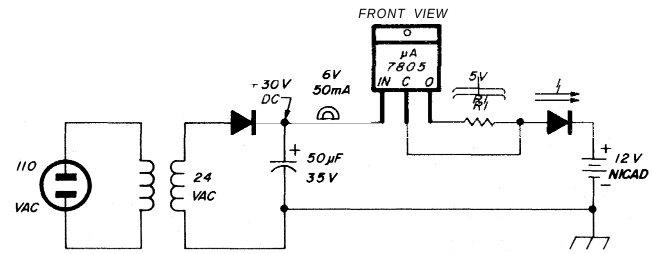


fig. 2. An excellent constant-current regulator using a popular 3-terminal voltage-regulator IC. Current source is determined by R1. For a 7805 IC, V is 5 volts. Thus for R 50 ohms, I – 50 mA constant current.

The use of three-terminal regulators is an excellent technique for defining a constant current level. The current level is easily adjusted by changing resistor $R1$ in the output circuit. The current is essentially constant regardless of the discharge state of the nicad battery.

voltage-doubler circuit

To use a constant-current source as described it's necessary to provide an input source voltage of at least 20 volts. In an automobile situation, a 20-volt source is not available, so the circuit of **fig.** 3 was designed. This circuit uses a NE-555 universal timer IC and two power transistors in a voltage doubling circuit. The output voltage is roughly twice the input voltage, minus any diode voltage drops. Thus, a 10-volt source will be converted to 19 volts and a 13-volt source to 25 volts. This doubled voltage is then used to drive a source current into a three-terminal current regulator.

The operation of the voltage doubler is as follows, referring to **fig.** 3.

The NE-555 is used in the common astable multivibrator configuration. The oscillation frequency is determined by $R1$, $R2$, and $C1$ and is equal to

$$F = \frac{1.44}{(R1 + 2R2)\,C1} \qquad \textbf{(2)}$$

To have a near 50 per cent duty cycle the ratio of $R1$ to $R2$ should be around 1 to 4 as shown.

The 555 astable oscillator drives a pair of complementary transistors. High-current power transistors were used to switch the large charge and discharge currents. Transistor Q1 charges capacitor C2 to the input voltage during the first part of the astable cycle. During the second half of the cycle, transistor Q2 puts the fully charged capacitor C2 in series with the supply voltage. Typical values for C2 are

$$C2 = \frac{I_{OUT}}{F} \qquad \textbf{(3)}$$

where F was defined earlier and I_{OUT} is the constant output current required. A diode and capacitor (C3) filter the pulsating dc to a value nearly equal to two times the supply voltage. The value of C3 should equal C2.

Transistors Q1 and Q2 are plastic-cased power transistors. Power transistors are used because of the high peak currents during the charge and discharge cycle of C2. Although MJE 2955 and MJE 3055 types are shown, most common power transistors will work. The transistors are not heat sinked because both are operated in the saturated mode, thus power dissipation is held to a minimum. The diodes are 1-amp silicon types and may also be substituted with similar devices.

The current regulator operates as the one previously described. It may be attached to a heat sink, but is not mandatory. If the regulator's input terminal voltage is 25 volts, and the output terminal is at 17 volts, only 400 mW of heat must be dissipated in the worst case. The oscillation frequency of the voltage-doubling circuit can be increased, but transistor switching becomes less efficient with the increased switching speed. A too-low frequency requires larger values of capacitance for C2 and C3. A good compromise is to make the frequency between 1 and 10 kHz.

In the circuit shown the switching frequency was set to 1.4 kHz. Since ten 500 mA-hr nicads were being charged, the chaiginy current (I_{OUT}) was set to 50 mA. For an input voltage range of 10-15 volts, the charging current was a constant 50 mA.

practical approach

To make these charging schemes work in a mobile charging system, charging current may be delivered

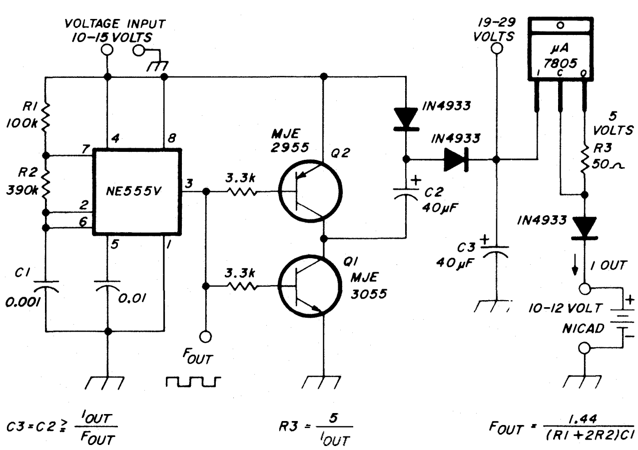


fig. 3. Voltage-doubler circuit suitable for developing enough voltage from an auto battery to run the current regulator. Power transistors may be substituted for those shown.

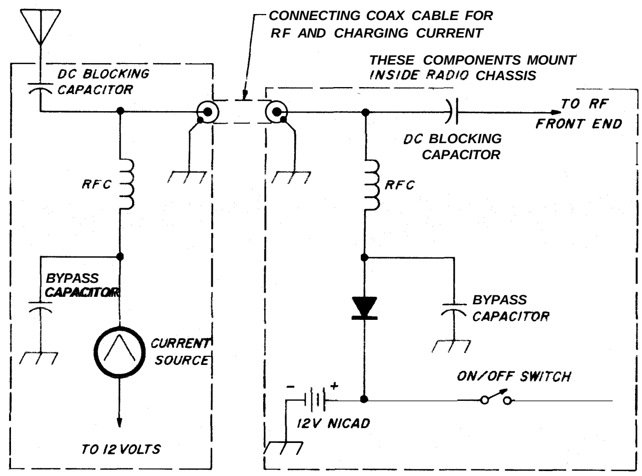


fig. 4. Use of the constant-current charging source requires only the connection of the antenna lead to the radio in a mobile or base station. The current source may be located inside your power amplifier or within an ac-operated charger. For 2 meters, the rf choke, RFC, is ten turns no. 26 AWG (0.3mm) copper wire on a 1-megohm 112-watt composition resistor.

to the portable unit through the antenna coaxial cable. This approach makes it easy to connect the handheld or other equipment to the charging source as well as to the antenna.

It's quite common to use small power amplifiers in a mobile system. This creates a convenient location for the current source circuits. **Fig. 4** shows a typical mobile setup. The current source is located inside the amplifier housing. A small rf choke isolates the incoming and outgoing rf energy from the current-source circuits. A small capacitor in series with the antenna confines any charging potential to the inside of the automobile. The coax connecting the handheld to the amplifier or antenna carries the charging current to the batteries through an isolating rf choke. A dc blocking capacitor should be used in series with the rf circuitry. Most transceivers employ such a capacitor; check your schematic. The charging current source should be connected across the battery at all times.

advantages

By charging the nicads through the coax line, the batteries can be maintained at full charge at all times.

By using the a-c operated charger and the automobile charger system, the nicads can be charged at any location. The current source could be switched to a lower trickle charge if desired. A 3-position switch could be used for this purpose and for turning off the regulator,

ham radio

how to modify linear amplifiers for full break-in operation

With the introduction of the Ten-Tec *Triton IV*, which provides full break-in operation on CW, I needed a companion linear amplifier which also had this very desirable operating feature. At the present time, however, there is only one commercially produced amplifier on the market which meets this requirement — and it's priced at almost $3000. The purpose of this article is to outline the theory and give some typical circuits which can be used to modify any power amplifier to provide full break-in capability. The circuitry is for use with grounded-grid triode linear amplifiers; class B or AB is assumed.

While class C might seem to be a better choice for a CW power amplifier, there are several reasons to maintain linear operation. First, and most important, since class C operation is not linear, the CW waveform supplied by the exciter will be distorted and the resulting output wave may produce serious key clicks and other unwanted spurious signals.

For the modification described here the grid of the amplifier tube must be at chassis ground, with a positive voltage applied as cutoff bias to the cathode. The classical biasing scheme, with the cathode at ground, fig. 1A, is rearranged to place the grid at both rf and dc ground, fig. 1B. The operating or cutoff bias is in the form of a positive voltage which is applied to the cathode; the high-voltage B+ plate supply is above ground by this potential.

When the amplifier is used for break-in operation, it is essential that the tube be biased to cutoff — with no plate current flowing — so the tube doesn't generate noise which would mask weak received signals.

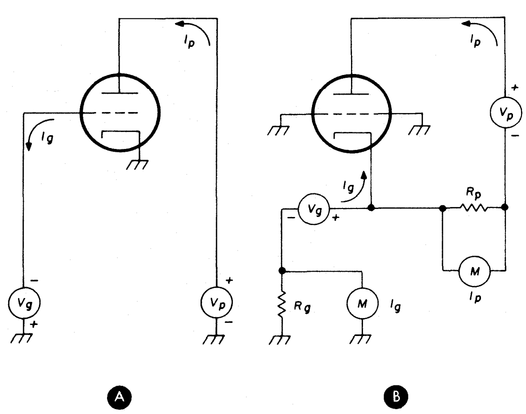


fig. 1. In the classical bias circuit, A, the cathode is grounded and a negative bias is applied to the grid; in the grounded-grid circuit a positive bias is applied to the cathode, B. In these circuits V_g is the bias supply, V_p is the plate supply, R_g is the grid meter shunt, and R_p is the plate meter shunt.

cutoff bias

Of the several types of tubes commonly used for linear amplifier service, there are two basic types: those with hot cathodes and those with indirectly heated cathodes (see fig. 2). For the purposes of this discussion the main difference is that the bias on a hot cathode is applied through the center tap of the filament transformer (fig. 2A); bias for indirectly heated tubes is fed directly to the cathode through an rf choke (fig. 2B).

Figs. 3 and 4 show two common ways to bias rf power amplifiers to cutoff. The circuit in fig. 3 (from the ARRL *Handbook)*, uses a 10k resistor in series with the cathode circuit to develop self bias. It is the least acceptable method, however, because it requires a certain amount of plate current flow to provide the bias. Therefore, the tube is not completely cut off.

Another cutoff bias circuit is shown in fig. 4. It uses a +150 volt bias supply which cuts off all plate current and powers the T/R relay, K1. Actually, for B+ supplies up to about 3500 volts, 75 volts of bias is enough to cut off a pair of 3-500Zs — 75-volt relays, however, are hard to find.

Fig. 5 shows a circuit which uses an rf-sensing transistor switch to remove the cutoff bias from the amplifier. Under standby conditions Q2 is an open circuit and 50 volts cutoff bias is applied to the cathode. When the drive signal is applied, Q2 turns on

By Dick Frey, K4XU, Chief Engineer, Ten-Tec, Inc., Sevierville, Tennessee 37862

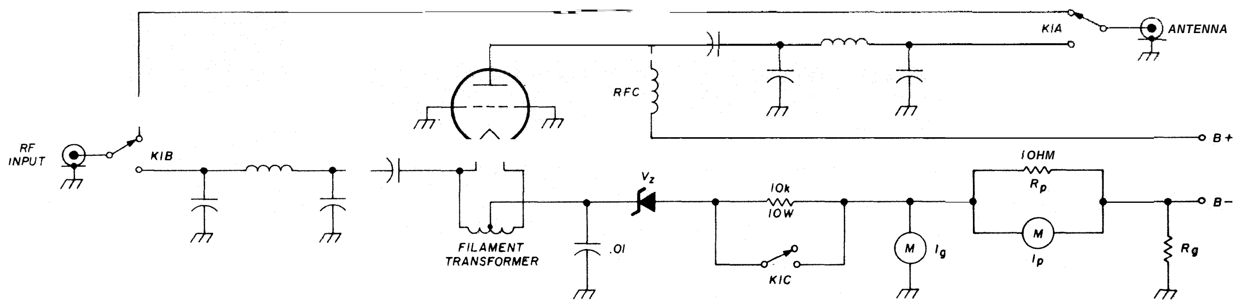


fig. 3. Typical linear amplifier bias and control using self bias developed across the 10k resistor in the B- line. Relay K1 is shown in the operating position with the amplifier on.

and the operating bias becomes the sum of the zener voltage, V_z, Q2's collector-emitter saturation voltage, $V_{CE(SAT)}$, and the voltage drop across the grid meter shunt, R_g. $V_{CE(SAT)}$ for the specified 2N5321 resistor is 0.8 volt at 500 mA.

When selecting Q2, two criteria must be met: a low $V_{CE(SAT)}$ because this adds to the operating bias, and a maximum collector-voltage rating that is greater than the applied voltage bias voltage. The Darlington configuration is used because transistors which meet the two voltage requirements seldom have the necessary current gain for this circuit. The power requirement for Q2 is quite low because the voltage impressed across it is low when it carries the grid current. A power dissipation rating of 10 watts is more than adequate. The resistance of the grid meter shunt resistor, R_g, should be as low as possible; usually 0.5 ohm is sufficient with a grid current meter which has a 1 mA movement.

The basic bias circuit of **fig. 5,** modified to accommodate the directly heated cathodes of two 3-500Zs, has been used successfully in a Heath SB-220. Note that the zener diode no longer has to pass the full plate current so it is permissible to use a less expensive 20-watt zener diode. An added feature of this circuit is that if any of the devices which supply operating bias fail open, cutoff bias will be applied to the tube — this prevents any damage that might otherwise occur.

The 1N4004 diode shown in the negative high-voltage line is a clamp which prevents the B− from going negative with respect to the chassis in case of bias supply failure.

When the linear amplifier is in the standby condition virtually no current is drawn from the bias supply. When Q2 turns on, resistor R1 limits the current flow. This current will show as grid current, so it's a good idea to keep R1 as large as possible. When testing this circuit I found it was useful to provide a means for turning on Q2 without applying rf drive power; S1 is a test button which does this.

In normal operation, the linear appears to be running class C because plate current is drawn only when an rf drive signal is present. The amplifier is not operating class C, of course, but since quiescent plate current is no longer being drawn under no-signal conditions, the average power dissipated by the tube due to quiescent bias (200 to 600 watts in a typical class AB amplifier) is greatly reduced; this increases tube life, improves reliability, and results in a cooler ham shack.

receiver switching

When the power amplifier can be operated without using relays to switch the bias, one step remains: receiver antenna switching. Electronic T/R switches have been available for years, but have never been widely accepted because, when the transmitter is tuned to resonance, the tank circuit acts as a "suck out" filter which reduces received signal strength. The electronic T/R switch was not the culprit — its placement in the system caused the problem. The solution is to place the electronic T/R switch at a high impedance point in the system. If you want to eliminate noisy relays which are slow and prone to failure and go to full break-in operation, the antenna

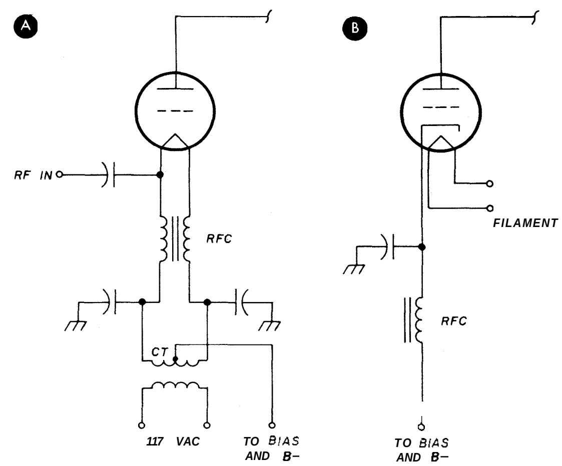


fig. 2. Grounded-grid biasing arrangements for hot cathode tubes, A, and indirectly-heated cathodes, B. Typical hot-cathode tubes are the 3-500Z, 3-1000Z, and 3CX1000; the *8877* and 8874 series of power amplifier tubes have indirectly-heated cathodes.

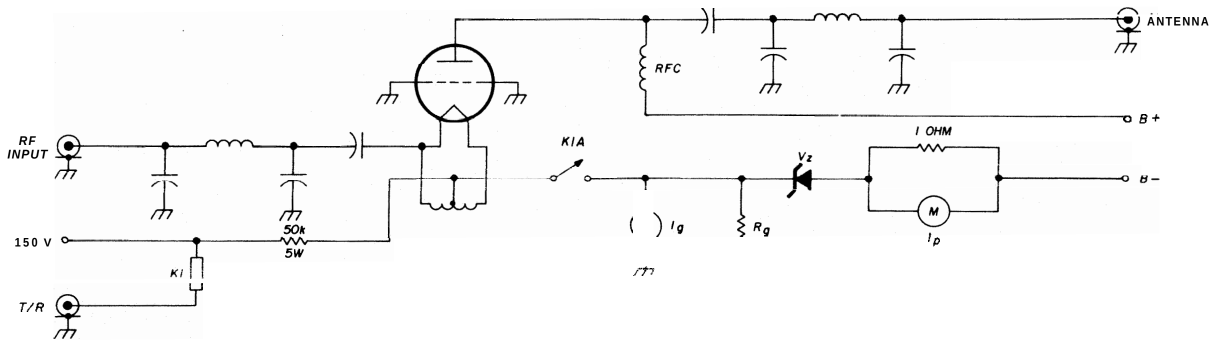


fig. 4. Linear bias and control using an external bias supply; this method provides better operation than the arrange rnent of *fig.* 3. Relay K1 is shown in the operating position with the amplifier biased to cutoff.

must be connected to the transmitter at all times. Thus, to find a high impedance point to connect an electronic T/R switch, the plate tank of the amplifier is a good choice.

Coupling to the plate tank is not without its problems, however. First of all, it's dangerous because of the high voltage that is present. Secondly, since there is a large amount of rf voltage present, the coupling capacitor C1 **(fig. 6)**, must be small both to reduce its effect on the amplifier tuning and to assure that the T/R switch is not overdriven. This presents a problem because the optimum value of C1 for receiving purposes, about 5 pF, is too much capacitance when used at the kilowatt level on transmit. Therefore, a compromise is made on the side of safety and reliability.

The capacitor specified is made from RG-8IU coaxial cable with the center conductor overlapping the braid for approximately 12 mm (1/2 inch). Two 25 mm (1 inch) square tabs, spaced 1 cm (3/8 inch), will also work if the mechanical layout of the amplifier will permit placement of this arrangement. Be sure that this capacitor is placed directly across the plate tuning capacitor *after* the plate blocking capacitor. This compromise results in a slight loss of gain on receive, but most transceivers have more than enough front-end gain to make up for the loss.

construction

Since most modern kilowatt linears use some sort of input tuning, the placement of S1 in **fig.** 6 is not too difficult. The switch wafer can usually be added to an existing switch shaft, or the shaft can be extended to provide room for the wafer. When pur-

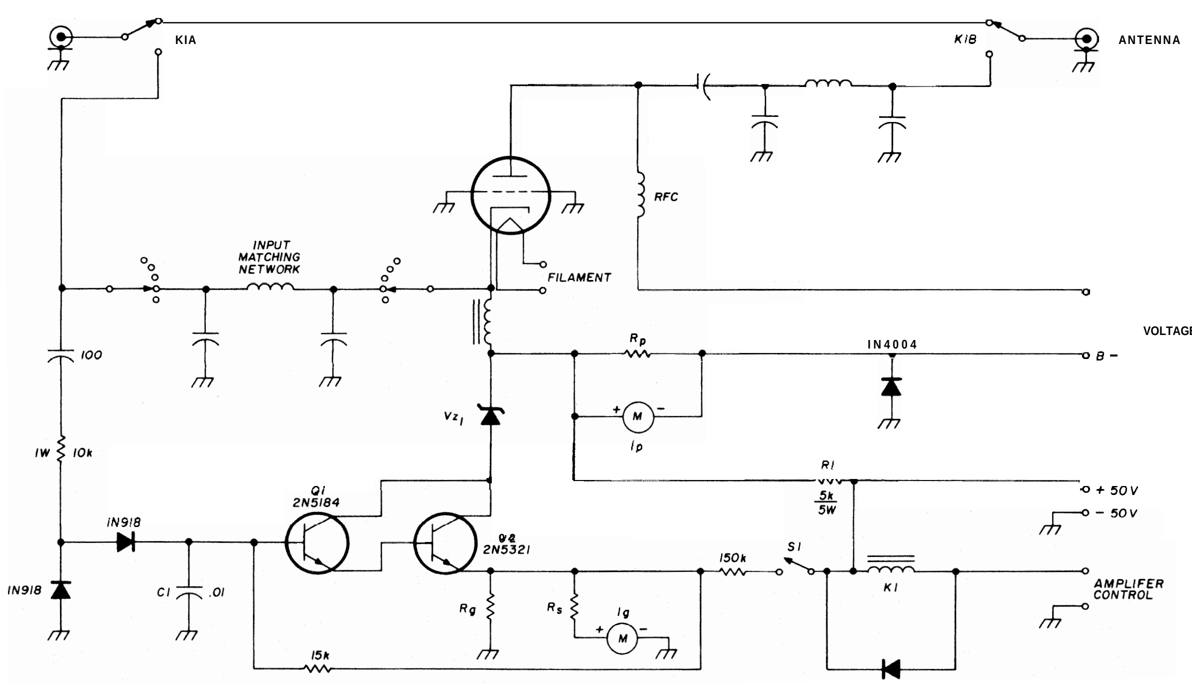


fig. *5.* Cutoff bias circuit for a linear amplifier which uses an rf sensing transistor switch. The zener voltage, V„ provides the proper operating bias and was part of the original amplifier circuit. Cutoff bias circuit operation is described in the text.

chasing the wafer switch for S1, be sure to note whether the amplifier uses 30, 45, or 60 degree indexing between switch positions.

The location of Q1 and its associated circuitry is not too critical except that it must be reasonably close to both C1 and S1 to minimize losses. It should be shielded if it is located inside the plate tank enclosure and all dc leads must be shielded. Since the dc current drawn is very small, the +18 volts can be obtained with a voltage doubler from the amplifier filament supply, or from a zener regulated drop from the positive bias supply. The switched coils form broadly resonant tuned circuits with the existing circuit capacitance, so some adjustment of the given values may be necessary.

Triton IV modifications

There are also modifications to the Triton IV which will improve performance when using a linear amplifier. First is a simple change to eliminate the delay of the control relay while using ssb. As built at the factory, the time constant capacitor, C3 in **fig.** 7 and page 3-34 of the manual, is tied directly to ground on the circuit board. By lifting the ground end of C3 and taking it to the unused center pin on the board, a wire can be run to the mode switch, S1E (CW1 and CW2 positions), which will activate the delay only when using CW. S1E is part of the rear wafer, closest to the chassis.

The second modification involves removal of the control relay, K1. This should be done only if the amplifier uses a positive voltage to key its control and

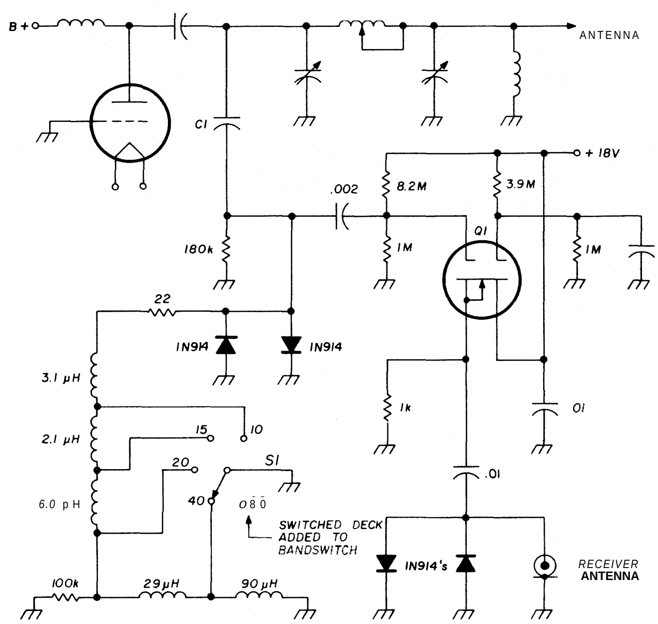


fig. 6. Electronic T/R switch for the receive antenna. The station antenna is connected to the linear at all times. Capacitor C1 is less than 1 pF and consists of 12 mm (1/2 inch) of RG-8/U coaxial cable (see text).

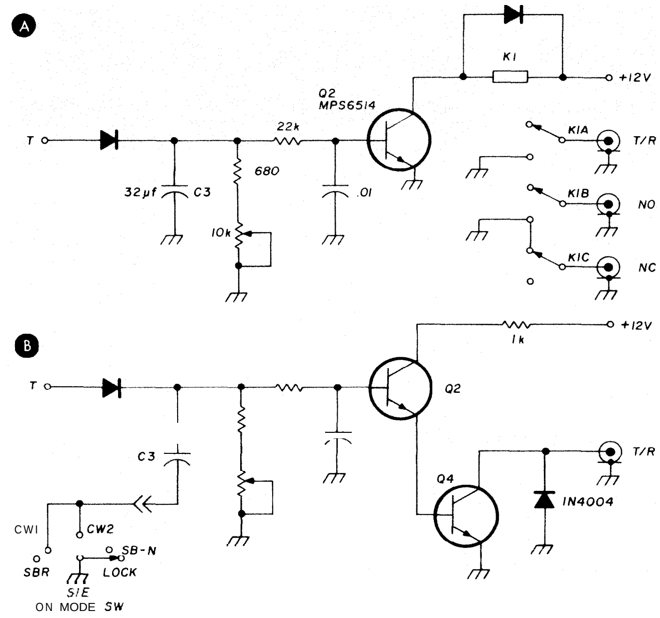


fig. 7. Modification of the Triton IV to remove the control relay and to eliminate unwanted delay times during CW break-in operation.

changeover relay. The reason for doing this is to eliminate the turn-on delay which is caused by using one relay to activate another. Relay K1 has no function in the operation of the Triton by itself. When the key is closed, it takes about 25 milliseconds for this relay to close, and perhaps another 23 ms for the amplifier relay to pick up, before the output of the Triton is amplified. This causes key clicks on the first dot. The transistor used for Q4 depends on the voltage used on the amplifier control relay and the relay current. With K1 removed, there is plenty of room for the added components on the board.

Some Triton users have complained of an ac hum in the receiver when using a linear amplifier — especially in the narrow CW-1 position. This is caused by ac ripple (or raw ac) on the relay control line. When attached to the Triton T/R relay's normally-open jack, unshielded wires going to K1 pass underneath the CW filter which picks up the hum. The solution is simple: reroute the wiring between K1 and T/R normally-open jacks on top of the chassis. This change has been incorporated in late production Tritons.

conclusion

What has been presented here are notes and basic technical information needed to modify an existing rf power amplifier for full break-in operation with a Triton IV; the circuits can also be adapted to other transceivers. Ten-Tec in no way assumes responsibility for the use or misuse of this information, nor for any damage to other manufacturer's equipment resulting from implementation of these circuits.

ham radio

how to design matching networks

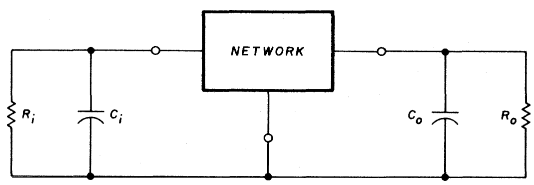
Six basic impedance matching networks and how to design them for your own applications

Common L, T, and pi networks work well for impedance matching but they lack the selectivity required for amplifiers or frequency-multiplier chains. Adding a component to the network and rearranging allows it to be selective on both sides of band center. Presented here are six simple matching circuits and their design equations that allow adjustment of design Q for selectivity and different source or load impedances.

Fig. 1 shows the basic network with its source and load interfaces. The assumption is that R_i is less than R_o. Simply reverse the input, output, and the network if R_o is less than R_i. Both source and load are assumed to have shunt capacitances; this is usually true and will include stray capacitance as well.

The value of Q is the design Q of the network and determines selectivity. It is *not* component Q which should be at least five times design Q. Selectivity around band center is treated the same as a tuned circuit with a certain Q. Each of the six networks has different attenuation far from center; examples of this are shown later.

The constants listed in **fig. 1** reduce the size (and complexity) of the network equations, and will be used with all six networks. It must be emphasized that all capacitive reactances must calculate *negative,* inductive reactances *positive.* Any exception to this rule with a network indicates that particular network cannot be used.



CONSTANTS

$X_i = -1/(2 \pi F C_i)$ $\qquad$ $X_o = -1/(2 \pi F C_o)$

F = CENTER FREQUENCY

$$N = \frac{R_i X_i}{R_i^2 + X_i^2} \qquad R_x = X_i N$$

$$Q_i = R_i + X_i Q \qquad Q_o = R_o + X_o Q$$

fig. 1. The basic matching network. Constants shown here will be used in calculating the values for the six different networks discussed in this article.

different sources and loads

Fig. 1 shows a parallel capacitive reactance across the input and output terminals. Sometimes the end impedances are given (or measured) in series form. This may be converted to parallel form either by impedance-to-admittance plus inversion of resulting conductance and susceptance. Any of the calculators with rectangular/polar conversion can handle this easily. The following conversion formula can be

By **Leonard H. Anderson,** 10048 Lanark Street, Sun Valley, California 91352

used if those functions are not available on your calculator:

$$R_p = R_s + (X_s^2/R_s)$$
$$X_p = X_s + (R_s^2/X_s)$$

where R_s = *Series reszstance*
$X,$ = *Serzes reactance*
R_p = *Parallel resistance*
X_p = *Parallel reactance*

Note that the sign of the reactance is preserved in conversion.

An inductive reactance is a special problem. Compensation of this is done by capacitive shunting so that the total reactance at the band center becomes capacitive. Some of the networks will have shunt inductors at the ends. This condition allows the physical inductor to be the parallel difference with stray capacitance (always present) forming C_i or C_o. If the end reactance changes rapidly around the band center, it is better to add a physical capacitor and use calculated inductance directly. In any case, end reactances must be capacitive.

handling data from the spec sheet

Transistor data is invariably given in admittance or S-parameters. Admittance is already in parallel form so the conductance and susceptance values are taken directly and inverted to yield end resistance and X_i or X_o (watch the sign of susceptance, it is positive when capacitive).

S-parameters are a bit different and are found on Smith chart representations. These are normalized to 50 ohms or 0.02 mho and can be taken directly,

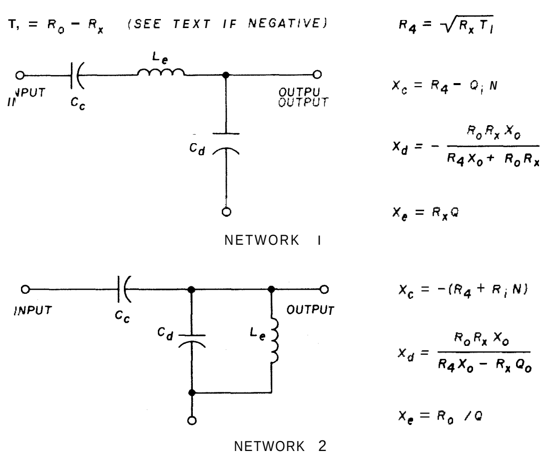


fig. 2. Networks 1 and 2.

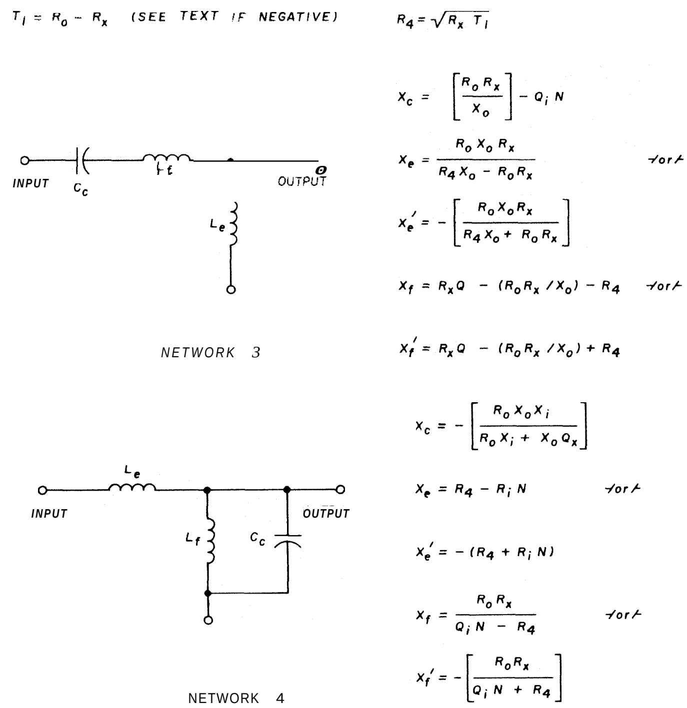


fig. 3. Networks 3 and 4

then un-normalized. S_{11} is the input impedance while S_{22} is the output impedance; both are complex numbers. Knowledge of Smith chart interpretation is required.[1]

Data in admittance or S-parameter form is invariably frequency sensitive. They also represent the average data of a production run so individual devices can vary. Under such conditions it is well to keep the design Q relatively low. There may be an advantage to swamping the end with a parallel resistance (small carbon composition resistor) to reduce sensitivity of values. This has two cautions: power loss and about 2 pF extra capacitance with each resistor.

examples

Numerical examples all assume matching a 50-ohm line to the input of a Motorola MC1590 amplifier IC (single-ended) at 7.15 MHz with a design Q of 20. Stray capacitance of 3 pF is assumed at each end; the MC1590 is insensitive with a parallel, single-ended impedance of 5k and 5 pF. The line is assumed perfect. **Fig. 1** constants are then:

$R_i = 50$	$R_o = 5000$
$C_i = 3\ pF$	$C_o = 8\ pF$
$X_i = -7419.81$	$X_o = -2782.43$
$N = -6.73841\ 10^{-3}$	$R_x = 49.9977$
$Q_i = -148346$	$Q_o = -50648.6$

An HP-25 calculator was used for these and following calculations.

networks 1 and 2

These are shown in **fig.** 2. Variable T_1 is a test value and is also used for Networks 3 and 4. If T_1 results in a negative, none of the first four networks can be used with the given constants. Design Q and/or end resistances can be changed to make it positive. The negative situation only comes about with low R_o:R_i ratios.

For the first four, $T_1 = 4950$ so $R_4 = 497\ 483$. Values of Network 1 are then:

$X_c = -502.135$	$C_c = 44.33\ pF$
$X_d = -613.263$	$C_d = 36.30\ pF$
$X_e = 999.555$	$L_e = 22.26\ \mu H$

Note that all reactance signs are correct. Values for Network 2 are:

$X_c = -497.146$	$C_c = 44.77\ pF$
$X_d = -605.847$	$C_d = 36.74\ pF$
$X_e = 250.000$	$L_e = 5.565\ \mu H$

Reactance signs are correct here, too.

networks 3 and 4

Test variable T_1 and R_4 apply here. Both inductors of both networks have two possible solutions. The reactance sign rule still applies so *both* X_e and X_f must be positive or *both* XI, *or* X_f' must be positive. Do not mix X_e' and X_f or vice-versa; use either all non-prime or all prime values. Network 3 will have:

$X_c = -1089.46$	$C_c = 20.43\ pF$
$X_e = 425.637$	$L_e = 9.474\ \mu H$
$X_f = 592.318$	$L_f = 13.18\ \mu H$

Non-prime values obey the rules as do Network 4 values:

$X_c = -274.782$	$C_c = 81.01\ pF$
$X_e = 497.819$	$L_e = 11.08\ \mu H$
$X_f = 497.853$	$L_f = 11.08\ \mu H$

The inductors in Network 4 came out very close to the same value. This happened with this particular example, but is not true of other conditions.

network 5

This one is shown in **fig. 4.** It must be noted that the two inductors have zero coupling and must be physically separate. The schematic may appear to be a conventional tapped-inductor circuit but such would need a different set of equations plus measurement of mutual coupling. Networks 3 and 4 must also have separate inductors.

Test variable T_2 cannot be negative. If T_2 is negative design Q or end resistances, and possibly input capacitance, may have to be changed. Example conditions fit alright so we get:

$R_5 = 22.0617\ 10^9$	
$T_2 = 827.815\ 10^9$	$R_4 = 64.3356\ 106$
$X_c = -274.782$	$C_c = 81.01\ pF$
$X_e = 227\ 822$	$L_e = 5.071\ \mu H$
$X_f = 28\ 7208$	$L_f = 0\ 639\ \mu H$

Non-prime inductor values were correct in this example. Note that C_c is the same as for Network 4.

network 6

Fig. 5 shows this to be the tuned circuit often found in receiver front ends. The test variable is T_3 and there is only one set of solutions. Our example condition results in:

$R_5 = 2.57302\ 10^9$	
$T_3 = 89.9416\ 10^9$	$R_4 = 21.2063\ 10^6$
$X_c = -250.921$	$C_c = 88.71\ pF$
$X_d = -32.9475$	$C_d = 675.60\ pF$
$X_e = 250.000$	$L_e = 5.565\ \mu H$

The inductor is the same value as in Network 2.

wideband response

Fig. 6 shows the frequency response over a two-decade range for the examples given. Voltage response has been calculated with constant end resistances. As such, it will be the same in either direction. Frequency is normalized to 7.15 MHz.

The joker in the deck is Network 3. The extra peak on the high side of resonance will vary in frequency and relative amplitude depending on the design Q and end impedances. A saving grace is that Network 3 has the best low-frequency attenuation. All networks will vary in other applications, primarily with different design Q; the general shape of the response curve, however, will still be the same.

Choice of a network depends on the application. A frequency-multiplier chain should consider Networks 2, 5, or 6 because of their better low-frequency attenuation. The last stage could use Networks 1, 4, or 5 in the output to reduce unwanted harmonics. An amplifier chain such as an i-f strip could alternate Networks 5 and 6 between stages for best skirt attenuation.*

'There are better ways; this would only be for miniature construction or multiple stages with degeneration of gain.

table 1. Design Q determination with fixed-value capacitors

Network 1, Fixed C_c:	$Q_1 = \dfrac{R_4 - X_c - R_i N}{X_i N}$
Network 2, Fixed C_d:	$Q_2 = (R_4/R_x) - \left[\dfrac{R_o(X_o + X_d)}{X_o X_d}\right]$
Network 3, Fixed C_c	$Q_3 = \dfrac{R_o R_x - X_o(X_c + R_i N)}{X_i X_o N}$
Networks 4 and 5, Fixed C_c	$Q_{45} = -\dfrac{R_o R_i(X_o + X_c) + X_c X_o R_i}{X_c X_o X_i}$

impedances presented to the source

The impedance presented to the source can be easily calculated.[2,3] Networks 1 through 4 have variations far from band center that might cause stability problems. Network 1 is inductive from the band center to about 20.75 MHz from the example. Parallel resistance climbs abruptly to about a megohm while parallel reactance changes swiftly to capacitive reactance. Network 2 has a similar, less abrupt change at about 9.76 MHz. Network 3 is also similar with the changeover coinciding with the voltage response peak at about 24.3 MHz.

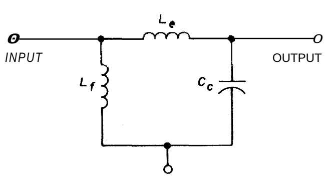
Network 4 becomes inductive below 4.9 MHz and capacitive above 28 MHz, resistance peaking to 4 megohms at that frequency. Networks 5 and 6 were much less susceptible to changes, and showed only slight variations at passband edges. All components were assumed losstess so a practical circuit would exhibit much less variation due to finite component Q.

The swamping-resistor method with transistor collectors works well from 6 meters and down with f_t of 150 MHz or greater. This increases a normally low-collector conductance but does have some power loss. A swamping resistor can also be used at the load since load changes reflect to the source. A

$$R_5 = X_i^2 + Q_i^2$$

$$T_2 = R_i R_5 - R_o X_i^2 \qquad \text{(SEE TEXT IF NEGATIVEJ}$$

$$R_4 = \sqrt{R_o T_2}$$



$$X_c = -\left[\frac{R_o X_o X_i}{R_o X_i + X_o Q_i}\right]$$

$$X_e = (X_i/R_5)(R_o Q_i + R_4) \qquad \text{+or-}$$

$$X'_e = (X_i/R_5)(R_o Q_i - R_4)$$

$$X_f = -\frac{R_o R_i X_i}{R_4 + R_o R_i} \qquad \text{+or-}$$

$$X'_f = \left[\frac{R_o R_i X_i}{R_4 - R_o R_i}\right]$$

fig. 4. Network 5.

rule of thumb is to add 2 pF for every resistor, using only carbon cornposition types.

Mismatch loss will add out-of-band attenuation when used with receiver front ends. An exception is where the input impedance of the network matches the antenna impedance out of band. This is rare, but it happens. A similar condition occurs with different equipment connected with coax cable: the line length may cause a match out of band.

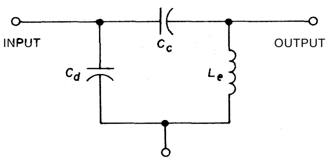
non-standard component values

A cure for non-standard component values is to change design Q slightly to accommodate at least one standard, fixed value. The other two compon-

$$R_5 = X_o^2 + Q_o^2$$

$$T_3 = R_i R_5 - R_o X_o^2 \qquad \text{(SEE TEXT IF NEGATIVE)}$$

$$R_4 = \sqrt{R_o T_3}$$



$$X_c = -(X_o/R_5)(R_4 + R_o Q_o)$$

$$X_d = \frac{R_o R_i X_o X_i}{R_4 X_i - R_o R_i X_o}$$

$$X_e = R_o/0$$

fig. 5. Network 6.

ents can be trimmable. **Table 1** is a tabulation of design Q for Networks 1 through 5 based on one specified reactance.

Our example for Network 2 gives a C_d value of 36.74 pF. A 39 pF mica is a standard value with a reactance of -570.755 ohms at 7.15 MHz. This X_d value is used and gives a design Qof 20.5074. The X, of Network 2 is unaffected by Q but X_f changes to 243.814 ohms or 5.427 μH.

Network 6 is a bit difficult to solve for Q but it can be done by programming an HP-25 or similar calculator to solve either X_c or X_d with manual Q input. An approximation that works in some cases is:

$$Q_6 \cong (R_o + R_i)/X_c$$

The Network 6 example had C_c at 88.71 pF. A fixed 82 pF capacitor will have -271.456 ohms so the approximate Q is 18.6034. Recalculating with Q gives:

$X_c = -272.564$	$C_c = 81.67\ pF$	(close!)
$X_d = -37.0998$	$C_d = 599.99\ pF$	
$X_f = 268.768$	$L_e = 5.983\ \mu H$	

C_c is within 0.4 per cent of desired value so it should work well.

Usual tolerances of fixed components are only 5 per cent. This will have little effect on matching when the other two components are trimmable. Lower design Q will show less sensitivity to tolerances.

applications and variations

Some situations cannot be met with given end resistances. This can sometimes be cured by using

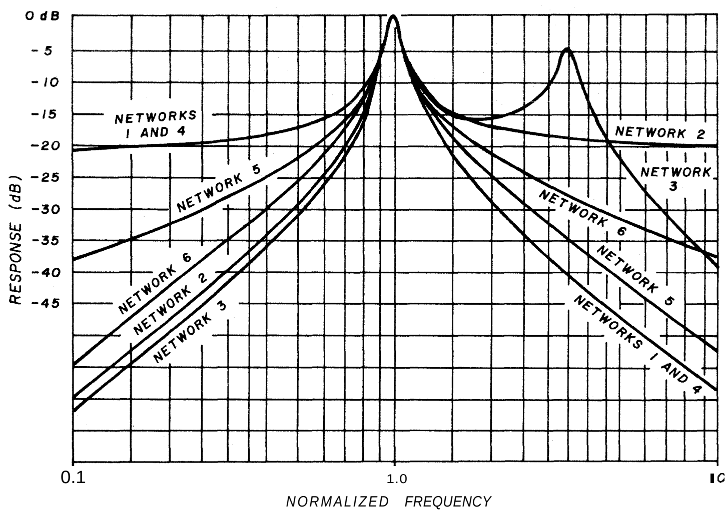


fig. 6. Frequency response of the six networks with the band center normalized to 7.15 MHz.

a broadband toroidai transformer.9 Impedance changes of up to 16:1 are possible. Remember that parallel reactance is also changed by the same ratio in this application.

Good bypassing in active-device applications cannot be over-emphasized. A poor bypass and decoupling not only cause trouble between stages but also become networks. Resonance of bypass capacitors with lead length is common at higher frequencies; one cure is to double up several lower-capacitance bypasses. Short lead lengths and a large ground plane should always be used.

An interesting application is the replacement of the preselector tuning used in 1960-era receivers such as the Heath SB-300 series. These all have one band per bandswitch position so the front end can be stagger-tuned over the desired portion of the band. Fixed capacitors replace the variable and Networks 5 or 6 can be used at the antenna input. The Heath design has enough room to add a couple of bandswitch wafers to allow selection of other matching networks for other antennas such as a longwire. The variable capacitor can be retained, insulated, and used as part of a Wien-bridge audio notch filter.

Antenna networks used in receivers should have a dc path to ground such as in Network 5. This avoids static build-up during electrical storms and potential arcing.

other networks

End impedance frequency sensitivity may require simpler networks. An excellent treatment of L-networks is found in the first reference along with proper use of the Smith chart. Tabulated values are available[4,5] and theoretical aspects can be found in texts.[6]

Application of transistor amplifiers and matching is well covered in reference 7. Access to a computer that speaks FORTRAN can use the program of reference 8 to calculate other networks and also determine amplifier stability.

references

1. Phillip H. Smith. Electronic Applications of the *Smith* Chart, McGraw-Hill, New York, 1969.
2. Leonard H. Anderson. "Admittance and Impedance, and Circuit Analysis," ham radio, August, 1977, page 76.
3. Leonard H. Anderson, "Calculator-Aided Circuit Analysis," ham radio. October, 1977, page 38.
4. ARRL Electronics Data Book, Chapters 3, 4, and 5, ARRL, Newington, Connecticut, 1976.
5. Frank Davis, "Matching Network Designs With Computer Solutions," Application Note AN-267, Motorola Semiconductor Products, Phoenix, Arizona.
6. Philip Cutler, Electronic Circuit Analysis; Volume 1, "Passive Networks."
7. Ralph S. Carson, High Frequency Amplifiers, John Wiley & Sons, New York, 1975.
8. Gunnar Richwell, "Design and Match RF Amplifiers," Electronic Design *10*, May 10, 1969, page 106.
9. John J. Nagle, "Use Wideband Autotransformers," Electronic *Design* 3, February 2, 1976 (reprinted in ham radio, November, 1976, page 10.

ham radio

"I DID clean up my desk . . . two years ago, just before I picked up that call from Russia."

overtone crystal oscillators without inductors

A discussion of overtone crystal oscillator circuits which don't require inductors

Until recently, all of the circuits for overtone crystal oscillators I've seen have included tuned LC circuits. It seemed necessary to have an LC resonator, tuned to the desired overtone frequency, to be sure the oscillator would operate at the proper overtone frequency and prevent operation at the crystal's fundamental frequency or some undesired overtone. It would be nice if the LC tuned circuit could be eliminated, of course, because it would simplify bandswitching of the crystal oscillator.

International Crystal has introduced a crystal oscillator circuit called the OF-I. Although the OF-I circuit has no inductor, and thus no LC tuned circuit, it can be used with crystals operating in the third-overtone mode. I found this quite interesting and did some relevant experimenting to satisfy my curiosity. My efforts are documented here for others who may share this interest.

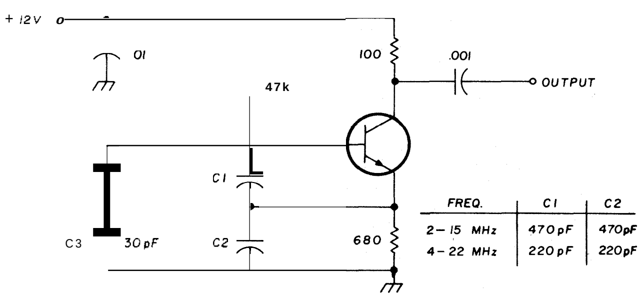


fig. 1. International Crystal OF-1 LO oscillator circuit for fundamental-mode crystals.

the circuit

International Crystal supplies two different kits of the OF-1 type. The OF-1 LO uses crystals operating in the fundamental mode from 2 to 22 MHz; fig. 1 is a schematic of the circuit. The OF-1 HI uses crystals

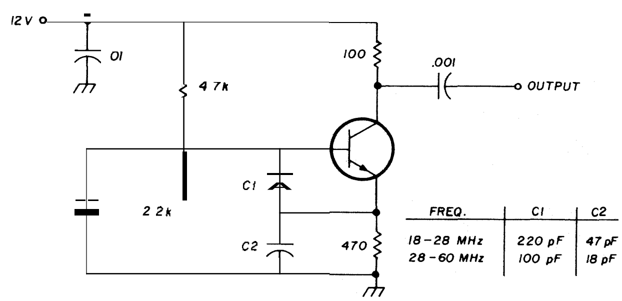


fig. 2. International Crystal OF-1 HI oscillator circuit for third-overtone crystals; note that no inductors are required.

operating in the third-overtone mode from 18 to 60 MHz; fig. 2 is a schematic of the OF-I HI. Notice that in the latter circuit capacitor C3 has been omitted; in the overtone mode, the crystal operates near series resonance.

I breadboarded and tested both of these circuits using a 2N4996 transistor. The crystal I used in all of my tests was a 28.3 MHz third-overtone type originally purchased for use in International Crystal's older OX oscillator circuit (which has an inductor). Using this crystal, the circuit of fig. 1 had an output frequency at the crystal's fundamental, or about 9.43 MHz. The circuit in fig. 2 produced the third-overtone frequency of 28.3 MHz when the smaller values shown for C1 and C2 were used; using the larger values for C1 and C2 given in fig. 2 produced oscillation at the fundamental frequency of 9.43 MHz.

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248

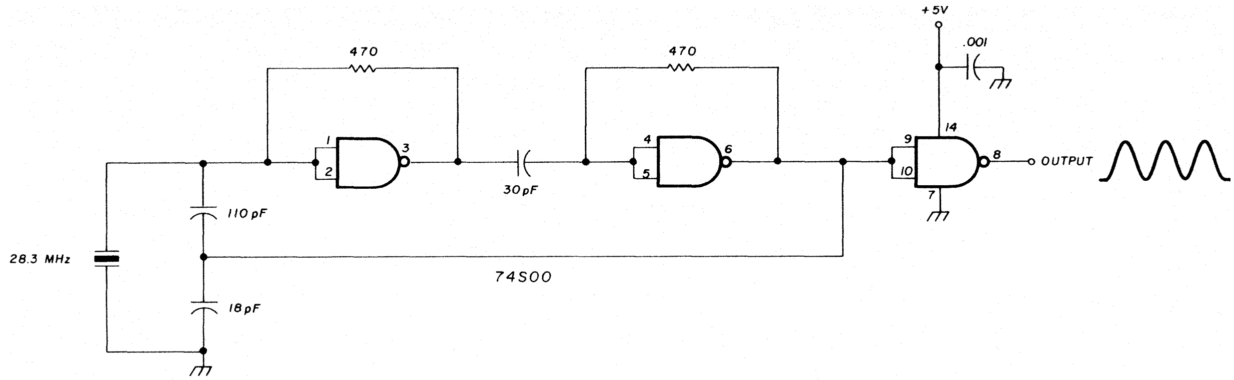


fig. 3. Third-overtone crystal oscillator which uses a 74S00 Schottky TTL gate; no inductors are required

TTL version

To obtain increased output amplitude I developed the circuit shown in **fig.** 3. The 74S00 quad NAND gate acts as both oscillator and output buffer. One of the gates in the IC is unused. A 74S00 is required because of the relatively high frequency; I don't believe the 7400, 74LS00, or 74H00 will work as well, if at all, at 28.3 MHz.

All wiring should be as short as possible, and the circuit should be shielded. The peak-to-peak output amplitude swings from about 0.4 to 3.5 volts, which are acceptable TTL levels. If the capacitor which couples the first two gates together (30 pF in **fig. 3**) is too large (1000 pF), the circuit's output frequency will drop down to the crystal's fundamental. Some experimentation with the value of this capacitor may be required for different crystals.

conclusion

The simplicity of these overtone crystal oscillators could make them desirable in many applications. I have not had the opportunity to investigate them to the degree I would like; therefore, I would welcome comments from interested readers on this type of overtone oscillator circuit.

ham radio

simple method for making printed-circuit boards

I have developed a method of making printed-circuit boards that, while not professional in appearance, work well and beat the mess that results from using wire-wrap sockets and soldering them in place.

I use paint (enamel) in a K&E *Leroy* pen, inserted in a hole of a test probe. When using paint, do not use the insert for the pen, which is used to prevent the ink from running through. The paint is not thinned but just as it comes from the can; the result is a slow writing-paint pen. I have found that a no. 2 or no. 3 *Leroy* pen is just about right. Cleaning is easy with pipe cleaners and paint thinner. I use an inking pen with paint for the long straight lines, and a small brush for the large areas.

To make a PC board, place the copper board under the layout, secure it so it won't slip, then use a sharp instrument such as a scribe and punch a prick hole through each place where there is a dot, or where there will be a hole. This gives the spacing you'll need to draw (free hand) the circuit on the board. After you have completed the hole punching (not through the board), remove the layout. Note that this does not destroy your layout. Now you can fill your pen and draw the circuit on the copper board.

I recommend you practice with simple circuits before you jump into a complex layout. After the paint has dried naturally or in an oven heated to 150 to 160 degrees, check to see that your layout is as you want it. If two lines are touching, the scribe will allow you to make a fine line between them.

I have tried two types of etching solutions: ferric chloride and an etch solution from Vector Electronics. The paint stood up well in both. First I tried two types of paint, Sears acrylic enamel, and Ace Hardware quick-drying black enamel. There was no apparent difference.

Drilling the board can be done before painting the circuit or after etching. The little prick marks show you exactly where the hole belongs. Removing the paint after etching is done with trichloroethane or other solvent that will not contaminate the board.

Robert H. Kernen, W4MTD

repeater interference:

some corrective actions

Suggestions for repeater associations to minimize the ever-increasing interference problem

Our local repeater, WR4ADC, operating on the 146.34/146.94 pair, has been experiencing interference on both the input and output frequencies. This article covers the technical corrective actions which have been considered for this interference. Some of these are now being used or are being installed for future use.

input channel anti-interference measures — lockout

Since repeaters operate on a fixed-frequency plan, one method of avoiding unnecessary call-up is to use a repeater lockout signal on transmissions not intended for repeaters. Technically, as in **fig. 1,** this would be simple. An NE-555 subaudible oscillator producing a small deviation would be required at the non-using transmitters. The repeater input receiver would need a tone detector and a relay contact in series with the carrier-operated relay (COR) to block operation when the tone is received.

While technically simple, this method of input interference prevention requires cooperation by the interfering stations. Lacking this, it is completely ineffective. It does seem to be a useful approach and could be adopted as a national standard. If done, the standard should include designation of the lockout tone frequency and of the deviation it produces.

While the lockout tone is not currently a useful technique, there are several technical anti-interference approaches available based on the characteristics of interfering signals. One of the most common of these is the frequency window, **fig. 2,** which locks out transmission if the incoming signal is appreciably off frequency. While this is usually done to avoid excessive distortion, it is a powerful antiinterference method. In its simplest form, it may be a Schmidt trigger on the dc voltage at the discriminator, set to the value corresponding to the selected value of allowable frequency error, with the trigger interrupting the COP.. Some filtering, on the order of a tenth-second time constant, is needed.

signal characteristic measures

Other anti-interference measures based on the characteristics of interfering signals include:

1. a-m rejection
2. Wideband fm rejection
3. Non-voice modulation rejection

The purpose of a-m rejection, **fig.** 3, is to prevent operation of the repeater COR by an a-m carrier. In simple form this requires a pickoff from the receiver ahead of the limiters, feeding to an agc-controlled stage and an a-m detector. The lockout circuit would require presence of carrier in the fm section and in the a-m section, plus presence of a-m exceeding some percentage, say around 25 per cent. Some protection against noise pulses or unmodulated pauses would probably be required, at least several seconds time constant of filtering.

Rejection of wideband fm, **fig. 4,** would require a wideband receiver, and a form of detection of the out-of-band energy component. A simple form would be to compare audio levels of the narrow and wideband discriminator outputs: if they approach equality, the incoming signal would be wideband. Another wideband modulation detector would beat the limiter output against a local oscillator, followed by a highpass filter. The filter cutoff frequency would be set to the desired peak deviation.

A lockout system based on the absence of voice modulation, **fig. 5,** would be a powerful anti-interference technique. In principle, this is relatively simple — just measure the ratio of peak-to-average power in the audio. The ratio varies from 3 dB for pure sine waves, to 13 dB for noise, and to about 16 dB for voice. Noise-operated squelch makes the distinction between the last two easy.

However, in current fm practice, the approach is not this simple, since most fm transmitters use a

By **R. P. Haviland, W4MB,** 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

limiter to keep the deviation high, and this reduces the peak-to-average ratio. Because of this, a peak-to-average detector would have to be set to a lower ratio. About 6 dB, or 2:1 in voltage, with a time constant of about one-tenth second should be good and would provide appreciable protection against non-voice signals.

wanted signal anti-interference measures

All of the above anti-interference measures are based on some characteristic of the unwanted signal. The other family of input frequency measures is based on characteristics of wanted signal. The two standard forms of this, fig. ***6,*** are the subaudible tone, or *private line,* required continuously for access; and whistle-on, a single tone or a Touch-Tone signal giving access. Access may be for a definite time period or until the COR is dropped for a time period. Various other forms have been used, including carrier-formed Morse and two-band interrogation.

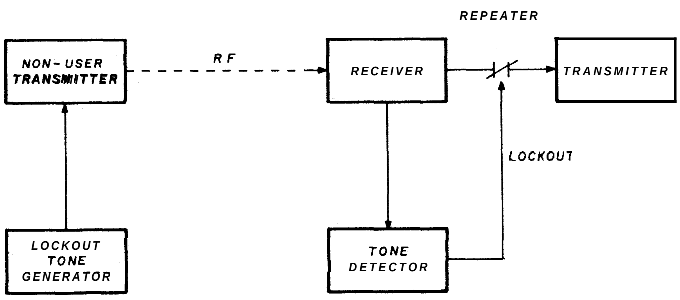


fig. 1. Elements of the repeater lockout system.

Most of these methods have the disadvantage of requiring additions to the using transmitter. This is not a severe problem for closed repeaters, where the number of users is small, but represents a serious drawback for open repeaters, especially those on the common frequencies or close to holiday areas or heavily traveled routes.

One method of partially overcoming this objection is to operate the repeater with carrier access only during no interference periods, and to switch to one of the other access methods when interference is experienced. This switch could be coupled with voice announcement on the identifier, giving instructions as to access method. A Touch-Tone access is probably best since such pads are common and since the signal should never be used in normal communication, except for signaling.

It would seem that this method would give good protection against incidental interference. Willful interference is another matter, since operation in accordance with the announcement instructions would turn the repeater on. However, malicious intent would seem to be proven if interference continues.

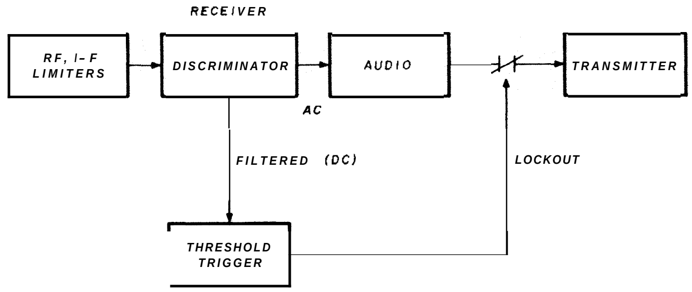


fig. 2. Elements of the frequency window, which locks out transmission if the incoming signal is appreciably off frequency.

output channel anti-interference measures – channel guards

Repeaters also cause interference, and some technique of guarding against the interference they cause may be desirable. Probably the best method of doing this is to guard the output channel, **fig. 7,** holding repeater operation in abeyance if the channel signal exceeds some predetermined value. To account for emergency operation, this guard could be combined with a timer identification announcement, giving instructions on the procedure to override the lockout.

The setting of the guard receiver would need to be based on a value of "signal to be protected," and a margin, or "protection ratio." These are common concepts in other radio services, but they have not been used in amateur operations. As a result, there are no accepted values for these quantities.

The closest other service is the Mobile Service, which includes land mobile. Since this was the source of many 2-meter repeater concepts and equipment, it should provide good guidance.

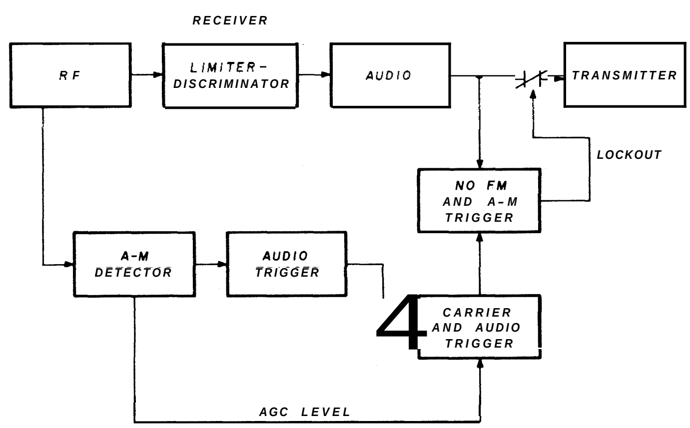


fig. 3. Principle of a-m lockout, which prevents operation of the repeater COR by an a-m carrier.

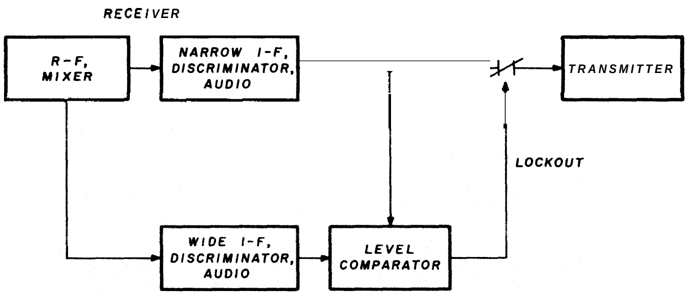


fig. 4. Principle of wideband fm lockout, which would require a wideband receiver and a form of detecting the out-of-band energy component.

The International Consultative Committee (CCIR) data which relates to this matter is in their Report 358. This report states that, not considering the effects of natural or man-made noise, the median signal levels to be protected for an average grade of mobile service are:

band (MHz)	signal (dB above 1 μV/m)
30-50	8
50-100	14
100-200	20
400-470	28

The report also states that the protection ratio, based on degradation from 20 dB S/N to 14 dB S/I, should be 8 dB for fm with fm interference, or 7 dB for fm with a-m interference. The report lists other ratios, but it has not been necessary to consider these, since the 2-meter National plan is based on narrow-band fm.

The 1971 Special Joint Meeting of the CCIR gives the following noise values:

typical urban noise, 150 MHz, +35 dB (KTB)
typical rural noise, 150 MHz, +3 dB (KTB)
maximum cosmic noise, 150 MHz, +8 dB (KTB)

These figures are based on omnidirectional antennas.

Many repeaters are located in urban areas and will see full urban noise levels; others are remotely located and will see less noise. Assume that the cosmic value, 8 dB, represents typical noise. Then the value of signal to be protected becomes 20+8 or 28 dB above 1 μV/m — about 25 μV/m. (This signal would give 20 dB output signal/noise in the average mobile installation and in a typical location.) The allowable value of interference that would give protection for this signal is 28-8, or 20 dB above 1 μV/m — just 10 μV/m. (Presence of the interference would reduce the 20 dB S/N to 14 dB S/I.)

It is easy to see the effect of this protection level if the values are translated to power and distance relationships. For average terrain, the effective radiated transmitter power to just reach the 10 μV/m signal, as a function of distance from the repeater, is approximately:

distance km	distance (miles)	transmitter antenna height: 100 ft (30m) or more transmitter power (W)	30 ft (9m) transmitter power (W)	10 ft (3m) or less transmitter power (W)
10	6.2	0.02	0.3	3.0
15	9.3	0.5	3.0	31.0
20	12.4	1.0	6.0	60.0
30	18.6	8.0	50.0	500.0
50	31.0	100.0	500.0	5000.0
100	62.1	3000.0	8000.0	—

The ERP is the product of the transmitter output and the antenna gain multiplied by 2.5 (to include the effect of ground reflection).

Stated another way, the 10-microvolt signal would be produced by a typical mobile at about 8 miles (13km) distance, by a typical base station at about 18 miles (29km), or by a DX station about 90 miles (144km) away and beaming toward the repeater. The signal would be the same as that of another typical repeater if it were some 50 miles (80km) away.

These values indicate that output channel guarding to the 10 μV level would not eliminate *all* interference. However, it would prevent interference to simplex operation within reasonable distances, to the point that any resulting interference would hardly be considered harmful. Accordingly it would appear that repeater associations wishing to minimize interference problems should consider this approach.

Construction of an automatic protection system should not be difficult. The output-channel monitor-receiver antenna could be mounted on the repeater tower at a height of 30 feet (9m). (This level is used because the height gain is zero.) Assuming use of a

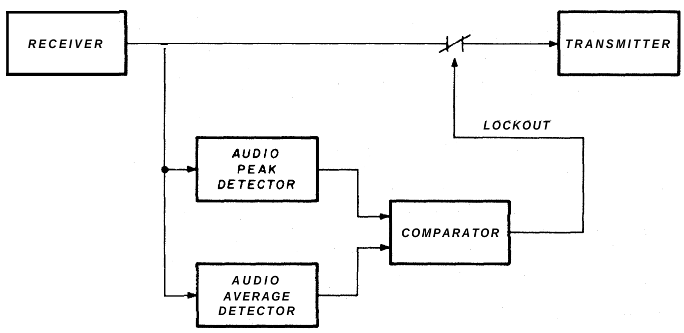


fig. 5. The non-voice lockout principle. Comparator measures peak-to-average power ratio in the audio, which varies from 3 dB for pure sine waves to about 16 dB for voice.

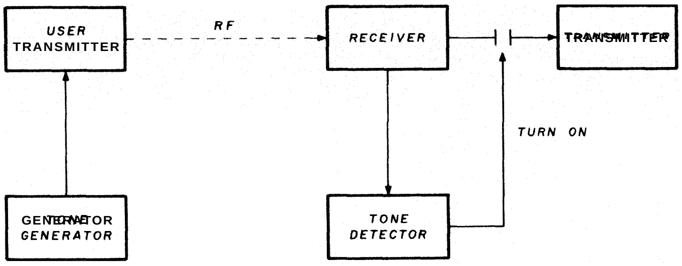


fig. 6. Elements of the tone-access system, which is based on characteristics of the wanted, rather than the interfering, signal.

dipole, its captive area is 0.54 m^2 (free space), or 1.35 m^2 over average ground. A signal of 10 $\mu V/m + 10$ dB above 1V corresponds to a flux level of -136 dB W/m^2, so the antenna captures about -133 dB relative to 1 watt. Assuming 3 dB line loss, this becomes -136 dBW at the receiver input, or about 0.2 microvolt.

Setting the squelch to operate at this signal level would allow use of the squelch circuit to produce the protect control signal. Suitable timers and desensitize circuits would have to be included. A five-minute delay after detection of the channel-occupied signal might be used. A voice announcement could be included to the effect that the output channel is occupied. Emergency over-ride instructions could also be included.

auto-QSY

Probably the most important HF method of interference control is to QSY. This is not really practical in repeater operation. However, there is a partial step which seems possible, **fig. 8,** and one which would make QSY for the station on the repeater output channel easier. This is to shift the repeater output frequency by a small amount; 5 kHz would seem to be suitable.

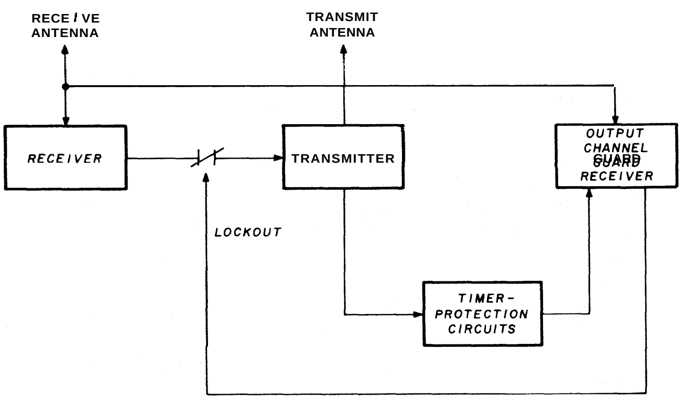


fig. 7. The output-frequency guard receiver principle, in which repeater operation is locked out if the channel signal exceeds some predetermined value.

A limited number of tests with this method indicate that the effect on normal repeater operation would be negligible – a small increase in distortion and a small decrease in range. With no other action, the effect on the other station on the channel would also be small. However, if the other station would also QSY by the same amount, but in the opposite direction, he would escape the interference completely. At the same time his operation would be affected only by a small amount. For the repeater users and the other stations involved, nearly full operation could be restored by retuning.

The signal for this scheme can be generated by the circuitry used for transmit lockout. (If desired, the channel-occupied receiver could be widened out a bit to give better detection of these slightly offset signals.)

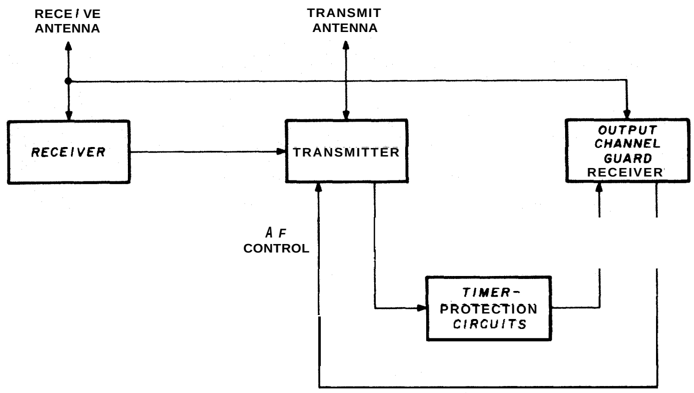


fig. 8. Principle of channel-occupied QSY method, which would make it worthwhile for the station on the repeater output to change frequency slightly.

It may be noted that this offset method is used in vhf television to reduce co-channel interference and to allow closer station spacings. It is a powerful anti-interference technique.

A number of other anti-interference measures have been considered with respect to WR4ADC interference problems. The other technical ones considered did not seem attractive and so are not reported. There were several operational ones of value, and these have been discussed along with the technical solutions above.

With increasing numbers of amateurs on the air, and especially with the continued expansion of repeater operation, interference is going to increase. It seems to be time for the repeater associations to consider the steps that should be taken to minimize those interference problems. This should include recommendation of standard techniques and of protection ratios.

ham radio

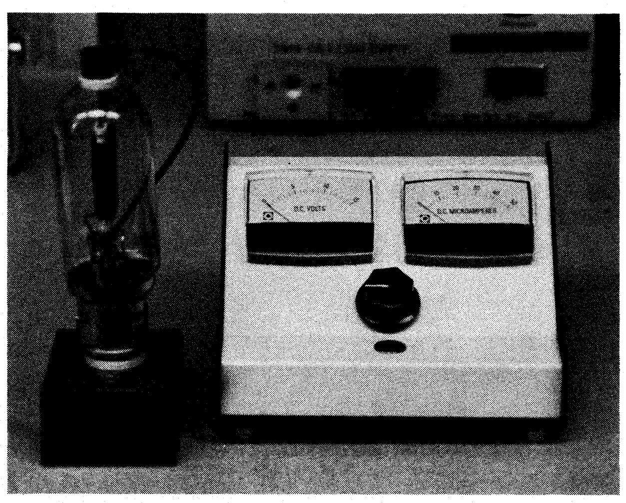
repair bench

Joe Carr, K4IPV

testing power tubes

Several months ago a reader of this column wrote me asking the question, "How do you test rf power tubes?" Not long afterward a close friend of mine, about to build a linear amplifier for 80 through 10 meters, bought a box full of used 4X150 power tetrodes at a hamfest. After building the amplifier he found that none of the tubes were good and that he would have been better off using another type of tube from his collection that was known to be good. He naturally wanted to know if there was a better way available to amateurs than testing by trying.

In my work, medical electronic servicing, we repair electrosurgical generators, which are rf power oscillators that produce up to several hundred watts in the 500-2500-kHz range. Some old but still perfectly good electrosurgical generators use the type UXCV-11 power triodes in a push-pull pair. These tubes cost about $120 each ($240 per pair!) unless you are clever enough to know about United Electronics in Newark, New Jersey. The cost makes stocking a number large enough to permit routine testing by substitution too costly — especially in a shop with a limited budget. Again, there "has to be a better way!"



Final project built by the author for testing high-power vacuum tubes.

Three times within the past six months people and situations have forced the issue of testing transmitting tubes, so perhaps it is time that we covered that point in this column. But before we answer the question posed originally, let's recap a little about elementary vacuum tubes.

A vacuum tube consists of an electron emitter called a cathode, several grids, and an electron collector called either a plate or an anode. The cathode can be either an incandescent filament, or an indirectly heated metal cylinder that has a heated filament at its center. The plate surrounds the cathode but is insulated from it. The various grids are placed in the space between the cathode and the plate.

The plate will have an electrical potential that is positive with respect to the cathode. Grid no. 1, the *control grid,* is given a negative potential with respect to the cathode so it can control the flow of electrons between cathode and plate. The second grid is called the screen grid or accelerator grid and is given a positive potential with respect to the cathode. The third grid, that nearest the plate (G3), is called the *suppressor grid* and may be either biased negative with respect to the cathode or (more commonly) tied directly to the cathode either internally or externally, as shown in **fig. 1.**

Variations in the grid voltage will produce variations in the plate current, which by Ohm's law, become variations in the voltage drop across the load resistor, R_L.

Several vacuum tube parameters are of interest when trying to ascertain overall quality. These are: amplification factor (mu or μ), plate resistance (R_p from the spec sheet or data book), and the transconductance (gm).

The amplification factor is defined as the change in output voltage caused by a given change in grid no. 1 (input) voltage. In fact, the amplification factor of any amplifying device can be given by:

$$\text{amplification} = \frac{E_{out}}{E_{in}} \qquad (1)$$

But for vacuum tubes specifically we can use the notation:

$$\mu = \frac{\Delta E_b}{\Delta E_c} \qquad (2)$$

where:

ΔE_b is the change in plate voltage
ΔE_c is the change in plate current

The transconductance rating relates a change in plate *current* (I_p) for a *small change* in grid voltage, with the plate *voltage* held constant. In other words:

$$gm = \frac{\Delta I_p}{\Delta E_c} \Bigg|_{E_b = constant} \quad (3)$$

where:

ΔI_p is the change in plate current expressed in amperes

ΔE_e is the small change in grid bias voltage expressed in volts

gm is the transconductance expressed in mhos

This relationship will lead us to a method for testing tubes easily and quickly.

tube tester configurations

Fig. *2* shows the basic circuit for a short-circuit tester. This will give only limited information but is useful when screening a large number of hamfest specials. Why perform a more time-consuming test on a tube that has a high resistance short between the filament and cathode, for example?

The tester circuit is nothing more than a series of several continuity testers arranged to ascertain the existence of any resistance paths between adjacent elements. Each continuity tester consists of a low-current filament transformer (*i.e.,* rated at less than 1 ampere) and a compatible lamp. **I** was tempted to specify the use of light emitting diodes instead of lamps but was quickly persuaded to use lamps because the LEDs made the device too sensitive. This problem causes apparent shorts, when all we are reading is electrons flowing from the heated filament to the electrode on half cycles when the ac is positive-going.

It is best to test for shorts after allowing the tube to warm up for a few minutes. Some shorts do not become apparent until after the tube has reached operating temperature. This is also the reason why the use of an ohmmeter is not recommended in this case.

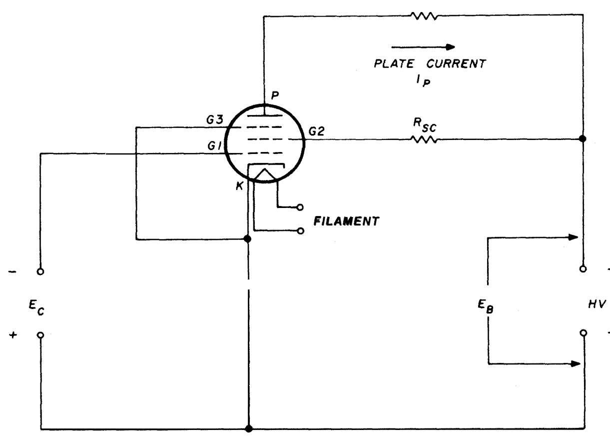


fig. 1. Fundamental vacuum-tube circuit.

The simplest tester circuit that gives us a qualitative insight into the worth of any given tube is the *emission tester* of **fig.** *3.* This circuit tests the tube for the emission of electrons from the cathode. Notice that the tube is connected in a diode configuration in which all elements except the cathode and filaments are connected to the plate.

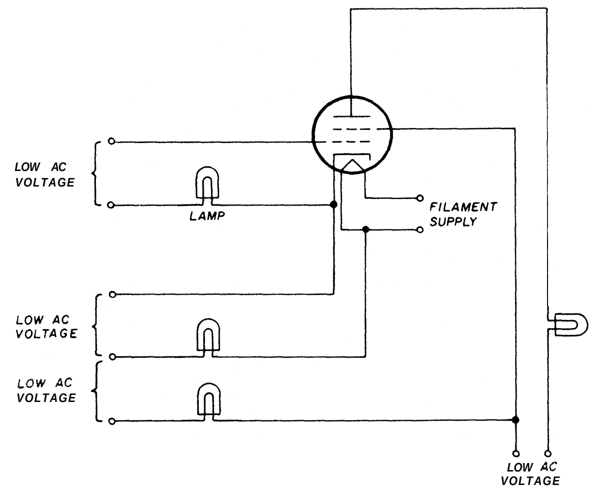


fig. 2. Short-circuit tester.

The tube in a circuit such as **fig.** 3 acts very much like the classic diode. The *emission current* is defined as the saturation current of the tube. If the plate voltage is increased from near zero, we find that the plate current will also increase in a nearly linear manner (except at very low plate potentials). But once a certain critical plate voltage is reached, we find that the plate will attract all the electrons that the cathode produces. Any further increase in plate voltage will produce very little, if any, increase in plate current. The current level at which this occurs is the *saturation* or *emission* current.*

The emission-type tube tester checks for this current, and if the current is low, it will indicate "reject." Some tubes cannot be tested at the actual saturation current either because it is inconvenient to do so, or because the current is too high and may damage the tube. These tubes are tested at a specific plate potential, at which a given current is expected. If the tube will not produce at least a certain predetermined percentage of that current (usually 80 to 90 per cent), then it is rejected. The emission tester can spot a grossly bad tube, but there are certain problems with

*An Interesting variation of the emission tester, which does not require high voltage but uses a low-voltage bias source, is described in reference 1 **editor.**

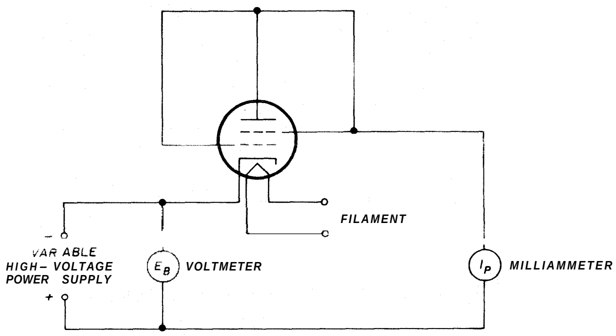


fig. 3. Circuit for simple emission-type tube tester.

this type. Note that most drug-store tube testers are emission types.

Defects that will prevent the tube from operating normally will not always show up on a simple emission tester, so some better means is necessary. The type of tester preferred by most professional servicers is the *mutual conductance* or *transconductance* tester. Examples of this type of circuit are given in **fig. 4.** The circuit shown in **fig. 4A** is a static transconductance tester using the *grid-shift* method, also called grid-level shift.

Switch S1 iis in position 1 at the beginning of the test, making grid voltage E_c equal to $E2$ alone. Both $E1$ and $E2$ are adjusted to produce a convenient (safe!) plate current. It is usually wiser to begin with the plate voltage (E1) at some level well within the range that can be tolerated by the particular tube being tested, and have the grid voltage at some value in excess of cutoff for that tube. This may be unnecessary much of the time but could save you some grief often enough to make it a good standard practice. You may then adjust the grid voltage downward until the plate current is at a convenient level. When this adjustment is completed, make a note of the values of E_b, I_p, and E_c.

To perform the test, place switch S1 in position 2. This operation makes grid voltage E_c equal to $(E2 + E3)$. The plate voltage is then measured and the supply readjusted if a change has been noted. The plate voltage at this point must be equal to the plate voltage that existed initially.

Now read the plate current, and find the *difference* between this reading and the initial reading. Plug the difference current into the formula:

$$gm = \frac{\Delta I_p}{1.5} \qquad \textbf{(4)}$$

where:

ΔI_p is in amperes

gm is in mhos

This answer is given in *mhos,* but most vacuum tube spec sheets list the transconductance in micromhos. 1 mho = 1-million μmhos, so multiply the answer by 10^6.

A *dynamic* transconductance tube tester is shown in **fig. 4B**. This circuit is the basis for most commercial tube testers. A low voltage ac transformer is connected in series with the grid bias power supply, so that the transformer secondary voltage forms the "*delta-E_c*." Actual transconductance is measured on the plate ac milliammeter, which is calibrated in units of conductance.

Fig. 5 and the photo show a test jig I built to test power tubes at work. It will serve equally well for amateurs. The circuit is the simple grid-shift method for finding the transconductance of the tube being tested. Note that it is not a real tube tester construction project because it lacks power supplies. Almost all amateurs can jury-rig adequate power supplies to make this test or can borrow bench supplies. Note also that it is not strictly necessary to use the full operating voltage of the tube to obtain meaningful results.

Please, be very careful when using this jig; high voltages will be exposed! If there is any doubt, place the tube socket subassembly inside an insulated or *grounded* metal enclosure. I may seem to harp on

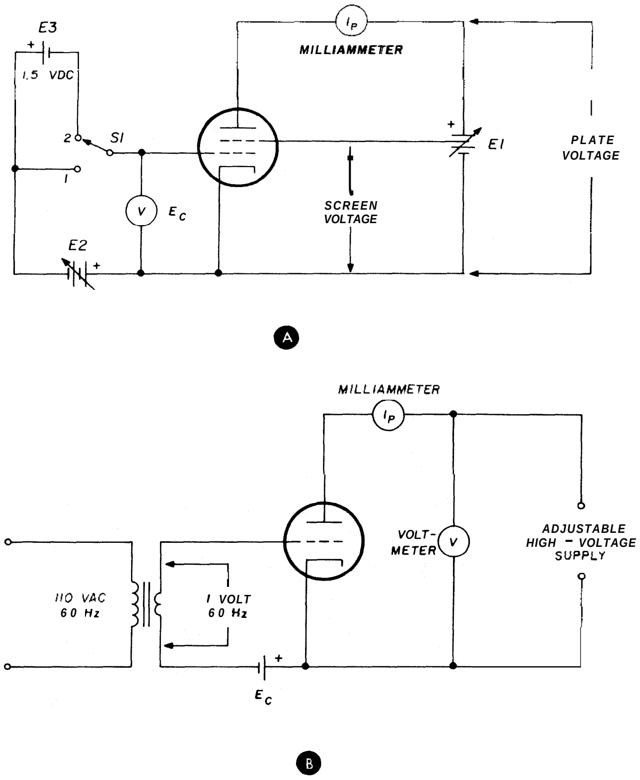


fig. 4. Static transconductance tester, A, and dynamic transconductance tester, B.

safety an awful lot, but it is so very important. Recall the note on one of my past articles which pointed out that an editor almost canned himself while working on a 4000-volt final amplifier power supply.

The test jig is built in two main parts, a main assembly and a tube-socket subassembly. A multiconductor cable connects the two parts of the jig. This design was selected so that the same mainframe can be used to accommodate a larger number of different tube types.

Jacks J1-J8 are heavy duty banana-jack binding posts, while J9/P1 are a mating pair of multipin connectors such as the circular MS or AN series. I used a high-voltage power supply that delivered 500 Vdc, but if greater potentials are needed (I doubt that they will) it would be wise to change J1-J4 to high-voltage chassis connectors.

Meter M1 is an appropriate high-voltage meter, although in my case a voltmeter was made from a suitable multiplier resistor (a pot and a fixed resistor) and an available 0-50-μA meter movement. Meter M2 is a dc voltmeter with a range suitable for the range of grid voltages expected.

Resistors R1-R3 were selected so that varying R2 produced approximately 1 volt of grid voltage change. R1 is used to allow precision trimming of that change. In my case, E_c' was 25 Vdc, but if your voltage is different, use the normal voltage divider equation to find appropriate resistor values.

Resistor R4 is made approximately equal to the plate resistance of the tube being tested, which in this case was about 3000 ohms.

I used an external plate current milliammeter because there was a digital multimeter available that would measure current very accurately.

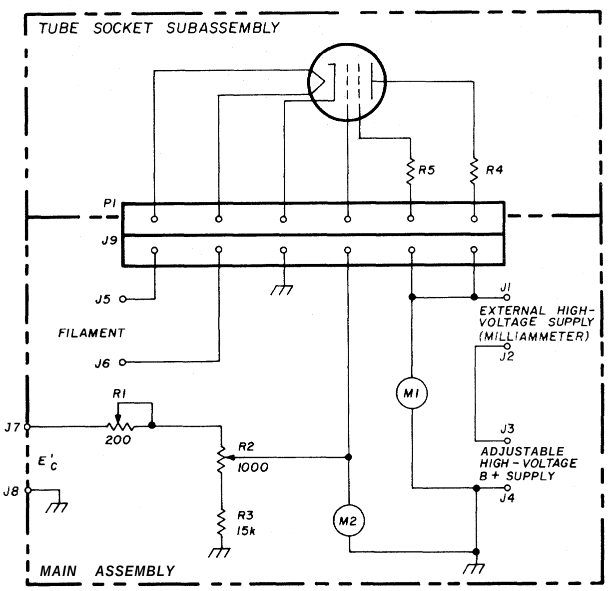


fig. 5. Schematic for a practical static (grid-shift) transconductance tube tester.

table 1. Transconductance measurements of type UXCV-11 tubes made by the author on the test jig shown in fig. 5.

bad tubes	ΔE_c (volts)	(amps)	gm (micromhos)
1	1	0.0013	1300
2	1	0.0015	1500
3	1	0.0014	1400
good tubes			
1	1	0.0045	4500
2	1	0.0039	3900
3	1	0.0043	4300

Table 1 lists the values of transconductance actually measured for both known good and known bad tubes. The UXCV-11 has a *mu* of 14 and a plate resistance of 3220 ohms. The spec sheet does not give the transconductance, but we may compute it from:

$$gm = \frac{\mu}{R_p} \tag{5}$$

$$gm = \frac{14}{3220}$$

$$gm \approx 0.00435 \ mhos$$

$$gm \approx 4350 \ \mu mhos$$

(Note that all the bad tubes had grossly lowered gm readings.)

This tester is not "scientifically" designed but is intended to allow amateurs to test power tubes on an occasional basis. Its saving grace is that it can be built inexpensively! Note that I defined "bad tube" by saving from my own work those tubes that had been found to produce customer complaints similar to, "It seems to work, but has low output." These observations were confirmed by an rf ammeter in series with my 500-ohm (not 50-ohm; these were medical rf generators) dummy load. The real value of this tester is that it will allow you to test hamfest specials or perform preventive maintenance on your equipment. Of course, if you have an rf power meter or rf ammeter in your feedline, then a low-power output coupled with seemingly normal drive will point the finger to the final amplifier tubes. But the thought of that poor guy building an entire linear amplifier around a whole box full of bad tubes seems to justify doing a little testing — at five minutes per tube, how could he have gone wrong?

references

1. Neil Johnson, W2OLU, "Testing High-Power Tubes," *ham* radio *(The Ham Notebook)*, March. 1972, page 64.

ham radio

microcomputer interfacing: interfacing a 50-bit DAC

A Digital-to-Analog Converter or DAC is an electronic device that converts digital signals into analog signals. A typical converter consists of an arrangement of "weighted" resistors, each controlled by a single bit of input data, that develops varying output analog voltages or currents in accordance with the digital input code.[1] You could use a DAC to provide a small analog error signal from a microcomputer used in a feedback circuit, to convert a sequence of bytes in memory into analog-vs-time data and thus simulate the output from an analog instrument such as a rotator control box for tracking OSCAR; to provide analog data for the two channels of an x-y recorder, or in general, to operate any device that requires an analog voltage or current and is interfaced to a digital device, such as a microcomputer.

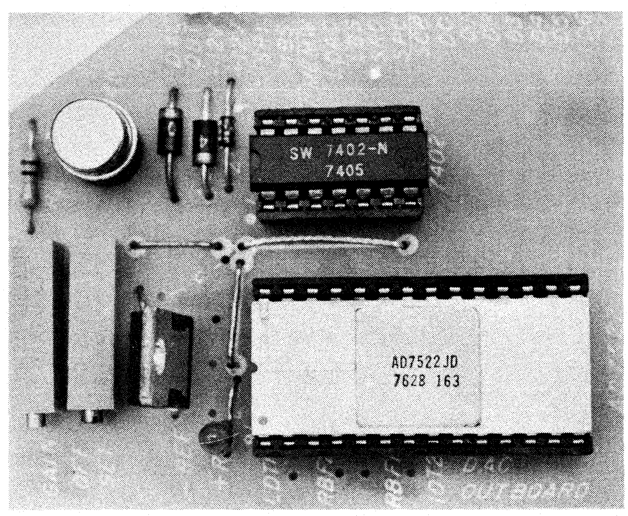
For a general discussion of the principles of analog/digital conversion, you should read the excellent Analog Devices conversion handbook[1] or the series of small pamphlets distributed by National Semiconductor Corporation.* Important terms and concepts associated with DACs include resolution, accuracy, scale error, gain error, offset error, linearity, differential linearity, settling time, slew rate, overshoot and glitches, temperature coefficient, supply rejection, conversion rate, and output drive capability. A few of the terms have been summarized in **table 1.**

By Peter R. Rony, Jonathan A. Titus, Christopher Titus, and David G. Larsen, WB4HYJ

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus, and Dr. Christopher Titus are with Tychon, Inc., Blacksburg, Virginia.

To help understand how you'd interface a DAC to an 8-bit microcomputer, **fig. 1** shows the connections between an Analog Device AD7522 and an 8080A-based microcomputer. An important feature of this specific DAC is the fact that it is double buffered: this means that there exist within the device

The **DAC *Outboard*** circuit board contains all the circuitry shown in ***fig.* 7. A** copy of the board layout is available from the authors.



two independent 10-bit registers, the DAC register and the two-bit and eight-bit shift registers (fig. **2**).

A DAC is an output device for a microcomputer, and thus data is strobed from the microcomputer data bus into the internal registers or latches, of the DAC. In fig. **1,** are shown the connections to the 8-bit bidirectional data bus, DØ through D7, the 8080A control signals OUT or MEMW, which are used with accumulator I/O or memory I/O data transfers,[3] and the channel select outputs 003 through 005 that are generated by a decoder tied to the microcomputer address bus.[4]

Since the AD7522 is a 10-bit DAC, it is not possible to simultaneously load all ten bits from an 8-bit microcomputer. The sequence that actually occurs can be summarized as:

1. The DAC input bits DBØ through DB7 are first strobed into the 8-bit shift register/latch using a positive device select pulse applied at pin 24, (LBS or Low Byte Strobe).

2. The most significant two bits, DB8 and DB9, are then strobed into the 2-bit shift register via the use of a device select pulse applied at pin 25 (HBS or High Byte Strobe).

3. Finally, a device select pulse applied at pin 22 (LDAC or Load DAC) transfers the ten bits of input data, DBØ through DB9, into the second buffer within the DAC chip, the DAC register.

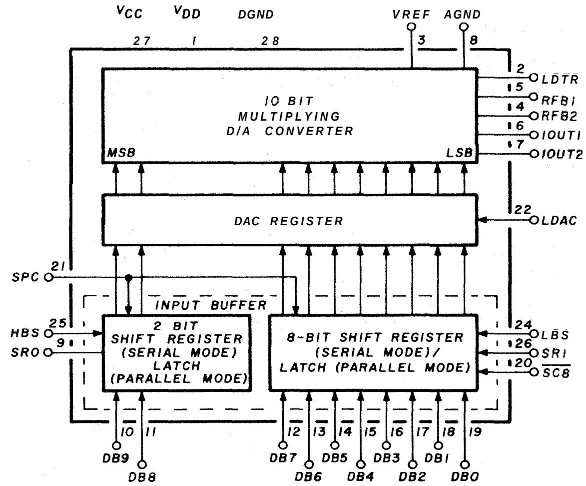


fig. 2. Functional diagram of the AD7522 IC. This figure is courtesy of Analog Devices.

The output current appears at IOUT1 and IOUT2 and is converted into a voltage with the aid of a 741 operational amplifier. The two most significant bits are loaded from the eight-bit microcomputer bus using any two bits. Generally bits DØ and D1 are chosen since it makes data formatting easier. Thus, the ten bits are transferred as eight bits DØ to D7 and as two additional bits, DØ and D1.

A simple program that exercises the DAC over its full operating range is provided in **table** 2. The program generates a slow linear ramp at the analog out-

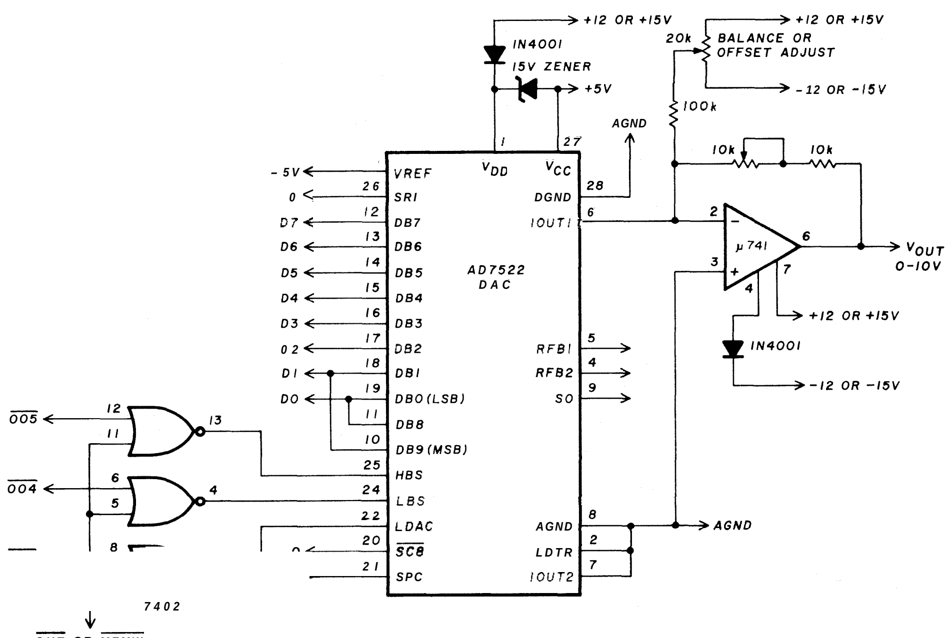


fig. 1. Schematic diagram of an interface circuit between an 8080A-based microcomputer and an Analog Devices AD7522 digital-to-analog converter.

table **1,** Important concepts and terms associated with digital-to-analog converters.

Resolution	The smallest standard incremental change in output voltage of a DAC A converter with *n* input bits can resolve one part in 2^n.
Accuracy	Describes the worst case deviation of the DAC output voltage from a straight line drawn between zero and full scale; it includes all errors.
Settling time	The elapsed time after a code transition for a DAC output to reach a final value within specified limits.
Conversion rate	The speed at which a DAC can make repetitive data conversions.
Nonlinearity	Error contributed by a deviation of the DAC transfer function from a best straight line function. Normally expressed as a percentage of full scale range,
Monolithic chip	An integrated circuit chip in which born active and passive elements are simultaneouslv formed in a single small silicon wafer via the use of diffusion and epitaxial processes Metallic stripes are evaporated onto the oxidized surface of the silicon to Interconnect the elements
Multiplying DAC	A digital-to-analog converter in which the output analog signal is the product of the number represented by the digital Input code and the input analog reference voltage, which may vary from scale to zero, and in some cases, even to negative values

table 2. Memory **I/O** program that generates a slow linear ramp. Execution starts at HI = **003** and LO = **000**.

LO address byte	instruction byte	mnemonic	comments
START: 000	042	SHLD	Strobe ten bits of digital data into the AD7522 DAC shift registers. The ten input data bits are contained in register pair H. The address select code for the LBS input is HI = 200 and LO = 004; the address select code for the HBS input is HI = 100 and LO = 005.
001	004	004	
002	200	200	
003	062	STA	Strobe ten bits of digital data from the input buffer into the DAC register within the AD7522 DAC. The address select code for the LDAC input is HI = 200 and LO = 003.
004	003	003	
005	200	200	
006	043	INX H	Increment register pair H
007	315	CALL*	Call 10 ms time delay routine, DELAY
010	277	277	LO address byte of DELAY
011	000	000	HI address byte of DELAY
012	303	JMP	Unconditional jump to START, where the input of new data into the DAC occurs
013	000	000	LO address byte of START
014	003	003	HI address byte of START

*On the 8080-based microcomputer that we use in our courses, a 10 millisecond time delay subroutine is located in EPROM starting at HI = 000 and LO = 277. Such a routine can be located anywhere in memory.

put of the AD7522. This can be observed on a Vom, digital muitimeter, or oscilloscope. The ramp is subdivided into 1024 small steps, each step being approximately 5 mV in magnitude. The total time required to change from 0.0 volts to +5.12 volts is 10.24 seconds. The SHLD <B2> <B3> instruction outputs two data bytes in succession, from register pair H, into the input buffer registers of the DAC. The contents of register L are transferred into the 8-bit shift register, while the least significant two bits in register H go into the 2-bit shift register. Note that the address is automatically incremented, and a second MEMW control pulse generated by the 8080A when the SHLD instruction is executed. The STA <B2> <B3> instruction provides only a strobe pulse at the LDAC input to the DAC; no data transfer occurs between the accumulator and the DAC.

Other small monolithic and hybrid DAC systems are available from different manufacturers. The Analog Devices converter was chosen because of the on-the-chip latches and double buffering registers. The use of a reference potential is common to many DAC modules. Perhaps in the future it, too, will be included in the module.

references

1. *Analog-Digital Conversion Handbook,* Analog Devices. Inc., Norwood, Massachusetts 02062.
2. "Specifying AID and D/A Converters," AN-156, and "Data Acquisition System Interface to Computers," AN-159. National Semiconductor Corporation, Semiconductor Division, 2900 Semiconductor Drive, Santa Clara, California 95051.
3. Peter R. Rony, *et al,* "Microcomputer interfacing: Accumulator I/O *vs* memory I/O," *ham radio,* June, 1976, page 64.
4. Introductory Experiments in Digital Electronics, 8080A Microcomputer Programming, and 8080A Microcomputer Interfacing, Bugbook V, E & L Instruments, Inc., Derby, Connecticut. 1976.

ham radio

let's reduce audio pollution in the vhf bands

How's your audio quality over the repeater? Here are some ideas for improvements

Much has been written about the control and power output of vhf fm equipment. But little has been written, that I can discover, about the input part — audio signal generation — particularly the how and why of obtaining audio quality throughout the entire chain of repeater operation.

As a one-time audiophile I have paid close attention, the past several months, to the audio quality of signals being retransmitted by amateur repeaters. It's been a pretty sad experience, but there's a glimmer of hope. I've also been hearing others commenting on the situation, and steps are being taken to remedy it.

Other than from equipment electrical malfunctions, bad audio is caused mostly by *close talking* into the dynamic microphone that usually comes with the radio. A couple of years ago I wrote an article on the subject of proximity effect for my club paper. The article was subsequently published in *ham radio*.[1] It pointed out that all conventional moving-coil microphones (including speakers used as microphones) accentuated rhe bass tones by a factor of 3 to 4 dB at 200 Hz (fig. 1). Such response can easily cause the first audio amplifier stage to overload and feed this signal into the transmitter.

I've heard many reasons why amateurs don't remain at least 5 cm (2 inches) from their microphones, none of which are valid. Many old wives' tales have sprung up regarding the use of microphones. The most ridiculous is that foreign equipment manufacturers have "set the audio stages of their radios for their countrymen." Sheer nonsense! It's usually not necessary to touch any internal controls in any radio, regardless of its origin, foreign or domestic. This statement is qualified by reminding you that, in most transceivers, audio is set using one audio frequency at a time from an audio oscillator, and that the radio and microphone are packaged in a carton as the equipment goes out the door.

talking across the microphone

The next fallacious theory to be laid to rest is the "talk-across-the-microphone" nonsense. *All* micro-

By Budd Meyer, K2PMA, 6505 Yellowstone Blvd., Forest Hills, New York 11375

phones are designed for speaking directly into the active element **(fig. 2)**. The sound waves should impinge onto the element perpendicularly; the microphone element should be in the same plane as your mouth. The microphone element is designed to react to, or transduce, all frequencies pretty much equally.

Unfortunately, microphone designers have a problem in that the higher audio frequencies are more directive. The further from a straight line access to the microphone element, the greater will be the loss of the higher audio frequencies reaching the microphone element. **Fig.** 3 gives an idea of what happens when an audio signal (your voice) impinges indirectly onto the microphone element. Audio frequencies above about 800 Hz travel almost directly toward the element, whereas lower audio frequencies have much less directivity and much more power per unit bandwidth.

So if you talk *across* the microphone, the bass voice frequencies can overload the audio chain in the radio; and the high frequencies, which are essential to intelligibility, are attenuated. I'd like to emphasize that the information above applies to *all* microphones whether crystal, dynamic, ceramic, or whatever. If you talk across the microphone, you're going to sacrifice the higher audio frequencies.

over-deviation

Another situation that should be remedied is the idea that increasing the fm deviation will increase power output. More nonsense! The rf output of an fm transmitter is constant, and no amount of "dinking" with the deviation control will increase power output. If you adjust the deviation control so that the radio is over-deviating, all you will do is increase distortion, since the transmitted signal bandwidth will be increased.

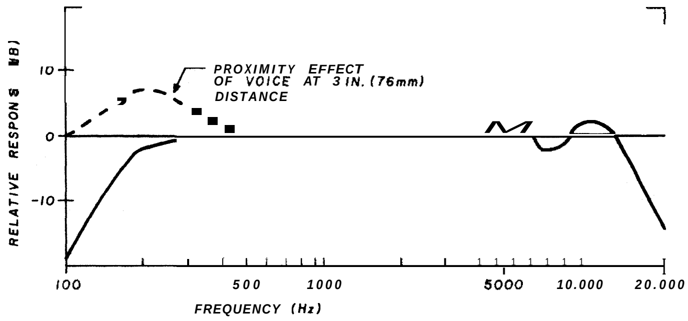
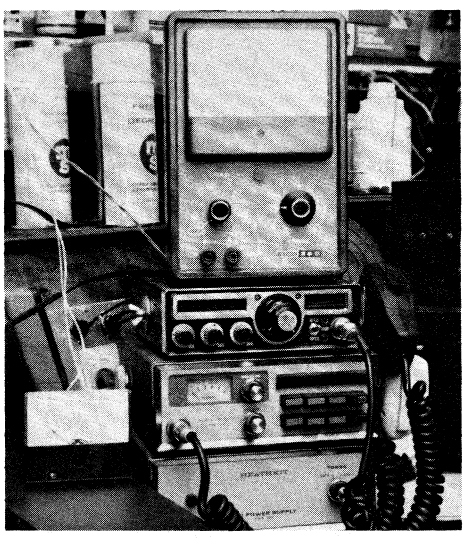


fig. 1. Generalized response curve of the dynamic microphone. Note the accentuated bass tones at 200 Hz (reference 1).



Test setup used by K2PMA to check VU meter readings with a calibrated audio power meter.

Most amateur repeaters have some form of deviation limiting control to prevent their signals from spreading out. If your radio's deviation setting is above this predetermined limit, you'll probably experience the phenomenon of "popping out the repeater." You won't gain anything but shrugs from other repeater users who know better. You may even get an admonishment to clean up your signal.

some answers to the problem

How do you obtain good audio quality through the repeater? Much of the answer lies in the desire to do so. A standard answer you'll get, as I have, when you mention to someone who is over-deviating that perhaps he might try backing away from the microphone, is "You're the first to complain." Don't let this answer put you off. Most amateurs will back away from the microphone if asked to do so. Of course, if you tell him he sounds better at *that* distance from the microphone, and he realizes he's a half-meter (2 feet) away, he just *might* consider adjusting his microphone gain control!

simple tests

Some time ago I came across a good buy in volume unit (VU) meters for my stereo system. I purchased an additional meter and connected it to the output of my vhf transceiver, which uses an 8-ohm speaker. I now had a reference of sorts – the fact that it was qualitative rather than quantitative served my purpose. At a comfortable listening level, the VU meter indicated average and peak audio output. To get quantitative values, I connected a calibrated audio-power meter to the output, which proved the validity of my VU-meter measurements.

Being aware that most amateurs don't have VU or audio-power meters, I connected a Simpson 260 Vom to the speaker output. Lo and behold — the VU-meter and Simpson 260 Vom readings matched! Such instrumentation can be used to advise the fellow on the other end that he's over-deviating, and you can prove it by comparison.

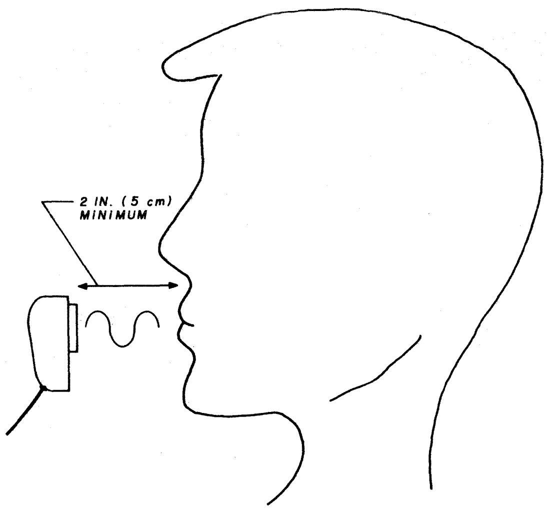


fig. 2. Most fm problems with poor audio can be remedied by speaking at least 5 cm (2 inches) from the microphone. Loud speech doesn't improve power output.

As a rough approximation, I suggest to the other fellow that he adjust his level to cause my VU meter, which has 300 ohms rather than the standard 600-ohm impedance, to average – 7 VU and peak at – 4 VU. (A correction factor of 15.75 dB must be added to the reading.) This comes out to about 7.5 mW average and 15 mW peak. These numbers are only guides; you can set your own standards based on your own equipment.

measurement problems

Audio measurements are subjective, which means that all such measurements obtained by instrumentation are subject to individual taste and hearing characteristics. I've listened to rigs set up by deviation meters and oscilloscopes and note that this method just doesn't accomplish the job. There are resolution problems with oscilloscopes and frequency problems with deviation meters. Personal habits and voice characteristics color the tests. Obviously an oscilloscope presentation will show clipping; however, by the time the clipping is discernible the distortion will probably be very high.

How can we obtain good audio? Probably 90 per cent of the distortion heard on repeaters can be eliminated if we all talked at least 5 cm (2 inches) away from the microphone. Also, accepting the fact that talking loudly does not equate to more output power will eltminate the other major culprit causing distortion, as noted previously.

I've no recommendation for setting up metered standards since an infinite number of variables must be considered. And no matter what one could come up with, we still must contend with the subjective aspect of audio measurements. I've heard many amateurs with unbearable distortion access a repeater and ask for an audio check, only to be told that they sound good. The smartest thing to do would be to listen to your favorite repeater for awhile and learn which people give accurate reports. If one fellow consistently *improves* the audio of others with honest reports, then he's your man.

level setting

Here are some ideas for setting the level of your radio. As an example of the typical imported fm transceiver, I've provided a couple of sketches showing where to find the deviation and microphone gain control **(fig. 4).**

First, you must have someone listen to your signal over a period of time, preferably on a simplex frequency, so that the repeater doesn't affect your audio. Start by practicing the technique of talking at least 5 cm (2 inches) from the microphone. Once you get the hang of it, this procedure will become second

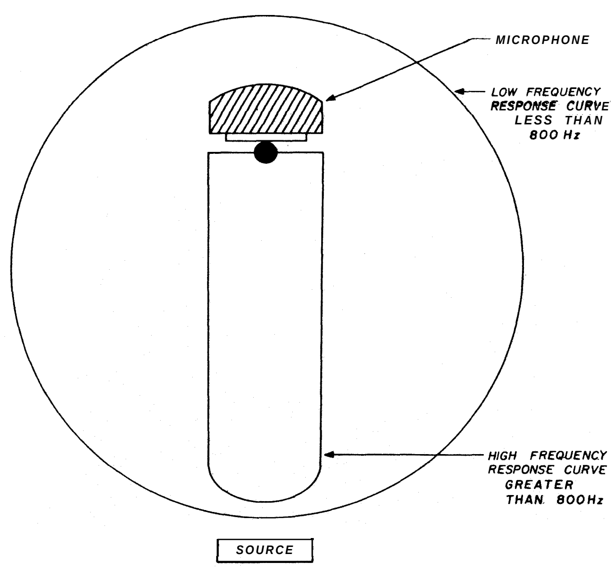


fig. 3. Response patterns of dynamic microphones. Higher audio frequencies, which convey intelligence, are more directional than the lower frequencies — a case supporting the theory for talking directly into the microphone rather than across it.

nature. Chances are you won't have to touch the radio's innards.

If you have to reduce the audio, locate the microphone gain control (not the deviation control), and turn it in small increments until your listener agrees that your audio is sufficient and clean — no rough edges. i don't beiieve that these tests can be done in committee fashion — stay with one man!

Only rarely is it necessary to touch the deviation control. What procedure do you use if there's only one control in your radio? Not much, really. The situation then becomes a tradeoff between microphone talking distance and level. Make every attempt to keep your audio clean. If your radio has only one control, not eating the microphone is bound to improve your signal quality. Some hand-held radios use the speaker as a microphone; not much can be done here.

If your radio has a deviation control but no microphone gain control, it's easy to adjust the deviation control so that your signal isn't two barn doors wide. As I've mentioned previously some repeaters are deviation limited, and if your fm signal swing is beyond the preset standard set by the repeater, the machine will pop out when you try to talk.

Locate the deviation control, which is generally near the end of the speech amplifier and audio chain. Ask your friend to listen to your signal while you adjust the deviation control for minimum swing with clean audio. Keep in mind that excessive deviation ***does** not* increase talk power or rf output. You can adjust the deviation control so that your signal will deliver all the talk power you can use without objectionable distortion.

the microphone

As a final attempt to clean up your signal, you can always buy a new microphone. Perhaps yours is out of spec, or you dropped it once too often. The thing to look for here is not so much impedance match (since most dynamic microphones will work with most transistorized amplifiers) but the microphone output level — somewhere in the vicinity of 10-50 mV across 2000 ohms. How do you ascertain this without meters? Buy it with return privileges.

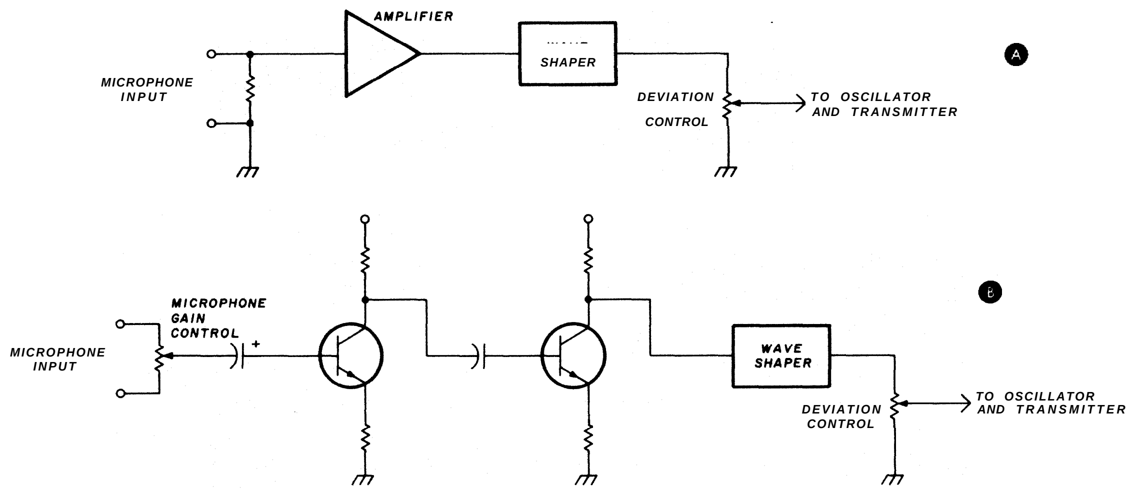


fig. 4. Generalized sketch of most fm vhf radios showing where to find the deviation control, A. and the microphone gain control, B.

closing remarks

Let us not forget that your transmitter is only one-third of the communications chain. Repeater operators should also get into the loop. Someone in the group should listen to the average quality of the repeater audio. Sad to say, many repeaters have all kinds of added exotic features, but repeated audio is poor because no one has taken the time to check it.

We have autopatch, beeps, timers, tone-encoded outputs, frequency checks, automatic signal reports, welcoming speeches, and the like, but bad audio in many cases. I earnestly believe that control stations should absolutely and positively comment on bad audio coming into the repeater.

I hope this article will help improve a long-neglected aspect of amateur radio. I've found most hams to be cooperative in helping to rid the airwaves of audio pollution.

reference

1. "Dynamic Microphones," ***ham radio***, (Circuits & Techniques), June, 1976, page 49.

ham radio

$1.50

ham radio magazine

MAY 1978

ham radio magazine

MAY 1978

volume 11, number 5

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

U.S. and Canada: one year, $12.00
two years, $22.00
three years, $30.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

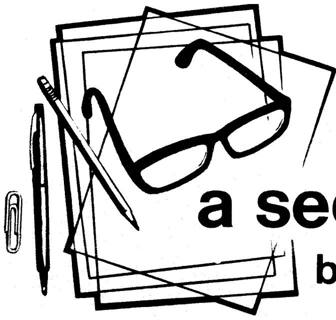
Foreign subscription agents are
listed on page 117

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a sec:ond look

by Jim Fisk

Let's say you have an Extra or Advanced-class license and are bored with working DX, handling traffic, or trying out new SSTV circuits. What next? Why not shift gears and put your station and expertise tc work helping out the Novice operators?

The other day I was tuning across the 40-meter Novice band and heard a cool, crisp, CW signal sending CQ. The band was full of shortwave-broadcast signals, and the CQ signal wasn't very strong. But it sounded like a tape recording – slow, steady, and with good spacing between characters. I listened. Nothing much but foreign-broadcast QRM and a few weak CW stations working each other. Then, down in the noise, I heard a fellow calling and answering the CQ. He was a WB9, sending very slowly with many errors, pauses, and lots of stops and starts. The station calling CQ answered the WB9, slowly and patiently.

What ensured was one of the more beautiful things that happen in amateur radio. The station calling CQ was an old timer, who decided to put his rig on the Novice 40-meter band. The Novice who answered his call, obviously brand new and unsure of himself, was glad to hear *any* station. I listened to the QSO. What happened renewed my faith in ham radio. The old timer was obviously ready to use his high-speed keyer – you could tell at the end of his transmission. But the old timer kept his cool and pounded away on his straight key, slow and steady.

It all ended with the usual amenities. The Novice wanted another QSO, but above all, he wanted a QSL card to confirm the QSO. No problem. Addresses arid names were exchanged. Hopefully, both exchanged QSL cards and both followed up to build a lasting friendship.

If you're an experienced ham operator and are interested in expanding your horizons, why not consider putting your rig on the Novice bands? The 40-meter Novice band is a good place to start. Many new hams obtain gear that is band limited, and most Novices start out on the low-frequency ham bands – 40 and 80 meters. Equipment is easy to get working on these bands, as most old timers will realize. Whatever Novice band you choose, bear in mind the following facts.

Most Novices have dipole antennas and rather unsophisticated transmit-receive facilities. Many use knife switches to transfer between receive and transmit. So if you decide to put your rig on the Novice bands, with all the new features, remember that the Novice is unaware of most modern developments. He's interested in receiving your transmission, simply and without flair.

Patience is the watchword when working Novices. Patience requires a certain discipline that pays off in genuine self satisfaction when you've completed a good QSO.

If you're an overseas amateur, your signal is more than welcome in the U.S. Novice bands. You'll find Novices that can handle Morse pretty well – they're almost ready for the next step – the exam for the General-class license. But don't overlook the vast majority of U.S. Novices. All would like a QSO with a DX station. Slow down from time-to-time and give the new fellows a chance.

Some additional tips if you're an old timer and want to work Novices: If you call a Novice station and a reply is not immediately forthcoming, listen for at least two minutes. Perhaps it's his first QSO. The new amateur who puts his rig on the air for the first time is usually nervous, anxious, and maybe somewhat confused. If you can remember your first QSO, I'm sure you know the feeling (the first time I attempted to answer another station I forgot to turn on the B+ supply to the transmitter, but that's another story). So if the Novice you call doesn't come back right away, tune around and wait. Chances are he'll answer after a few minutes. If not, call him again. He's probably overwhelmed that someone answered his call.

It's a great feeling to be able to help a newcomer. All you need is a little patience and perseverance. If you do it right, you'll be surprised. You'll have a pile-up of Novices trying to work you – it's almost like being a rare DX. Try it, you'll like it. . . but be patient.

Jim Fisk, W1HR
editor-in-chief

Windom antennas

Some facts and fiction about the Windom antenna, and how it measures up in terms of bandwidth and input impedance

One of the simplest and most economical antennas suitable for multiband amateur use is the single-wire, off-center-fed antenna, often called the Windom antenna.

Although this antenna in its present form was first described in the professional literature in 1929 by Everitt and Byrne,[1] and in the amateur literature by Loren Windom, W8GZ/W8ZG,[2] very little new or original theoretical or design data has been published since that time. To the best of my knowledge, no information on impedance or bandwidth measurements on the antenna has ever been published.

In this article I will begin by giving the history of the antenna, then a theoretical discussion of the antenna and single-wire transmission line, followed by a method of adjusting the antenna, and finally, the results of measurements I have made on an actual antenna.

history

The single-wire feedline has been credited to Frank Conrad, 8XK, of Westinghouse who used it in the broadcast band to feed a quarter-wavelength grounded (Marconi) antenna. The next step was taken by V. D. and E. B. Landon, 8VN, who connected the single feedwire to the junction of the antenna and counterpoise.[3] In a later article by Howard M. Williams, 9BXQ, the counterpoise was stretched out and made a part of the antenna with the feedpoint still off-center.[4] **Fig. 1** shows a summary of these early developments.

In the middle 1920s Loren Windom operated 8GZ/8ZG in the Columbus, Ohio, area. Windom ran high power for that time (250 watts) and was considered the technical bellwether of Columbus hams, as he was an active experimenter, working with antennas among other things.

In July, 1926, the technical editor of QST, Robert S. Kruse, published an article which gave a roundup of various methods of feeding an antenna as they were understood at that time.[5] A discussion by Windom on how to adjust the off-center-fed antenna was a part of that article; the procedure consisted of placing a light bulb in the center of the antenna **(fig. 2**) and adjusting the feedpoint for maximum lamp brilliance (maximum current). This procedure was also published in the first three editions of the ARRL *Radio Amateur's Handbook.*

John Byrne, then 8DKZ, became associated with Windom; 8GZ/8ZG QSL cards, circa 1925, carried the names of both Windom and Byrne. Both were students at Ohio State University — Byrne in electrical engineering and Windom in law.

It was customary in those days for senior engineering students to do a thesis for graduation; Byrne and his thesis partner, E. F. Brooke, 8DEM, chose the single-wire transmission line as their topic. They carried out a considerable amount of research on the

By John J. Nagle, K4KJ, 12330 Lawyers Road, Herndon, Virginia 22070

subject under the guidance of their faculty advisor, W. L. Everitt, who was then Associate Professor of Electrical Engineering. Everitt will be remembered for his book, *Communication Engineering,* which was the standard college textbook on the subject in the 1930s and early 1940s.

Windom, although not formally associated with the project, assisted from time to time in rigging or making measurements on his lunch hour or after classes, so he was up-to-date on the progress of the work. In fact, he incorporated many of the more promising facets at his own station.

Byrne's investigation was incomplete when he graduated in June, 1927; however, he returned as a graduate student for the 1927-28 academic year and continued work on the single-wire feeder as his Master's thesis with A. B. Crawford as his partner. The work thus accomplished was published in the *Proceedings of the IRE* in October, 1929, with Everitt and Byrne listed as authors.[1] It was standard practice then, as now, that when a student's research work is formally published, the faculty advisor is listed as senior author.

The work of Everitt, Byrne and Brooke showed that the Windom procedure of 1926 was incorrect. Windom agreed in his 1929 article, and stated that his earlier method should not be used; he went on to describe the Everitt-Byrne method that I will explain later. Beginning with the seventh edition of the *ARRL Handbook,* the old, Windom graphs for the single-wire feeder were presented which give the length of the antenna as a solid line and the location of the tap as a dashed line for each amateur band. The Everitt-Byrne procedure has not been published in any of the ARRL handbooks or antenna manuals and is not widely known.

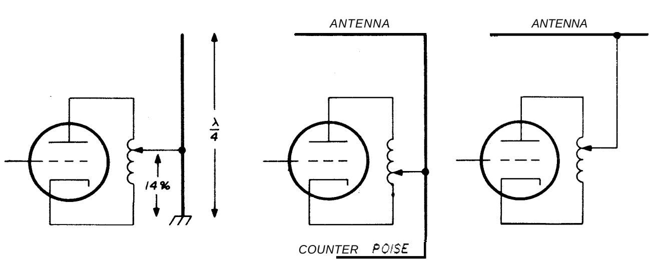


fig. 1. Early development of the off-center fed antenna beginning with the single-wire feedline used on the broadcast band by Frank Conrad, 8XK, to feed a quarter-wavelength vertical (left). Later the Landon brothers, 8VN, connected a single feedline to the junction of the antenna and counterpoise (center). Howard Williams, 9BXQ, stretched out the counterpoise and made it part of the antenna with the feedpoint still off-center (right).

Windom recognized the utility of the method to the amateur community and encouraged Byrne to write it up for QST. Byrne declined, suggesting that Windom write it up himself as Windom was familiar with the work. Windom's article was published in the September, 1929, issue of *QST,*[2] a month before the Everitt-Byrne article. The delay in the Everitt-Byrne

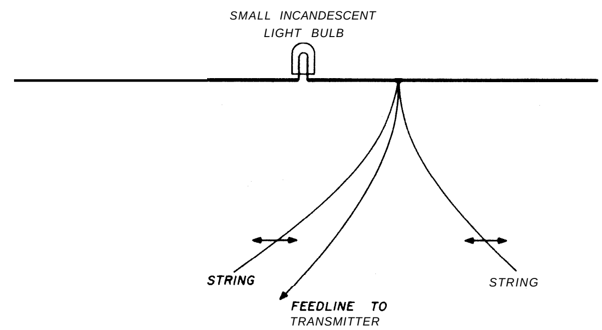


fig. 2. Windom's 1926 procedure for adjusting the feedpoint of an antenna as described in QST. The feedline was moved along the antenna until the bulb glowed the brightest. Later experiments at Ohio State University showed this method to be in error.

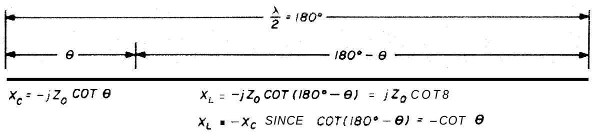
paper is ascribed to the more extensive editorial review required by the *Proceedings of the IRE;* Windom, to his credit, stated in an early paragraph that he had not done any of the work himself — that he was only reporting the work of others. The fact that Windom's article was published a month before the Everitt-Byrne paper did cause considerable consternation, however.

Byrne, having completed his studies, turned the project over to John Ryder, W8DQZ. Ryder was later to become Professor of Electrical Engineering at Michigan State University and the author of numerous electronics textbooks (he is not the Rider of *Rider's Manuals* fame). When Byrne left Ohio State University, he went to work at Bell Telephone Laboratories and gave up amateur radio. I cannot help but feel that amateur radio lost a valuable member by this decision.

The first use of the name "Windom" appears to be by the Wireless Institute of Australia which in 1930 published an article which was substantially a reprint of Windom's *QST* article. Shortly after World War II, the Radio Society of Great Britain ran an article, "Why Not a Windom?" It thus appears that the name Windom was imported into the United States from overseas. Rightly or wrongly, the name has stuck because it is much simpler than the more technical term. "off-center-fed Hertz antenna."

antenna theory

One of the most important characteristics of an antenna, when trying to couple energy into it, is its input impedance which, of course, is composed of both resistance and reactance The resistance and reactance both depend on a large number of factors



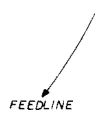


fig. 3. The reactance at any point along a half-wavelength antenna is zero. The situation is analogous to the parallel-tuned circuit of fig. 4.

such as height above ground, diameter of the antenna, nearby obstructions, and length of the antenna compared to a half-wavelength. While the input impedance can be calculated in the ideal case, the feedpoint impedance of an actual antenna must usually be measured, especially under amateur conditions.

Let's first talk about the antenna's reactance; to simplify things, we'll assume the antenna is exactly one-half-wavelength long or, as the old timers would say, "operating on its fundamental wave."

As is well known, the reactance looking into the center of a half-wavelength antenna is zero. Not as well known is the fact that the reactance at any other point on a half-wave antenna is also zero. Consider fig. 3 where the antenna feedpoint is off-center. Looking into the short end, we have a transmission line with the far end open-circuited; the reactance looking into an open-circuited transmission line is given by

$$X_C = -jZ_o \cot \theta \qquad \textbf{(1)}$$

Where

$\theta =$ the electrical length of the transmission line in degrees

$Z_o =$ the characteristic impedance of the line.

If the antenna is exactly one-half-wavelength long, the length of the long end will be $180° - \theta$ and its reactance will be

$$X_L = -jZ_o \cot (180° - 8) \qquad \textbf{(2)}$$

From trigonometry it can be shown that

$$\cot (180° - 8) = -\cot 8 \qquad \textbf{(3)}$$

regardless of the value of θ. Combining **eqs. 1,** 2, and 3 gives

$$X_L = -X_C \qquad \textbf{(4)}$$

As the reactances are equal in magnitude but opposite in sign, the input reactance is zero. Since this is true regardless of the length of the line (value of θ), the reactance will be zero regardless of where the tap is located, provided only that the antenna is exactly a half-wavelength long. This sitaation is analogous to the parallel-tuned circuit shown in fig. **4.** At frequencies below resonance, the inductive reactance will be less than the capacitive reactance so the current in the inductive branch will be greater than the current in the capacitive branch. Similarly, at frequencies above resonance, the current in the capacitive branch will be greater. At resonance the two currents will be equal. This fact provides an excellent means of determining the resonant frequency of any antenna that is an integral number of half-waves long, as discussed later.

We will now discuss the resistive component of impedance. For a half-wavelength dipole, the resistive component at the center is usually considered to be 72 ohms. This is true when the antenna is in free space or at certain heights above ground. The input resistance of the antenna at any location along its length can be easily determined in terms of its input resistance at the center. When the antenna is very thin and not terminated (as in a rhombic antenna), the current distribution along the antenna is essen-

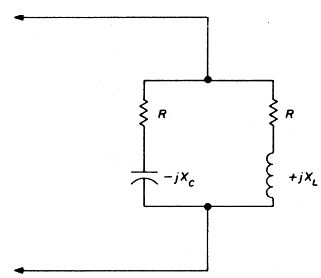


fig. 4. The parallel-tuned circuit analog of the off-center fed antenna.

tially cosinusoidal, as shown in fig. 5. The power applied to the center of the antenna is

$$P_o = I_o^2 R_o \qquad \textbf{(5)}$$

Where

P_o = power applied to center of antenna

I_o = rms current at center

R_o = radiation resistance at center

The power given by **eq. 5** must be equal to the power at any other point x along the antenna; therefore

$$I_o^2 R_o = I_x^2 R_x \qquad \textbf{(6)}$$

Since the current distribution is assumed to be cosinusoidal, ***I*,** $= I_o \cos 8$ where 8 is the distance along the antenna in electrical degrees from the center. Substituting eq. **6**

$$I_o^2 R_o = (I_o \cos\theta)^2 R_x$$

so that

$$R_x = \frac{R_o}{(\cos\theta)^2} \qquad (7)$$

From eq. 7 we see that *theoretically* the input resistance of a half-wavelength antenna goes from a nominal 72 ohms at the center to infinite ohms at the ends. In practice, the antenna current does not drop to zero at the ends, so the resistance does not become infinite; the resistance can become very high, however. If the characteristic impedance of the single-wire feeder is between 72 ohms and "very high," we should be able to find a point on the antenna that will match the feedline characteristic impedance.

transmission line theory

When discussing a transmission line of any type, probably the first question to be asked is what its characteristic impedance is (and how it is measured or calculated). With conventional two-conductor transmission lines, such as coaxial cable, one method to determine the characteristic impedance is to measure the impedance seen looking into a length of the line with the far terminals both open- and short-circuited; the square root of the product of these two measurements is the characteristic impedance of the line. However, in the case of the single-wire line, what do you short-circuit it to?

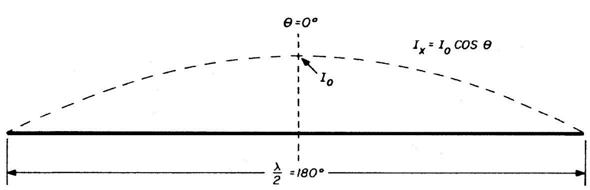


fig. 5. The current distribution along a half-wavelength antenna is essentially cosinusoidal as shown here.

All right, you say, if you can't measure it, can you calculate it? The usual way to calculate the characteristic impedance of a transmission line is to calculate its inductance per unit length and the capacitance between the conductors per unit length; then divide the inductance by the capacitance and take the square root:

$$Z_o = \sqrt{\frac{L \text{ per unit length}}{C \text{ per unit length}}} \qquad (8)$$

It is not difficult to calculate the inductance per unit length of a single-wire line, but the capacitance per unit length is another matter. The capacitance to what? There is no other conductor, so this leaves ground. However, the capacitance of the single-feed wire to ground can be expected to vary over a fairly wide range as the feedline leaves the station — relatively close to ground — and winds its way up to the antenna high in the air. And as the capacitance varies, so does the characteristic impedance of the line. Hence, the calculation approach is not practical.

table 1. Characteristic impedance of long wires.

conductor diameter mm	inches	AWG	characteristic impedance, Z_o (ohms) 3.5 MHz	7.0 MHz	14.0 MHz	28.0 MHz	56 MHz
12.5	0.500	—	560	518	475	435	393
6.5	0.250	—	600	560	518	475	435
3.3	0.128	no. 8	641	600	559	516	474
2.6	0.102	no. 10	654	613	572	530	490
2.1	0.080	no. 12	669	628	586	545	503
1.6	0.064	no. 14	684	643	600	560	517

This can be seen more graphically by taking a slightly different approach. Consider **fig. 6** which shows the single-wire feeder and its ground image; this converts the single-wire transmission line into a balanced line. This might be used to calculate the characteristic impedance of the line, but half the distance between the actual line and its image is in the ground whose dielectric constant and other electrical parameters are not accurately known. This makes it difficult, if not impossible, to accurately calculate the characteristic impedance.

Using the representation of **fig. 6** does point up the fact that the single-wire feedline can be explained as one-half of a variable spaced balanced line, with two wires of the balanced line being closer together at the bottom and gradually increasing in spacing as they approach the antenna. Tranmission lines of this types are known as "tapered lines." Since the wire diameter is the same all along its length, the characteristic impedance will gradually increase as the line approaches the antenna. This also means that, assuming the feedline is matched to the antenna, the current measured along the line will not be constant, but will slowly decrease as one moves toward the antenna. Therefore, the current measured at the input end of the feedline should be greater than the current measured at the antenna, even when the system is matched.

After explaining so carefully why it is so difficult to calculate the characteristic impedance of a single-wire feedline, I was rather nonplussed to find the following equation in an old publication:[7]

$$Z_o = 138 \left[\log \frac{0.56\lambda}{2\pi r}\right] \qquad (9)$$

Where

Z_o = characteristic impedance of single-wire line
r = radius of conductor
λ = operating wavelength

Table **1** gives several values of characteristic impedance, taken from reference 8. As no sources or references are given, I do not know what approximations or assumptions were made in this formula's derivation and can not attest to its accuracy. Since wavelength is included in the calculation, note that characreristic impedance varies with frequency!

The fact that it is not practical either to measure or calculate the characteristic impedance of the single-wire feedline does not mean that the line does not have a characteristic impedance. It does — we just don't know what it is. While this will certainly affect the procedure used to match the feedline to the antenna, it should not stop us from obtaining a match.

bandwidth considerations

Although the bandwidth of the Windom antenna has always been assumed to be large, I have never seen it discussed in the literature. My own experience indicates that while an off-center-fed antenna can be used over a relatively wide range of frequencies, it operates as a true Windom only over a very narrow bandwidth.

As shown in **fig. 4,** the Windom antenna is analogous to a parallel-tuned circuit The Q of a parallel-tuned circuit can be defined as the ratio of either the inductive or capacitive branch current (at resonance, the two are equal) to the line current. Therefore, the Q of a Windorn can be easily measured by using the circuit of **fig.** 7.

If the antenna operates as a true Windom over a relatively narrow bandwidth, why does the antenna work so well as a wide band antenna? Let's look at **fig. 8A.** As the frequency is increased above the

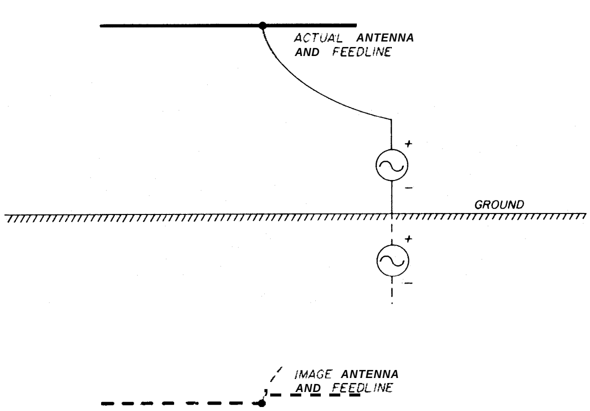


fig. 6. A single-wire feedline and its ground image.

Windom frequency, the current in the long end decreases while the short-end current increases. The long end loses its effectiveness as a radiator and the Windom antenna degenerates to a random-length, single-wire antenna with the single-wire feedline operating as part of the radiating system. Below the

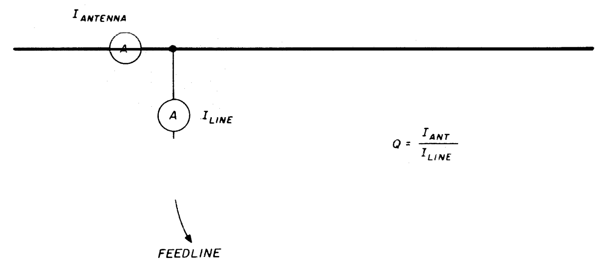


fig. 7. Test circuit for measuring the Q of an off-center-fed Windom antenna.

Windom frequency the current in the long end predominates and the antenna appears as in **fig. 8B.** Again the feedline will radiate.

harmonic operation

The Windom antenna has a theoretical advantage over the balanced, center-fed dipole because the Windom will resonate on even-order harmonics or, more accurately, on *approximate* harmonics. Because of the *end effect,* harmonic resonant frequencies are not integral multiples of the half-wave resonant frequency. An antenna that is half-wave resonant at *3.525* MHz, for example, will have harmonic resonant frequencies at *7.235,* 14.656, 22.077, and 29.498 MHz. Note that some of these frequencies are not in an amateur band. Thus, if multiband operation is desired, the antenna must be operated off-resonance on some bands and might not load up well.

As mentioned in the bandwidth discussion, the antenna may operate satisfactorily at other than its Windom frequency, but as a random-length antenna. It is also doubtful whether the feedline tie point will be sufficiently accurate on the harmonic bands to provide a good match. My own measurements indicate that it will not.

Those amateurs who are contemplating harmonic operation of the Windom antenna should read Wrigley's excellent article on harmonic operation of dipoles.8 My own experience with the Windom, though limited, bears out Wrigley's comments. The antenna will operate on harmonics, probably not as a Windom, but rather as a random-length antenna.

efficiency and radiation

The final factors I will discuss are the efficiency and radiation of the single-wire feedline. I have not attempted measurements of this type myself, but will mention the results from reference 1 which describes the work done by J. D. Ryder and E. D. Shipley. Ryder and Shipley report that for a 365-meter (1200-foot) feedline driving a 15-meter (50-foot) antenna,

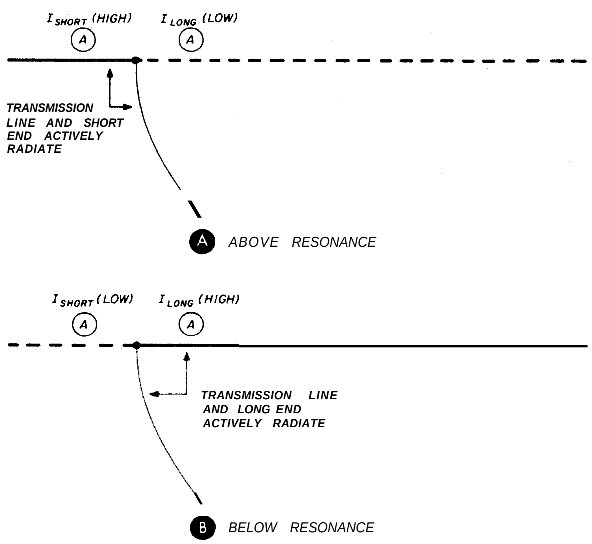


fig. 8. Operation of the Windom antenna above resonance, A, and below resonance, B. Above resonance the current in the longer leg loses its effectiveness as a radiator because rf current decreases and the antenna degenerates to an end-fed single-wire antenna with the feedline as part of the radiator. Below resonance the rf current in the short end decreases and the antenna again operates as a random-length end-fed antenna.

all the measured losses in the system could be accounted for by the *PR* loss of the feedline. Radiation losses were too small to measure. The accuracy of the power measurements was estimated to be within 4 per cent. Therefore, it appears that the radiation from a single-wire feedline operating at its matched (Windom) frequency is negligible, while at other frequencies feedline radiation may be appreciable.

adjusting the antenna

The biggest problem I found with Windom's QST article was his failure to recognize the very complex variation of antenna impedance with height and local ground conditions; he assumed that the feedline tap could be placed on the basis of distance measurements alone. This myth has been perpetuated by the ARRL *Handbooks* and *Antenna Manuals,* and by the *Radio Handbook* as well.

With this in mind, the most important question facing the amateur who installs a Windom is "How do I adjust the antenna?" Basically, there are two adjustments to make: 1) the antenna must be cut to the desired frequency, and 2) the single-wire feedline must be connected to the proper place on the antenna to provide a good match for the transmission line.

The first problem must be solved first – the antenna must be cut to the desired frequency (or the actual resonant frequency of the antenna must be accurately known). This is necessary to insure that the antenna will present a pure resistive load.

The method for finding resonance recommended by Everitt-Byrne proves to be very simple, yet very exact. As discussed earlier for the analog with a parallel-tuned circuit, when the tuned circuit is at resonance, the currents in the inductive and capacitive branches are equal. In the Windom the rf currents can be measured by placing rf ammeters in the short and the long sides of the antenna as shown in **fig.** 9. The antenna is resonant at the frequency where the rf currents in the two ends are equal.

If resonance at a specific frequency is desired, begin by cutting the antenna about 1 meter (3-1/2 feet! too long (on 80 meters). Set the transmitter to the desired frequency and prune the ends of the antenna until the current in the two sides is equal. I found I had a strong psychological urge to cut the antenna length from the end with the larger current; actually, the antenna may be cut from whichever end is more convenient.

If a knowledge of the actual resonant frequency of the antenna is all that you want, connect the ammeters as shown in **fig. 9** and vary the transmitter frequency until the two currents are equal.

From a practical point of view, the situation is not as simple as depicted in **fig.** 9. The stresses in a wire antenna are considerable and would probably pull most meter cases apart. Therefore, a means must be devised to take the mechanical stresses off the meters. I mounted the two meters in a piece of plexiglass and suspended the plexiglass from the antenna with spring clips of the type used at the end of dog leashes. The antenna is broken with an insulator and the single-wire feedline is supported by a second insulator. The photographs show the plexiglass meter bracket and how it is suspended from the antenna. I

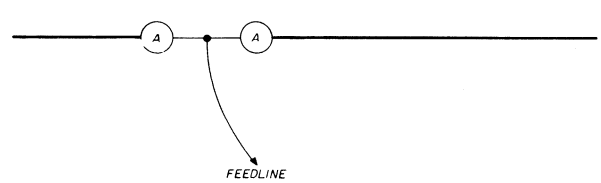
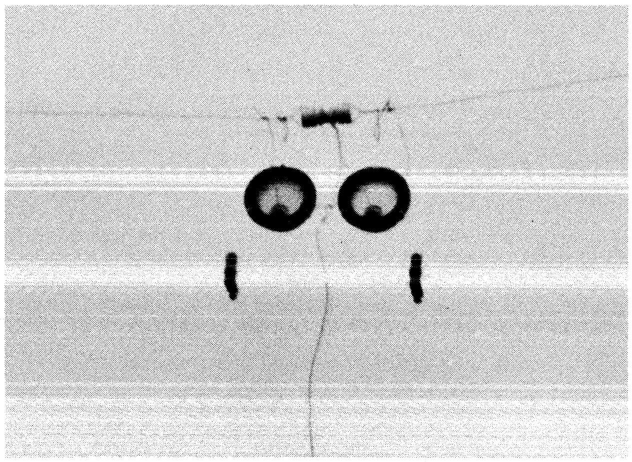


fig. 9. Test circuit for determining resonance of an off-center-fed antenna. When the rf current in the two ends is equal, the antenna is resonant. If the current in the long end is higher than in the short side, the antenna is too long; if the short side current is greater, the antenna is too short.



The test meters installed in the antenna as seen from the ground. The antenna is not quite at resonance since the currents are not equal. This was taken with a 200mm lens.

used l-ampere meters with an output power of about 90 watts. The resonant currents were 0.76 ampere on 80 meters. This value will depend on how far off-center the feedline is initially attached.

When measuring the rf current, if the current in the short side of the new antenna is lower than in the long side, the antenna is too short; if the long side current is greater, the antenna is too long. See fig. **10.**

I highly recommend that the transmitter power be brought up slowly and that you have a friend watch the meters to make sure they don't go off scale; thermoammeters are not noted for their tolerance to overloads. The meters, of course, must be read from a distance. I used a 25-power spotting telescope. Higher magnification might be better, depending on how high your antenna is — try to pick a windless day so the meters don't bounce around while you try to focus on them!

Another possibility is a telephoto lens on a 35 mm single-lens reflex camera. However, I found my 135 mm lens was much too short to do any good; I estimate a lens 300 mm or longer would be needed if your antenna is exceptionally high.

Using this method of adjusting the antenna length, I was able to notice a change of as little as 5 to 7 cm (2 to 3 inches) near the resonant point or a frequency change of less than 10 kHz (on 80 meters). The primary factor in limiting the accuracy of this method is the problem of reading the meters at great distances. Much greater accuracy could be obtained by using remote thermocouples — mount the thermocouples on the antenna and run leads to the meters on the ground — or better yet, at your operating pcsitioa. Unfortunately, remote-reading rf ammeters are very expensive and hard to find!

It is interesting to note that the two-meter method of determining antenna resonance is not limited to the Windom antenna; it can be applied to other dipoles as well. Simply short-circuit the center insulator, insert a second insulator at some convenient place, install an off-center feedline and the meters, and check. The method can also be used to determine harmonic resonant frequencies.

I am surprised that the antenna manual publishers have not presented this procedure before; it is relatively simple and straightforward and has been available since 1929. To the best of my knowledge, the only discussion of this technique to appear in the amateur literature, since Windom's 1929 article, was written by Paul Rockwell, W3AFM, in 1963.[9]

adjusting the feedline

After the antenna has been cut to the proper length, the correct feedpoint can be determined. At first glance this appears to be a formidable problem since we only approximately know the characteristic impedance of the feedline and only approximately know how the radiation resistance varies along the antenna. However, we do know certain characteristics of transmission lines which are helpful; namely, that when a transmission line is terminated in its characteristic impedance, 1) the input impedance of the line is a pure resistance, and 2) the current or voltage is constant along the line (neglecting the impedance taper effect along the line). Either of these facts may be used to find the proper feedpoint.

It is absolutely necessary to make all feedline tests

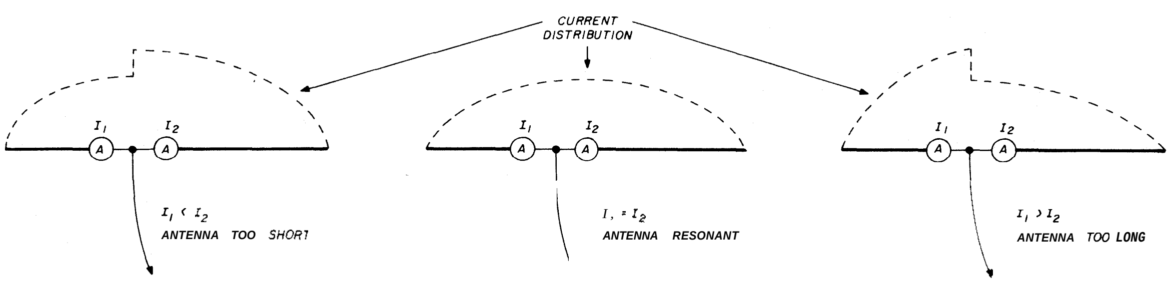


fig. 10. Determining resonance of the off-center-fed antenna with ammeters in each leg. Current distribution is shown by dashed lines.

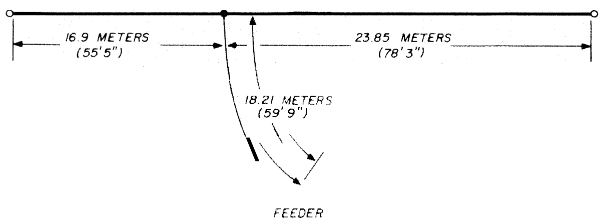


fig. 11. Dimensions of the experimental Windom antenna built by K4KJ. Resonant plots of this antenna for 80, 40, and 20 meters are shown in figs. 12, 13, and 14, respectively.

at the resonant frequency of the antenna. Only at this frequency will the input resistance of the antenna be purely resistive.

For my own tests I used a General Radio 916A rf impedance bridge to measure the input impedance of the line.[10] This measurement could also be made with an RX noise bridge such as the one described in reference 11. Simply measure the input impedance of the line at the antenna resonant frequency, and adjust the feedline tap along the antenna until the reactance component of the input impedance is zero.

If an rf impedance bridge is not available, try the method recommended by the old timers: if the feedline is on the order of a half-wavelength long, insert four rf ammeters in the lower half of the line and adjust the tap for identical current on all four meters. In this case, it will be assumed that the impedance taper along the lower half of the line is negligible so that line current will be constant.

If rf ammeters are not readily available, try soldering neon bulbs along the feeder and adjust the tap point for a constant brightness. This is best done at night, but be prepared for startled neighbors!

results

My Windom antenna was configured as shown in **fig. 11.** With this arrangement the results of my 80-meter measurements are plotted in **fig. 12** which

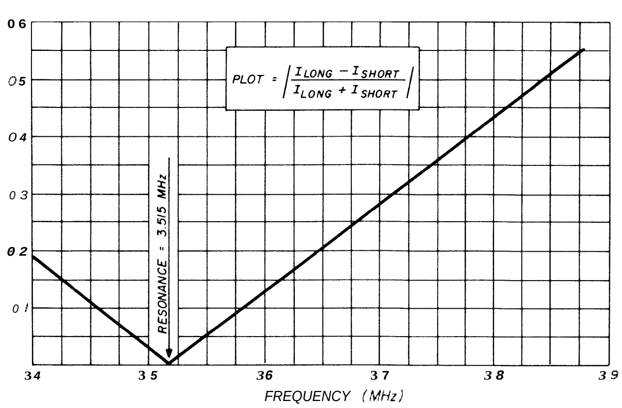
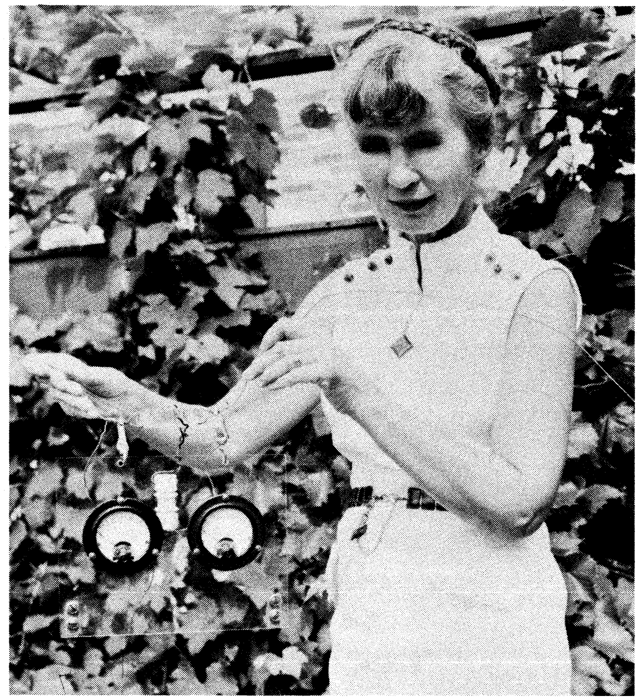


fig. 12. Resonance of the 80-meter Windom antenna of fig. 11, as determined by the method described by Everitt and Byrne (see text).

shows the difference between the antenna current in the two legs. To remove any variations due to changes in rf power input, I referenced everything to the sum of the two currents. Taking the absolute magnitude of the differences eliminates the problem of algebraic sign.

$$plot = \left| \frac{I_{long} - I_{short}}{I_{long} + I_{short}} \right| \tag{10}$$

The resonant frequency is where the plot equals zero. **Figs. 13** and **14** show the same factor for the same antenna on the 40- and 20-meter bands, respectively. Notice that the resonant 20-meter frequency is above the amateur band.



K4KJ's wife holding the plexiglass meter panel as installed in the antenna. The feed-line runs out the bottom.

The input impedance to the feedline is shown in **fig. 15** for 80 meters; the input impedance is anything but constant. The reactance is zero a little below 3.525 MHz at which point the input resistance is 590 ohms. **Fig. 16** shows the same factor over the 40-meter band. Note that the zero reactance point falls above the band limits at which frequency the input resistance is more than 1000 ohms. This indicates to me that the feedline is not matched at any frequency. The input impedance on 20 meters is graphed in **fig. 17** and the same general comments apply.

conclusions

As a result of my experiments, I have reached the following conclusions concerning the off-center-fed antenna:

1. The Windom antenna is very simple and economical to build and, if the proper procedures are used, relatively simple to adjust.

2. The bandwidth as a true Windom antenna is relatively narrow The antenna will operate relatively well, over a much wider frequency range as a random-length antenna, however.

3. Based on the above data, it is my opinion the Windom antenna can be properly matched for only one amateur band. On other bands, the comments of item 2, above, apply. After reviewing the comments of Windom and Ryder to an early draft of this article, however, I am not convinced that the tap is at the optimum point. It may be possible to improve the match on 40 meters.

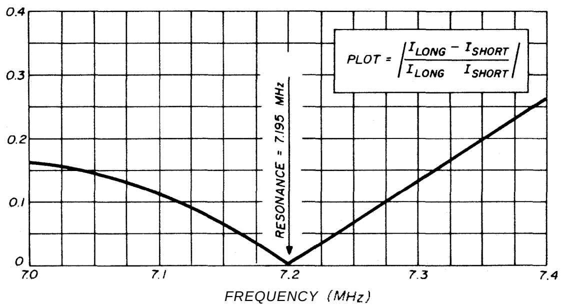


fig. 13. Resonant curve of the Windom antenna of fig. 11 on 40 meters.

4. Those amateurs whose antenna requirements are modest and whose circumstances dictate a simple and inexpensive antenna will continue to find that the off-center-fed antenna will serve their needs.

acknowledgement

Some of the historical data on the Windom antenna was obtained from Paul Rockwell, W3AFM, who very kindly made his files available to me. I would like to thank Messers Everitt, Ryder, Byrne, and Windom, who very kindly made extensive comments on a draft of this article. In addition, Professor Ryder provided previously unpublished technical data which space does not permit me to include.

postscript

The Windom antenna, one of the oldest antennas developed for amateur use, has had a complicated history, one as interesting as the theory of the antenna itself. And the people who have contributed to its development are equally intriguing:

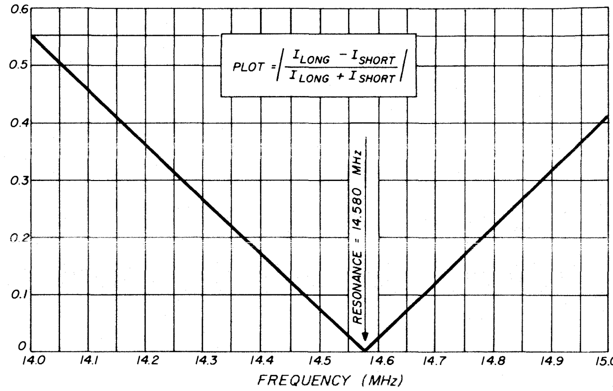


fig. 14. Resonant curve of the Windom antenna of fig. 11 on 20 meters. Note that the antenna is resonant above the amateur band.

Frank Conrad, 8XK, who started it, is called the father of broadcasting. His amateur station became KDKA in Pittsburgh, the first broadcasting station in the country.

Vernon D. Landon, one-half of 8VN, went on to become an eminent scientist with RCA. He has contributed widely to the advancment of electronics with his many very readable papers published in the *Proceedings* of the *IRE* and in the RCA Review.

Loren "Windy" Windom, 8ZG, for whom the antenna was named, did not, oddly enough, pursue electronics professionally, but became a lawyer in Ohio. He is still an active amateur.

John Ryrne, who with his thesis partners, E. F. Brooke and A. B. Crawford, did much of the actual work on developing the antenna, became an outstanding research engineer and educator.

John Ryder, now K4IHX, who also worked on the antenna as a thesis project, was to become a well-

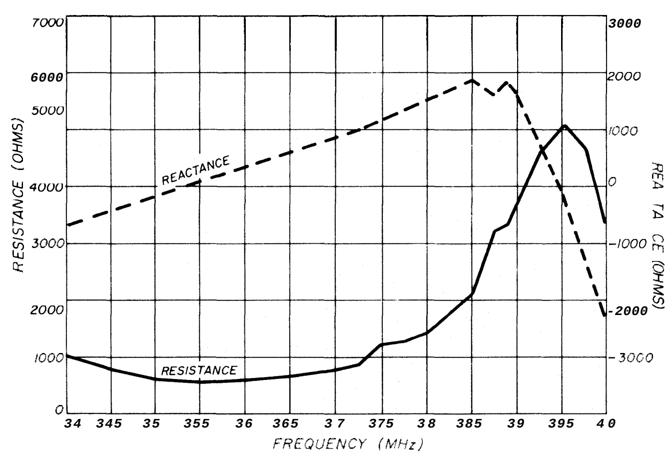


fig. 15. Input impedance of the single-wire feedline to the Windom of fig. 11 on 80 meters, as measured with a General Radio 916A rf impedance bridge. The reactance is zero at 3.520 MHz.

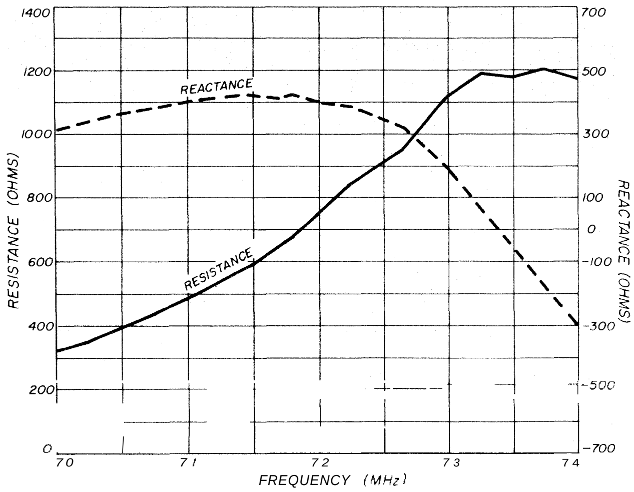


fig. 16. Feedline input impedance on 40 meters as measured with an rf impedance bridge. The reactance is zero at about 7.338 MHz; the resistance at this frequency is more than 1000 ohms.

known educator, an author of several electronics textbooks, and a president of the IRE. He is an active amateur in Florida.

William L. Everitt, as faculty advisor to the students who worked on the antenna, contributed much to its development. Everitt began his amateur career in 1914 as 2ABI; in 1921 he became 8CRI. When he decided to go into communications professionally, he dropped amateur radio because he did not want to have the same vocation and avocation. Everitt was later to become a prominent author and educator. He retired as Dean of Engineering at the University of Illinois and is now Dean Emeritus at that university.

Unfortunately, Dr. Everitt has been ill-treated by the amateur community. His work on the off-center-fed antenna is largely unknown and certainly unrecognized. Unhappiiy, the same thing happened to him a second time. Everitt was the first to describe the use of a pi network as a coupling device. He published this work in the Proceedings of the IRE in 1931[12] and in *Communications*;[13] the pi network is also described in his book, Communication *Engineering*.[14] Arthur Collins, W9CXX, of Cedar Rapids, Iowa, recognized the advantages of the pi network to couple the output stage of a transmitter to a transmission line, and used it in his transmitters. This application played an important part in establishing the reputation that Collins equipment will load up to "anything." Collins described the pi network to the amateur community in a *QST* article[15] and in a similar article *Radio*,[16] and the network became known by old-timers as the "Collins Coupler" instead of, perhaps, the "Everitt Easy Loader."

I believe these are excellent examples of how people who have made notable contributions to the advancement of electronics developed their interest in electronics through amateur radio. It would be interesting to be able to look into a crystal ball to see how the many young people who are today joining the ranks of amateur radio through high school science classes, or as CBers, will go on to make significant contributions to future electronics.

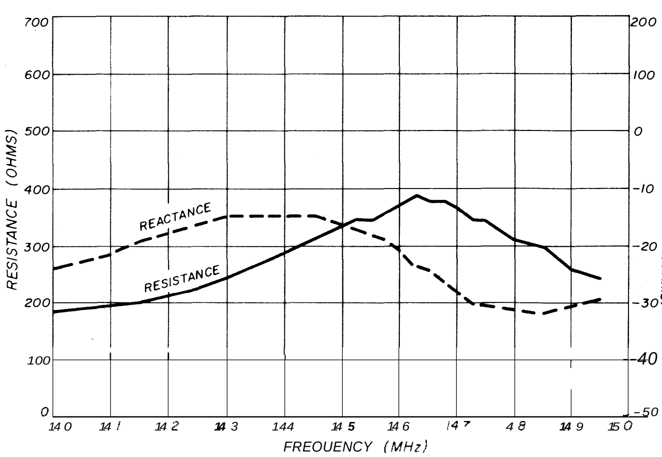


fig. 17. Feedline input impedance on 20 meters as measured with rf impedance bridge. The reactance never falls to zero over the frequency range from 14 to 15 MHz.

references

1 W. L. Everitt, arid J F Byrne, "Single-Wire Transmission Lines for Short-Wave Antennas," *Proceedings of ?he IRE*, October, 1929, page 1840 (A more detailed version of this paper was published by The Engineering Experiment Station, Ohio State University as Bulletin No 52. February, 1930.1

2. L G Windom, W8GZ/W8ZG, "Notes on Ethereal Adornments," *QST*, Vol. XIII, No 9, September, 1929, pp. 19-22

3. V. D and E. B Landon, 8VN, "Capacity Coupling to Operate the Antenna at its Fundamental," *QST*, June, 1923, page 22.

4. Howard M Williams. 9BXQ, "The Hertz Antenna at 20 and 40 Meters," *QST*, July, 1925, page 24.

5. Robert S. Kruse, "Feeding the Antenna," *QST*, July, 1926, page 11

6. John D. Kraus, *Antennas*, McGraw-Hill, New York, 1950, page 147.

7. *The Radio Antenna Handbook*, edited by Frank C. Jones and W. W. Smith, 1936, page 32.

8. William B Wrigley, "Impedance Characteristics of Harmonic Antennas," *QST*, February, 1954, page 10.

9. Paul Rockwell, W3AFM, "Further on the Windom," 73, August, 1963, page 76.

10 John J Nagle, "Impedance Measurements at RF," *CQ*, Vol. 30. No. 11, November, 1974, pp. 18-24

11. R. A. Hubbs, W6BXI, and A F. Doting, W6NKU, "Accurate RX Noise Bridge." *ham radio*, February, 1977, page 10.

12. W. L. Everitt, "Output Networks for Radio-Frequency Power Amplifiers," *Proceedings of the IRE*, May, 1931, page 725.

13. W. L. Everitt, "Coupling Networks," *Communications*, September, 1938.

14. W. L. Everitt, *Communication Engineering*, McGraw-Hill, New York, 1937, page 263.

15. A A Collins, W9CXX, "A Universal Antenna Coupling System for Modern Transmitters," *QST*, February. 1934, page 15.

16 A A Collins, W9CXX, "The Answers to the Flood of Inquiries Regarding the Collins Antenna System," *Radio*, March, 1934, page 5.

ham radio

selective receiving antennas:
a progress report

An active receiving antenna system useful from 75 through 10 meters capable of directing a null toward the interference source

The concept that man-made interference is a receiving problem that can best be handled by improved receiving techniques rather than increased power, voice processing, and similar transmitting enhancements, can hardly be disputed. While the high-power advocate may seem to be solving his own difficulties, he can become a major part of the problem to the rest of us.

What to do? One can't always outshout the opposition. No amount of racket on the frequency can prevent signals from point A reaching point B. Only conditions of propagation can do that. What such racket does do, however, is drown out and prevent copy of the signals from A, unless means can be devised at B to make the interference self-destruct. The signal *strength* produced by a given receiving antenna is of little importance compared to the signal-to-noise ratio it delivers. And you'd better believe that all forms of man-made interference must realistically be classified as noise!

Therefore, under interference conditions, it behooves us to consider the use of a separate specialized receiving antenna system. If its signal level production should be less than that of the transmitting antenna, the difference can usually be made up by suitable amplification, without loss of signal quality.

This article describes an active receiving antenna system, useful from 75 through 10 meters, with the ability to direct a null toward the source of the interference. The rejection of unwanted signals that can be obtained can often make the difference between solid copy and losing the battle with the interference.

design consideration

What sort of antenna is indicated for this application? Obviously the major requirement is not forward gain but the exact opposite: a broad coverage with a deep null in the response pattern, which can be directed toward rhe source of interference. Furthermore, you can get many more dB difference in a null than in forward gain. Thus, our problem is to devise a simple inexpensive system having just those characteristics.

The familiar phased array with quarter-wavelength spacing, although capable of an excellent null, is far too cumbersome as a specialized receiving antenna for the lower frequency bands. If its elements should be brought close together for compactness, mutual coupling between the elements increases to a point where phasing and power distribution go haywire and complicate the problem beyond reason.

What to do? In the first place, coupling between the antenna pickup elements should be minimized by making the elements nonresonant. Secondly, an isolating preamplifier following the antenna probe will ensure independence of each unit. Thus, each pickup element acts as a probe in the electrostatic field of the passing wave front.[1]

In an earlier antenna study, a wide range phase control (phasor) was developed for pattern control of

By **Henry S. Keen, W5TRS, Fox,** Arkansas, 72051

a pair of vertical antennas.* This phasor, which provided 180° of continuous phase control, was later incorporated in a feasibility study directed originally at the 40-meter foreign broadcast problem.3 The resulting conclusion was that a useful degree of rejection was possible and that further study would be justified.

All-band system. In this effort, signals from two small, fixed, active antennas with close spacing had been combined in a quadrature hybrid, with control of phase and amplitude. It soon became apparent that, because of the close spacing, 180° phase control was overkill, and the single-band frequency limitations imposed by the phasor and hybrid were undesirable. Therefore, attention was directed toward the development of an all-band receiving antenna system without these restrictions.

Preamplifiers. The first requirement was the development of improved preamplifiers to be used with the antenna probes. They must be wide band, have low noise response, and be stable. In addition to having high gain, they must also have very high input impedance, compatible with the tiny antenna probes with which they would be used (2 foot or 0.6m whips).

To obtain this high impedance, an MPF102 fet is employed as a source follower, with feedback to the bias network. Although this type of input stage may seem to lose signal, the actual power gain is considerable as the impedance level is reducec! to several hundred ohms. This stage in turn drives a bipolar transistor stage, which is coupled to the coaxial output line through a 3:1 stepdown transformer trifilar wound on a T50-6 toroid core.

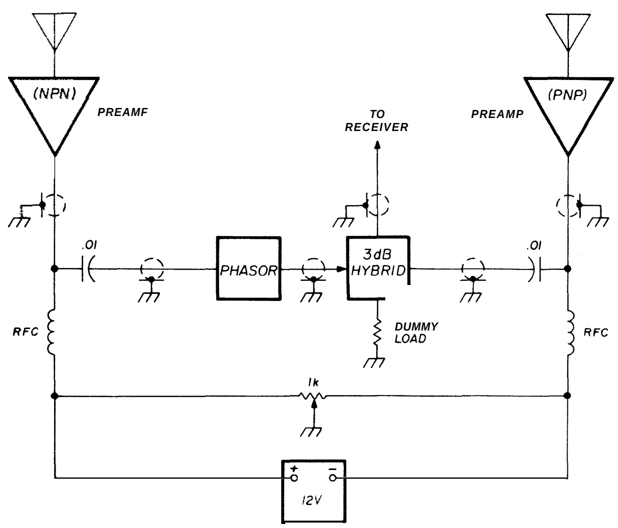


fig. 1. Method of balancing gain in the signal preamplifiers. Balance potentiometer bridges the dc supply.

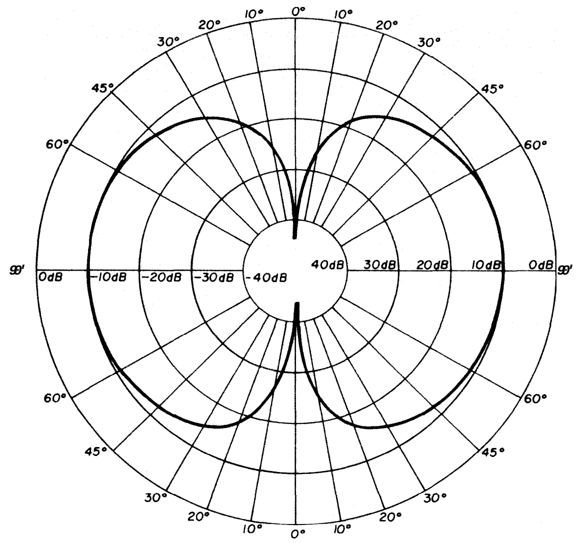


fig. 2. Antenna Figure-8 pattern, calculated for 0.05-wavelength spacing between antennas.

Various means were tried to balance the signal amplitude of one preamplifier against the other, but the method finally chosen was that which had been used in reference 3; that is, by controlling the dc voltages applied to each one **(fig. 1)**. In this method the two preamplifiers are in series, one being fed a positive voltage to its npn output stage while the other gets a negative voltage to its pnp output stage.

Identical lengths of coaxial cable were used between the antenna modules and the signal balancing device, so that phasing was obtained mechanically by rotating the boom upon which the modules were mounted. This eliminated the phasor and resulted in a Figure-8 pattern, **(fig. 2)**.

Antenna pattern. The very sharpness of the null offers both advantages and disadvantages. The rejection of a local amateur signal, for example, was a walloping 60 dB, assuming 6 dB per S-unit. In a crowded area where one has problems from "the ham down the street" this arrangement may offer much-needed relief. Rejection of skip signals was less effective, as the multipath propagation effects accompanying such signals introduced variations in both phase and amplitude relationships of the signals. This had the effect, particularly on distant signals, of making the *direction* of arrival appear to vary.

It would seem that a reduction in the sharpness of the null would improve this situation. Analysis of the Figure-8 pattern indicates that signals that deviate from the direction of the null will vary as the sine of the angle of deviation. A cardioid pattern, on the other hand, **(fig. 3)** should give a response based upon the delay between the time when a signal strikes the first antenna and when it reaches the second. This parameter should vary as the cosine of the angle of deviation, and is therefore less critical.

Signal combiner. Many different methods of combining signals from the two preamplifiers were also tried and compared. For example, a CA3028 differentiai amplifier was built with a baiancea output transformer, but was found to be unsatisfactory. Spurious signals caused by IMD made it clear that signals from the antenna modules would have to be balanced out before further ampiification could be used. Too much continuous amplification, particularly in a very broad spectrum system such as this, pushes the final stage beyond linearity. Signal reduction inherent in the phasing-out process keeps the amplitudes within limits.

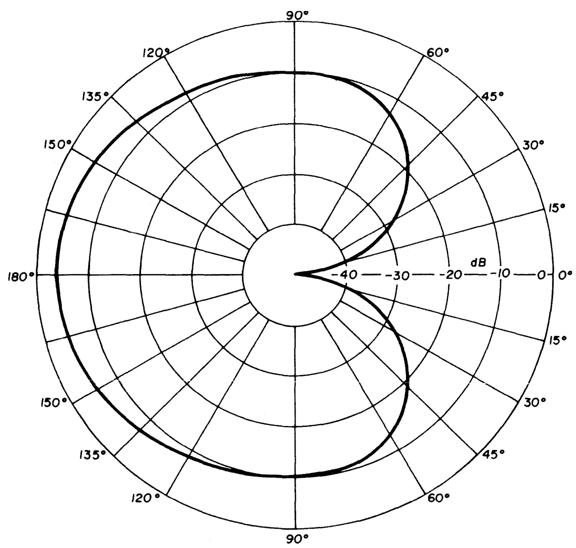


fig. 3. Cardioid pattern, calculated for 0.05-wavelength antenna spacing.

The most effective combiner, other than the frequency-limited quadrature hybrid previously used, was a toroid input transformer with separated primary and secondary windings. A bifilar primary was used for improved balance. The five-turn secondary was close wound and separated as far as possible from the primary (fig. 4).

Each coaxial line from the antenna modules is terminated by a 51-ohm resistor to reduce mismatch reflections that might affect the phase of signals presented to the differential transformer.

Common amplifier. The amplifier that follows the differential transformer was originally a broadband unit similar to the preamplifiers. It was found that spurious IMD signals were present, particularly on 20

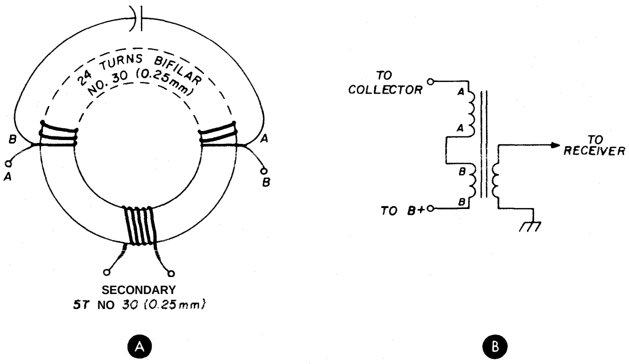


fig. 4 Method of constructing transformer T2 (A) and schematic of output transformer T3 (B).

meters, caused by out-of-band commercial stations causing overloading. At the same time, the overall gain left something to be desired. The cure to both problems was to place a tuned circuit ahead of the common amplifier. The Q of the tuned circuit raised the gain to an acceptable level, while at the same time rejected signals responsible for the IMD.

For operation on different bands some means must be provided for bandswitching this tuned circuit. This is what you must do with your transceiver, so it doesn't represent any additional operating hardships.

Phasing method. The boom upon which the antenna modules were mounted was approximately two meters in length — an arbitrary choice. To obtain a cardioid pattern, the coaxial line from one module to the common amplifier should be longer than the other by approximately 2/3 of the spacing between the antenna probes (allowing for the velocity of propagation in the coaxial lines). Once adjusted, phasing should be independent of frequency, right? Well, not exactly, for allowance is made for the fact that signals arrive at different angles from the ionosphere, and a minor compensation should be provided. Either the phasing extension of the coaxial line can be shortened, or the boom can be tilted to correspond to the angle of arrival of the signal. A compromise angle of 25 degrees was chosen.

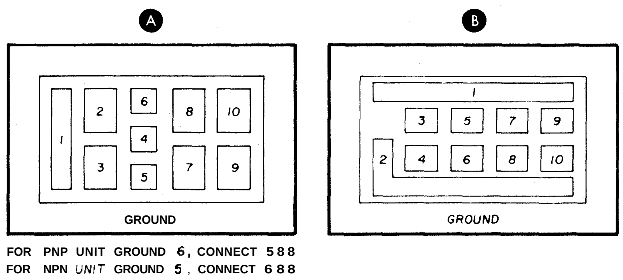
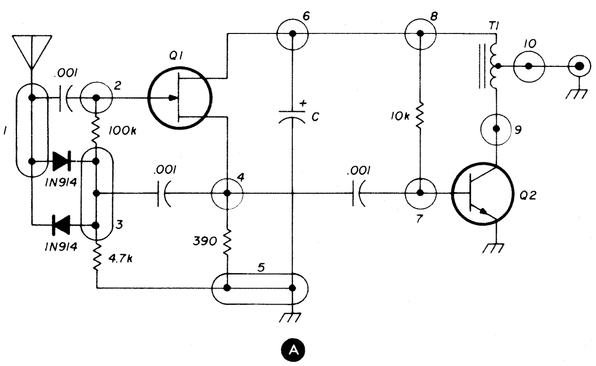
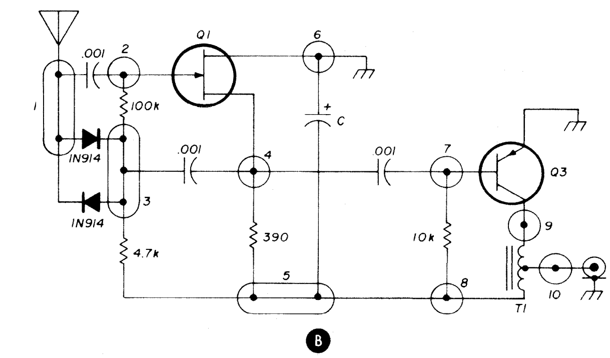


fig. 6. PC-board layout for the preamplifiers (A), and common amplifier (B).

Strictly speaking, signals will arrive at different angles on different frequency bands, with different heights and with different conditions of propagation. However, the exact angle does not seem to be too critical. Phase differences in the two preamplifiers will likely have a greater effect. With the Figure-8 pattern, there should be two nulls 180° apart. If not, the change to the cardioid pattern will not be very effective.

To be able to use either pattern, the extension





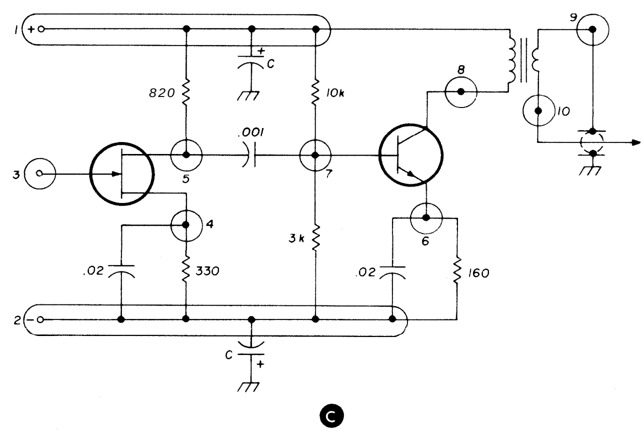


fig. 5. Amplifier schematics with PC-board mounting keys. The npn preamp and pnp preamp are shown in (A) and (B); the common amplifier is shown in (C).

should be an extra piece, which can be added to either of the feed lines. The angle of signal arrival does not affect the Figure-8 pattern.

As with all antennas, the higher and more in the clear you can get it, the better. Other antennas or structures in the vicinity can distort the pattern obtained.

construction

The amplifier modules were built on PC boards as shown in **figs.** 5 and ***6.*** Transistor sockets were used to mount the fets, which were inserted after all connections were completed. Although sockets may be dispensed with, if care is used in the assembly process, the ability to quickly select and substitute individual fets to help balance the gain of each module is worth the extra trouble. Where balance is as important as it is in this application, bargain-basement transistors are not recommended. Procedures in making PC boards have been well covered in the literature and are not repeated here. Perf board may make an equally effective substitute if you're not too fussy.

The preamplifier transformers use 22 turns, trifilar wound, of no. 30 (0.25mm) enameled wire on a T50-6 core; the windings are connected in series. Toroids are conveniently mounted by inserting a small rolled cylinder of paper through the center and applying a drop or two of cement in the right places. The preamplifiers are mounted in Miniboxes with a porcelain feedthrough antenna terminal on one end and a UG-625/U BNC connector at the other. Because of their light weight they can be supported by the antenna and output leads.

The common amplifier floats across the power supply, operating from the full 12 volts, neither terminal of which is grounded. Note that the output transformer of this amplifier has an isolated secondary winding, rather than being an autotransformer, as used in the preamplifiers.

The differential input transformer and the tuned circuit combination is placed in a shield box, made by soldering together bits of scrap PC board. At present, operation on 20 and 40 meters is covered by the tuning range of the trimmer capacitor, although a complete bandswitching arrangement is scheduled for the near future. The coil could be tapped, or shunt capacitors or inductors switched in for operation on the other bands. The overall schematic of the entire system is given in fig. **7.**

preliminary tests

The common amplifier is first checked out by connecting a receiver to its output terminal and a signal generator to first one input terminal and then the

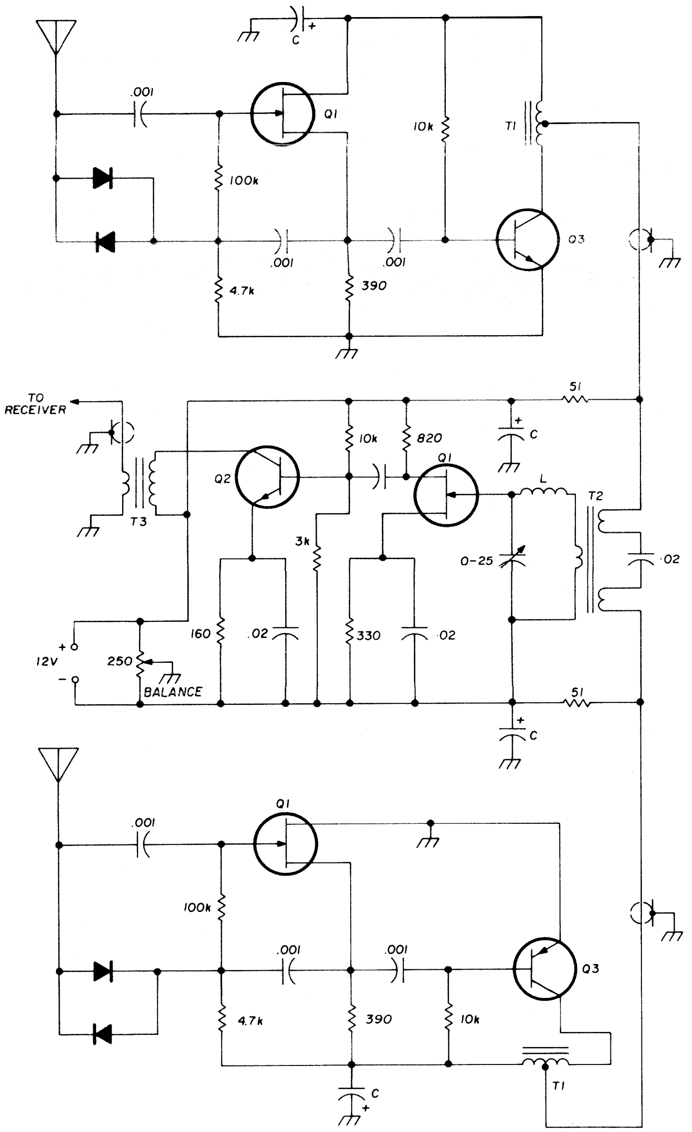


fig. 7. Overall schematic of the selective receiving system

other. Use a series capacitor to avoid shorting the dc on the input connectors. The tuned circuits should be checked to ensure peaking on each band.

Dc voltages on the input connectors should be checked as the balance potentiometer is rotated. And don't forget to mark the polarity at each connector so that the right preamplifier will be connected to each.

An antenna preamplifier is now connected to the proper input terminal with a random length of coaxial cable. A short piece of wire clipped to the antenna connector should pick up strong signals. Repeat the test with the other preamplifier.

When both antenna modules have passed this preliminary test, you are ready for the final checkout. Both antenna modules are connected to the common amplifier with equal lengths of coaxial cable. Their Miniboxes should be clipped together and the antenna terminals fed through a capacitive tee junction, as in **fig. 8.** The two antenna terminals should not be directly connected together because of the dc potential difference between the npn and pnp preamplifiers.

The signal thus injected into each preamplifier is identical in both phase and amplitude, and the balance potentiometer is adjusted for minimum signal. To get a comparison first adjust for minimum signal then disconnect one of the preamplifiers. A signal increase of from 40 to 50 dB should be obtained. If not, you will have to do some troubleshooting. Ideally the dc voltages at balance should be equal on each preamplifier, but 2:1 may be par for the course.

This null depth test should be made on all bands where the antenna system will be used. If you can't get a deep null at this stage, you are not likely to get it later.

Having passed the null depth test, the system is ready to be installed. Mark the coaxial lines and amplifier terminals, if it hasn't already been done, so that there will be no polarity mixup. I found during the null depth tests that rejection was better if the outer conductors of the coaxial lines were tied together at intervals of several feet (about 1 meter). All these things are better done on the ground before the assembly is raised.

operation

Tuning up and adjusting a system of this kind can be a frustrating experience if you don't go about it the right way. For example, you tune in a good strong signal and start making adjustments only to find that it's in an ssb net, operating VOX, and the signals bounce back and forth so fast that you don't know which one you are hearing. Or else rapid fade makes you think you have a null when you haven't. The foreign broadcasters are particularly troublesome in this respect. If you try watching the S-meter while tuning the antenna, that's no answer either. The trick is to reduce the receiver rf gain enough to disable the agc, and then you can do it by ear. A CW or SSTV station may be a better choice. You may also find that a test signal fed into a separate antenna is a useful aid in adjustment. A TV birdie sometimes provides a good test signal.

The CW man can buy a great deal of interference relief with appropriate "filter systems, but the ssb operator, with his wider bandwidth requirement must resort to other means. Some little operating tricks are helpful. For example many 40-meter hams, operating lower sideband among the broadcasters, will operate several hundred to a thousand Hz

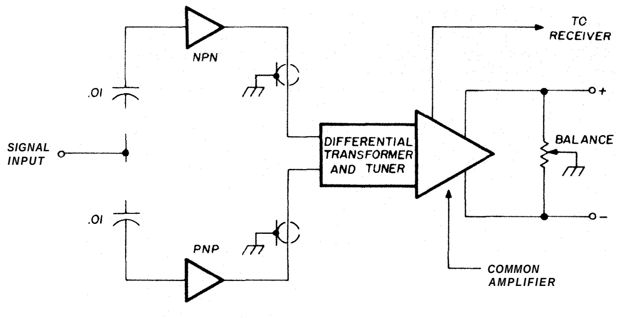


fig. 8. Setup for the null-depth test,

below beat, letting the steep skirt of the crystal filter take out the heterodyne and much of the monkey chatter. If the broadcaster is not too strong, zero-beat works fairly well, particularly with the receiver's rf gain turned down so that the agc doesn't fight you. Often overlooked in interference situations is the simple expedient of switching sidebands. After all, what is so sacred about lower sideband on 75 and 40 or upper sideband on the higher frequency bands? Frequently it's not only easier to switch sidebands than to hunt for a clear frequency but there's a lot less chance of losing your contact in the process.

closing remarks

In our battle for better radio contacts much development work remains to be done. The directional receiving antenna, of which this system is only one possible variation, appears to be a promising area for study. It is still in the experimental stage and there are probably better ways of doing it. Perhaps we owe it to ourselves to investigate some of them.

The major problem in this approach is the method by which we balance out one signal against the other. Alternative methods, such as the phasor-hybrid combination of reference 3, worked very well on a single band. Another method that worked

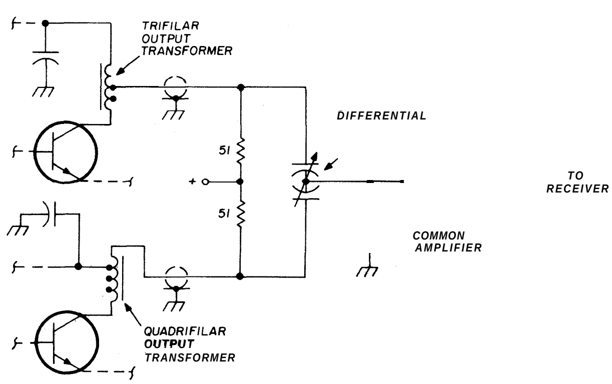


fig. 9. Alternative balancing method, which has been tested. The balancing potentiometer is replaced by a differential capacitor.

quite well was to obtain a phase reversal in one of the preamplifiers by means of a reversed winding on one of the output transformers, coupled with a differential capacitor ahead of the common amplifier **(fig. 9).** This circuit acted as an amplitude control and eliminated the need for the npn-pnp combination of preamplifiers with the balancing potentiometer. It is probably of equal merit with the system described in this article, but most of us don't have differential capacitors in our junk boxes.

Yet another possibility, which has not been tested, is shown in **fig. 10.** In-phase components would be out of phase when arriving at resistor R, while out-of-phase components would add, as in a lattice filter. It would seem that the value of R would have to be quite low for good rejection, but the tuning network should retrieve a useful part of the loss.

No single piece of equipment can solve all of our problems, and this is no exception. There are times when it is very impressive, as well as times when conditions of propagation make it less effective. Ad-

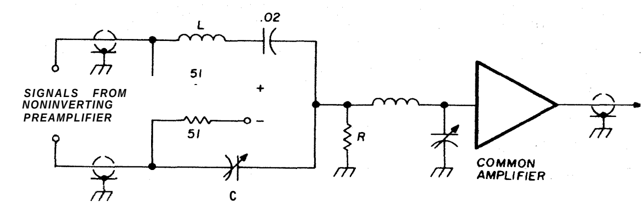


fig. 10. Another alternative balancing system, which has not been tested. In-phase signals would be out of phase at resistor R, while out-of-phase signals would add. The tuning network should compensate for signal loss because of the low value of R.

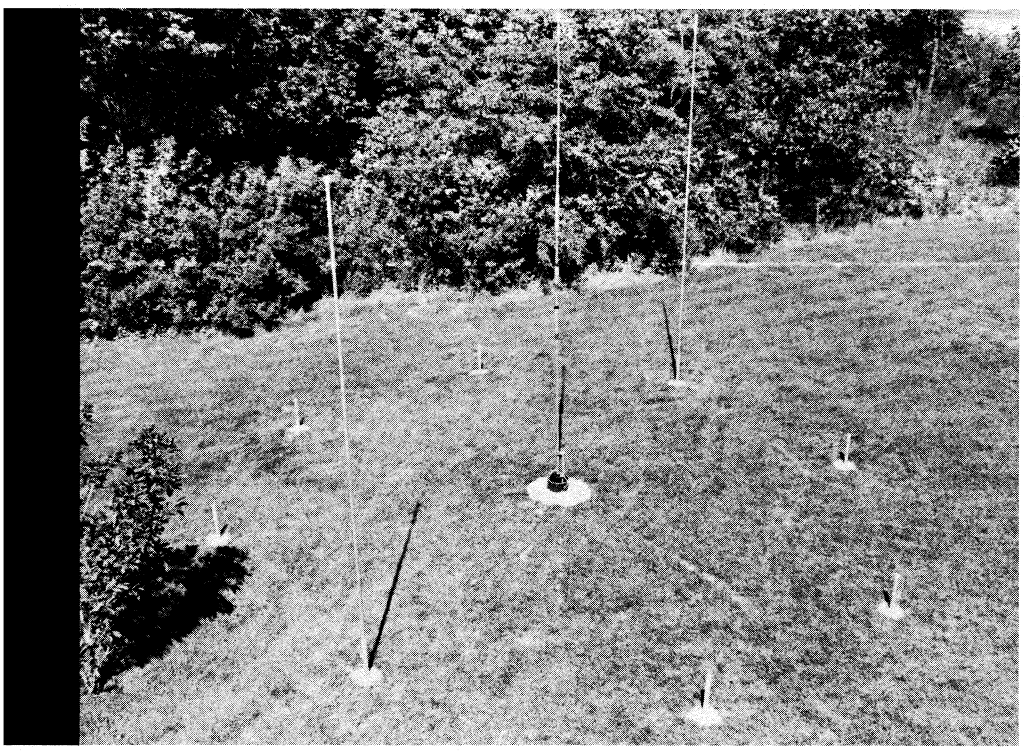
vantages of as much as 30 dB have been obtained when conditions were favorable. This has the effeet of reducing kilowatts to watts, a worthwhile gain. During a contact, when someone opens up too close to your frequency, it's most gratifying to turn the receiving antenna and hear the interfering signal fade into the background. It doesn't always happen that way as the interference may be in line with your contact; however, it's a satisfying feeling when it does.

The forthcoming WARC confab is no guarantee that we'll be any better off when it's over. Can we afford to just sit around and wait?

references

1. Jim Fisk. W1DTY, "The Voltage Probe Antenna," *ham radio,* October, 1970, page 20.
2. H. S. Keen, W5TRS, "Electrically Controlled Phased Array." *ham radio,* May, 1975. page 52.
3. H. S Keen, W5TRS, "Selective Antenna System," *ham radio,* May, 1976, page 28.

ham radio



multiband vertical antenna system

A good ground system and parasitic elements enhance the performance of a commercial trapped multiband vertical

This project started out as a simple installation of a Hy-Gain 18AVT/WB vertical antenna. The 18AVT/WB is a trapped vertical for 80-10 meters that uses top-loading to achieve 80-meter operation. The ground system began as a single 2-1/2 meter (8 foot) ground rod driven in at the base of the antenna with the base of the antenna mounted 30 cm (12 inches) above the ground. The performance was good enough to arouse my curiosity to see what improvements I could experiment with to make the system work even better.

Since a vertical antenna is no better than its grounding system, that was a good starting point. I began by researching the *ARRL Handbook*[1] and *QST* for grounding information. K4ERO's article on ground-

By Ladd Seaberg, WØNCU, Route 1, Atchison, Kansas 66002

ing systems for vertical antennas? provided the information I was looking for.

The ground system grew slowly, from four radials, each 10.7 meters (35 feet) long, to the final system of 72 radials shown in fig. **1.** Theoretically this configuration has a 2 dB power loss on 80 meters at 0.125 wavelength for 24 radials; a 1 dB loss on 40 meters at 0.2 wavelength for 60 radials, and slightly less than 1 dB loss on 20, 15, and 10 meters. As explained by K4ERO in his article, it does little good to increase the radial length unless you also increase the number of radials.

The radial system was surveyed using a 1910 vintage transit with a large magnetic compass. In my location, magnetic north is 8 degrees east of true

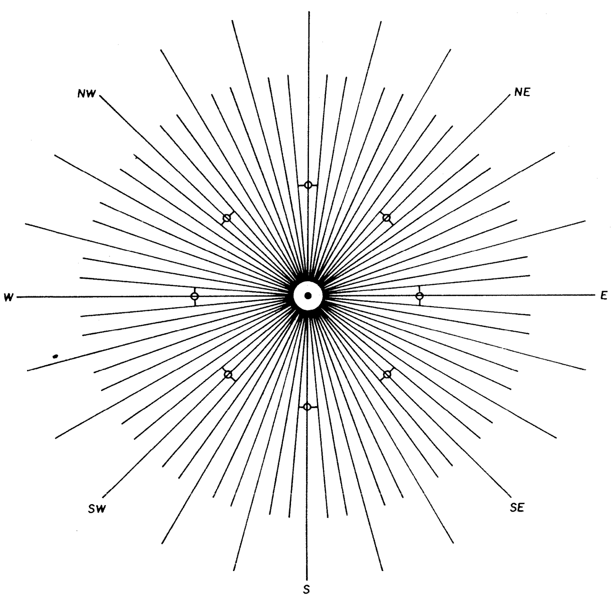


fig. 1. Layout of the radial system used with WØNCU's vertical antenna system for the five high-frequency amateur bands. The system uses a total of 72 radials: 24 are 11 meters (35 feet) long; the other 48 are 8.5 meters (28 feet) long.

north. Eight degrees was subtracted from magnetic headings so the cardinal radials are on true compass headings, 45 degrees apart.

Long before I finished the ground system, the idea of making the antenna system directional kept popping into my mind. I looked into both phasing[3,4] and parasitic arrays.[5] I wanted to keep the system simple but I didn't want to lose its multiband capabilities. I finally settled on a three-element parasitic array; it would be directional on 20, 15, and 10 meters and built in such a way that I could remove the director and reflector for omnidirectional use on 80 and 40 meters.

construction

The ground system was built from scrap no. 14

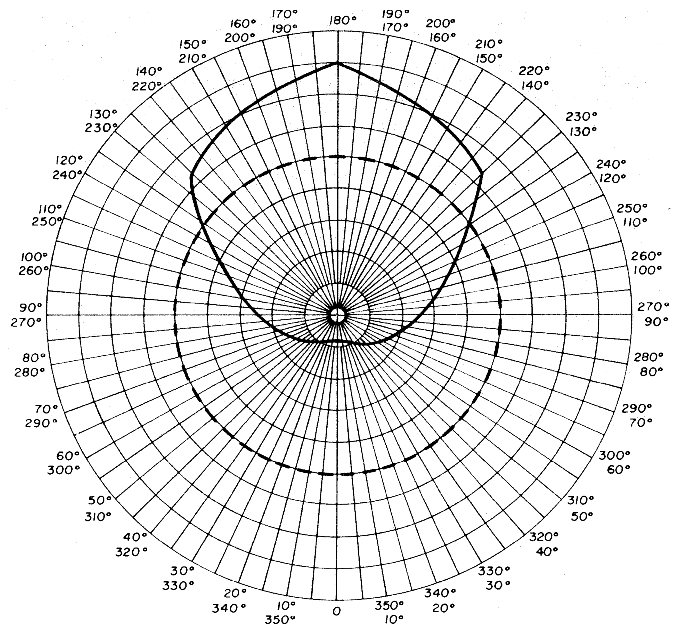
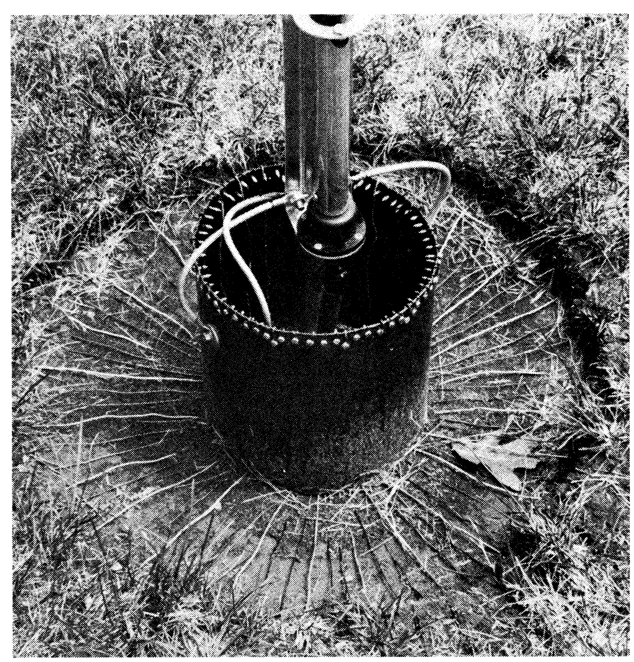


fig. 2. Relative field strength of the three-element vertical beam on 20 meters. The dotted line shows the radiation of the center vertical by itself. Front-to-back ratio is 20 dB; gain is 4.1 dB.

(1.6mm) and no. 16 (1.3mm) copper wire. An edging spade was used to cut a small slit in the grass; then the wire was pushed down into the slit. Using this method, I didn't have to dig up the whole yard. All the wires were terminated on a copper pipe, 30 cm (1 foot) in diameter and 30 cm (1 foot) tall which was lo-



Base of the driven element showing the length of large diameter copper pipe used for terminating the radials. Narrow slits were cut in the lawn to place 72 radials around the base of the antenna.

After the antenna was installed and tuned, the base was covered with plastic sheeting and crushed stone to give it a nice appearance.

cated at the base of the antenna. (The copper pipe was a scrap piece from an old vodka still at the distillery where I work as a chemical engineer; perhaps this antenna should be called the "moonshine vertical!")

The reflectors and directors are built of 25 mm (1 inch) aluminum conduit with 19 mm (3/4 inch) aluminum conduit slipped inside to make a nice fit. The eight ground array supports are made of 3.2 cm (1-1/4 inch) galvanized conduit. They are 1.2 meter (4 feet) long with 75 cm (30 inches) driven below ground. A 6.5 mm (1/4 inch) hole is drilled in the supports at ground level to drain out water. The array elements slip easily into the ground supports; good electrical connection is made with a 6.5 mm (1/4 inch) bolt in the elements which slips into a spade lug on the ground support (see photograph). A wing nut is used to quickly tighten the connection. The spade lug on the ground support is in turn wired to the radial running immediately below it and to the two adjacent radials. The connections below ground are wrapped and soldered.

The RG-8/U coaxial feedline is in a conduit which runs underground approximately 11 meters (35 feet) from the shack and comes up through the copper ground-wire termination pipe.

tuning and measurements

The central driven element was tuned for minimum swr on all bands starting with 28 MHz and working down to the lower frequencies, one band at a time. The antenna dimensions are quite different than those recommended by Hy-Gain because the grounding system changes the resonant characteristics of the antenna.

After this radiator was tuned up, the 20-meter elements were installed on the ground supports. The director was raised and lowered for maximum gain as

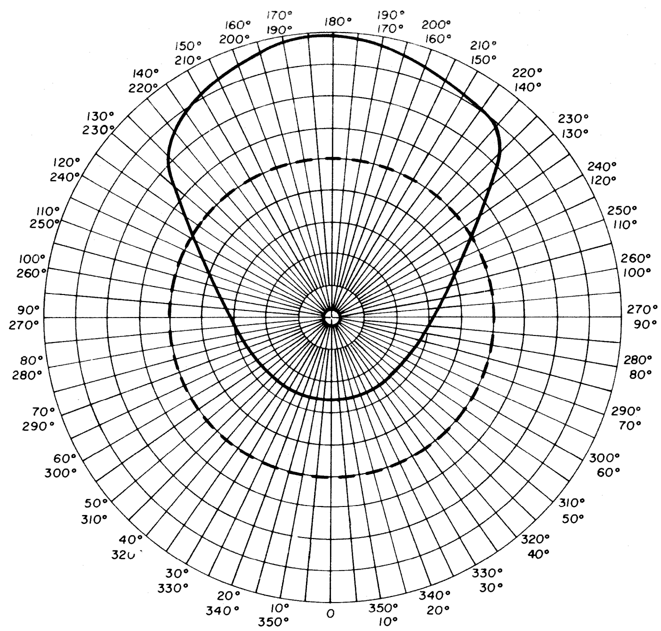


fig. 3. Radiation pattern of the three-element vertical beam on 15 meters, as determined by field strength measurements. The dotted line shows the radiation pattern of the driven element by itself. Front-to-back ratio is 11 dB; gain is approximately 5 dB.

indicated by a field-strength meter; the reflector was raised and lowered for maximum null. The front-to-back ratio was 20 dB as indicated by the field-strength measurements, with 4 dB gain over the vertical without the parasitic elements. This was later confirmed by W9HF when I repeatedly turned the

table 1. Dimensions of the multiband vertical beam and its performance characteristics with the radial ground system shown in fig. 1. The base of the driven element is 30 cm (12 inches) above ground. The height of the antenna on 40 meters is 5.11 meters or 16 feet, 9 inches (from antenna base to bottom of the top hat); the height of the antenna on 80 meters is 6.41 meters or 21 feet, 1/2 inch (from base of antenna to antenna tip).

band	director length	driven element	reflector length	element spacing	front-to-back ratio	gain
20 meters	4.67 m (15′4″)	3.71 m (12′2-1/4″)	5.59 m (18′4″)	0.18λ	20 dB	4.1 dB
15 meters	2.90 m (9′6″)	2.58 m (8′5-3/4″)	3.81 m (12′6″)	0.27λ	11 dB	5.0 dB
10 meters	2.13 m (7′)	2.10 rn (6′10-5/8″)	2.74 m (9′)	0.36λ	6.4 dB	5.7 dB

beam to and from him as he gave me reports.

The distance from the driven element to the parasitic elements is 3.8 meters (12-1/2 feet) as recommended by W2FMI for his 20-meter vertical beam.4 With this element spacing fixed, the front-to-back ratio falls off as the frequency is increased, but the gain over the single vertical increases.

Thus, the front-to-back ratio is 11 dB on 15 meters with 4.95 dB gain over the vertical. The front-to-back ratio on 10 meters is 6.35 dB with 5.68 dB gain over the center vertical by itself (see table 1). Even though this performance is a compromise between

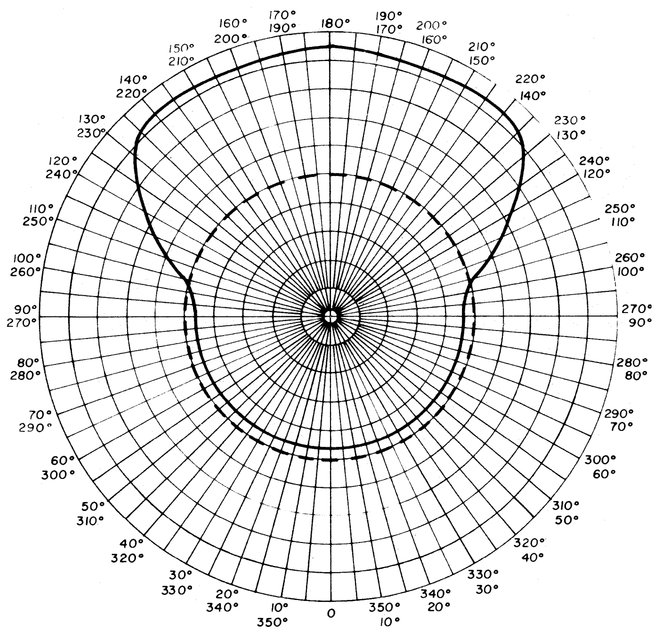


fig. 4. Relative field strength of the three-element vertical beam on 10 meters. Front-to-back ratio is 6.4 dB; gain is about 5.7 dB.

bands, it works out nicely since selectivity is more important for a crowded 20-meter band, and gain is more important for 10-meter DX work. (Comparative field-strength measurements for 10, 15, and 20 meters are shown in **figs. 2,** 3, and **4.**) The 10- and 15-meter elements were tuned in the same manner as the 20-meter elements using trial and error while taking field-strength measurements.

WBØSOT helped man the station while I took the field measurements and made adjustments to the elements. When Joe went back to college in the fall, I ran 75 meters (250 feet) of extension cord from my operating position out to the measurement site so I could key the transmitter and take field-strength readings. The readings were taken with a Heath HD-1426 relative field strength meter located about 50 meters (175 feet) from the antenna.

It is possible to build a combination antenna which works well as an 80- and 40-meter vertical and as a

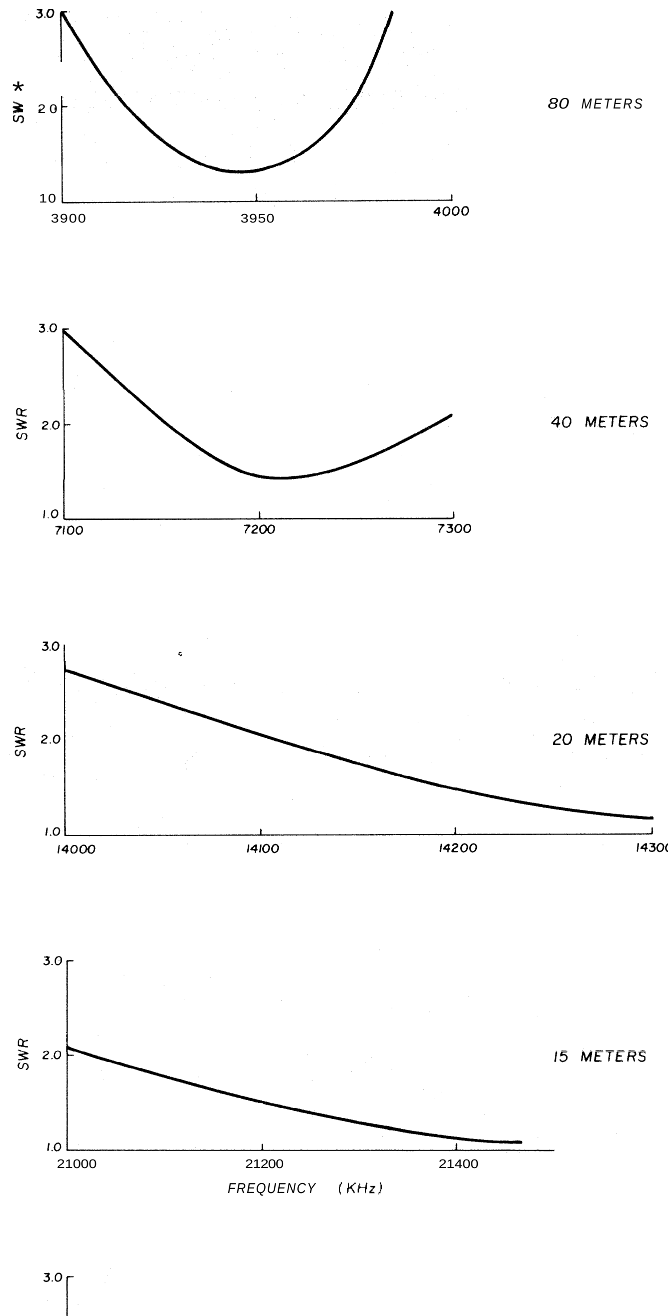
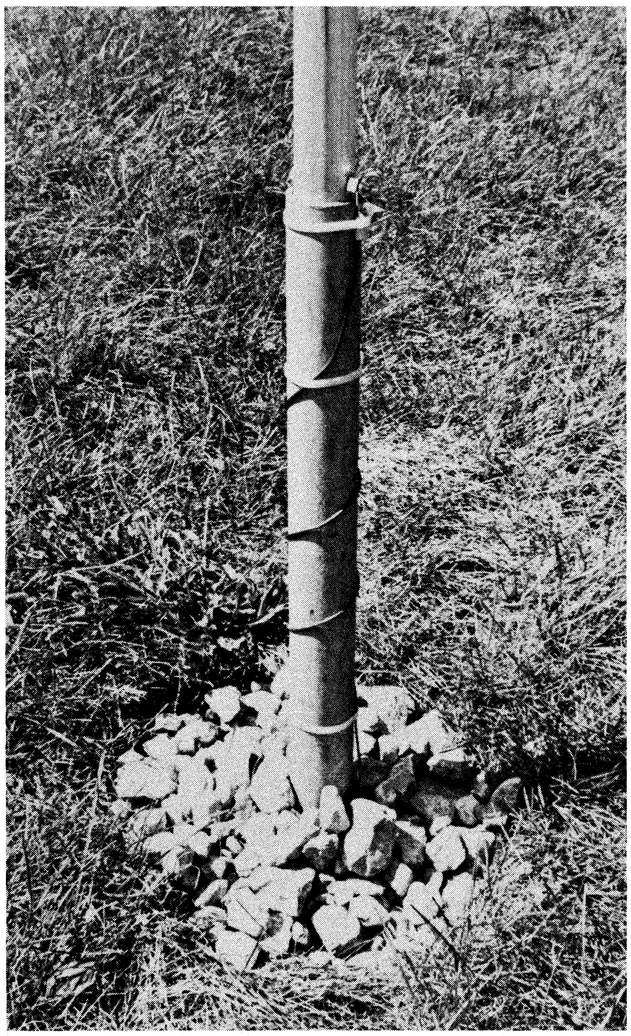


fig. 5. SWR curves for the multiband vertical antenna for each of the five high-frequency amateur bands. The antenna system could also be resonated for use on the CW ends of the bands if desired.

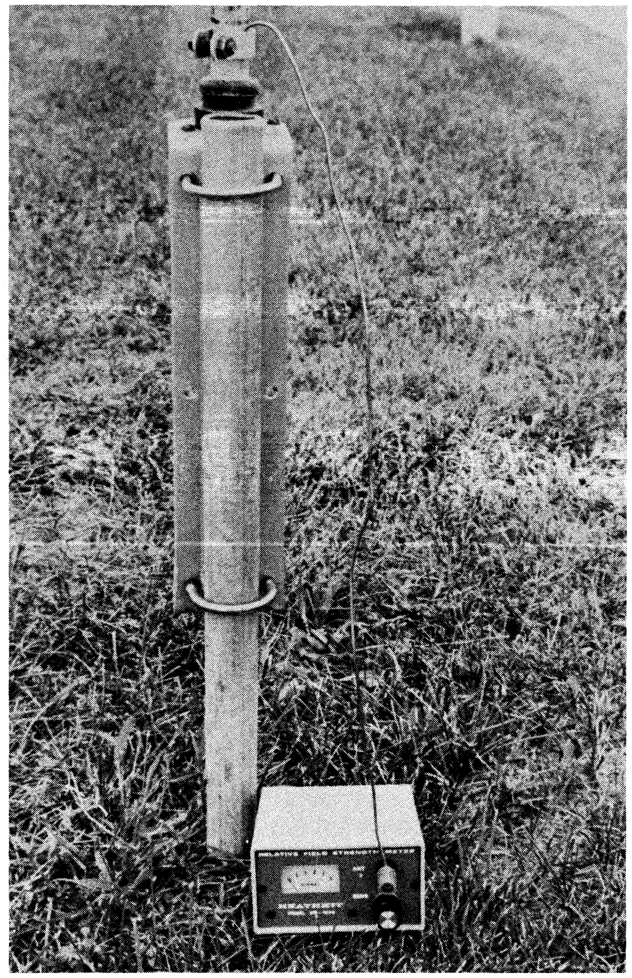
20-, 15-, and 10-meter rotatable vertical beam. It competes with the towers even though it is very close to the ground. In my opinion, the low-loss radial ground system is responsible for its good performance. Since there are no conspicuous towers, esthetically the antenna rates very high.

Many times I use the antenna as a single vertical, even on 20, 15, and 10 meters. If I hear weak DX stations, I install the director and reflector in the appropriate direction. It only takes three minutes to go from the shack to the garage, where I have a rack for the elements, out to the antenna, and back to the shack. The beam can be moved only in multiples of 45 degrees However, this works nicely because the direction beamwidth is approximately 80 to 90 degrees.

In the near future I plan to remove the top hat section and instal! a 40-meter trap. ! wi!! then add another section (approximately 6 meters or 20 feet) for 80 meters. The completed antenna would be guyed at the 40-meter trap level with nylon ropes. This modification would provide greater bandwidth on 80 meters. I would also like to try traps in the director and reflector.



One of the ground mounts for the parasitic elements. The wire wound around the mount is soldered to three radials under the ground; it is attached to the base of the parasitic element with a wing nut.



Set up for making field-strength measurements using a Heath HO-1426 field-strength meter. The field-strength antenna was located 53 meters (175 feet) from the antenna site.

I was recently talking to a New Zealander on 10 meters who said, ''It sounds like a tidy little system. You can take down the director and reflector and play croquet or soccer right there on the lawn." I assured him that my antenna was very versatile.

If you talk to someone who says, "Standby while I run outside to move my director and reflector," maybe you've found someone else crazy enough to try a multiband beam only 30 cm off the ground! The antenna has been fun to work with and I welcome any suggestions for improvement.

references

1 *The* Radio Amateur's Hanclbook, 54th edition, ARRL, Newington, Connecticut 06111, 1977.

2. John Stanley, K4ERO, "Optimum Ground System for Vertical Antennas," *QST,* December, 1976, page 13.

3 Richard Fenwick, K5RR, and R.R. Schell. "Broadband. Steerable Phased Array," *QST,* April, 1977, page 18.

4. Private correspondence from Hy-Gain Electronics Corporation, May, 1977.

5. Jerry Sevick. W2FMI, "The W2FMI 20-Meter Vertical Beam," *QST,* June, 1972, page 14.

ham radio

antenna bridge calculations

Using a hand-held programmable calculator to increase the utility of the RX noise bridge

A true picture of antenna impedance is important for maximum radio communications capability. The RX Noise Bridge is the most effective and simple measurement tool for the task.[1] Whether you use this or any other instrument, the final key to successful measurement is understanding and applying the complex impedance data.

The Smith chart is the traditional graphic tool for data reduction. It provides easy visualization of impedance and vswr as well as telling you what happens at the other end of the transmission line.[2] The modern programmable pocket calculator and basic transmission line formulas, however, will yield the same information with ease and better accuracy.

*Nationwide discount house price for the HP-25 was $100 in early 1978

Using all of these tools results in a better understanding of your antenna and transmission line.

Hewlett-Packard HP-25 calculator programs are provided for all formulas. Other types may be programmed by following the equations. A programmable calculator is recommended as a basic station tool; the HP-25 was considered the most cost effective and easiest to use."

RX noise bridge review

The basic circuit of the RX noise bridge is shown in **fig. 1.** A noise source serves as a wideband, untuned generator, and the null detector is a frequency-accurate receiver. Balance occurs when parallel resistance and reactance of both arms are equal. Signal output is then at a minimum. A fixed capacitor at the unknown connection allows the variable capacitor to balance with both inductive or capacitive reactances.

The RX noise bridge actually measures admittance but is calibrated in terms of resistance and reactance. This is possible from duality in expressing series and parallel forms of impedance. Admittance is a parallel of conductance G and susceptance B while impedance is a series of resistance R and reactance X. Admittance is the complex inverse of impedance and vice-versa. In addition, the reciprocals of G and B can be used to express admittance as parallel resistance and reactance. The fundamental relations are

$$Z=R+jX \qquad Y=G+jB \qquad Z=\frac{1}{Y}$$

By **Leonard H. Anderson,** 10048 Lanark Street, Sun Valley, California 91352

Remember that the "j" signifies the imaginary part of the complex quantity pair. Impedance can be expressed in terms of conductance and susceptance by

$$Z = [G/(G^2 + B^2)] - j[B/(G^2 + B^2)] \qquad \textbf{(1)}$$

These are awkward to handle in actual values so we can use fundamental identities of $R_p = 1/G$ and $X_p = 1/B$. The "p" subscript signifies *parallel* resistance and reactance. The equation can now be expressed as

$$Z = \left[\frac{R_p X_p^2}{R_p^2 + X_p^2} \right] - j \left[\frac{R_p^2 X_p}{R_p^2 + X_p^2} \right] \qquad \textbf{(2)}$$

Note the sign of the imaginary terms. At first glance, this might appear that a capacitive susceptance has an inductive reactance dual and vice-versa. Not so. A capacitive susceptance and inductive reactance are positive quantities, inductive susceptance and capacitive reactance are negative.

It does not matter which way the parallel susceptance/reactance sign is used: as long as a parallel capacitor has a series capacitor dual and a parallel inductor has a series inductor dual, the noise bridge can be calibrated to your choice. The RX noise bridge by W6BXI and W6NKU uses a negative inductive susceptance calibration and this is carried through in the calculator programs which follow.

simplifying complex number operations

Here is where the scientific calcultor shines; the ability to convert from rectangular form to polar form and back again makes things easy. **Equations** 1 and 2 are unwieldy in that the same values must be reentered more than once for conversion of parallel to series. A better way is to use polar form in division and multiplication. If:

$$Z_a = Mag_a \angle Pha_a \quad \text{and } Z_b = Mag_b \angle Pha_b$$

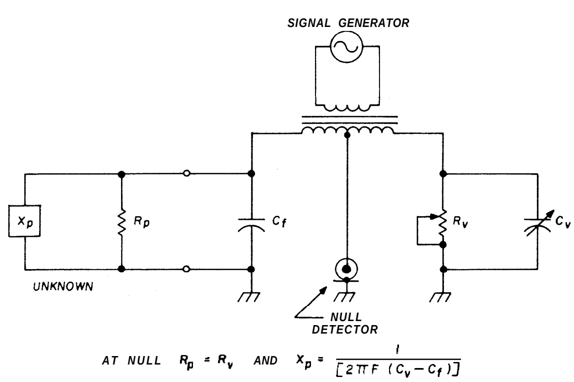


fig. 1. Basic diagram of the RX noise bridge. A complete description of this simple but useful instrument is given in reference 2.

where Mag is the magnitude and Pha is the phase angle, then

$$Z_a \cdot Z_b = (Mag_a \cdot Mag_b) \angle (Pha_a + Pha_b) \qquad \textbf{(3)}$$

$$\frac{Z_a}{Z_b} = \frac{Mag_a}{Mag_b} \angle (Pha_a - Pha_b) \qquad \textbf{(4)}$$

Another useful property to be used later is

$$\sqrt{Z_a} = \sqrt{Mag_a} \angle \frac{Pha_a}{2}$$

If your familiarity with rectangular and polar forms of complex numbers is a bit rusty, the rules are:

$$Z = R + jX \text{ or } Mag \angle Pha$$

where

$Mag = \sqrt{R^2 + X^2}$	(magnitude)
$Pha = Arctangent\ (X/R)$	(phase angle)
$R = Mag * Cosine(Pha)$	(real part)
$X = Mag \cdot Sine(Pha)$	(imaginary part)

A real and imaginary part is the rectangular form while a magnitude and phase angle is the polar form. They are just different ways to express the same thing. A calculator can convert forms in a single keystroke.

It may not be clear how all this will help you, so let's examine the problem of finding the parallel resistance and reactance after you have read the bridge dials. The steps are:

1. Calculate conductance and susceptance.

2. Enter G and B into the calculator in proper order.

3. Convert to polar form using the built-in function.

4. Invert the magnitude and change the angle sign.

5. Convert to rectangular form using the built-in function.

6. Read out real and imaginary part values R and X.

Measured values are entered just once. The calculator does the rest.

Still unconvinced? Suppose you get a bridge reading of 20 ohms R_p and −10 ohms for X_p at a particular frequency. Since these are parallel values, invert 20 ohms for 0.05 mho conductance and −10 ohms for −0.10 mho susceptance. Go through the calculator steps and you will get an R, (series resistance) of 4 ohms and an X_s (series reactance) of 8 ohms at **step** 6. The s subscript is used for series values of impedance so you can keep them separated from parallel duals.

Eq. 2 will give equal results. R, would come out as $2000/500 = 4$ and X_s would be $4000/500 = 8$. The dif-

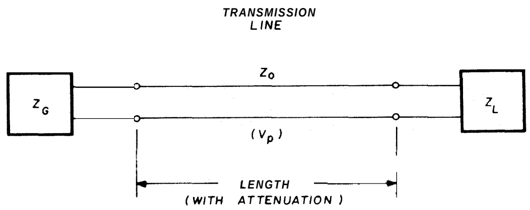


fig 2 Basic transmission line configuration Z_G is the impedance of the generator (transmitter); Z_L is the impedance of the load (antenna)

ference is that you have to enter R_p and X_p several times during the calculation. Try it both ways with a set of readings. Although polar/rectangular conversion is easier, you can make the programmable calculator do most of the work.

the first program

Program 1 is used to convert parallel bridge readings into series values of impedance. Steps 01 through 16 take care of this task. Steps 17 through 43 are optional and yield vswr at the point of measurement (more on vswr later).

Parallel-to-series value conversion uses the polar/rectangular conversion functions of the calculator as described before. Susceptance is calculated as $2\pi fC$. Since MHz and picofarads are assumed, the constant $2\pi \cdot 10^{-6}$ is stored in Register 6. Data entry steps are:

1. Press GTO, O, R/S (only for first set)
2. Key in frequency, press ENTER
3. Key in R_p, press ENTER
4. Key in C_p, press R/S

The first step is used only to tell the calculator where to start in the program. In this and subsequent programs, steps are arranged so that the last value will be displayed after a loop back to program start; you just enter the data for the next set and start again.

Frequency, resistance, and capacitance use the stack entry procedure at a single stop. Each time the ENTER key is pressed, the stack moves up once. The last item of a data set requires only the R/S key to begin the program; the last item is already in position after key-in.

The first stop **(step 14)** will display R_s. If you used an extender resistor, subtract its value at this stop. You would do this anyway and subtraction here will make the vswr values correct. Pressing R/S will display X_s at **step 16.** Pressing R/S once more will tell the program to proceed with vswr calculations which are displayed after looping back to the stop at **step 01.** The calculator is ready to take another stack entry for the next data set.

vswr equations

Voltage standing wave ratio or vswr is defined as3

$$VSWR = \frac{1 + |\rho_t|}{1 - |\rho_t|} \qquad (6)$$

where

$$\rho_t = \frac{Z_t - Z_o}{Z_t + Z_o} \qquad (7)$$

ρ_t complex reflection coefficient at an arbitrary point t on the line

$Z_o =$ transmission line characteristic impedance

$Z_t =$ impedance of line at point t

The difference between Z_o and Z_t is that Z_t has some sort of load on each end. If the load exactly equalled Z_o, then Z_t would equal Z_o and the vswr would be 1:1. Note that **eq. 6** requires the magnitude of ρ_t.

Transmission lines can usually be considered to be resistive-only at vhf and below. This makes $Z_o = R_o + j0$ so, with a bit of algebraic manipulation

$$VSWR = \frac{1 + \rho}{1 - \rho} \qquad (8)$$

when

$$\rho = |\rho_t|$$

with

$$\rho = \sqrt{\frac{(R_t - R_o)^2 + X_t^2}{(R_t + R_o)^2 + X_t^2}} \qquad (9)$$

Where R, and X_t are the real and imaginary parts of Z_t in the usual series form.

HP-25 Program

SWITCH TO PRGM MODE PRESS ⊡ PRGM THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	C_p	R_p	F		STACK ENTRY
02	32	CHS	$-C_p$	R_p	F		
03	21	x↔y	R_p	$-C_p$	F		
04	15 22	g 1/x	G_p	$-C_p$	F		
05	23 03	STO 3	G_p	$-C_p$	F		
06	22	R↓	$-C_p$	F			
07	61	X	$-FC_p$				
08	24 06	RCL 6	$2\pi 10^{-6}$	$-FC_p$			
09	61	X	$-B_p$				
10	24 03	RCL 3	G_p	$-B_p$			ADMITTANCE
11	15 09	g →P	1/MAG	-PHA			POLAR CONVERT
12	15 22	g 1/x	MAG	-PHA			
13	14 09	f →R	R_s	X_s			SERIES VALUES
14	74	R/S	R_s	X_s			MANUAL SUBTRACT
15	21	x↔y	X_s	R_s			OF EXTENDER-R
16	74	R/S	X_s	R_s			
17	23 01	STO 1	X_s	R_s			BEGIN VSWR
18	23 61 01	STO X1	X_s	R_s			
19	21	x↔y	R_s	X_s			
20	23 00	STO 0	R_s	X_s			
21	14 05	RCL 5	R_o	R_s	X_s		
22	23 51 00	STO +0	R_o	R_s	X_s		
23	21	x↔y	R_s	R_o	X_s		
24	41	−	R_{SM}	X_s			
25	31	ENTER↑	R_{SM}	R_{SM}	X_s		
26	61	X	R_{SM}^2	X_s			
27	24 01	RCL 1	X_s^2	R_{SM}^2			
28	51	+	M_N^2				NUMERATOR^2
29	24 01	RCL 1	X_s^2	M_N^2			
30	24 00	RCL 0	R_{sp}	X_s^2	M_N^2		
31	31	ENTER↑	R_{sp}	R_{sp}	X_s^2	M_N^2	
32	61	X	R_{sp}^2	X_s^2	M_N^2		
33	51	+	M_D^2	M_N^2			DENOMINATOR^2
34	71	÷	ρ^2				
35	14 02	f $\sqrt{x}$	ρ				
36	23 02	STO 2	ρ				
37	01	1	1	ρ			
38	51	+	$1 + \rho$				
39	01	1	1	$1 + \rho$			
40	24 02	RCL 2	ρ	1	$1 + \rho$		
41	41	−	$1 - \rho$	$1 + \rho$			
42	71	÷	VSWR				
43	13 01	GTO 01	VSWR				DISPLAY VSWR
44							AT STOP OF
45							STEP 01
46							
47							
48							
49							

MEMORY REGISTERS
R0 R_s
$R_o + R_s = R_{sp}$
R1 X_s
X_s^2
R2 ρ
R3 G_p
R4
R5 R_o (CONSTANT)
R6 $2\pi\ 10^{-6}$ (CONSTANT)
R7

Program 1. Converting noise bridge readings to impedance and vswr.

Program 1 uses **eqs. 8** and **9** in **steps 17** through **42** and requires the value of R_o (characteristic impedance) stored in Register 5 as a constant. The program also uses R, in place of *R*,, X_s in place of X_t since the vswr point is no longer arbitrary.

Some of the program steps may appear confusing due to use of register arithmetic functions. This is used to accumulate X_s^2 in Register 1 and the value R, +R, in Register 0. If needed, consult the *HP-25 User's* Manualfor these functions.

A separate program for vswr could be made by using only **steps 16** through **43. Step 16** would become **step 01**, **step 43** would become **28,** and no step commands would be changed. Stack entry would require only R_s and X_s.

rotation equations

These equations are used to calculate an unknown impedance at the opposite end of a transmission line. The word "rotation" comes from Smith chart usage where the measurement point is rotated around the chart by a specified wavelength fraction. The wavelength is dependent on frequency, physical length, and velocity of propagation of the line.

The basic equations depend on which way you are looking along the line. **Fig. 2** shows the general case where Z_L is the load, Z_o is the line's characteristic impedance, and Z_G is the impedance seen at the generator end due to mismatch between Z_L and Z, Best power transfer occurs when the transmitter or generator has a source impedance equal to Z_G. From reference 3 the equations are

$$Z_G = \left[\frac{Z_L + Z_o Z_A}{Z_o + Z_L Z_A}\right] Z_o \quad \textbf{(10)}$$

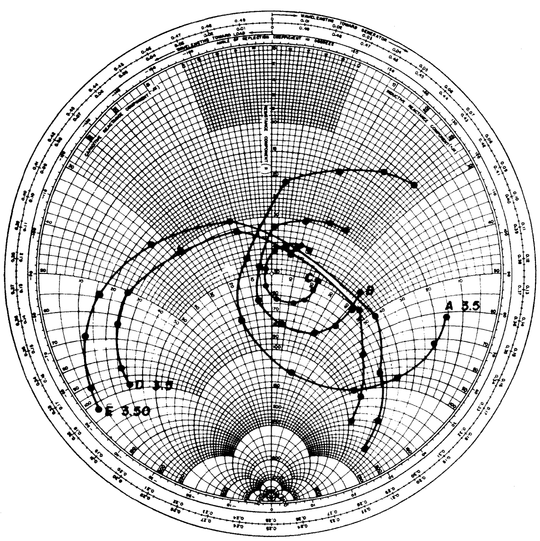


fig. 3. Impedance plots on a 50-ohm Smith chart. Each of the curves is discussed in the text.

$$Z_L = \left[\frac{Z_G - Z_o Z_A}{Z_o - Z_G Z_A}\right] Z_o \quad \textbf{(11)}$$

Eq. 10 is the same as clockwise Smith chart rotation and marked there as "wavelengths toward generator." **Eq. 11** is counter-clockwise rotation and marked "wavelengths towards load" on the Smith chart. Notice that a new complex value of Z_A has been added. It has the value of $tanh(a+jb)$ where $a=$ attenuation in nepers or decibels/8.686, and b is line wavelength expressed as an angle.

If you have followed this far, you are probably a bit confused. Either equation is enough to make a professional engineer seek computer help. Now you can understand why Phillip H. Smith invented his famous chart in the 1930s when computers were non-existent. Don't be afraid of the math because you now have calculator help.

The complex hyperbolic tangent of Z_A can be broken down to manageable terms by using some identities:

$$tanh(a+jb) = \frac{sinh(2a) + jsin(2b)}{cosh(2a) + cos(2b)}$$

Since

$$sinh(2a) = \frac{e^{4a} - 1}{2e^{2a}} \text{ and } cosh(2a) = \frac{e^{4a} + 1}{2e^{2a}}$$

Then

$$Z_A = \left[\frac{e^{4a} - 1}{D}\right] + j\left[\frac{2e^{2a} sin(2b)}{D}\right] \quad \textbf{(12)}$$

where

$D = e^{4a} + 1 + 2e^{2a} cos(2b)$

$e = 2.718282 \ldots$ (base of natural *logarithms)*

Z_A has now become manageable and has a real part R_A and an imaginary part X_A. Each part can be worked on separately and can use built-in calculator functions.

Many calculations can use rotations with lossless lines so let's find out what happens when attenuation, a, becomes zero. Recall that anything raised to a zero power becomes one. Under that condition, the R_A value becomes zero via its numerator and $X_A = tangent(b)$.

Before going further, **eqs. 10** and **11** have to be in a more manageable form. We can use the fact that Z is purely resistive at lower frequencies so

$$Z_G = \left[\frac{(R_L + R_A R_o) + j(X_L + R_o X_A)}{(R_o + R_L R_A - X_L X_A) + j(R_A X_L + R_L X_A)}\right] R_o \quad \textbf{(13)}$$

$$Z_L = \left[\frac{(R_G - R_A R_o) + j(X_G - R_o X_A)}{(R_o - R_G R_A + X_G X_A) - j(R_A X_G + R_G X_A)}\right] R_o \quad \textbf{(14)}$$

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	X_G	R_G	F		STACK ENTRY
02	23 01	STO 1	X_G	R_G	F		
03	23 02	STO 2	X_G	R_G	F		
04	22	R↓	R_G	F			
05	23 00	STO 0	R_G	F			
06	23 03	STO 3	R_G	F			
07	22	R↓	F				
08	24 04	RCL 4	K	F			
09	61	X	KF				
10	14 06	f TAN	X_K				
11	23 61 02	STO X2	X_K				PARTIAL PRODUCT
12	23 61 03	STO X3	X_K				PARTIAL PRODUCT
13	24 05	RCL 5	R_o	X_K			
14	61	X	R_oX_K				
15	24 01	RCL 1	X_G	R_oX_K			
16	21	x↔y	R_oX_K	X_G			
17	41	–	IM_N				NUMBER IMAG PART
18	24 03	RCL 3	$-IM_D$	IM_N			BEGIN DENOM.
19	24 05	RCL 5	R_o	$-IM_D$	IM_N		
20	24 02	RCL 2	X_GX_K	R_o	$-IM_D$	IM_N	
21	51	+	RE_D	$-IM_D$	IM_N		
22	15 09	g →P	M_D	$-P_D$	IM_N		POLAR CONVERT
23	23 02	STO 2	M_D	$-P_D$	IM_N		
24	22	R↓	$-P_D$	IM_N			
25	23 03	STO 3	$-P_D$	IM_N			FINISH DENOM.
26	22	R↓	IM_N				
27	24 00	RCL 0	RE_N	IM_N			NUMERATOR
28	15 09	g →P	M_N	P_N			
29	21	x↔y	P_N	M_N			
30	24 03	RCL 3	$-P_D$	P_N	M_N		BEGIN DIVIDE
31	51	+	PHA	M_N			
32	21	x↔y	M_N	PHA			
33	24 02	RCL 5	R_o	M_N	PHA		RESTORE FROM
34	61	X	M_N'	PHA			NORMALIZED FORM
35	24 05	RCL 2	M_D	M_N'	PHA		
36	71	÷	MAG	PHA			FINISH DIVIDE
37	14 09	f →R	R_L	X_L			RECT.CONVERT
38	74	R/S	R_L	X_L			DISPLAY REAL
39	21	x↔y	X_L	R_L			
40	13 01	GTO 01	X_L	R_L			DISPLAY IMAG.
41							PART AT STOP
42							OF STEP 01
43							
44							
45							
46							
47							
48							
49							

MEMORY REGISTERS
R0 R_G
R1 X_G
R2 X_G
X_GX_K
M_D
R3 R_G
R_GX_K
$-P_D$
R4 K
(CONSTANT)
SEE TEXT
R5 R_o
(CONSTANT)
R6
R7

Program 2. Lossless line impedance point rotation due to line length.

These may look worse, but they can now be mechanized on the calculator. The numerators and denominators are complex numbers but a program can use the simpler polar division operation for the answer. Term grouping and the signs suggest that a calculator program can be used for both expressions with only minor changes.

lossless line rotations

This should be limited to low frequencies or short line lengths since attenuation is considered zero. It has been established that only X_A is finite at zero loss and is simply the tangent of line wavelength expressed as an angle. Using degrees and one wavelength equal to 360 degrees, we can set up a constant

$$K = 0.366013\,\frac{L_{ft}}{V_p} = 1.20083\,\frac{L_M}{V_p} \qquad \textbf{(15)}$$

where

L_{ft} = line length in feet
L_M = line length in meters
V_p = velocity of propagation of the line

Velocity of propagation is usually the reciprocal of the square-root of the dielectric constant in coax. It is different for twin-lead or open-wire line. These can be found in handbooks or transmission line tables.4

Wavelength is frequency sensitive. Angular wavelength uses **eq. 15** multiplied by the frequency in MHz. A lossless line situation is a special case so we can use $X_A' = X_K = tangent(Kf)$ with f as the frequency. The lossless line rotation equations are simplified

$$Z_{GL} = \left[\frac{R_L + j(R_oX_K + X_L)}{(R_o - X_LX_K) + j\,R_LX_K} \right] R_o \qquad \textbf{(16)}$$

$$Z_{LL} = \left[\frac{R_G + j(X_G - R,,}{(R_o + X_GX_K) - j\,R_GX_K} \right] R_o \qquad \textbf{(17)}$$

the second program

This was written for **eq. 17.** Program start and stack entry procedures are the same as Program 1. Data set entry is frequency in MHz, R_G, and X_G, in that order. The ENTER key is not pressed after keying in X_G since it is the last item; simply press R/S to start.

Constant K from **eq. 15** must be preloaded into Register 4. R_o must also be preloaded into Register 5. The first stop at **step** 38 will display R_L; X_L is displayed after loop-back to the stop at the first step. Another stack entry may be done after reading X_L.

Programming is straightforward but **step 31** might seem to violate polar division rules in that the angles are added instead of subtracted. **Step 12** accumulates the denominator imaginary part (R_GX_K) as a positive quantity in Register 3. After recall at **step 18** and conversion to polar form at **step** 22, the denominator phase angle sign is changed. Rather than using an extra CHS command, **step 31** is an **ADD** (subtraction of a negative number is the same as addition of a positive number).

Steps 33 and **34** multiply the numerator magnitude by R_o to achieve the same purpose as the right bracket multiply in **eq. 17.** Deleting these two steps would yield normalized impedance values for the result although the data input must be in conventional un-normalized form. Program 2 can be written to accept a normalized impedance input if the equation is re-arranged.

Program 2 can be easily modified to solve **eq. 16;** just change **step 17** to **ADD, steps 21** and **31** to SUB-

table 1. Calculated values obtained from original bridge readings.

f, MHz	R_p	C_p	Z_G	vswr	Z_L
3.50	149"	– 163	15.93 + j 61.92	8.15	30.57 – j 95.96
3.55	164*	– 163	20.99 + j 72.14	7.63	26.95 – j 85.04
3.60	202"	– 141	42.75 + j 91.97	5.81	24.29 – j 64.62
3.65	240"	– 98	85.91 + j100.28	4.41	24.11 – j 50.14
3.70	129	– 66	124.14 + j 24.57	2.60	28.59 – j30.76
3.75	78	152	72.35 – j 20.21	1 .	32.49 – j 10.05
3.80	144*	20	43.32 – j 9.86	1.29	41.17 + j 7.54
3.85	121"	– 9	20.92 + j 3.19	2.40	48.68 + j 44.61
3.90	118"	– 39	16.52 + j 13.14	3.26	64.52 + j 69.58
3.95	116"	– 65	12.08 + j 20.97	4.91	78.74 + j106.83
4.00	117*	– 84	10.27 + j 27.24	6.36	98.80 + j141.22

"indicates extender-resistor of 100 ohms used

TRACT, and use load impedance at stack entry. This can be seen by inspecting the signs of each equation.

practical example

The noise bridge dial readings of the W6BXI/W6NKU article are used to illustrate operations with Programs 1 and 2; the results are shown in **table 1.** R„ X„ and vswr at the generator are from Program I while antenna impedance R_L and X_L were derived via Program 2. The original 60-foot (18-meter) loss-line line of 50 ohms characteristic impedance and v_p of 0.66 were used for rotation. Program 2 data entry used two-decimal values, quite adequate for normal use.

Curves **A** and D of the Smith chart in **fig. 3** are plotted from **table 1.** Curve **A** is the measurement (or generator) end, and curve D is the antenna end. There is a slight difference which is due mainly to the resolution of the chart when rotating manually. Resolution and manual errors are minimized with the calculator programs; wavelength is automatically calculated instead of being a separate operation.

Note that **fig.** 3 is a 50-ohm Smith chart and not the usual normalized version. This is slightly better *if* you are using a 50-ohm transmission line. Normalized impedances are found by simply dividing real and imaginary parts by R_o.

Impedance will change rapidly around resonance. For best accuracy, several more readings could be taken and calculated where the vswr appears low. Two to five times as many readings are possible with an accurate receiver frequency calibration. The calculator programs make short work of data reduction.

table 2. Calculated values from the first example with 0.8 dB line loss and the same impedance readings at the generator end of the line.

f, MHz	Z_G	generator vswr	R_A	X_A	Z_L	load vswr
3.50	15.93 + j 61.92	8.15	0.4474	− 1.9267	8.25 − j103.21	32.02
3.55	20.99 + j 72.14	7.63	0.4015	− 1.8021	8.59 − j 90.87	25.18
3.60	42.75 + j 91.97	5 81	0.3628	− 1.6889	12.03 − j 69.13	12.26
3.65	85.91 + j100.28	4.41	0.3299	− 1.5855	15.23 − j 53.94	7.27
3.70	124.14 + j 24.57	2.60	0.3018	− 1.4907	23.30 − j 33.85	3.29
3.75	72.35 − j 20.21	1.64	0.2774	− 1.4033	29.45 − j 11.27	1.83
3.80	43.32 − j 9.86	1.29	0.2563	− 1.3225	39.44 + j 8.74	1.36
3.85	20.92 + j 3.19	2.40	0.2379	− 1.2473	43.65 + j 52.91	2.96
3.90	16.52 + j 13.14	3.26	0.2217	− 1.1771	55.37 + j 86.95	4.52
3.95	12.08 + j 20.97	4.91	0.2075	− 1.1114	55.13 + j137.54	8.76
4.00	10.27 + j 27.24	6.36	0.1949	− 1.0497	54.91 + j189.95	15.08

operations with lossy lines

Fig. 4 shows the attenuation in dB per 100 feet (dB per 30.48 meters) of common coaxial lines. These are nominal since there is slight variation from one production run to another, and between manufacturers. You can see that attenuation becomes more pronounced at higher frequencies.

Complex attenuation/phase factor Z_A must be precalculated to use **eqs. 13** and **14** on the HP-25. This is a simple task when Program 3 is used and the output data can be tabulated with four-decimal accuracy. This data will be used with Program 4 following.

Program 3 takes advantage of the fact that attenuation variations are small over a narrow frequency band. Attenuation is treated as a constant and expressed as nepers (dB/8.686) stored in Register 7 at twice its value. Wavelength constant K is preloaded

table 3. Calculation of changes due to introducing attenuation pads between the noise bridge and the line in the first example.

	no pad		3 dB pad		6 dB pad	
f, MHz	Z_G	vswr	Z_G	vswr	Z_G	vswr
3.50	15.93 + j 61.92	8.15	44.15 + j39.56	2.29	51.09 + j20.20	1.49
3.55	20.99 + j 72.14	7.63	50.67 + j41.99	2.25	54.72 + j20.03	1.48
3.60	42.75 + j 91.97	5.81	65.48 + j40.45	2.10	60.37 + j16.88	1.43
3.65	85.91 + j100.28	4.41	78.26 + j30.64	1.92	63.97 + j11.59	1.38
3.70	124.14 + j 24.57	2.60	77.71 + j 6.51	1.57	62.27 + j 2.52	1.36
3.75	72.35 − j 20.21	1.64	60.82 − j 8.13	1.28	55.28 − j 3.66	1.13
3.80	43.32 − j 9.86	1.29	46.81 − j 5.27	1.14	48.45 − j 2.73	1.07
3.85	20.92 + j 3.19	2.40	32.99 + j 2.19	1.52	40.70 + j 1.31	1.23
3.90	16.52 + j 13.14	3.26	30.60 + j 9.43	1.72	39.54 + j 5.78	1.31
3.95	12.08 + j 20.97	4.91	28.59 + j15.72	1.99	38.85 + j 9.80	1.40
4.00	10.27 + j 27.24	6.36	28.66 + j20.64	2.15	39.51 + j12.79	1.49

into Register 6 as twice its value also. Doubling of the constants saves program steps.

The only stack entry is frequency in MHz at **step 01**. Real part R_A is displayed at **step 23** and imaginary part X_A displayed after loop-back to the first step. Just key in the next frequency and press R/S for the next Z_A set.

Program 3 follows **eq.** 12 but may be confusing due to extensive use of register arithmetic functions. It is a fairly good example of program optimization and is worth study just for that reason.

Change the attenuation constant if tabulating Z_A for more than one band. Attenuation is proportional to physical length and can be found easily from **fig. 4.**

rotations with lossy lines

Program 4 is written to mechanize **eq. 14.** There are four items in stack entry: R_G, X_G, R_A, and X_A, in that order. Remember that only the R/S key is pressed after keying in X_A. The first display is R_L and another stack entry can be done after display of X_L. Constant R_o must be preloaded in Register 5.

We can use the previous example to show attenuation effects. Assume the same bridge readings with the same line length and v_p. Now add a total attenuation of 0.8 dB to the line, calculate Z_A with Program 3, and calculate the new antenna impedance with Program 4. The data is given in **table 2** and the new impedance is curve E of **fig. 2.** Note that **E** is more reactive than curve D.

If the antenna impedance was actually curve D, then the measurements would show another curve at the generator end that is closer to the center of the Smith chart. This would be a "masking" effect on impedance due to attenuation.

Program 4 can be altered to find the generator impedance from **eq. 13** by simply changing **steps 25, 32,** and **35** to ADD, **step 41** to SUBTRACT. Stack entry would use R_L in place of R_G, and X_L in place of X_G: R_A and X_A would remain the same.

The Z_A tabulation used a constant of $(0.8/4.343) = 0.184204$ for a in Register 7 of Program 3; the $2K$ constant was $2(0.366013 \times 60/0.66) = 66.5478$ in Register 6. Six-place constants were used for accuracy. Output data for Z_A can be to four places.

impedance masking by attenuation

This has already been stated but another example is in order. Let's take the first example again and use only the measurement end data. Now assume that short pads of 3 dB and 6 dB were inserted between the noise bridge and line. To calculate this condition, Program 3 used a zero line length. X_A became zero and R_A was frequency-insensitive with a value of *0.332275* for 3 dB, and *0.598474* for 6 dB. Program 4 was modified to fit **eq. 13** and results are shown in **table 3** and **fig. 2** as curve **B** (3 dB) and curve ***C*** (6 dB).

The original 8.15:1 vswr at 3.5 MHz dropped to 2.29:1 at 3 dB and only 1.49:1 at 6 dB. This not only soaks up power but could fool a swr meter installed at the transmitter. Suppose you had a 10-meter rig with a total of 60 meters (200 feet) of RG-58/U coax feeding the antenna. Total line attenuation from **fig. 4** indicates a line loss of 4.5 dB. An antenna-end mismatch giving an 8:1 vswr would show up as less than

table 4. Solution of attenuation and test angle or unknown transmission line by open- and short-circuiting the load end of the line.

	shorted load end			open end load			Z_u solution	
f, MHz	R_p	C_p	Z_G	R_p	C_p	Z_G	dB	ϕ, degrees
21.0	190	125	17.56 + j 55.03	310	– 75	29.78 – j91.34	1.3315	37.4396
21.1	220	110	19.48 + j 62.50	280	– 85	25.56 – j80.64	1.3319	41.1636
21.2	250	100	20.68 + j 68.86	250	– 100	20.68 – j68.86	1.2854	45.0000
21.3	280	85	25.12 + j 80.02	220	– 110	19.15 – j62.02	1.3197	48.8323
21.4	310	75	28.78 + j 89.96	190	– 120	18.27 – j56.02	1.3396	52.0243
21.5	340	65	34.30 + j102.40	160	– 140	15.75 – j47.67	1.3150	56.2338

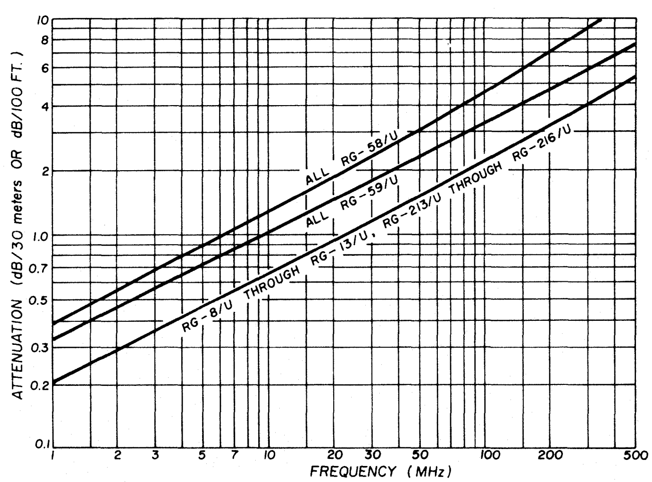


fig. 4. Attenuation versus frequency for common coaxial transmission lines.

2:1 at the transmitter end. Only a calibrated noise bridge and careful calculations tell the truth.

Larger coax for less loss is best at higher frequencies. It also improves antenna measurements since there is less "masking" of impedance; the impedance points move closer to the center of the Smith chart with increased line attenuation.

Impedance masking by attenuator pads is an old technique used in microwave bench measurements to improve the "flatness" of the source impedance; it's also a good idea at lower frequencies. Less expensive generators vary widely so a pad would make the device under test "see" a better impedance.

Comparison of the plot points of curves **A,** B, and ***C*** in **fig.** 3 will show no change in rotation position. The same is true when comparing curves D and E. Attenuation changes only the chart position away from center even though the tabulated impedance seems to change rotation. Rotation is strictly a wavelength function and is due to frequency, line length, and velocity of propagation.

transmission line quality

It is common to run across some really juicy bargains in coaxial cable at surplus houses; it is just as common to find out there is no known identification and the cable probably has different characteristics than advertised. Is it good or bad? What is the characteristic impedance? Velocity of propagation? All are unknown. On the other hand, perhaps your four-year-old feedline is going bad due to weather.

Fortunately there is a way to find out, and fairly simply with an accurate RX noise bridge and another program. Just check out the line itself with a short and an open at the other end. This sounds too simple so let's investigate the theory.

Call the measured impedance with an open circuit Z_{oc} and the measured impedance with a short circuit Z_{sc}. From reference *3* we know that

$$Z_o = \sqrt{Z_{oc}\, Z_{sc}} \qquad \textbf{(18)}$$

This calculation is easily done manually in polar form using **eq. 5.** Do this over several frequencies and average the results. Averaging will dilute errors in calibration and dial readings.

The next problem is to find attenuation and the velocity of propagation. You can take a guess at v_p just from examining the dielectric. The most common type is polyethylene ($v_p = 0.6594$); other solid dielectrics are close to this value. Foamed dielectrics will vary a great deal. The v_p guess is best since the line wavelength should be an *odd* multiple of 1/8th wavelength.

Choice of length is better explained by calling attention to the bridge measurement limit contour. The shorted-end measurements will be directly opposite the chart from the open-end measurements. The RX noise bridge cannot measure impedances falling at the high-resistance end. The first step is to calculate

$$Z_u \text{ (polar form)} = \frac{1+\sqrt{Z_{sc}/Z_{oc}}}{1-\sqrt{Z_{sc}/Z_{oc}}} \qquad \textbf{(19)}$$

From this we can find

$$dB\ attenuation = 4.343\ Log_e\ (Z_u\ magnitude) \qquad \textbf{(20)}$$

$$\emptyset, \text{ the test angle} = \frac{(Z_u \text{ phase angle})}{2} \qquad \textbf{(21)}$$

Eq. 20 uses natural logs, not the base 10 or common logarithm. We are stuck with a "test angle" since wavelength is not precisely known. Velocity of propagation is found from:

$$v_p = \frac{0.366013\, f\, L_{ft}}{\emptyset + 360n} = \frac{1.20083\, f\, L_M}{\emptyset + 360n} \qquad \textbf{(22)}$$

where

f = measurement frequency in MHz
L_{ft} = physical line length in feet
L_M = physical line length in meters
n = any integer number including zero

We can pin down the value of v_p by using ***eq.*** 22 with several values of n at each frequency. **Eqs. 21** and 22 assume the test angle to be in degrees and only one n value will be correct; this can be found by inspection.

Program 5 was written to mechanize **eq. 19.** Stack entry requires four values; these are obtained from Program 1 — ignore the vswr data in this case. The first display stop of Program 5 is the attenuation in dB. The next display is test angle $\emptyset$ after looping back to the first program step.

example

You have a 20 meter (65 foot) length of coaxial cable that appears to be RG-58/U but the markings are unknown. The dielectric appears to polyethylene so you can estimate wavelength using **eq. 15.** K will be 36.0759 so it seems that multiplying this by 15-meter frequencies will be correct. At 21.0 MHz you get 757.670° (2.104 wavelengths) and 21.5 MHz gives 775.710° (2.155 wavelengths). Measurements are taken every 100 kHz.

table *5*. Velocity of propagation test at different values of *n*.

	v_p at different n values				
f, MHz	**n=0**	**n=1**	**n=2**	**n=3**	**n=4**
21.0	13.3439	1.2571	0.6596	0.4471	0.3382
21.1	12.1949	1.2513	0.6595	0.4477	0.3389
21.2	11.2081	1.2453	0.6593	0.4483	0.3396
21.3	10.2081	1.2395	0.6591	0.4489	0.3404
21.4	9.7863	1.2357	0.6595	0.4497	0.3412
21.5	9.0960	1.2289	0.6590	0.4502	0.3419

The results of bridge measurements and data from Programs 1 and 5 are given in **table 4.** Note: This particular noise bridge had a 500-ohm potentiometer instead of the original 250-ohm pot.

Characteristic impedance is found by manual calculation with **eq. 18** and is

74 49 / 0.1800° at 21.0 MHz
74.42 ∠0.1378° at 21.1 MHz
71.90 ∠0.0000° at 21.2 MHz
73.78 L – 0.1345° at 21.3 MHz
74.60 ∠0.1612° at 21.4 MHz
73.63 ∠ – 0.1177° at 21.5 MHz

The average Z_o is 73.80 ohms over six frequencies. The small angle residue comes trom minor calibration and measurement errors can be ignored. If it is over 5° you have done something wrong or the bridge limits have been exceeded.

Average attenuation is found to be 1.3205 dB. Trials of n values for v_p are given in table **5.** This table shows that low n values have a steadily decreasing v_p with increasing frequency while high n values have steadily increasing v_p. The correct n value has v_p bouncing around some average value. Data at $n=2$ is steadiest and the average is 0.6593. The guess at polyethylene dielectric was correct.

This cable example was actually RG-59/U and a manufacturer's reject sold as surplus. A check using an HP 8507A Automatic Network Analyzer over many more frequencies showed that the cable Z_o was actually 73.0 ohms and attenuation measured 1.30 dB. At only six frequencies the noise bridge was 1.1 per cent high for Z_o and 1.6 per cent high for attenuation, a very good score for equipment costing one-thousandth of laboratory instruments!

jumping over the measurement contour

The noise bridge is limited at high R, or large X_s values. A novel suggestion by Dean Straw, NGBV, is to use a fixed resistor in *shunt* with the potentiometer.5 This could be plugged into the TUNING arm of the bridge and still have the option of using the extender resistor on the UNKNOWN arm. A problem is that calibration must be done *very* carefully.

NGBV cites the case of a marine whip antenna having an impedance of 10-j400 ohms at 2 MHz. Even with the extender resistor, a null is not possible. The R_p value would have to be 1565 ohms. Placing a 220-ohm fixed resistor across the potentiometer allows a null at an R_p of 193 ohms. C_p would be – 185 pF and at the capacitor limit.

The need for accurate calibration can be seen by comparing the impedance derived from measurements versus required values. A dial reading of 193 ohms R_p and – 185 pF C_p will calculate as 21.24 – j392.72 ohms. Correct dial readings *should* have been 192.88 ohms for R_p and – ***184.96*** pF for C_p. A very small change in R_p gives a very large change in R,

Another method, more accurate with calculator help, is to use an *extender line.* This is used between the noise bridge and the unknown. Length should be between 1/8th and 3/8th wavelength to bring you back inside the limit contour. The extender line can be measured using the short/open method described before, then tagged for reference.

A small pad can also be used but should not exceed about 3 dB to retain accuracy. Program 4 is then used with the pad added to line attenuation. Eighth-watt carbon composition resistors are suitable for the pad and the finished unit must be checked carefully.

using an extender line

The procedure for using an extender line is simple. Take bridge readings with the extra line between bridge and unknown. Check for good nulls; a poor null will tell you that the extra line is either too short or too long. Try an extra extender if necessary. Find the total line length.

The first step is to use Program 1 to obtain "extended" R_G and X_G values. The second step is to

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM.

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	F				ENTER FREQUENCY
02	24 06	RCL 6	2K	F			
03	61	X	2KF				
04	23 03	STO 3	2KF				
05	In 05	f COS	COS 2KF				
06	24 07	RCL 7	α	COS 2KF			
07	15 07	g	e^a	COS 2KF			
08	23 00	STO 0	e^a	COS 2KF			FORM $2e^a$
09	23 51 00	STO +0	e^a	COS 2KF			IN R0
10	23 01	STO 1	e^a	COS 2KF			IN R1 e^{2a}
11	23 61 01	STO X1	e^a	COS 2KF			
12	31	ENTER↑	e^a	e^a	COS 2KF		
13	51	+	$2e^a$	COS 2KF			
14	61	X	$2e^a$COS				
15	24 01	RCL 1	e^{2a}	$2e^a$COS			
16	51	+	DEN.-1				
17	01	1	1	DEN.-1			
18	23 41 01	STO -1	1	DEN.-1			
19	51	+	DENOM.				
20	23 71 00	STO ÷0	DENOM.				REGISTER
21	23 71 01	STO ÷1	DENOM.				ARITHMETIC
22	24 01	RCL 1	RA				
23	74	R/S	RA				DISPLAY REAL PART
24	24 03	RCL 3	2KF				
25	14 04	f SIN	SIN 2KF				
26	24 00	RCL 0	IM/SIN	SIN 2KF			
27	61	X	X_A				
28	13 01	GTO 01	X_A				DISPLAY IMAG.
29							PART AT STOP
30							OF STEP 01
31							
32							
33							
34							
35							
36							
37							
38							
39							
40							
41							
42							
43							
44							
45							
46							
47							
48							
49							

MEMORY REGISTERS
R0 e^a
$2e^a$
IMAG/SIN 2KF
R1 $e^a \rightarrow e^{2a}$
e^{2a} -1
REAL
R2
R3 2KF
R4
R5
R6 2K
(CONSTANT,
SEE TEXT)
R7 α=dB/4.343
(CONSTANT)

Program 3. Complex attenuation-wavelength factor.

"look backward" on the total line to find impedance at the desired point. This can be done with either Program 2 or 4, each modified to solve for Z_L. Length for the second step is that of the extender alone, not the total line length. The data output is the impedance at the desired point.

Puzzled? The first step brought you within the measurement limit contour but the length was longer than actual; the impedance was rotated too far in a counter-clockwise direction. The second step found the correct impedance by clockwise rotation of a wavelength determined by the extender line.

Characteristic impedance of the extender line must be equal to the measured line for the two-step procedure. A solution with different impedances is possible but the math is lengthy. A direct measurement without transmission line would use the Z_o of the extender line.

An extender line can be of different dimensions as long as the equal Z_o rule is met. A word of caution with RG-58/U: Common RG-58/U and RG-58B/U have a Z_o of 53.5 ohms, while RG-58A/U is 52.0 ohms; RG-58C/U is 50.0 ohms. RG-58C/U is recommended as a general-purpose 50-ohm line since it is extra flexible and more durable for test work.

keeping tabs on that new antenna

Weather takes its toll on all antenna installations.

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM.

LINE	CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	X_A	R_A	X_G	R_G	STACK ENTRY
02	23 03	STO 3	X_A	R_A	X_G	R_G	
03	23 07	STO 7	X_A	R_A	X_G	R_G	
04	22	R↓	R_A	X_G	R_G	X_A	
05	23 02	STO 2	R_A	X_G	R_G	X_A	
06	23 06	STO 6	R_A	X_G	R_G	X_A	
07	22	R↓	X_G	R_G	X_A	R_A	
08	23 01	STO 1	X_G	R_G	X_A	R_A	
09	21	x↔y	R_G	X_G	X_A	R_A	
10	23 00	STO 0	R_G	X_G	X_A	R_A	
11	23 61 06	STO X 6	R_G	X_G	X_A	R_A	PARTIAL PRODUCT
12	23 61 07	STO X 7	R_G	X_G	X_A	R_A	ACCUMULATE
13	22	R↓	X_G	X_A	R_A		
14	61	X	X_GX_A	R_A			
15	23 41 06	STO − 6	X_GX_A	R_A			
16	21	x↔y	R_A	X_GX_A			BEGIN DENOM.
17	24 01	RCL 1	X_G	R_A	X_GX_A		
18	61	X	X_GR_A				
19	24 07	RCL 7	R_GX_A	X_GR_A			
20	51	+	$-IM_D$				IMAGINARY PART
21	24 05	RCL 5	R_o	$-IM_D$			
22	23 61 02	STO X 2	R_o	$-IM_D$			PARTIAL PRODUCT
23	23 61 03	STO X 3	R_o	$-IM_D$			ACCUMULATE
24	24 06	RCL 6	R6	R_o	$-IM_D$		
25	41	−	RE_D	$-IM_D$			REAL PART
26	15 09	g →P	M_D	$-P_D$			POLAR CONVERT
27	23 06	STO 6	M_D	$-P_D$			
28	21	x↔y	$-P_D$	M_D			
29	23 07	STO 7	$-P_D$	M_D			FINISH DENOM.
30	24 01	RCL 1	X_G				BEGIN NUMERATOR
31	24 03	RCL 3	R_oX_A	X_G			
32	41	−	IM_N				IMAGINARY PART
33	24 00	RCL 0	R_G	IM_N			
34	24 02	RCL 2	R_oR_A	R_G	IM_N		
35	41	−	RE_N	IM_N			REAL PART
36	15 09	g →P	M_N	P_N			POLAR CONVERT
37	24 05	RCL 5	R_o	M_N	P_N		RESTORE FROM
38	61	X	M_N'	P_N			NORMALIZED FORM
39	21	x↔y	P_N	M_N'			
40	24 07	RCL 7	$-P_D$	P_N	M_N'		
41	51	+	PHA	M_N'			
42	21	x↔y	M_N'	PHA			
43	24 06	RCL 6	M_D	M_N'	PHA		
44	71	÷	MAG	PHA			
45	14 09	f →R	R_L	X_L			RECT. CONVERT
46	74	R/S	R_L	X_L			DISPLAY REAL PART
47	21	x↔y	X_L	R_L			
48	13 01	GTO 01	X_L	R_L			DISPLAY IMAG PART
49							AT STOP OF STEP 01

MEMORY REGISTERS
R0 R_G
R1 X_G
R2 R_A
R_oR_A
R3 X_A
R_oX_A
R4
R5 R_o
(CONSTANT)
R6 $R_A \rightarrow R_GR_A$
$R_GR_A - X_GX_A$
M_D
R7 $X_A \rightarrow R_GR_A$
$-P_D$

Program 4. Impedance point rotation on lossy lines due to line length.

A good way to locate potential problems is to periodically measure the line and antenna. Take readings in both good and bad weather; snow, ice, or rain can cause impedance changes on good antennas.

You might be able to catch a line break before it happens. A pooriy sealed coax can cause erosion of the center conductor, changing both the impedance and attenuation at one end. Direct cable feeds to rotating antennas can bend and stretch the coax out of shape.

The near-field or Fresnel Zone (about five wavelengths from center) can include trees and poles. This may cause impedance changes with rotary antennas. Try measurements at different azimuths. Impedance changes will warn you to expect different transmitter loading at certain antenna directions.

Fixed antennas are not immune to change. Trees grow slowly and old ones may be cut down. House remodelling can cause changes, too. Replacement of rain gutters can change the local ground of a rooftop mount. In one observed change, a 40-meter vertical mounted on a back wall was changed by adding an aluminum patio roof. The pattern was also changed, but the noise bridge can't measure that.

the forgotten receiver

Receiver agc can fool the best of us even if the input impedance is grossly mismatched (many receivers are faulty in this respect). A good communications link needs best performance in *both* directions so it's a good idea to check your receiver, too.

If vou have one accurate-freauencv receiver and it was used as the bridge detector, you have to substitute. A general-purpose communications receiver is good enough and you can calibrate it using received signals from local stations.

Most receiver designs use preselector tuning of the front end. Don't forget to peak the preselector; noise from the noise bridge generator is good for this. An off-peak preselector will reflect a different impedance. A known input impedance and thorough schematic study will be enough to tell you what to do for a best match.

cautions with high vswr

High vswr may be unavoidable, but peak voltages and currents must be considered in relation to the line and matching network. The following relations help

$$E_{peak} = E_{nominal}\sqrt{2\ VSWR} \tag{23}$$

$$I_{peak} = I_{nominal}\sqrt{2\ VSWR} \tag{24}$$

Nominal values are rms at a perfect match.

HP-25 Program

SWITCH TO PRGM MODE, PRESS f PRGM THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	X_{oc}	R_{oc}	X_{sc}	R_{sc}	STACK ENTRY
02	21	x↔y	R_{oc}	X_{oc}	X_{sc}	R_{sc}	
03	15 09	g →P	M_{oc}	P_{oc}	X_{sc}	R_{sc}	POLAR DENOM.
04	15 22	g 1/x	$1/M_{oc}$	P_{oc}	X_{sc}	R_{sc}	
05	23 00	STO 0	$1/M_{oc}$	P_{oc}	X_{sc}	R_{sc}	
06	22	R↓	P_{oc}	X_{sc}	R_{sc}		
07	32	CHS	$-P_{oc}$	X_{sc}	R_{sc}		
08	23 01	STO 1	$-P_{oc}$	X_{sc}	R_{sc}		
09	22	R↓	X_{sc}	R_{sc}			
10	21	x↔y	R_{sc}	X_{sc}			
11	15 09	g →P	M_{sc}	P_{sc}			POLAR NUMERATOR
12	23 61 00	STO X 0	M_{sc}	P_{sc}			POLAR
13	21	x↔y	P_{sc}	M_{sc}			MULTIPLY IN
14	23 51 01	STO + 1	P_{sc}	M_{sc}			R_o & R_1
15	24 01	RCL 1	2P				
16	02	2	2	2P			
17	71	÷	P				
18	24 00	RCL 0	M^2	P			
19	14 02	f √x	M	P			$(Z_{sc}/Z_{oc})^{1/2}$ TERM
20	14 09	f →R	RE	IM			RECT.CONVERT
21	23 02	STO 2	RE	IM			
22	01	1	1	RE	IM		
23	51	+	RE + 1	IM			
24	21	x↔y	IM	RE + 1			
25	23 03	STO 3	IM	RE + 1			
26	21	x↔y	RE + 1	IM			
27	15 09	g →P	M_N	P_N			POLAR NUMER.
28	23 00	STO 0	M_N	P_N			
29	21	x↔y	P_N	M_N			
30	23 01	STO 1					
31	24 03	RCL 3	IM				
32	01	1	1	IM			
33	24 02	RCL 2	RE	1	IM		
34	41	−	1-RE	IM			
35	15 09	g →P	M_D	$-P_D$			POLAR DENOM.
36	23 71 00	STO ÷ 0	M_D	$-P_D$			POLAR
37	21	x↔y	$-P_D$	M_D			DIVIDE IN
38	23 51 01	STO + 1	$-P_D$	M_D			R_o & R_1
39	24 00	RCL 0	M_U				Z_U MAGNITUDE
40	14 07	f LN	$LN(M_U)$				
41	24 04	RCL 4	4.343	$LN(M_U)$			
42	61	X	dB				
43	74	R/S	dB				ATTENUATION
44	24 01	RCL 1	P_U				Z_U PHASE ANGLE
45	0 2 1 2	2	2	P_U			
46	71	÷					TEST ANGLE
47	13 01	GTO 01	Ø	—			DISPLAY Ø
48							AT STOP OF
49							STEP 01

MEMORY REGISTERS	
R0	$1/M_{oc}$
	M^2
	M_N M_U
R1	$-P_{oc}$
	2P
	P_N P_U
R2	RE
R3	IM
R4	4.343
	(CONSTANT)
R5	
R6	
R7	

Program 5. Calculating properties of unknown transmission line from open- and short-circuited measurements.

Assume perfect conditions with a 400-watt transmitter. Line voltage with a 50-ohm load is 141.42 rms or 200 volts peak. Line current is 2.8284 amps rms or 4 amps peak. Now change the load so that the vswr is 4:1.

At a certain line wavelength, the transmitter may see an impedance of 200 + j0 ohms. At maximum real part and minimum imaginary part of impedance, **eq. 23** will apply. Peak voltage is 400 or twice the perfect condition. Changing the line length by exactly a quarter wavelength will change the load to 12.5 + j0 ohms. **Eq. 24** indicates a peak current of 8 amperes. Peak values increase by the square-root of vswr at both lengths. *Both the transmission line and matching network must handle these peaks or breakdown occurs.*

Other line wavelengths will present different loads to the transmitter. This can be seen by following the vswr circle on a **Smith** chart. A long transmission line can have both peaks; voltage breakdown rating is the prime consideration. Power handling capability is next. It is a good idea to check both from cable tables.[4,6]

If at all possible, try to match *at the antenna.* The reason is peak line current. Center conductor heating in coax will increase attenuation in peak current regions; heating increases conductor resistance for more loss. It is worst in warm weather. Balanced lines have a similar condition but it's less pronounced — impedance is generally higher and the current is divided equally in each wire.

High power impedance *may* be different at high vswr than that measured with low power bridge readings.

baluns and balanced lines

A balun transformer allows you to check twinlead and open-wire lead with the RX noise bridge. However, the balun itself must be carefully checked since it becomes part of the bridge. Checks and calibration are done in the same manner as original bridge calibration.

Impedance-change baluns will affect bridge readings. Both real and imaginary series values must be multiplied by the impedance-change value. All other programs use the balanced line characteristic impedance.

Remember to keep away from the field around balanced lines. Measurement errors can be made if you are too close to the balun attachment point. You can also use Program 5 with balanced lines. A suggestion is to suspend the line to be measured by string from a rope. A I-meter (three foot) distance above ground should be sufficient.

other uses

The RX noise bridge lends itself to any frequency in the high-frequency band. It is well within the FCC rules for incidental radiation. N6BV cited its use with a marine whip; it can also be used to check an SWL's long-wire.

Your CB neighbor might even be won over to amateur radio. Offer to check out the CB antenna and give suggestions if the readings aren't good.* The double handful of bridge and calculator, plus scratchpad and this issue of *ham radio* (for the programs, of course) can be impressive. Just be prepared for a lot of questions if friendly neighbor starts reading the magazine!

*Tape the transmit switch to receive-only or you can damage the bridge and don't forget you need a commercial ticket to adjust the CB transmitter.

references

1. Robert A. Hubbs, W6BXI, and A. Frank Doting, W6NKU, "Improvements to the RX Noise Bridge," *ham radio,* February, 1977, page 10.
2. James Fisk, W1DTY, "How to Use the Smith Chart," *ham radio,* November, 1970, page 16 (reprinted March, 1978, page 92).
3. Robert A. Chipman, *Theory and Problems of Transmission Lines,* Schaum's Outline Series, McGraw-Hill. New York, 1968 (still sold at large paperback stores and a bargain at $5.95. Math heavy but gives all the answers. Recommended for serious study).
4. Doug DeMaw, W1CER, ARRL *Electronics Data Book,* The American Radio Relay League, Newington, Connecticut. 1976.
5. R. Dean Straw, N6BV, letter to Charles Carroll. K1XX, June 1, 1977.
6. RF *Transmission Line Catalog and Handbook,* Catalog TL-3, 1970, Times Wire and Cable, Wallingford, Connecticut, 1970.

ham radio

noise bridge calculations with Texas Instruments Programmable 58159 calculators

The description of the RX noise bridge which recently appeared in *ham radio*[1] came at a propitious time — I wanted to build a three-element vertical phased array for 20 meters and faced the prospect of tuning the system with a dip meter! A capable noise bridge seemed an ideal answer and the version described by W6BXI and W6NKU was successfully built and tested.

Only one problem remained: I had no Smith charts to convert the series impedance values at the end of the feedline to series impedance values at the antenna feedpoint. A little study also revealed that it would be difficult to read or enter values with any precision in certain regions of the Smith chart. Finally, since I normally have five or six frequency-impedance combinations to process for each trial antenna length (times three for my phased array), the graphical construction techniques using a Smith chart would take a very long time.

the solution — a calculator program

My solution to this dilemma appeared in the form of a newly-purchased Texas Instruments *Programmable 58* calculator. This remarkable calculator packs up to 480 program steps or up to 60 memories in a small hand-held unit. (The user may partition the available memory between program and data storage to suit his needs.) More importantly, the 25-program "Master Library Module," a small plug-in element containing the equivalent of 5000 program steps, provides extensive vector algebra capability; — this is precisely what I needed to perform Smith chart calculations. (If you can afford it, the *Programmable 59* possesses even more memory and provides the capability for storing and entering programs using small magnetic cards; this is especially useful for long programs.)

A little digging in a local library provided the necessary formulas;[2] some rearrangement to suit the problem led to eq. 1, below. This calculates the equivalent series impedance Z_t of the termination of a transmission line (characteristic impedance, Z_o) of a given electrical length A', when the equivalent series input impedance Z_i is known.

$$Z_t = \frac{Z_o\,(Z_i \cos A' - jZ_o \sin A')}{(Z_o \cos A' - jZ_i \sin A')} \qquad (1)$$

Note that all impedances must be treated as *vector* quantities. Once the impedance at the termination (the antenna) is known, an additional calculation quickly provides the vswr at the antenna

$$vswr = \frac{1 + |\rho'|}{1 - |\rho'|} \qquad (2)$$

$$\rho' = \frac{Z_t - Z_o}{Z_t + Z_o}$$

Again, the impedances must be treated as vector quantities.

The program presented in fig. 1 handles all these complex calculations in a matter of seconds, including the series-parallel impedance conversions and the range extender correction described in the original noise bridge article. This makes it easy to perform many trials on an antenna system in a very short time. While the program is long (184 steps), the time spent entering it pays large dividends in time saved during antenna adjustments. (For the *Programmable 59* owners, magnetic card storage makes this a one-time concern.)

program details

The flow chart (fig. 2) depicts the various modules of the program. You should consult the noise bridge article for the equations for calculation of X_p, conversion of parallel to series impedance, and the range extender correction.

The correction factor for the range extender can vary with frequency, so register 19 is set aside to contain that value. When you have determined the exact value, simply store it in register 19 when you initialize the calculator.

Since you will normally be working with many bridge readings taken with the same piece of transmission line, each case after the first does not require the input of the line length or its characteristic impedance.

You will notice that the electrical length of the line is expressed in radians instead of degrees. When the calculator enters the "Master Library Module" programs 04 and 05 for the vector calculations, it is set in the radian mode and remains there upon its return. If the equation for the electrical length was expressed in degrees, each case after the first would be in error.

By T. J. Anderson, WD4GRI, 1907 Lodgepole Avenue, North Augusta, South Carolina 29841

TITLE NOISE BRIDGE CALCULATIONS

PROGRAMMER WD4GRI

Partitioning (Op 17) 2,3,9,2,9 Library Module Master Library Printer ____ Cards ____

TI Programmable **Program Record**

PROGRAM DESCRIPTION

Calculates antenna impedance and VSWR given noise bridge readings.

USER INSTRUCTIONS

STEP	PROCEDURE	ENTER	PRESS		DISPLAY
1	INITIALIZE: 2nd Rad; -159000 STO 15; 0.006386/v STO 16; extender value STO 19		CLR	RST	0
2	Enter line impedance.	Z_o	R/S		Z_o
3	Enter line length in feet.	ℓ	R/S		ℓ
4	Enter frequency in MHz.	f	R/S		f
5	Enter bridge resistance.	R_p	R/S		R_p
6	Enter bridge capacitance.	C_p	R/S		C_p
7	Enter "1" if range extender used, "0" if not.	"1" or "0"	R/S		R_s (ant)
8	Read R_s (ant)		R/S		X_s (ant)
9	Read X_s (ant)		R/S		VSWR
10	Read VSWR; If another case with same line, go to STEP 4. If not, press CLR and RST & go to STEP 2.				

USER DEFINED KEYS: A, B, C, D, E, A', B', C', D', E'

DATA REGISTERS (INV ■)

0		10	Z_o
1		11	λ'
2		12	f
3		13	R_p
4		14	$C_p \rightarrow X_p$
5		15	-159000
6		16	0.006386/v
7		17	R_s
8		18	X_s
09	ℓ	19	Extender value

LABELS (Op 08)

FLAGS 0 1 2 3 4 5 6 7 8 9

TITLE NOISE BRIDGE CALCULATIONS

PROGRAMMER WD4GRI

TI Programmable **Coding Form**

LOC	CODE	KEY	COMMENTS
0	42	STO	
1	10	10	
2	91	R/S	
3	42	STO	
4	09	9	
5	91	R/S	
6	42	STO	
7	12	12	
8	91	R/S	
9	42	STO	
10	13	13	
11	91	R/S	
12	42	STO	
13	14	14	
14	91	R/S	
15	94	+/-	
16	42	STO	
17	17	17	
18	43	RCL	
19	19	19	
20	49	2nd Prd	
21	17	17	
22	43	RCL	
23	09	9	
24	65	x	
25	43	RCL	
26	12	12	
27	65	x	
28	43	RCL	
29	16	16	
30	95	=	
31	42	STO	
32	11	11	
33	43	RCL	
34	15	15	
35	55	÷	
36	43	RCL	
37	12	12	
38	55	÷	
39	43	RCL	
40	14	14	
41	95	=	
42	42	STO	
43	14	14	
44	65	x	
45	43	RCL	
46	13	13	
47	71	SBR	
48	02	2	
49	00	00	
50	42	STO	
51	18	18	
52	43	RCL	
53	13	13	
54	65	x	
55	43	RCL	
56	14	14	
57	71	SBR	
58	02	2	
59	00	00	
60	44	SUM	
61	17	17	
62	43	RCL	
63	17	17	
64	65	x	
65	43	RCL	
66	11	11	
67	39	2nd cos	
68	95	=	
69	42	STO	
70	01	1	
71	43	RCL	
72	11	11	
73	39	2nd cos	
74	65	x	
75	43	RCL	
76	18	18	
77	75	-	
78	43	RCL	
79	11	11	
80	38	2nd sin	
81	65	x	
82	43	RCL	
83	10	10	
84	95	=	
85	42	STO	
86	02	2	
87	43	RCL	
88	11	11	
89	39	2nd cos	
90	65	x	
91	43	RCL	
92	10	10	
93	85	+	
94	43	RCL	
95	11	11	
96	38	2nd sin	
97	65	x	
98	43	RCL	
99	18	18	
100	95	=	
101	42	STO	
102	03	3	
103	43	RCL	
104	11	11	
105	38	2nd sin	
106	65	x	
107	43	RCL	
108	17	17	
109	95	=	
110	94	+/-	
111	42	STO	
112	04	4	
113	36	2nd Pgm	
114	04	04	
115	18	2nd C'	
116	65	x	
117	43	RCL	
118	10	10	
119	95	=	
120	91	R/S	
121	32	x⇄t	
122	65	x	
123	43	RCL	
124	10	10	
125	95	=	
126	91	R/S	
127	43	RCL	
128	10	10	
129	49	2nd Prd	
130	01	1	
131	49	2nd Prd	
132	02	2	
133	43	RCL	
134	01	1	
135	42	STO	
136	03	3	
137	43	RCL	
138	02	2	
139	42	STO	
140	04	4	
141	43	RCL	
142	10	10	
143	22	INV	
144	44	SUM	
145	01	1	
146	44	SUM	
147	03	3	
148	36	2nd Pgm	
149	04	04	
150	18	2nd C'	
151	36	2nd Pgm	
152	05	05	
153	12	B	
154	42	STO	
155	01	1	
156	85	+	
157	01	1	
158	95	=	
159	55	÷	
160	53	(	
161	01	1	
162	75	-	
163	43	RCL	
164	01	1	
165	54	)	
166	95	=	
167	91	R/S	
168	61	GTO	
169	00	0	
170	06	6	
200	33	x^2	
201	55	÷	
202	53	(	
203	43	RCL	
204	13	13	
205	33	x^2	
206	85	+	
207	43	RCL	
208	14	14	
209	33	x^2	
210	54	)	
211	95	=	
212	92	INV SBR	

MERGED CODES: 62, 63, 64, 72 STO, 73 RCL, 74 SUM, 83 GTO, 84, 92 INV SBR

TEXAS INSTRUMENTS INCORPORATED

fig. 1. TI-58/59 program for calculating antenna impedance and vswr from RX noise bridge measurements.

Eq. 3 gives the electrical length, λ', in radians.

$$\lambda' = \frac{0.02095lf}{v}\ (meters) = \frac{0.006386lf}{v}\ (feet) \qquad (3)$$

Where

l = physical line length in meters or feet
f = frequency in MHz
v = line velocity factor

The factor ($0.02095/v$ or $0.006386/v$) is entered in register 16. For RG-58/U ($v = 0.66$), this factor is 0.03174 for metric lengths (0.009675 for feet). A subroutine (steps 200-212) is used in the parallel-series conversions to reduce the length of the program.

using the program

You should become fairly familiar with your calculator's programming manual before you attempt to enter this program. The *TI-58* is a very complex logic system. When you are ready, do the following:

1. Turn the calculator on and enter the program. The machine's extensive editing capabilities and the program key codes will assist you if difficulties arise.

2. Convert the calculator to the radian mode (2nd Rad).

3. Enter the appropriate constants (see program) in registers 15, 16, and 19.

4. Press CLR and RST.

5. In order: enter characteristic line impedance in ohms, line length in meters (feet), frequency in MHz, bridge resistance, bridge capacitance, and range extender status, pressing R/S after each entry. (Range extender status = 1 if the extender was used, 0 if not).

6. After a few seconds the calculator will display the series resistance at the antenna; record the value and press R/S.

7. Next the calculator will display the series reactance at the antenna; record the value and press R/S.

8. Finally, the calculator will display the vswr.

9. If you are using the same transmission line at a new frequency, enter the new frequency in MHz, the bridge resistance, the bridge capacitance, and the range extender status pressing R/S after each entry. The results are displayed as before. If you are using a different transmission line go to **step 4.** (Remember to change register 16 if the velocity factor of the line is different.)

It is very important to verify correct entry of the program before use. You can do this by runnning a couple of cases from **table 4** of the noise bridge article.[1] Choose one case using the adaptor and one

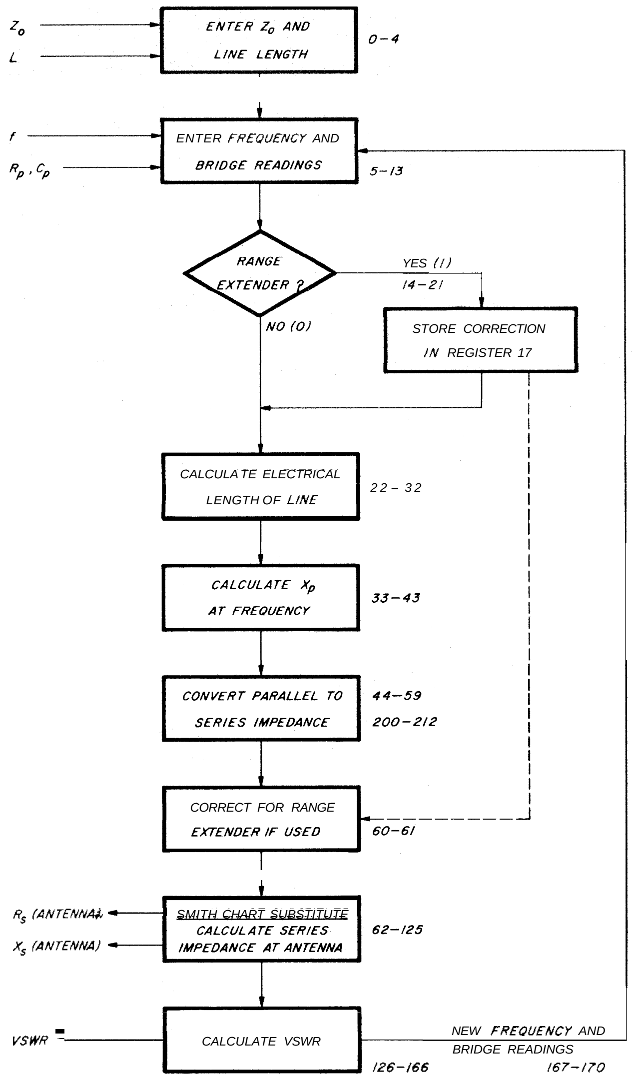


fig. 2. Flow chart far the TI-58/59 program for calculating antenna impedance and vswr.

without. If you run into trouble, check the key codes versus those appearing in the program.

summary

I have used this program with great success to tune my 20-meter array. What would have taken weeks to accomplish with Smith charts required only a couple of evenings with the Programmable *TI-58* calculator. As you become familiar with the calculator and the program, you can adapt them to your own special needs. I'm sure you'll be very impressed with the calculating power this system provides.

references

1. Robert A. Hubbs, W6BXI and Frank Doting, WGNKU, "Improvements to the RX Noise Bridge." *ham radio,* February, 1977, page 10.
2. Henry Jasik. *Antenna Engineering Handbook,* 1st edition, McGraw-Hill, New York. 1961, pages 31-3 and 31-8.

ham radio

tree-mounted ground-plane antenna for 80 meters

Using elevated ground-plane antennas you eliminate the need for an extensive ground radial system

Even though winter has passed, and solar activity is on the rise, 80 meters always provides excitement for the nocturnal amateur who craves DX. Many amateurs have a desire to erect an antenna that will punch through the pile-ups better than the standard dipole or inverted V. And, people's thoughts ultimately turn to the vertical antenna with its inherent low angle of radiation. However, most amateurs discard this idea because of the necessity for an extensive ground radial system. Having faced the same perplexing problem, a viable solution was found in Orr's *Radio Handbook*' — his description of a low-band ground-plane antenna. Its simplicity and four elevated ground radials readily pointed to tree mounting the antenna, and with several tall trees on the property, my solution was near at hand.

My initial fear of rf absorption by the foliage and oblique branches was dispelled by the realization that at the 80-meter frequencies the wavelength is much too long. There would be some inductive coupling to the tree trunk, which is conductive to a degree, especially when the sap is flowing, but the effect is negligible.

The first hurdle was mounting the 18.9 meter (62 foot) vertical section 4.6 meters (15 feet) above the ground in a maple tree only 18.3 meters (60 feet) tall. I came to the conclusion that since Marconi's original grounded antenna was in the shape of an inverted L, my tree-mounted version could have a bent top section as well. An added benefit of the bent top is the additional capacitance which makes the antenna appear electrically longer.

construction

The first ground plane was constructed in my basement using no. 12 AWG (2.1 mm) insulated wire for the vertical section and ground-plane radials **(fig. 1)**. From past experiences, I'd learned that insulated wire considerably reduces precipitation static and has no noticeable effect on the radiation or reception of signals.

I designed the antenna for 3.8 MHz by using the standard quarter-wavelength vertical formula of:

$$\textit{vertical element length} = \frac{71.3}{f(MHz)} \text{ meters or } \frac{234}{f(MHz)} \text{ feet}$$

Since the horizontal portion tended to top load the antenna, the antenna resonated at 3785 kHz, slightly lower than the design frequency. This was of little consequence because of the broadband characteristics of the antenna.

The four radials were cut 2-1/2 per cent longer than the vertical radiator. Each was soldered to a mounting hole of an SO-239, with the quarter-wave vertical section soldered to the center pin of the connector. For mounting and tying down, small ceramic insulators were attached to the ends of the ground radials and quarter-wavelength vertical section.

mounting and installation

Mounting the antenna in the maple tree was much easier than I had anticipated. After deciding where the antenna would fit with no obstructions to the spreading ground radials, I used the bow and arrow technique to thread a small fishing line over the appropriate branch. Adding weight to the tip of the

By Joseph D. Liga, K2INA, 1 Stirrup Lane, Eatontown, New Jersey 07724

arrow will ensure a vertical descent and prevent hang-up in the branches.

With 36.5 meters (120 feet) of plastic clothes line attached to the monofilament line and also the radiator of the antenna, it was then easy to pull the rope through and subsequently raise the antenna base 4.6 meters (15 feet) above the ground. As a result, the top portion of the quarter-wave section extends about 4.6 meters (15 feet) from the top branches, slanted towards the point of tie down. I attached the clothes line as high as possible in a nearby tree to minimize the slant of the bent top.

Since most of the antenna current flows in the lower half of the vertical quarter-wavelength section, the entire upper half can be bent over without serious degradation. It should be noted that the longer the flat top or bent portion, the greater will be the top loading, and consequently, the lower the resonant frequency.

The radials were slanted to tie points about 1.5 meters (5 feet) above the ground, and spaced approximately 90 degrees apart. They also serve as guys to stabilize the base of the antenna, but should not be brought too close to the ground because the electric field must be confined between the vertical radiator and the groundplane, rather than the lossy ground. The return current will then flow in the highly conductive groundplane radials with little power

The first ground plane antenna, but with six radials instead of the original four. The spiral at the base of the antenna is plastic line taped to the base portion for reinforcement.

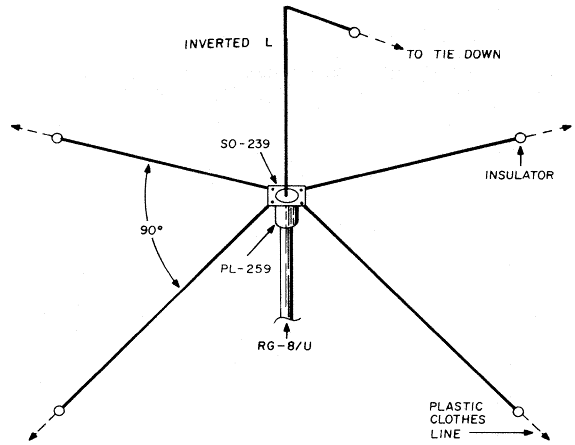


fig. 1. Diagram of one of the tree-mounted ground-plane antennas. The radials are spaced approximately 90 degrees apart and slope at an angle of about 20 degrees from the horizontal.

loss. This is an important consideration close to the base of the antenna, since the current is at a maximum. If the base of the groundplane must, from necessity, be located a short distance above the ground, additional radials (at least four more) must be installed for efficient operation of the antenna. Even spacing of the radials provides a more effective groundplane.

A few words of caution are in order here. The end of each radial is at a high rf voltage point. The position where a radial is tied down should not be in a location where it can be accidentally touched.

base impedance

The impedance at the base of a ground-plane antenna is about 35 ohms with horizontal groundplane radials, but will increase as the radials are sloped downward. A slope of 45 degrees from the horizontal will raise the base impedance to about 50 ohms. However, a 45-degree slope is quite difficult to achieve in an 80-meter tree-mounted version. My radials slope at an angle of 20 degrees, producing a base impedance of 40 ohms. Attempts to lower the radials or increase the droop did not significantly change the impedance. By using a 50-ohm, low loss, coaxial cable, the 1.25:1 mismatch is of little consequence.

initial testing

The tree-mounted, ground-plane antenna performed beyond my expectations. It loaded up beautifully and the swr at the end of a one-half wavelength of transmission line was close to the expected 1.25:1. At the band edges, the swr did not exceed 2:1, which verified the broadband characteristic of the antenna.

On the first night of extensive on-the-air tests,

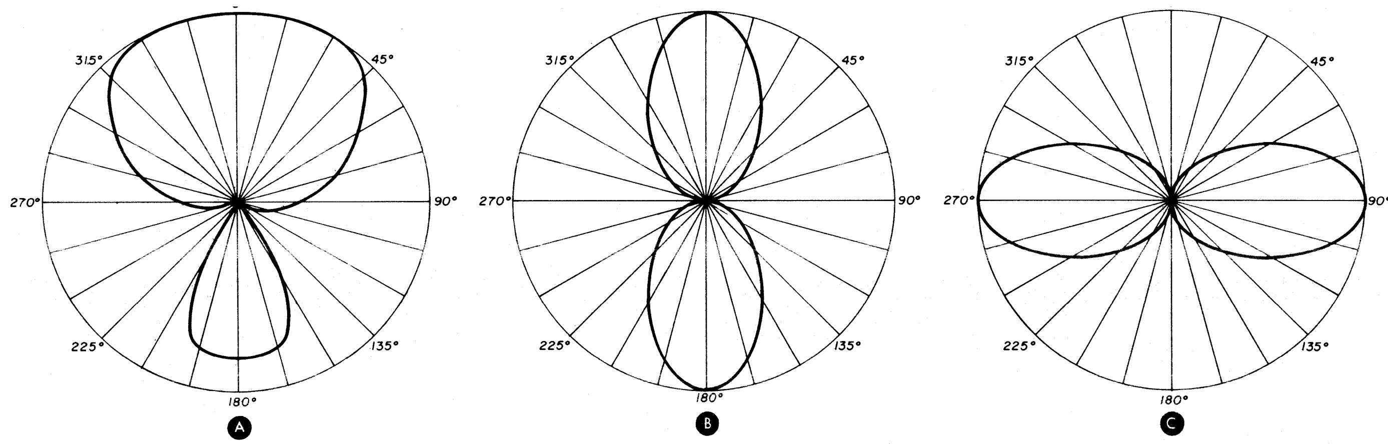


fig. 2. Horizontal radiation patterns for the pair of phased ground planes, spaced one-half wavelength apart. Pattern A is with a phase difference of 90 degrees, B with 180 degree phase difference, and C, no phase difference.

contacts into Europe showed the ground-plane to be one S-unit stronger than my sloping quad antenna.[2] Subsequent European contacts, and a later contact with ZL2BT, confirmed the superior performance of the ground-plane antenna.

During several weeks of operating with a single ground-plane antenna, I found that the horizontal portion of the vertical element provided an advantage I had not anticipated. It contributed sufficient high angle radiation to permit good short skip contacts. Although the sloping quad was the superior antenna for short and intermediate skip, the ground-plane held its own.

phasing distance

I followed the reasoning that if one ground-plane antenna worked well on **80** meters, then two antennas should perform better, especially a pair properly phased. Using this as the basis, I began pacing distances between the tall trees on the property, and found another maple **39.6** meters **(130** feet) from the tree supporting the ground-plane antenna. At **3.8** MHz, **39.6** meters **(130** feet) is approximately one-half wavelength since:

$$\lambda/2 = \frac{150}{f(MHz)} \text{ meters or } \frac{492}{f(MHz)} \text{ feet}$$

The second maple tree was over **16.5** meters **(55** feet) tall, and before long, another ground-plane antenna was constructed and installed in the same manner as the first.

phasing the antennas

Since simplicity was the byword in constructing and mounting the antennas, I planned a simple system of phasing. By design, I cut the length of the two transmission lines to make them **90** degrees out of phase. In this case, the length of coaxial line to the first antenna was **112** wavelength at **3.8** MHz, and the length of line to the second was **314** wavelength at the same frequency. After initially cutting the coaxial lines a little long, I grid dipped and pruned the lines to their exact electrical lengths.[3] With both coaxial lines fed from a common feedline, the antennas were then **90** degrees out of phase. This relationship produced a pattern with a broad forward lobe to the Northeast, and a minor lobe to the Southwest as shown in fig. **2**.[4] Now it was just a simple matter of cutting a half-wave phasing line to make the antenna system perform the same in the opposite direction.

One method of phasing antennas is to have equal electrical lengths of coaxial line from each antenna to the operating position. Then it is a simple procedure to use a quarter wavelength coaxial line to switch directivity. With the high cost of coaxial cable, it may be desirable to space the antennas one-quarter wavelength apart, and use a quarter wavelength of coaxial line from each antenna to a coaxial relay. This method permits the use of a single coaxial line from the relay to the operating position. Also, when quarter-wave spacing is used, the horizontal plane radiation is in the form of a cardioid (broad heart-shaped) pattern, providing a better front-to-back ratio as shown in fig. 3.

matching

As mentioned earlier, the impedance measured at the base of each antenna was **40** ohms. Feeding the antennas with 50-ohm coaxial lines presented no major problem when each antenna was used separately. However, when the two coaxial lines were joined at the coaxial-tee connector, the resultant impedance was about 20 ohms. The 2.5:1 mismatch presented no problem in antenna operation or amplifier loading. In addition, the loss caused by an swr of 2.5:1 in 30 meters (100 feet) of RG-8/U at 4 MHz, is a fraction of a dB and can be ignored.5

For those who do have loading problems, it might be desirable to use a 70-ohm coaxial line from each antenna to the coaxial tee connector. This line must be one-quarter wavelength long or odd multiples of a quarter wavelength. Each line acts as a quarter-wave matching transformer (*Q* section). Assuming a 50-ohm base impedance, the matching sections transform this up to 100 ohms, giving you a 50 ohm impedance when the lines are joined together.

performance

After three years of constant use, the phased ground-plane system consistently outperformed my sloping quad and new delta loop.[6] Despite the minor lobe to the rear, I've experienced a 10 to 15 dB front-to-back ratio. The very broad major lobe provides approximately 120 degrees of effective coverage. The minor lobe to the rear, although contributing to some QRM and noise from the undesired direction, can be used to an advantage when rapid switching of direction is not possible.

Receiving with the phased ground-planes was a new experience. Unreadable signals on the loop antennas became Q-5 on the phased ground-planes. Precipitation static was never bothersome during the heaviest rain or snowfall. The capability of the antennas to accept high-angle as well as low-angle signals provided additional versatility when simultaneously operating with stations close by and far off. The very broad major lobe contributes little front-to-side rejection. This characteristic provides advantages as well as disadvantages. I used a quarter-wavelength phasing line to change the radiation pattern to an in-phase broadside array, or a 180 degree out-of-phase, end-fire array as shown in **fig.** 2.

A loop antenna normally accepts less atmospheric noise than a dipole or similar antenna, due to the noise cancelling effect of the closed loop. However, over a period of three noisy summers, I experienced a greater signal-to-noise ratio on the phased ground-planes than with the loop antennas. The phased ground-planes are certainly more susceptible to man-made noise, due to its predominant vertical polarization.

conclusion

As can be expected with any antenna system, improvements are attempted. Some succeed, while others do not. I found that increasing the number of radials from four to twelve slightly improved antenna performance when transmitting. On receiving, the improvement was quite noticeable from a standpoint of signal-to-noise ratio. The base impedance of the antennas decreased slightly due to the increased current in the additional radials.

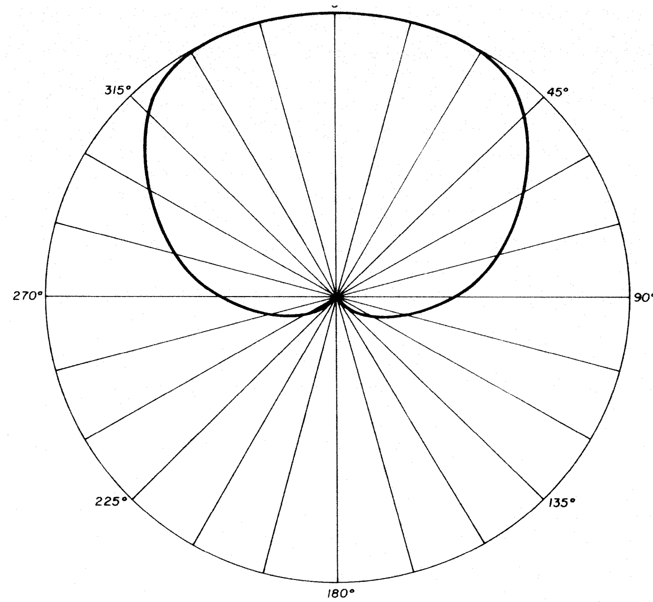


fig. 3. The unidirectional cardioid pattern is produced by spacing the antennas one-quarter of a wavelength apart and using a 90 degree phasing line.

I also tried connecting the ends of the sloping radials where they almost overlap between antennas. The rationale was to provide a common ground-plane. This ,experiment was a failure, since each antenna must operate independently, coupled only by their fields.

It would appear that the antennas are a bit unsightly with the radials fanning out in all directions. On the contrary, I had to point out the antennas to several visiting amateurs who had difficulty spotting them from the operating position.

And finally, I want to express my gratitude to all the US and foreign amateurs who patiently contributed their time during the months of checking and testing the antenna system.

references

1. William Orr, W6SAI, editor, *Radio Handbook,* Howard *W.* Sams and Co , Inc , Indianapolis, Indiana, 19th edition, 1972.
2. Joseph Liga, K2INA, "A Sloping Quad for 80 Meters," *CQ,* April, 1974, page 46
3. Kenneth Glanzer, K7GCO, "Determining Resonant Lengths of Transmission Lines," *CQ,* May, 1974, page 37.
4 G H Brown, "Directional Antennas." *Proceedings of the IRE,* January, 1937.
5. Walter Maxwell, W2DU, "Another Look At Reflections," ***QST,*** April, 1973, page 35.
6 L. V Mayhead, G3AQC, "Loop Aerials Close to the Ground," *Radio Communication,* May, 1974, page 298.

ham radio

design of
omega-matching networks

A straightforward procedure for designing antenna networks

The omega network has been used for years, mostly by amateur experimenters, to match a coaxial line or open-wire feedline to a thin, linear antenna element as shown in fig. **1.** Although this matching network looks somewhat like the more common gamma network, two capacitors are used. In this illustration L is the length of one-half of the driven dipole element (or total length of a monopole element). The dimension l_Ω is the length of the omega-matching rod and is the same as a gamma matching rod; S is the center-to-center spacing between the omega rod and a parallel driven element.

The reactance X_Ω is the shunt omega capacitor at the omega-rod feedpoint. X_C is the series capacitor used to tune out any reactance in the antenna system's input impedance. R_o is the characteristic resistance of the coaxial transmission line, which is assumed, for the purposes of this discussion to be lossless.

The practical omega match shown in fig. 1 (less capacitor X_C) can be represented by the simple block diagram of fig. 2. In this diagram H_z is the impedance step-up ratio, Z_a is the antenna impedance, X_S is the reactance of the omega rod, and X_Ω is the reactance of the omega capacitor. The antenna impedance step-up ratio, H_z, is a function of the diameters of the omega rod and driven element, as well as the spacing between them and is given by the following formula:[1]

$$H_z = \left[1 + \frac{\left(\cosh^{-1} \dfrac{4S^2 - D^2 + d^2}{4Sd}\right)}{\left(\cosh^{-1} \dfrac{4S^2 + D^2 - d^2}{4SD}\right)} \right]^2$$

where D is the diameter of the driven element, d is the diameter of the omega rod, and S is the spacing between the two. This factor is plotted in normalized form in fig. 3 for omega match designs. In plotting this graph it was assumed that H_z is at least 4:1, a realistic assumption in most amateur applications.

The quantity Z_a is one-half the total input impedance of a balanced antenna and is equal to

$$Z_a = \frac{Z_a'}{2} = R_a \pm jX_a \ ohms$$

where the quantity Z_a' is the antenna driving point impedance.

The reactance of the omega rod, X_S is equal to $jZ_o \tan kl_\Omega$ ohms when the quantity kl_Ω is less than 90 electrical degrees as will be assumed here and $k = 2\pi/\lambda$ radians per meter. The impedance of the omega rod in air, Z_o is given by

$$Z_o = 60 \cosh^{-1} \frac{4S^2 - D^2 - d^2}{2Dd} \ ohms$$

where

D = diameter of antenna driven element
d = diameter of omega rod
S = center-to-center spacing between the omega rod and the parallel antenna driven element

Fig. **4** shows a graph of Z_o plotted as a function of normalized values of S/D and D/d.

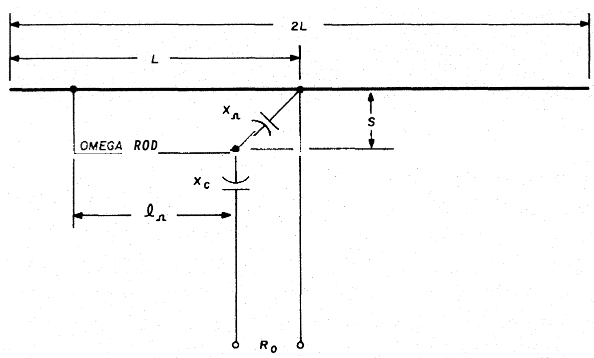


fig. 1. Layout of the basic omega match. L is one-half the length of the driven dipole. X_Ω is the omega capacitor, and X_C compensates for the inductive reactance component appearing at the omega rod input point.

By Harold F. Tolles, W7ITB, Box 232, Sonoita, Arizona 85637

The length of the omega rod is approximately

$$l_\Omega - \frac{(kl_\Omega)^o}{0.01233 f_{MHz}} \text{ cm} = \frac{(kl_\Omega)^o}{0.03188 f_{MHz}} \text{ znches} \quad \textbf{(1)}$$

From **fig.** 2 the complex antenna system input impedance, Z_i, may be written as

$$Z_i = M \angle \psi \text{ ohms} \quad \textbf{(2)}$$

Where

$$M = \frac{(X_\Omega X_s H_z \sqrt{R_a^2 \; t \; X_a^2})}{\sqrt{[H_z R_a(X_\Omega + X_s)]^2 + [H_z X_a(X_\Omega + X_s) + X_\Omega X_s]^2}} \text{ ohmr}$$

$$\psi = \tan^{-1}\left[\frac{(X_\Omega + X_s)H_z(R_a + X_a)}{X_\Omega X_s R_a} + \frac{X_a}{R_a}\right] \text{ degrees}$$

For **eq.** 2 to match a high-frequency lossless coaxial transmission line with a characteristic resistance, $R_{,,}$

$$M \cos \psi = R_o \text{ ohms} \quad \textbf{(3)}$$

By using the above substitutions and manipulating, the solution may also be written as a quadratic in Q whose positive discriminant root is

$$Q = A + \sqrt{A^2 + B} \text{ ohms} \quad \textbf{(4)}$$

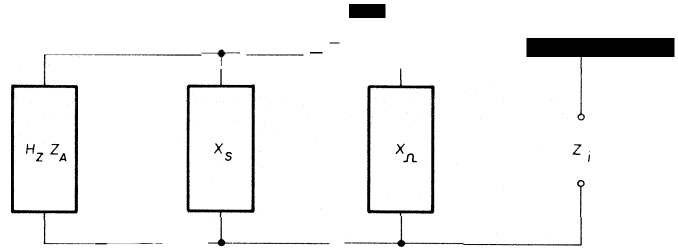


fig. 2. Equivalent circuit of the omega-matching system shown in fig. 1. Z_a is one-half the dipole input impedance, H_z is the impedance step-up ratio, X_s is the reactance of the omega rod, and X_Ω is the reactance of the omega capacitor.

Where

$$Q = \frac{X_s X_\Omega}{H_z(X_s + X_\Omega)}$$

$$A = \frac{R_o X_a}{H_z R_a - R_o}$$

$$B = \frac{R_o(R_a^2 + X_a^2)}{H_z R_a - R_o}$$

Equation 4 has the restriction that the negative sign of X_Q must be larger than the positive sign of X_s (in this assumed case) and that

$$H_z > \frac{R_o}{R_a} \text{ numeric}$$

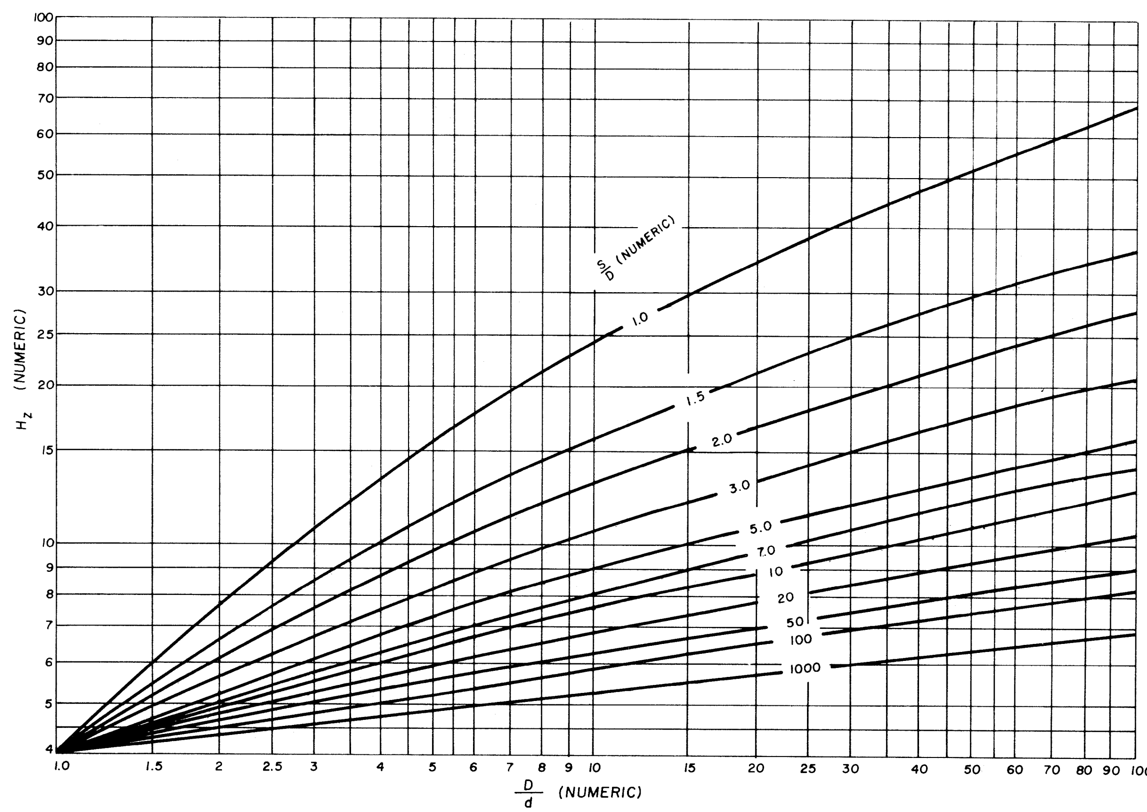


fig. 3. Matching network's impedance step-up ratio, H_z, as a function of element diameters and spacing. D is the diameter of the dipole element, d is the diameter of the omega rod, and S is the center-to-center spacing between them.

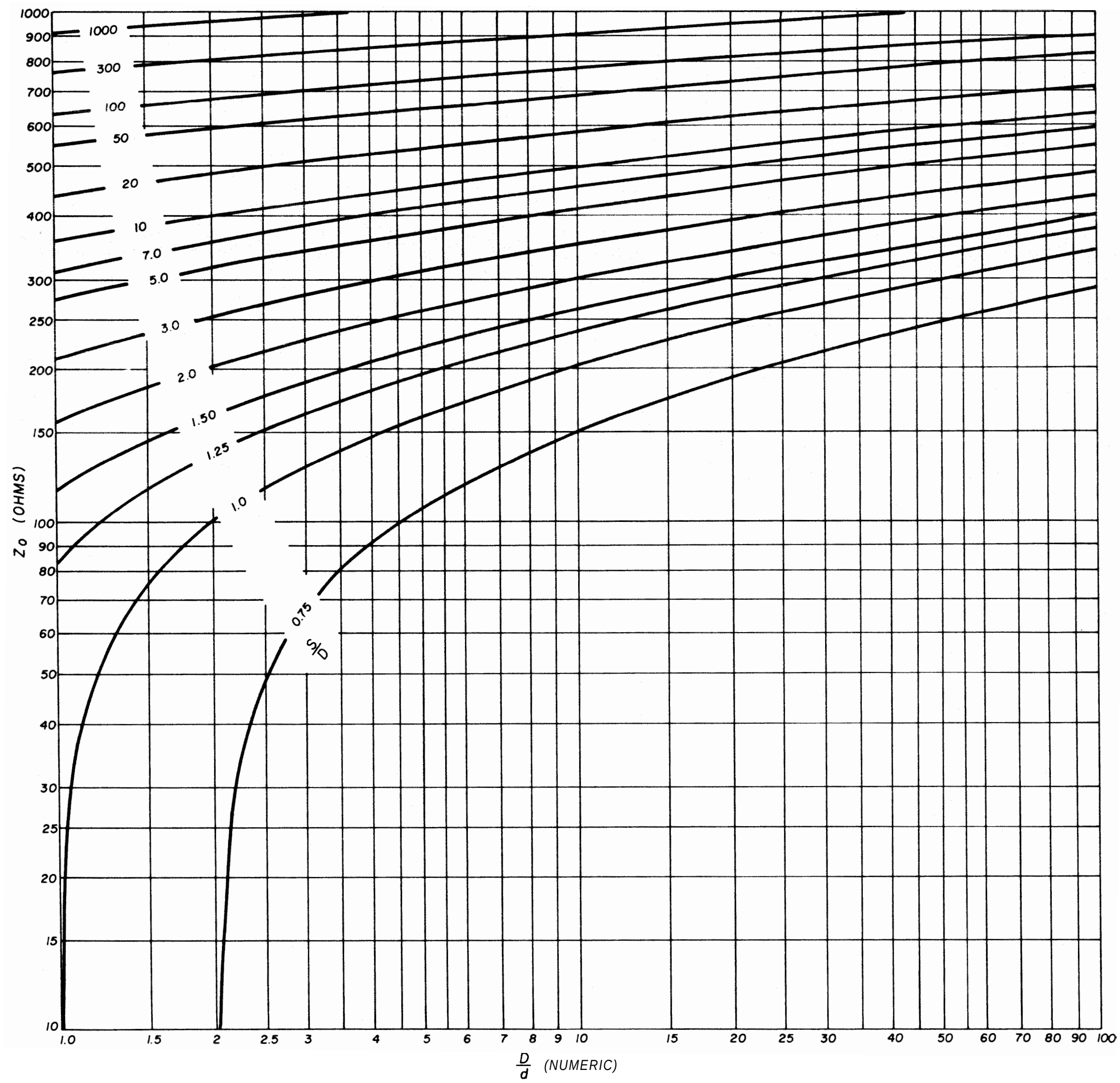


fig. 4. Characteristic impedance of the omega rod, Z as a function of element diameters and spacing.

In the above definition of Q, when X_Ω is infinite (open circuit), the solution of Q is exactly the same as that for a gamma-matching network[1] because it can be seen from inspection that when $X_\Omega = \infty$, the gamma and omega matching networks are identical.

Once Q in **eq. 4** has been determined as a function of A and B (i.e., R_o, H_z, R_a, and X_a) the solution for X_Ω and X_s appears to be an arbitrary one of two circuit elements X_Ω and X_s in parallel or

$$H_zQ = k' = \frac{X_sX_\Omega}{X_s + X_\Omega} \text{ ohms } |X_\Omega| > |X_s| \qquad \textbf{(5)}$$

A graph of this equation, for various values of H_zQ, is plotted in **fig. 5.** In **fig. 6** is plotted a graph of the omega rod reactance X_Ω as a function of the impedance of the omega rod, Z_o, the omega rod length kl_Ω.

For **eq.** 2 to provide a reactive component equal in magnitude but opposite in sign to X_C in **fig. 1**

$$M \sin\psi = -X_C \text{ ohms} \qquad \textbf{(6)}$$

Combining this with **eq.** 3 yields

$$X_C = -R_o \tan\psi$$
$$= -\frac{R_o}{R_a}\left[\frac{(X_\Omega + X_s)H_z(R_a^2 + X_a^2)}{X_\Omega X_s} + X_a\right] \text{ ohms}$$

And, combining with 4,

$$X_C = -\frac{1}{2\pi f C_C}$$
$$= -\frac{R_o}{R_a}\left[\frac{R_a^2 + X_a^2}{Q} + X_a\right] \text{ ohms}$$

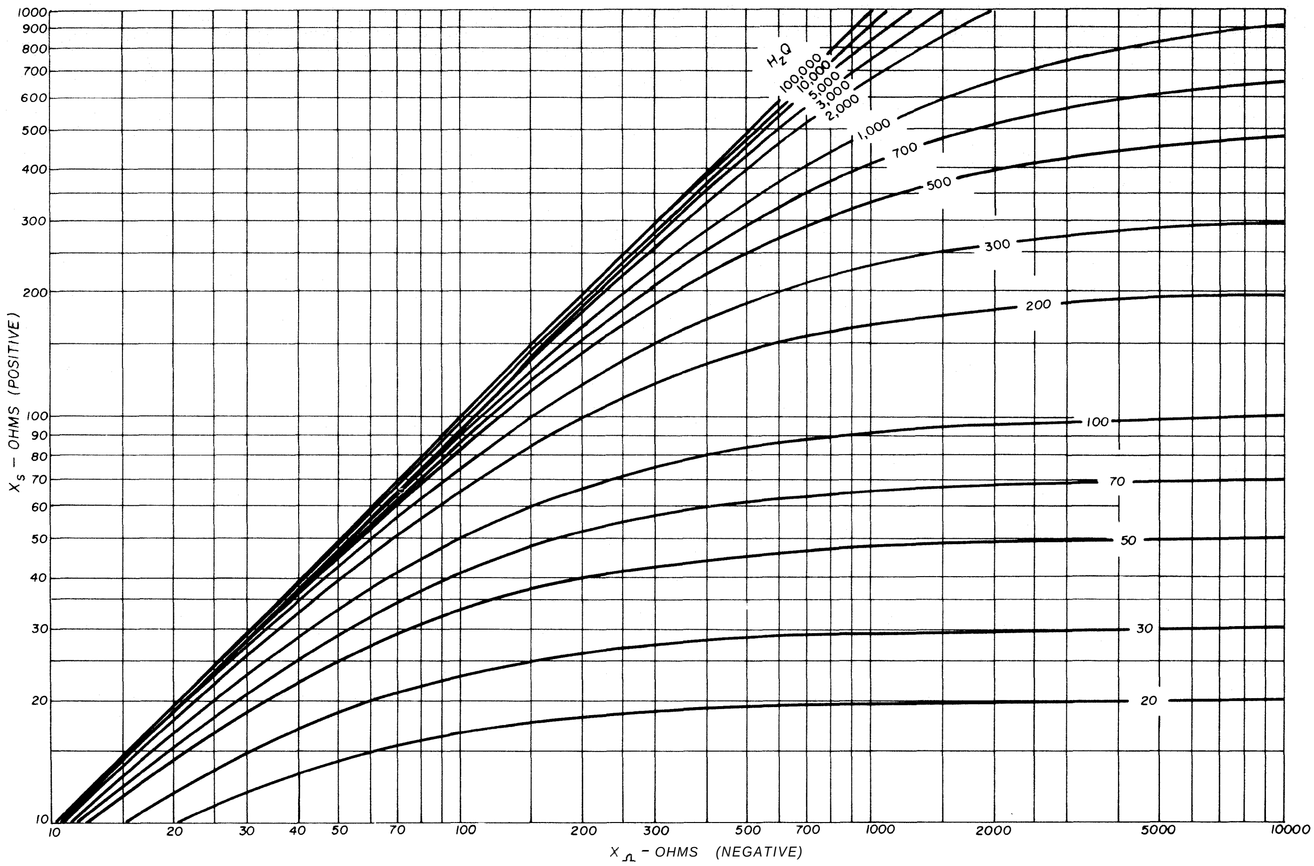


fig. 5. Relationship of the reactance of the omega rod, X_s, and reactance of the omega capacitor, X_Ω, as function of the H_zQ product (see text).

From which

$$C_C = \frac{10^6}{2\pi(E+F) f_{MHz}} pF \tag{8}$$

Where

$$E = \frac{R_o(R_a^2 + X_a^2)}{R_a Q} \; ohms$$

$$F = \frac{R_o X_a}{R_a} \; ohms$$

design procedure

When designing an omega-matching network it might seem that the following procedure could be used with good success:

1. Use the graph of **fig.** 3 to select a transformation ratio H_z *vs* the desired S/D and D/d where $H_z > R_o/R_a$.

2. Use **fig. 4** to determine the impedance of the omega rod, Z_o from the S/D and D/d ratios selected in **step 1.**

3. Use **fig.** 6 to determine a practical, *realistic* value of omega rod reactance, X_s vs the rod's impedance, Z_o, and length, kl_Ω.

4. When kl_Ω is known in degrees from **fig.** 6, the physical length of the omega rod, l_Ω, may be calculated from **eq. 1.**

5. Calculate Q from **eq. 4.** When this value of Q is multiplied by H_z, you can select the correct H_zQ line on **fig.** 5. This provides the reactance of the omega capacitor, X_Ω, at the bottom of the graph.

6. The value of the omega capacitor in pF is calculated at the frequency of interest using **eq. 8.**

The problem with this design procedure is in the selection of a "realistic value" of the omega rod reactance in **step** 3. That is, should X_s be selected large so X_Ω is large, or should X_s be selected small so X_Ω is small? In gamma-matching networks (the same omega-matching networks with $X_\Omega = \infty$, as noted above), it's desirable to make X_s large to minimize the IR losses. In the omega network the current I is

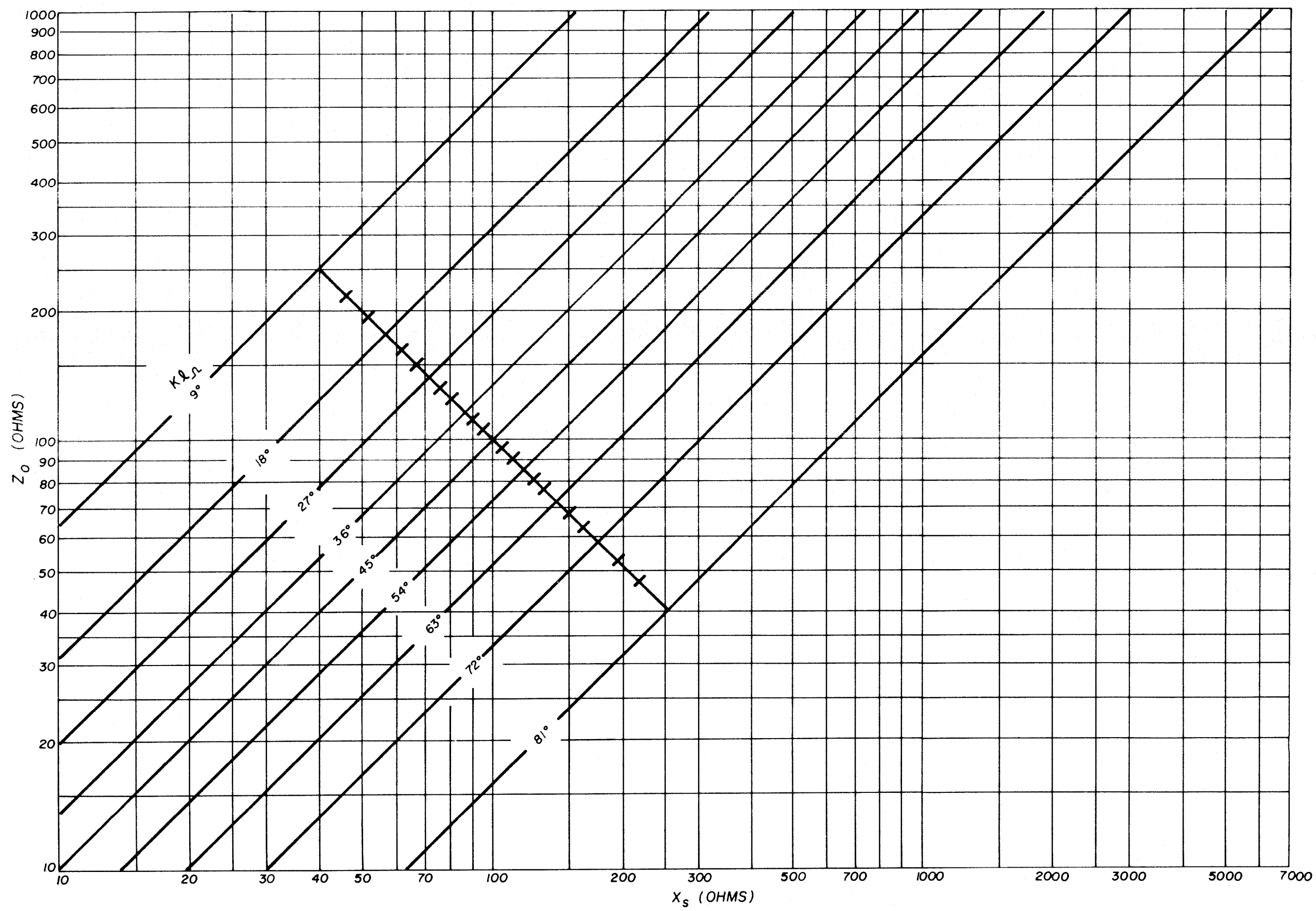


fig. 6. Reactance of the omega rod, X_Ω, as a function of the rod's characteristic impedance, Z_o, and length, kl_Ω.

divided between X_s and X_Ω so the answer may not be so readily apparent.

The answer will become apparent, however, if two vector graphs are plotted so the currents can be examined for the two cases. The two vector graphs, **figs. 7** and **8,** assume the same rf power, antenna driving point impedance, Z_a, as well as H_z and R_o. **Fig. 7** uses a gamma-matching network as an example of $X_\Omega = \infty$, and **fig.** 8 shows an omega-matching network with $|X_\Omega| = |2X_s|$.

These two graphs give complete insight into the operation of both the gamma- and omega-matching networks. For any particular antenna impedance H_zZ_a to resistively match a transmission line through a series capacitive reactance, it's necessary that the antenna impedance, together with the impedances of its shunt correcting elements, provide a complex input impedance with a positive angle; also, the magnitude of the real part must be equal to the characteristic resistance of the transmission line, and the imaginary part must be equal but opposite in sign to the series capacitive reactance. With this in mind it shouldn't be surprising to see that the magnitudes of V_{in}, I_{in}, V_{Xs}, V_{HzRa}, V_{HzXa}, and ■ of the gamma match in **fig. 7** are the same as fhose for the omega match in **fig.** 8.

The difference between the two matching networks is that I_{Xs} in the gamma network is the vector sum of I_{Xs} and $I_{X\Omega}$ in the omega network. From this it is obvious that I_{Xs} in the gamma network is less than I_{Xs} in the omega network, and that the magnitude of X_s in the gamma network is greater than the magnitude of X_s in the omega network.

The omega network is easier to design than the gamma network because the only restriction upon X_s in the omega network is that it must be *less* in magnitude than it is in the equivalent gamma network. In the omega network the difference between the two values of X_s can be compensated by adjusting the value of X_Ω. Because of this, the rod length kl_Ω in the omega network can be made shorter than kl_Γ of the equivalent gamma network.

At the lower operating frequencies where the length of the rod is apt to be very long, the omega network has merit. However, it is likely that the efficiency and bandwidth of the omega network will be less than the equivalent gamma network because of the higher IR losses in omega network's reactance

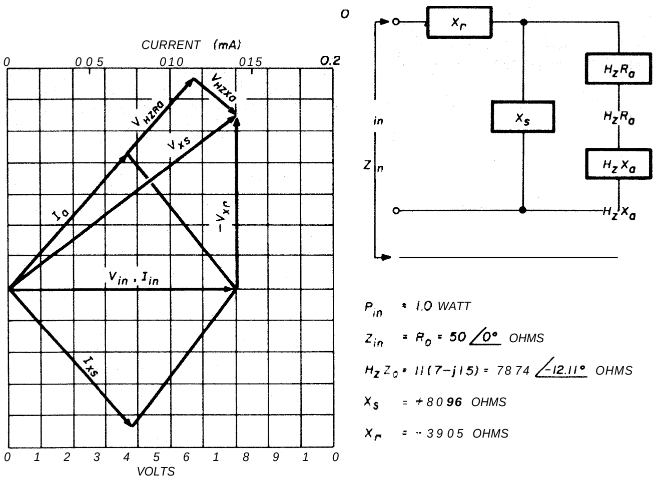


fig. 7. Voltage and current distribution of a **gamma**-matching network designed to match a 50-ohm coaxial line to a 10-element beam ($Z_a = 7 - j1.5$ ohms). Voltage and current values are based on a power input of ▮ watt.

X_s. Any predication on the fact that these reactive circuit elements are lossless is rather naive!

The double omega (or what I call the theta network) can be used to match a balanced two-wire transmission line to the driven element of a balanced dipole. In this case the balanced line characteristic resistance R_o' is halved to R_o, and the solutions to *each* arm are X_r, X_Ω, and X_C in terms of R_o, H_z, R_a, and X_a. The results are merely imaged or flipped over to the other arm of the theta network.

reference

1. Harold Toiles, W7ITB, "Gamma Match Design," *ham radio,* May, 1973, page 46

ham radio

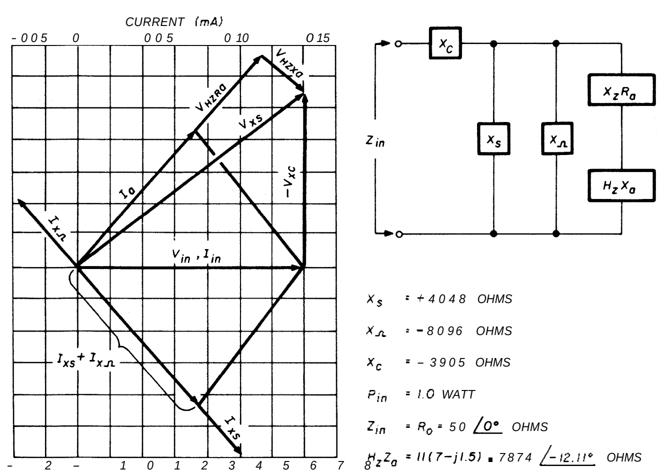


fig. 8. Voltage and current distribution of an **omega**-matching network designed to match a 50-ohm coaxial line to the same load as the gamma match in fig. 7 ($Z_a = 7 - j1.5$ ohms). Vnltage and current values are based on a power input of 1 watt. Note that I_{xs} in the omega network is greater than I_{xs} in the gamma network when the rod is too short to use a gamma network. So, current (and IR loss) is greater in the omega match.

improved indicator system

for the Hy-Gain 400 antenna rotator

With this circuit you can read antenna pointing direction during antenna rotation

The designers of the Hy-Gain 400 rotator apparently didn't realize that most amateurs want to know which way their antenna is pointing at any moment, and an indicator was provided that didn't provide this information during antenna rotation. The *ARRL Antenna Book* (13th edition) contains information on a replacement directional indicator that certainly will work. However, it requires an 8-wire control cable whereas the original indicator required a 5-wire cable and Hy-Gain supplied such a cable suitable for the 110 volts that the rotator motor requires. So it was rather frustrating to read about an approach that's available and convenient for those who had not yet installed the rotator on a tower, but less so for those who had already installed the rotator and wished to change the indicator.

This article describes a replacement direction-indicating control unit that involves no modifications of the rotator and is compatible with the original 5-wire cable. It has an additional feature that permits control from multiple locations — a convenience for those who operate from several locations as I do. The ideas in this article are adaptable to other types of rotators where multiple control is wanted. Further, it's one way to get a direction indication from a rotator if one end of the indicator pot has become disconnected or if a failure has occurred in the indicator circuit. The latter failure is probably rare but is mentioned as an emergency or permanent solution that could still make use of the original indicator meter.

The crucial matter is that the indicator resistor in the Hy-Gain 400 rotator was wired as a rheostat (even though it was physically a three-terminal pot). If the familiar ohmmeter circuit is used to read the rheostat setting, and hence antenna heading, the scale will be grossly nonlinear and not very satisfactory. The cure is, therefore, to supply the rheostat with a constant-current source and then read the voltage across it, which will then be directly proportional to the angle of rotation.

constant-current supply

A recent advertisement gave the clue to a conveniently available constant-current source, the popular voltage-regulator ICs, which are used to provide a regulated 5, 12, or 15 volts. A single resistor connected between what is normally the output terminal and the normal ground — now using the ground as the constant-current output — does the trick. The current is constant *only* if the supply voltage is high enough. In this case, 24 Vdc was required to supply 10 mA constant current through a resistor varying between 0-1000 ohms. (This is the value of the rheostat in my Hy-Gain 400; but watch it, as the *ARRL Antenna Book* cites 5k and there may have been a design change.) A value of 1.2k between terminals 2 and 3 of a 7805 IC establishes the 10-mA current level. This turned out to be a very convenient value, as it causes the voltage across the rheostat to vary between 0-10 volts.

Next, to preserve linearity, a high-impedance 10 Vdc meter is needed; I borrowed a simple vtvm circuit using a single fet from Radio Shack's book *Transistor Projects Vol. 2.* Using a 1-mA meter and a 9-volt battery, this unit reads full scale at 10 volts.

indicator circuit

The complete indicator circuit is given in **fig. 1.** The spdt center-off switch for rotation could be replaced by push-button switches. All parts for this indicator are available from Radio Shack stores. Note the arrangement for zero and full-scale adjust of the vtvm. This can easily be dispensed with, as the calibration of the vtvm is quite stable. The vtvm isn't difficult to recalibrate with a short circuit for zero and a 1000-ohm resistor for full scale. A semi-adjustable pot is used with the regulator IC to set current at precisely 10 mA.

Fig. 1 also includes circuitry for a control and indicator for a second position. This could be multiplied for a third or more positions. For the indicator at the

By Guy Black, W4PSJ, 12317 Hanger Road, RR5, Fairfax, Virginia 22030

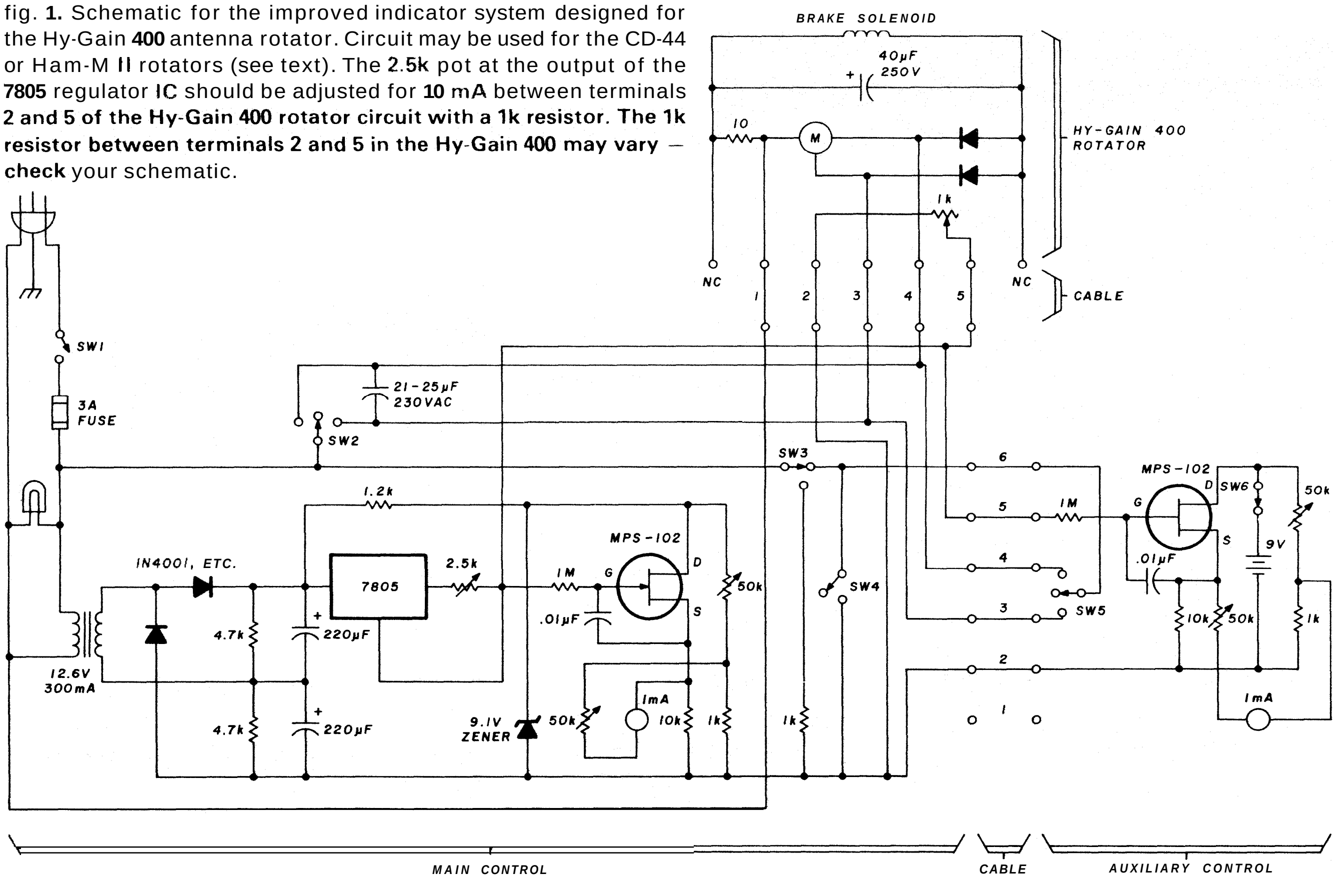


fig. **1.** Schematic for the improved indicator system designed for the Hy-Gain **400** antenna rotator. Circuit may be used for the CD-44 or Ham-M **II** rotators (see text). The **2.5k** pot at the output of the **7805** regulator **IC** should be adjusted for **10** mA between terminals **2 and 5 of the Hy-Gain 400 rotator circuit with a 1k resistor. The 1k resistor between terminals 2 and 5 in the Hy-Gain 400 may vary — check** your schematic.

quantity	part
1	Hy-Gain 400 rotator
1	transformer 115-12.6 Vac at 8.3 amp RS 273 1385 (a 24-Vac transformer could be substituted with a bridge rectifier and single filter capacitor)
2	diodes 1N4001 to 1N4007 RS 276 1101
1	voltage regulator IC such as 7805, 7812, 78L12AC (the latter is cheapest as RS 276-801)
1	zener diode, 9.1 V 1 watt 1N7439 RS 276 562
1 (2)	fet transistor MPS 102 or RS 276 2028
1	semiadjustable potentiometer 2.5k RS 271 228
2 (4)	semiadjustable potentiometers 50k RS 271 219
2	resistors 2.7k
1	resistor 1.2k

quantity	part
1 (2)	resistors 1 meg
2 (3)	resistors 1k
1 (2)	resistors 10k
2	capacitors 220 μF 35 Vdc
1	capacitor 21-25 μF 230 Vac (can be taken from original Hy-Gain 400 control)
1 (2)	.Ot mfd μF ceramic
1 (2)	milliameter 1 mA RS 22052
2	switch spst main power (SW1 & SW4)
3	switch spdt center neutral (SW2 & SW5) RS 275 325
1	switch spdt RS 275 326

Notes.

1. Quantities in parentheses are needed only if an auxiliary indicator is desired.

2. Part numbers preceded by RS are available from Radio Shack stores.

auxiliary positions, a 9-volt battery is practicable, although an ac supply may be preferred, especially by those who want a continuous reading.

Note that the ac line that goes to the auxiliary control is not the ac line that goes to the rotator itself. Allocating the rotator ac to terminal 1 on the terminal switch and the auxiliary-unit ac to terminal 6 will help avoid confusion and fireworks.

other rotators

Some change in values might be needed if the circuit were to be adapted to the CD-44 or Ham-M II rotators, which have 500-ohm pots connected to terminals 3 and 7 of their rotator units with the sliding contact grounded and connected to terminal 1. I haven't tried the circuit on these units. Although it appears feasible, it would seem to be worthwhile only if multiple-position control were wanted, in which case the vtvm circuit shown here might be used to read the voltage between terminals 1 and 3, or 1 and 7 of the CD rotators. It also seems feasible to use the meter and case of the original control and put the few extra parts inside.

ham radio

calculating antenna bearings

for geostationary satellites

Using a pocket calculator to accurately determine antenna bearings and range to geosynchronous satellites

Many members of **the** amateur community have recently "discovered" the geosynchronous communications satellite. This surge of interest is in response to a program of weather satellites now affording an outstanding view of the earth from a 22,000 mile (35,000 km) vantage point.[1] It is further spurred by the recent availability of low cost, high quality microwave receiving equipment which enables the individual to recover not only this weather data,[2] but promises reception of direct-broadcast closed-circuit TV programs.

Although a great deal of material has been published on generating tracking information for polar orbiting spacecraft such as Oscar 7,[3,4] as well as the highly elliptical orbit planned for AMSAT Phase III,[5] little has appeared in the amateur magazines regarding the geostationary, or earth-synchronous, orbit. Peter Thompson has published a set of generalized equations which could be applied to tracking satellites in a variety of orbits,[6] but the calculations are unnecessarily cumbersome when considering the simple geostationary case. Ralph Taggart has outlined an appealing method for estimating azimuth and elevation bearings with a globe and string,' but like most graphical plotting techniques, this one offers resolution limited by the finite size of the globe. The equations presented here can be readily solved on a pocket calculator and afford bearing accuracy which is limited only by your ability to point the antenna.

'If the computed value for L is greater than 180°, correct it by subtracting 360°; if the computed value is less than −180°, correct it by adding 360°.

If an imaginary line is drawn which connects the center of a satellite with the center of the earth, that line intersects the earth's surface at a location known as the Sub-satellite Point or SSP (**fig. 1**). Generating *azimuth* angles to a satellite is a function only of the location of the observer and the sub-satellite point, and is completely independent of the satellite's altitude (altitude does, of course, determine the elevation bearing, which is computed later). Hence, the azimuth data is based solely upon terrestrial coordinates; the law of cosines for determining distance and bearing coordinates in a great circle will apply.

Jerry Hall reported the relationships in his article on terrestrial great-circle computations;* those same equations may be used to calculate satellite azimuth angles:

$$\cos D = \sin A \sin B + \cos A \cos B \cos L \qquad (1)$$

$$\cos C = \frac{\sin B - \sin A \cos D}{\cos A \sin D} \qquad (2)$$

Where A = latitude of Point 1 (the observer)

B = latitude of Point 2 (the sub-satellite point)

L = longitude of Point 1 *minus* the longitude of Point 2*

D = distance between Point 1 and 2, degrees of arc

C = true bearing if L is positive

or 360 − C = true bearing if L is negative

Interestingly enough, the geostationary satellite presents a special case because the latitude of the sub-satellite point is always approximately zero degrees; under these conditions B in the above formulas is 0.

Since $\sin B = 0$ and $\cos B = 1$, these equations simplify to:

$$\cos D = \cos A \cos L \qquad (3)$$

$$\cos C = -\frac{\sin A \cos D}{\cos A \sin D} \qquad (4)$$

By trigonometric identities, **eq. 4** further simplifies to

$$\cos C = -\frac{\tan A}{\tan D} \qquad (5)$$

Thus, by solving **eqs.** 3 and **5** for D and C, respectively, antenna azimuth information can be found.

By H. Paul Shuch, N6TX, Microcomm, 14908 Sandy Lane, San Jose, California 95124

calculating elevation

If a straight line is drawn between a satellite and the center of the earth, and another straight line is used to connect the observer with the earth's center, the angle formed at the intersection of these two lines represents D, the distance between the observer and the sub-point, in degrees of arc, just computed **(fig. 2).** If the observer happened to be situated very near the center of the earth, angle D would relate directly to elevation angle; that is, ***D*** would be the angular *displacement from vertical* for aiming the antenna. Since elevation angles are generally specified with respect to the horizontal, *uncorrected* elevation data is computed from

$$EL\ (assumed) = 90 - D \qquad \textbf{(6)}$$

This assumes that the observer is located at or near the center of the earth. Since this is obviously impossible, let's consider corrections to the assumed elevation which would apply to an observer on the earth's surface.

If **eq. *6*** is used, the error is negligible when R_s, the radius of the satellite's orbit, is at least two orders of magnitude greater than r_e, the radius of the observer's orbit (the radius of the earth). Thus, **eq. *6*** is used in determining elevation information for radio astronomy, where the object being tracked is very far from earth. Under these conditions the observer *is* near the earth's center, *relative* to the object being tracked. **Eq. *6*** has also been used by EME operators for tracking the moon; although the radius of the lunar orbit is only about 50 times the radius of the earth, elevation data calculated from **eq. *6*** is correct to within a fraction of a degree.

It is interesting to note that in both of the above applications, the location of the object being tracked is generally specified in Greenwich Hour Angle (GHA) and declination. These coordinates correspond exactly to the longitude and latitude, respectively, of the sub-satellite point shown in **fig. 1.**

What do you do to correct **eq. *6*** for geostationary satellites, whose orbital radius is *not* significantly greater than the radius of the earth? A correction formula published by L. R. Larson[9] applies:

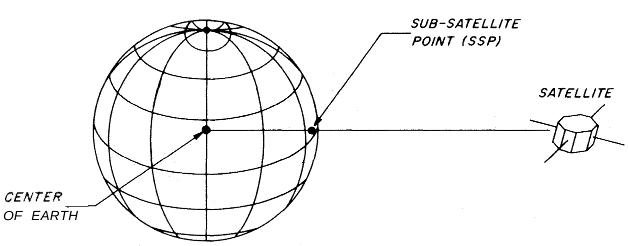


fig. 1. When a straight line is drawn from the center of the earth to a satellite, the intersection of the line with the earth's surface is known as the Sub-satellite Point or SSP. The SSP is used to calculate the azimuth bearing to the satellite.

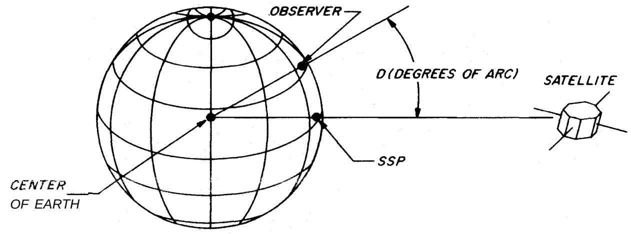


fig. 2. The observer's position on the earth's surface, in relation to the imaginary straight line shown in fig. 1, yields the angle *D*, which is used to calculate antenna elevation angle as discussed in the text.

$$tan\ EL = tan\ (90 - D) - \frac{1}{K\ cos\ (90 - D)} \qquad \textbf{(7)}$$

where K is the ratio of satellite orbital radius to earth radius

$$K = R_s/r_e$$

Since the mean radius of the earth is approximately 3444 nautical miles, and the mean radius of geostationary orbits is approximately 22,766 nautical miles, I use the value $K = 6.61$ when working with geostationary orbits. In the interest of minimizing computational steps, **eq.** 7 may be restated

$$tan\ EL = \frac{sin\ (90 - D) - (1/K)}{cos\ (90 - D)} \qquad \textbf{(8)}$$

where $\frac{1}{K} = r_e/R_s = 0.1513$ **(9)**

Eq. 8 may be further simplified

$$tan\ EL = \frac{cos\ D - (1/K)}{\sqrt{1 - (cos\ D)^2}} \qquad \textbf{(10)}$$

calculating slant range

System performance predictions, which include link calculations of signal margin, require a knowlege of the exact distance from the ground station to the orbiting satellite. One convenient equation for determining slant range is stated as follows[10]

$$Range = R_s\ x\ \sqrt{1 - 2(1/K)\ cos\ D + (1/K)^2} \qquad \textbf{(11)}$$

If you want to calculate the range in nautical miles, use 22,766 nautical miles for R_s. If you would rather compute the range in kilometers, use 42,166 km for R_s. The ratio K, of course, is the same regardless of the units used to express R_s and r_e.

Example. My station is located approximately 37.3° N, 121.9° W;* I wish to receive signals from SMS-2,

*If you know your latitude and longitude to within 1/10 degree, you have pinpointed your location to within about 6 miles!

in geostationary orbit above 135° W. The longitude difference is thus 121.9 − 135 = −13.1°. Note that L is *negative;* this information will be used later.

The distance D is found from **eq.** 3 to be 39.22°

From **eq. 5,** C is found to equal 158.99°. Since L is negative, true azimuth bearing is 360 − C, or 201.01°.

From **eq. 10,** the corrected elevation bearing equals 44.61°.

Eq. 11 yields a slant range of 20,215 nautical miles.

These computations agree well with my actual experience in tracking SMS-2.

calculator programs

The above relationships, with their conditional branching requirement depending on the sign of L, are ideal candidates for solution on a programmable calculator. I use a Hewlett-Packard 25 calculator which, with its 49 available program steps, demands a degree of programming ingenuity. After several false starts, I came up with the program listed in **table 1,** which appears to be valid for any point on the Earth*. Note that southern latitudes and eastern longitudes must be entered as *negative* numbers.

The following steps show how to use the program.

1. Key in the program, switch to RUN, depress f PGM.
2. Store the constant 180 in Register 3 (180 STO3)
3. Key in 3444 (for km, key in 6368)
4. ENTER
5. Store 22766 in Register 4 (for km store 42166)
6. Depress divide (÷) key
7. Store answer in Register 2
8. Store latitude of observer in Register 0
9. Depress *f* COS keys
10. Store answer in Register 1
11. Key in longitude of observer.
12. ENTER
13. Key in longitude of SSP.
14. Depress R/S key
15. Calculator displays range in nautical miles (or kilometers)
16. Depress rolldown key (R↓)
17. Calculator displays elevation angle in degrees
18. Depress RCL 6

*In reviewing this manuscript WB8DQT observed that the program will display ERROR if the observer is located directly under the satellite (on the equator at the SSP). Under these conditions the required azimuth bearing is undefined (the bearing is 90 degrees and the slant range is equal to $R_s - r_e = 19{,}322$ *nautical miles* [35,798 *km*]).

HP-25 Program

SWITCH TO PRGM MODE, PRESS f PRGM, THEN KEY IN THE PROGRAM.

DISPLAY LINE	CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	41	−					
02	23 05	STO 5					
03	14 06	TAN					
04	32	CHS					
05	24 00	RCL 0					
06	14 04	SIN					
07	71	÷					
08	15 06	TAN^{-1}					
09	24 03	RCL 3					
10	51	+					
11	23 06	STO 6					
12	24 00	RCL 0					
13	15 51	X ≥ 0					
14	13 17	GTO 17					
15	24 03	RCL 3					
16	23 51 06	STO+6					
17	24 06	RCL 6					
18	24 03	RCL 3					
19	02	2					
20	61	X					
21	14 51	X ≥ Y					
22	13 24	GTO 24					
23	23 41 06	STO−6					
24	24 05	RCL 5					
25	14 05	COS					
26	23 61 01	STO X1					
27	24 01	RCL 1					
28	24 02	RCL 2					
29	41	−					
30	01	1					
31	24 01	RCL 1					
32	15 02	X^2					
33	41	−					
34	14 02	$\sqrt{X}$					
35	71	÷					
36	15 06	TAN^{-1}					
37	01	1					
38	24 02	RCL 2					
39	24 01	RCL 1					
40	61	X					
41	02	2					
42	61	X					
43	41	−					
44	24 02	RCL 2					
45	15 02	X^2					
46	51	+					
47	14 02	$\sqrt{X}$					
48	24 04	RCL 4					
49	61	X					

MEW" REGISTERS

R0 Latitude observer

I K)

R3 ---

R4 Rs (22766 naut miles)

R5 Longitude of observer minus Long. of sat.

R6 Azimuth

R7

table 1. HP-25 program for calculating antenna azimuth and elevation angles and slant range to a geostationary satellite from any point on earth. Note that southern latitudes and eastern longitudes must be entered as negative numbers.

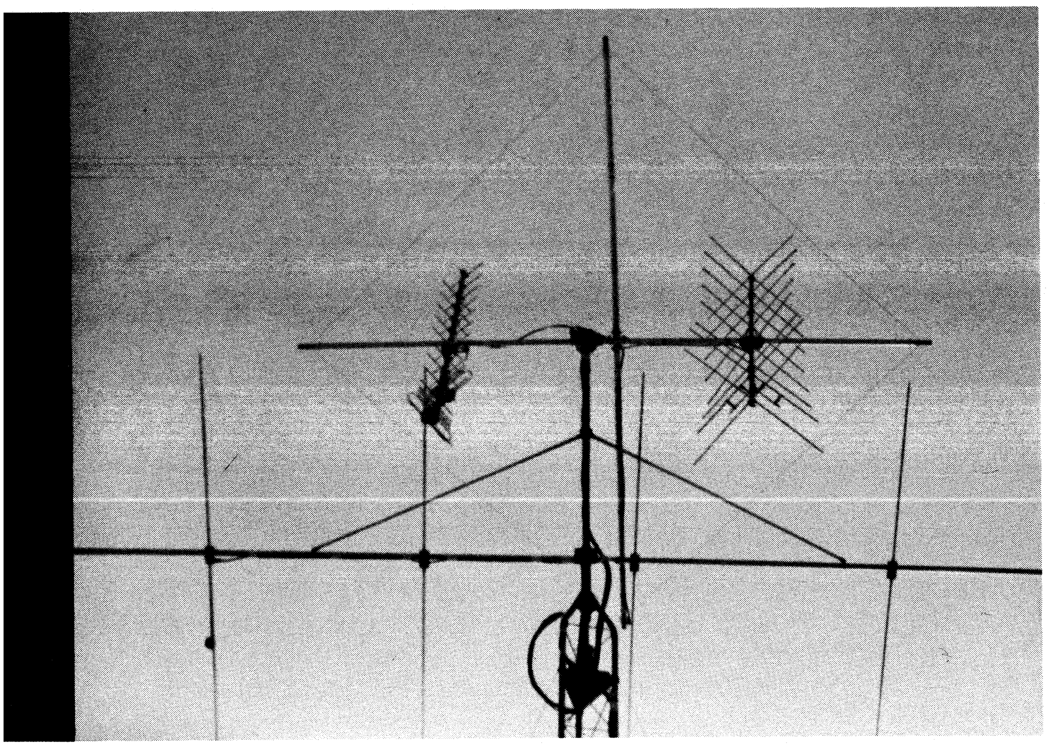
19. Calculator displays azimuth bearing in degrees
20. To perform further calculations, return to **Step 8.**

I would like to thank Glenn Thomas, WB6YZI, for his assistance in evaluating these computations. Glenn works at a satellite tracking facility and says, "Look-angles for geosynchronous satellites are my stock in trade." Believe me, he really makes it all look ridiculously easy!

references

1. *SMS/GOES User's Guide,* U.S. Department of Commerce, National Oceanographic and Atmospheric Administration, National Environmental Satellite Service. Washington, DC, 1977.
2. H. Paul Shuch, "A Cost-Effective Modular Down Converter for S-Band WEFAX Reception," *MTT-S Digest,* June, 1977.
3. W. Danielson and S. Glick, "Australis-OSCAR 5, Where It's At," *QST,* October, 1969. page 54.
4. Earl F. Skelton, WA3THD, "Individualized AZ/EL Coordinates in Minutes." *Amsat Newsletrers,* March, 1976, page 18.
5. Otmar K. Popp, DL3SX, "An Impression of Elliptical Satellite Orbit Tracks," *AMSAT Newsletter,* September, 1976, page 5.
6. Peter Thompson, Jr., "General Technique for Satellite Tracking." *QST,* November, 1975, page 29.
7. Ralph Taggart, WB8DQT, "Weather Satellite Handbook," *73 Publishing,* Peterborough. New Hampshire, 1976.
8. Jerry Hall, K1PLP, "Bearing and Distance Calculations by Sleight of Hand," *QST,* August. 1973, page 24.
9. L. R. Larson, WØDET, "Bearing and Elevation Calculations for Satellites," *QST,* Technical Correspondence, May 1974, page 57.
10. Wolfhard Vogel, "Graph Points Toward Any Synchronous Satellite," *Microwave Journal,* October, 1977. page 56.

ham radio



Oscar
az-el antenna system

Construction details for a complete antenna system for tracking Oscar including antennas for 28, 144 and 432 MHz

After checking all **your** charts and graphs, and double-checking all the calculations, you're left with only one problem, how to point your antenna at the satellite, and how to keep it accurately pointed as the bird moves through the sky. Obviously, some type of antenna tracking system is very desirable. Unfortunately, however, a check of commercial manufacturers did not yield any system that provided full frequency capabilities for working Oscar (the lack of 10-meter receiving antennas was the biggest stumbling block). After talking with many other amateurs, I decided to build my own tracking antenna system.

One of the methods used by others used some form of counterbalance to offset the antenna weight, permitting the rotator to elevate the antennas with a minimum of torque. This type of system tends to be impractical, however, because you should try to eliminate as much weight and wind resistance as possible. Also, the counterbalance system can become a mechanical nightmare if you try to design it to track from horizon to horizon in elevation (180-degree coverage). With these thoughts in mind, I decided that my tracking system should have the following features:

1. Capabilities for 28, 144, and 432 MHz
2. Full horizon-to-horizon elevation tracking
3. Cross polarization on both up and down links to prevent signal nulls
4. Elimination of the counterweight

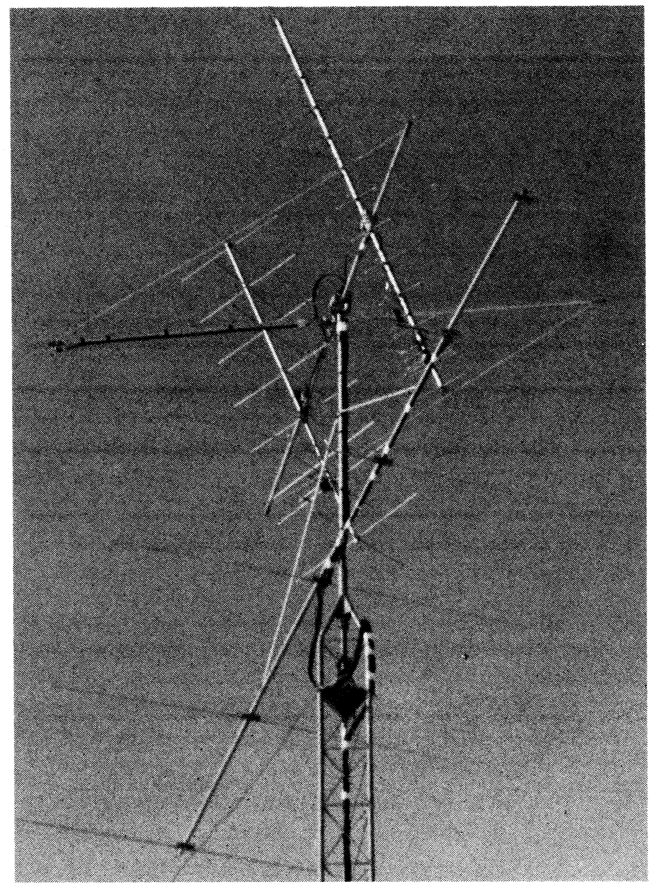
By Al Greenwood, WA1NXP, RR1, Greenville Road, Mason, New Hampshire 03048

construction details

Cross frame. The cross frame is the heart of the entire antenna system, supporting both the 144- and 432-MHz beams and also acting as spreaders for the 10-meter quad loop. The basic layout of this assembly is shown in **fig. 1.** The horizontal and vertical members are made from two lengths of aluminum tubing, each with an outside diameter of 35 mm (1-3/8 inches) and 3.7 meters (12 feet) long. The plate which holds the two spreaders together can either be purchased ready-made* or built from 6.5 mm (1/4 inch) thick aluminum plate.

Phenolic insulators are inserted into the ends of the tubing to insulate the quad loop from the spreaders (see **fig.** 2). Each insulator is held in place with a 1/4-20 (M6) bolt through the tubing and insulator. Another 6.5 mm (1/4 inch) hole is drilled through the insulator to hold the quad-loop wire.

After the two spreaders are fastened together and the insulators inserted in the tubing, the horizontal spreader is inserted into the U-100 rotator. Before tightening the clamps, you should run the rotator



Satellite tracking antenna at WA1NXP combines antennas for 10 meters, 144 MHz, and 435 MHz on one tower. The six-element beam located below the OSCAR antenna is for 50 MHz.

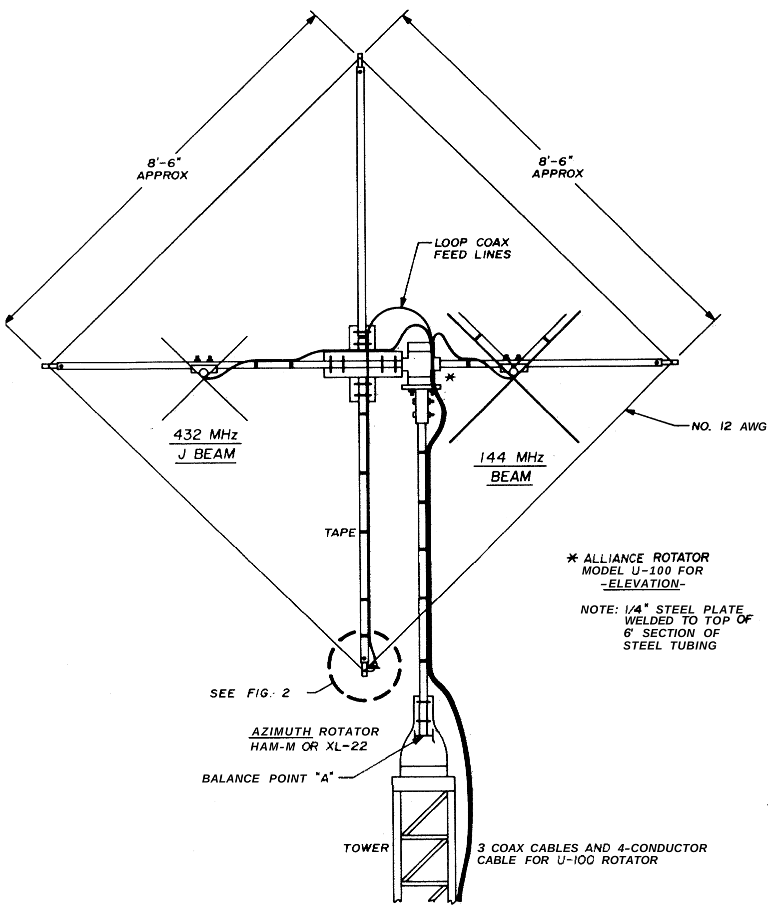


fig. 1. Head-on view of the az-el antenna system. The 144- and 432-MHz beams should be positioned along the horizontal spreader to balance the array at the base of the mast. The quad loop is made from number 12 AWG (2.1-mm) wire.

from stop to stop to make sure that you have it in the desired position for assembly.

The final part of the cross frame is the adapter between the elevation rotator and the mast from the azimuth rotator. This was fashioned by drilling a piece of 6.5 mm (1/4 inch) thick steel plate with four holes that matched the bolt pattern of the U-100 rotator. To the bottom of the steel plate I welded a 15 cm (6 inch) piece of steel tubing. The inside diameter of the tubing must be large enough to accommodate the mast from the azimuth rotator.

Two-Meter Antenna. The 144-MHz cross-polarized Yagi consists of two 8-element beams mounted on a common boom (elements 90 degrees apart). The dimensions are standard and can be found in any antenna manual. The elements are mounted through a 32 mm (1-1/4 inch) aluminum boom, with the two beams displaced 12.5 mm (1/2 inch) along the boom.

*A parts kit including X bracket, spreaders, phenolic insulators. U-100 welded mounting bracket, coax bracket, and wire is available from Alden Engineering Company, P.O. Box 493, Greenville, New Hampshire 03048.

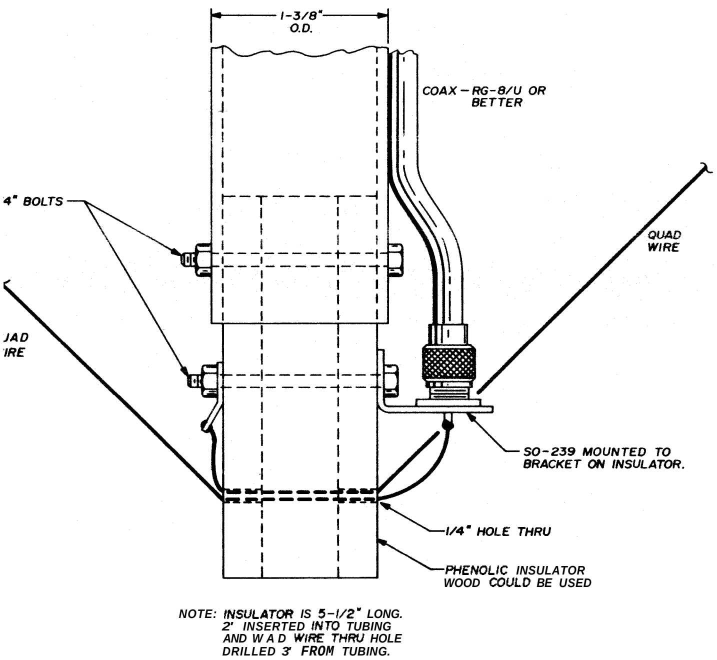
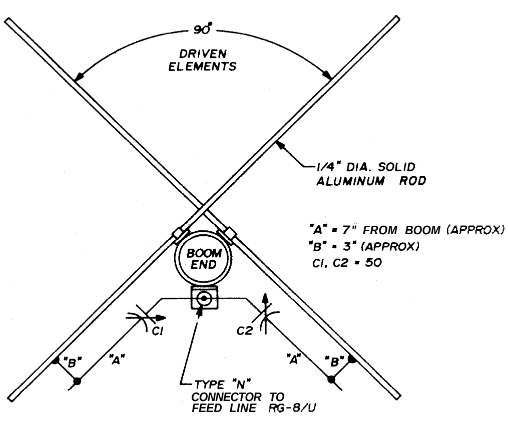


fig. 2. Detail of the four phenolic insulators which are used to support the 10-meter quad loop. The insulator is 14 cm (5-1/2 inches) long; the wire passes through the insulator approximately 7.5 cm (3 inches) from the end of the tubing.

The feed system for the beams is shown in **fig.** 3. The capacitor used for the gamma match is a split stator unit with approximately 50 pF per section. Be sure to provide an insulated mount for this capacitor. Prior to mounting the beam on the horizontal spreader, the longitudinal balance point of the beam must be determined. By attaching at this point, the elevation rotator will not be excessively strained.

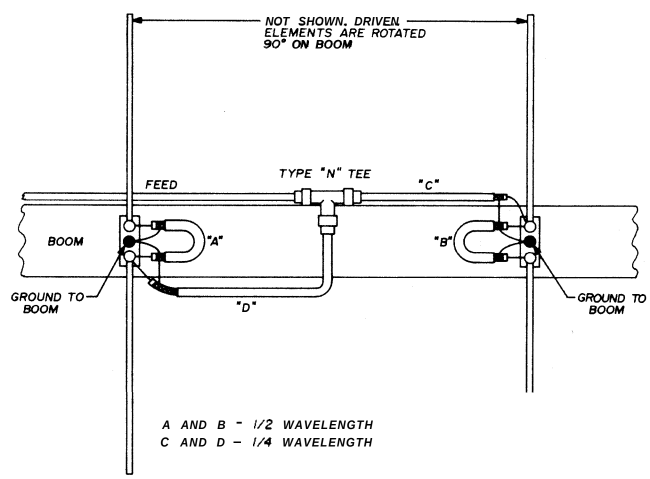
Seventy-Centimeter Antenna. For 432 MHz, I used a commercial 26-element J *Beam.* As with the two-meter beam, this antenna is also mounted to provide cross polarization. **Fig.** 4 shows the baluns and the matching necessary to simultaneously feed both antennas from one 50-ohm feedline. You should also determine the balance point for this beam before attaching it to the horizontal spreader.



144 MHz MATCH

fig. 3. Gamma match for the two 144-MHz beams. The gamma capacitor is a split-stator capacitor with approximately 50 pF per section. After final adjustment, the entire unit was fiberglassed for weather proofing.

Ten-Meter Antenna. As a final step, the wire for the quad loop is strung through the holes in the phenolic insulators. With the dimensions shown, the loop should resonate at approximately 29.6 MHz. When using this system, the pattern exhibited by the loop tends to be sharper than would be expected, or desired. As a remedy, I opened the loop at both ends of the horizontal spreader, effectively producing a dipole which has a wider beam-width than the loop.



432 MHz MATCHING HARNESS

fig. 4. The two 432-MHz beams require a balanced 300-ohm feed. With this arrangement, the 112-wavelengths of coax act as 4:1 baluns, while the 114-wave sections step the impedance up to approximately 100 ohms. The two beams then closely match a 50-ohm feedline.

Either version can be used; it depends only on your personal preference and the resolution of your azimuth rotator.

summary

In the described configuration, the U-100 rotator easily moves the array through a full 180 degrees elevation. As an added feature, the quad wire acts as a mechanical stop when it hits the azimuth rotator mast, stalling the elevation rotator. One drawback to this system is the resolution of the rotators. For extremely accurate pointing, infinite control is a necessity. Unfortunately, this is not provided by either rotator.

ham radio

high-gain 1296-MHz antenna

For those interested in an easy-to-build antenna for the 1215-1300 MHz amateur band, the design described here might be the answer. It features high gain, good capture area, and shielding from high-power local signals. The design also includes both vertical and horizontal polarization and, with appropriate phasing equipment, right and left circular polarization. The simplicity of the feed system and overall antenna design should make it a good candidate for just about any requirement in the 1215-1300-MHz band.

I have received signals as far as 80 km (280 miles) away with this antenna. With a motorized tilt control, the antenna has possibilities for use in amateur satellite work.

The design features a waveguide-type can, which bypasses strong local signals and shields direct radiation interference from the 1/4-wavelength dipoles. The addition of 10-degree tubes doubles the gain, concentrates the signal, and increases signal capture area.

construction

The antenna design is shown in fig. 1. To simplify matters, I had a local tinsmith make a 178 mm (7 inch) ID x 406 rnm (16 inch) long round can of 24-gauge galvanized tin, with a solid back, crimped and soldered. (Brass or copper could be used.)

Four aluminum tubes are required, measuring 12.7 mm (1/2 inch) ID x 16 mm (5/8 inch) OD x 140 cm (55 inches) long. Don't bend the tubes and don't substitute any other diameter or length.

Next, you'll need four lengths of galvanized-steel water pipe. The dimensions are 6.4 mm (1/4 inch) diameter x 457 mm (18 inches) long. Bend at the 305-mm (12-inch) point (sharp bend) at a 10-degree angle (fig. 1). Aluminum tubing measuring 13 rnrn (1/2 inch) ID x 16 mm (5/8 inch) slips snugly over the extending 152-mm (6 inch) portion of the water pipe. The aluminum tubes are secured by a self-tapping screw, one screw per tube.

Drill a 9.5-mm (3/8 inch) hole in the top, and drill another hole 90-degrees away for the TV F-type coaxial fittings. The holes should be drilled 54 mm (2 1/8 inch) from the back of the can. Two 54 mm (2 1/8 inch) lengths of 1-mm (18-AWG) wire should be soldered to the TV F-type fittings for the dipole radiators.

I obtained a piece of steel measuring 1.6 x 102 x 229 mm (3/16 x 4 x 9 inches) from a local metal shop, which I used to mount the antenna to a 25-mm (1-inch) diameter mast made of galvanized-steel water pipe. A steel plate was drilled for three TV mast clamps. These clamps were placed 51 mm (2 inches) apart. Fig. **1** gives the

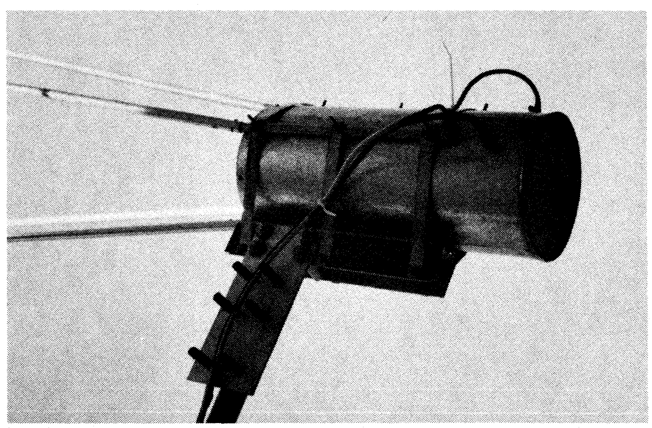
The can has a piece of clear plastic measuring 178 x 6.4 mm (7 x 1/4 inch) mounted snugly in the open end. See fig. 1. Drain holes should be drilled into the bottom of the can.

The TV-type chimney mast straps are used for reinforcement, as shown in fig. 1. These reinforcement straps are absolutely necessary.

When, completed, the distance between the ends of each set of tubing (horizontal or vertical) should be between 686 and 711 mm (27 and 28 inches).

weather protection

I used polyurethane liquid plastic coating for weather protection. However, almost any good spray enamel could be used. Extra coats of paint on



Complete antenna showing the can-type resonator and feed-system details. Two coaxial cables are used.

the steel plate are desirable to inhibit corrosion. (Don't paint the plastic cover for the cans.) Liberal use of PVC tape over the coaxial-cable fittings where they fasten to the cans will help to make waterproof joints.

ham radio

By Paul F. Magee, W3AED, RR2, Box 432, Berlin, Maryland 21811

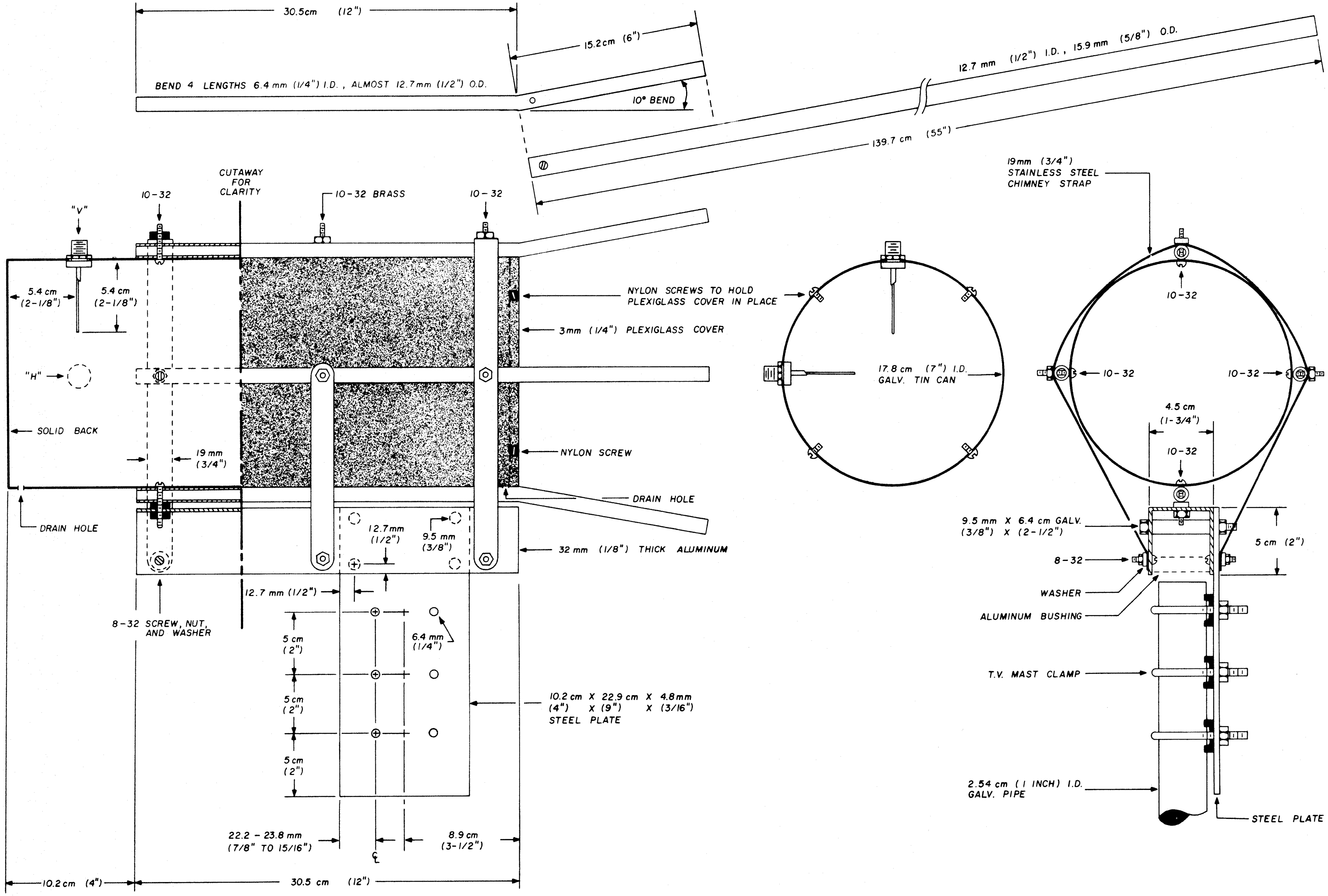


fig. 1. Construction details of the high-gain 1215-1300-MHz antenna. Design features both vertical and horizontal polarization and an enclosed, weather-proof feed system.

simplified antenna gain calculations

A straightforward approach for comparing the gain of a number of popular amateur antennas

It is well recognized by amateurs that improvements in an antenna installation can usually do more to enhance a station's capabilities than any other improvement. The purpose of this article is to help you choose or optimize an antenna arrangement from the standpoint of gain. Of course, many other factors have a large influence on antenna performance including ground conditions, ohmic losses, height above ground, polarization, bandwidth, multiband operation, effects of nearby objects, and so on. This article will not deal with these topics, but will attempt to consolidate some results of gain calculations made on some of the more popular antennas in use today. Along the way, a relatively straightforward approach for making such calculations will be described so that you can make similar calculations for configurations in which you are interested. It should be observed, however, that the methods described are applicable to only some antenna types. For example, multi-element arrays with parasitic elements are not suited to the simple technique described here. However, this technique does allow you to gain a significant amount of insight into a number of different antennas.

basic assumptions

As it turns out, very few assumptions must be made to compare the gains of the antenna types which are considered here. These assumptions are listed below:

1. The feedpoint is at a point of maximum current.

2. The feedpoint resistance (at resonance) is known.

3. All power fed into the antenna is radiated.

4. The current distribution for wire resonant antennas is sinusoidal and for antennas containing inductive loading is approximately trapezoidal.

5. The current in each part of the antenna is either in phase or 180° out of phase with the current in all other parts of the antenna, with phase reversals occurring where the current goes through a

By Frank Witt, W1DTV, 20 Chatham Road, Andover, Massachusetts 01810

minimum. Although this assumption is not 100 per cent valid,[1] the results should be close enough to reality to provide insight for those of us who enjoy experimenting with different antennas.

All gain calculations will be made relative to a half-wavelength dipole. With the exception of the monopole example given later, the antennas will be assumed to be in free space."

Although the calculations are made for transmitting antennas, the results apply equally to receiving because of the principle of reciprocity.

theory

The field strength due to a part of an antenna is proportional to the magnitude of the current flowing in that part. This dependence is shown in fig. **1.** It is assumed, of course, that the observer is many wavelengths from the antenna; all parts of each of the antennas considered here will be essentially the same distance from the observer. By "essentially the same" I mean that any differences in distance from the observer are very small compared to a wavelength. A surprisingly large number of practical antennas exhibit maximum gain in a direction where this distance constraint is valid.

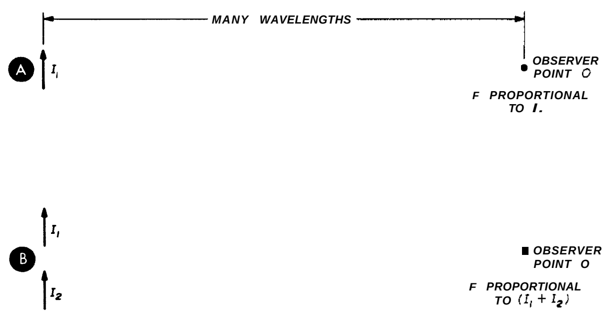


fig. 1. Field strength due to a part of an antenna is proportional to the magnitude of the current flowing in that part. At A, the field strength at the observer (point O) is due to the current element I_1. Since I_1 and I_2, B, have the same polarization and are the same distance from the observer, the field strength F is the sum of the field strengths due to each current element.

Another simplifying assumption is that in the direction for which the gain is to be calculated some radiation cancellation takes place, as seen in fig. **2.**

'An assumption of free-space conditions simplifies the calculations, but many would argue that it's both unrealistic and misleading to discuss the radiation characteristics of either straight or bent dipoles for use in a ground-oriented communications link if the effects of ground reflections are ignored. Nevertheless, the gain technique discussed by the author provides a basis of antenna comparison. It should be kept in mind, however, that the gain and radiation patterns presented here are not necessarily representative of those found in a real life, near ground situation. **Editor.**

feedpoint resistance

To determine the relative magnitude of the current flowing in an antenna, the feedpoint resistance must be either assumed, measured, or calculated. Through the use of an impedance bridge, using either an oscillator or noise generator as a source, quite accurate measurements of feedpoint resistance may be made (see references 2 through 4). Anyone who is seriously interested in optimizing an antenna installation should have a means for measuring feedpoint resistance.

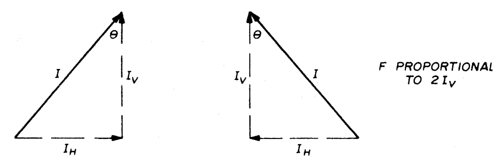


fig. 2. The observer in this case is looking into the plane of the page. Note that the vertical components, I_V, add while the horizontal components, I_H, cancel. Field strength F is proportional to $2I_V$.

Direct calculation of antenna feedpoint resistance is a complex mathematical process except for the simplest of structures. The engineering literature yields some useful results, but much more work needs to be done in this area.

Fig. 3 shows how the currents are related for two antennas with different feedpoint resistances when you assume that the same amount of power is fed into each antenna. Thus G_R, that part of gain due to feedpoint resistance, is given by

$$G_R(dB) = 20 \log \frac{I_2}{I_1} = 20 \log \sqrt{\frac{73}{R}} = 10 \log \frac{73}{R}$$

where R is the feedpoint resistance of the antenna under consideration (73 ohms is the free-space feedpoint resistance of a half-wavelength dipole).

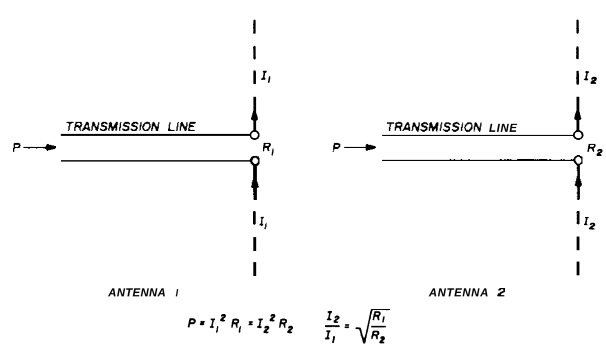


fig. 3. Effect of feedpoint resistance on current when the same power P is fed to two different antennas.

current distribution

The currents which affect the field strength vary along the length of the antenna. At the open ends of a wire antenna (e.g., the ends of a dipole), the current is practically zero. It is customary to assume that the current varies along a resonant antenna sinusoidally and this will be done here; **fig. 4** shows the assumed current distribution along a half-wavelength dipole.

Since it will arise frequently, it's useful to compute the field strength due to a sinusoidal current distribution along a straight wire. **Fig. 5** shows the general result for the field strength due to a section of antenna wire of length x_1 with a sinusoidal current distribution as shown. For a half-wavelength dipole, $x_1 = \lambda/4$, and since the fields radiated from the two

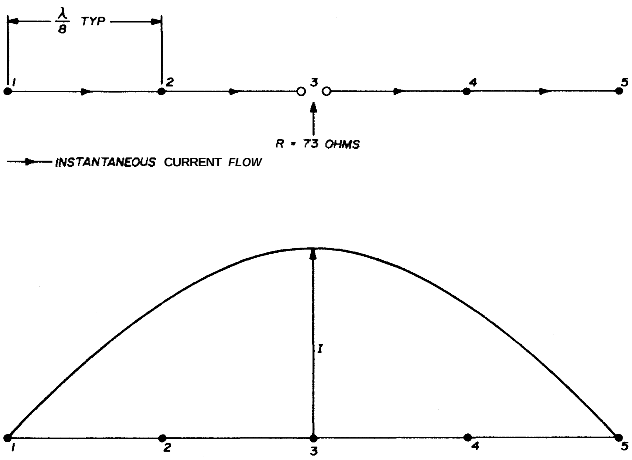


fig. 4. Current distribution along a half-wavelength dipole.

halves add in phase (in the plane normal to the radiator, at its center)*

$$F_R = 2\frac{k\lambda I}{2\pi}\, sin\, \frac{2\pi}{\lambda} \cdot \frac{\lambda}{4} = \frac{k\lambda I}{\pi}$$

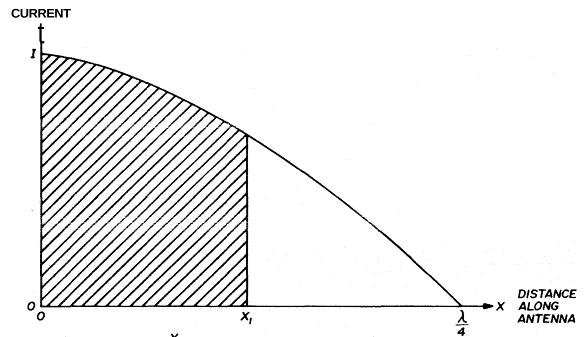
This is the reference field strength due to current distribution for gain calculations. G_C is defined as the part of the gain due to current distribution and is given by

$$G_C\,(dB) = 20\, log\, \frac{F}{F_R} = 20\, log\, \frac{\pi F}{k\lambda I}$$

where F is the field strength of the antenna under consideration. Thus the gain relative to a dipole in dB is given by

$$G = G_R + G_C = 10\, log\, \frac{73}{R} + 20\, log\, \frac{\pi F}{k\lambda I}$$

*Off the plane normal to the radiator the fields add out of phase, to a degree related to the angle off the plane, resulting in the dipole radiation pattern. **Editor.**



$$F = k\int_0^{x_1} I \cos \frac{2\pi}{\lambda}\, x\, dx = \frac{k\lambda I}{2\pi}\, sin\, \frac{2\pi}{\lambda}\, x_1$$

fig. 5. Field strength, F, due to a sinusoidal current distribution; k is a constant of proportionality.

Since k, λ, and I will cancel in the subsequent calculations, it is clear that if R and F can be determined, the gain relative to a dipole may be found.

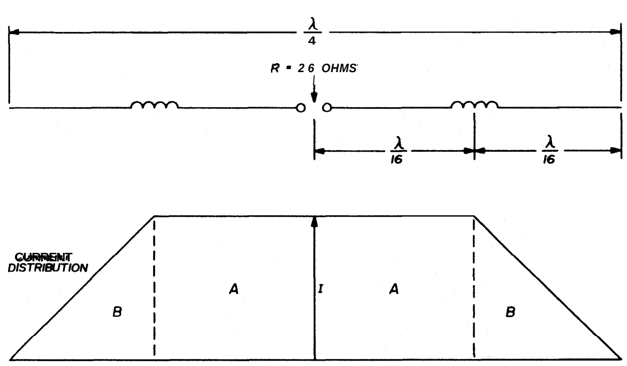
Another current distribution of interest is the trapezoidal distribution, which is a good approximation for inductively-loaded, short antennas.5 The only antenna having this current distribution that will be considered here is the quarter-wavelength dipole with loading coils located as shown in **fig. 6.**

other factors

As seen in **fig.** 2, in some cases the current in the direction of the wire is not the appropriate current to use. What is required is the current in the direction of the polarization being considered. Hence, for **fig.** 2, the current component of interest is

$$I_V = I \cos \theta$$

Another consideration is the number of radiating elements. In this case, you must account for the fact



$$F = \underbrace{k \cdot \frac{\lambda}{8} \cdot I}_{A} + \underbrace{2 \cdot k \cdot \frac{1}{2} \cdot \frac{\lambda}{16} \cdot I}_{B} = \frac{3}{16}\, k\lambda I$$

$$G = 10\, log\, \frac{73}{26} + 20\, log\, \frac{3}{16}\, k\lambda I \cdot \frac{\pi}{k\lambda I}$$

$$= 4.48\ dB - 4.60\ dB = -0.12\ dB$$

fig. 6. Gain of a quarter-wavelength inductively-loaded dipole.

that the power is divided among these elements. For the two-element broadside and collinear arrays considered here, the power is equally divided between the two radiators. Hence, the multi-element factor G_M, which must be included in the gain calculations, for the two-element case is

$$G_M = 10 \log \frac{1}{2} = -3.01\ dB$$

With these preliminaries disposed of, we can now consider several specific antenna types.

inductively-loaded dipole

Consider the inductively loaded dipole shown in fig. 6. The feedpoint resistance is assumed to be 26 ohms.[5,6] Further verification of this value is required, however. The fact that the gain is essentially that of a full-size dipole is consistent with the popularity of this antenna for many sloping-dipole installations on 80 meters. However, it is hampered somewhat by its relatively narrow bandwidth.

inverted vee

Fig. 7 shows the instantaneous currents in a dipole having legs formed as an inverted-V. When the legs form an angle less than 180 degrees, the radiation resistance is reduced, resulting in a corresponding reduction in feedpoint resistance.[7,8] However, the current distribution is practically the same as that for the straight dipole of fig. 4. In the broadside direction the fields resulting from the vertical current components cancel; however, the horizontal field components add, but they are diminished by the folding. On the other hand, this reduction is almost completely compensated for by the overall increase in magnitude of *all* the field components, resulting from the higher antenna current I, due to the lower feedpoint resistance. When the included angle is 90°, the feedpoint resistance in free space is typically around 40 ohms, and the gain relative to a straight, 180-degree dipole is approximately −0.4 dB.

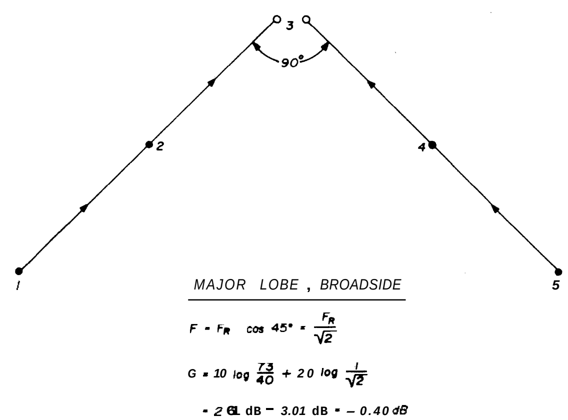


fig. 7. Gain of the inverted-V dipole. Observer is facing the page, and polarization is horizontal. An observer in the plane of the paper, looking into the antenna from left to right would see a small component of vertical polarization due to uncancelled vertical components radiated from each **leg**. The vertical components cancel in the broadside direction.

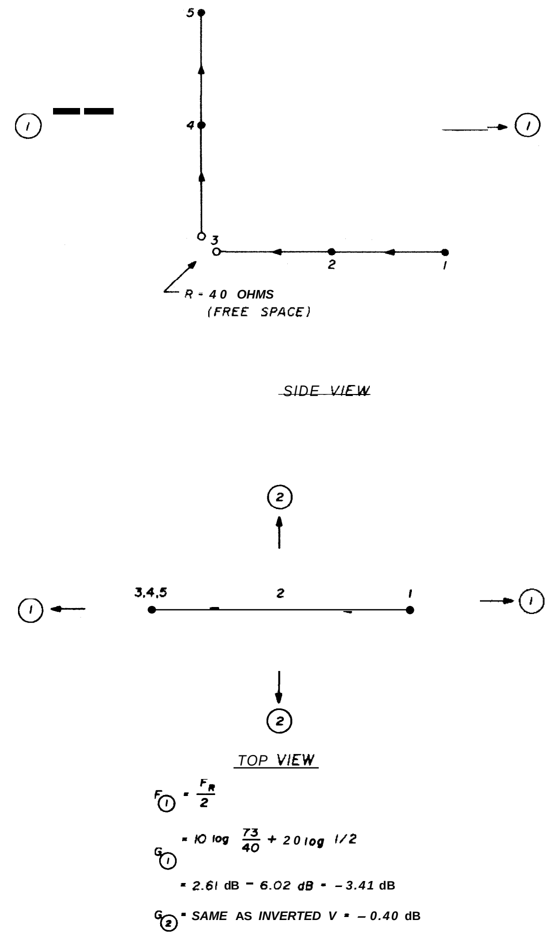


fig. 8. Gain of the L antenna. All points on the equatorial plane are broadside to the antenna. In the polarization plane normal to the equatorial plane the gain is uniformly maximum everywhere on the equator. Maximum gain is in direction 2 and is the same as an inverted V with a 90 degree included angle.

L antenna

The L antenna shown in **fig. 8** is of interest because it has been used where height is limited. In direction **1** the polarization is vertical, but in direction 2, where the gain is maximum, the polarization is half horizontal and half vertical. In direction **1** only half the antenna is seen by the observer and the gain is

-3.41 dB (free space). In direction 2, the gain is the same as that for the inverted V, which is obvious when you remember that the L is a rotated V. In direction 3 the gain in only slightly less than in direction 2, but will not be caiculated here.

multi-element driven arrays

You can use the methods described here to determine the relationship between gain and feedpoint resistance for certain types of driven arrays. Such data are available in the ARRL *Antenna Book* for ideal situations, but the method given here wil! enable you to infer gain directly from feedpoint resistance measurements.

The cases considered are the collinear and broadside arrays, where two in-line or parallel half-wavelength dipoles are fed in phase as shown in fig. 9. The current distribution is the same as for the reference dipole but since there are two radiators, $G_C = 6.02\ dB$. The power is split between the two radiators, however, so the field strength is modified by $G_M = -3.01\ dB$. Thus, as seen in fig. 9, the gain

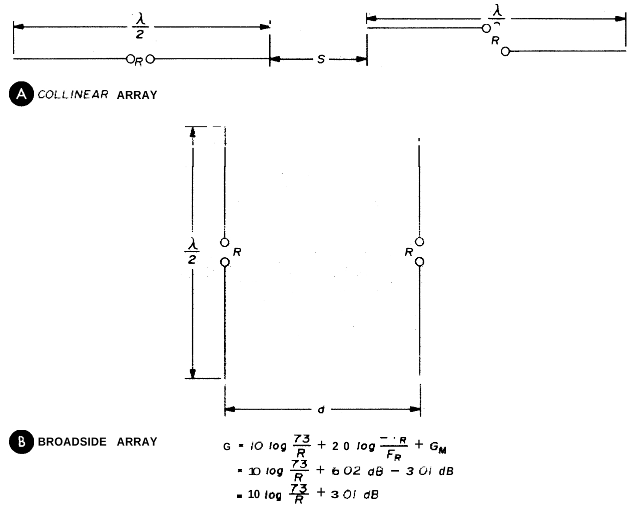


fig. 9. Gain calculation for two in-phase parallel half-wavelength dipoles in a collinear arrangement, A, or broadside array, B.

table 1. Calculated gain of collinear array.

spacing s	feedpoint resistance for each dipole R (ohms)	gain G (dB)
0	94	1.9
$\frac{\lambda}{8}$	87	2.2
$\frac{\lambda}{4}$	76	2.8
$\frac{\lambda}{2}$	70	3.2
λ	73	3.0

is related very simply to the feedpoint resistance by

$$G = 10\log\frac{73}{R} + 3.01\ dB$$

This relationship is plotted in fig. **10.**

The feedpoint resistance and gain for several spacings of collinear dipoles are shown in **table 1.** The resistance values were obtained from *The ARRL Antenna Book*2 (page 134); the calculated gain values using fig. **10** agree exactly with those given in the *Antenna Boak*.

The *Antenna Book* does not give the feedpoint resistance for the broadside array, but it does give gain as a function of spacing. Again fig. **10** was used, but this time to determine the feedpoint resistance of each dipole. The results, shown in **table** 2, agree very well with those given in reference 9.

monopoles

A popular vertical radiator is the monopole, or quarter-wavelength half-dipole over ground. At higher frequencies, resonant radials are used if the antenna is elevated well above ground, at least a quarter wavelength. At lower frequencies elaborate radial systems (usually buried a few inches) are often employed. Such a ground system is *required* if optimum radiating efficiency is to be obtained. An infinite, perfectly-conducting ground is assumed in fig. **11.** The feedpoint resistance of the quarter-wavelength half-dipole, in conjunction with its ground-reflected image is 36.5 ohms, half that of a complete dipole. In this case, the total field strength results from the in-phase addition of the direct radiation from the monopole and the radiation from the ground image. The total field is therefore twice that radiated directly from the monopole. If the reference dipole antenna is

table 2. Calculated feedpoint resistance of the broadside array.

spacing d	feedpoint resistance for each dipole R (ohms)	gain G (dB)
0	146	0
$\frac{\lambda}{8}$	136	0 3
$\frac{\lambda}{4}$	116	1.0
$\frac{\lambda}{2}$	59	3.9
$\frac{3}{4}\lambda$	51	4.6
λ	77	2.8

considered to be in free space then it would appear from fig. **11** that the monopole has +3.01 dB gain, but this is misleading. The reason for the apparent gain is that the monopole radiates the power only in the "half" space (or half hemisphere) above ground,

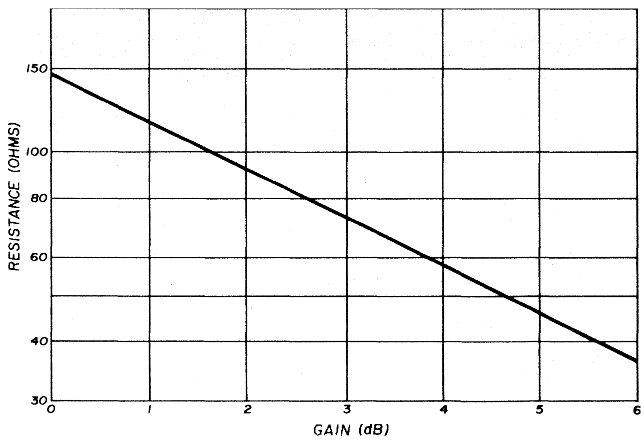


fig. 10. Feedpoint resistance versus gain for parallel half-wavelength dipoles fed in phase.

while a free-space dipole radiates the same amount of power in all space.

A similar calculation can be used to determine the gain of a half-wavelength ground-mounted vertical dipole. Since this antenna plus its image is like a collinear array with zero spacing, table **1** may be used to obtain the feedpoint resistance of 94 ohms (the feedpoint is at the current maximum, one-quarter wavelength above ground). This calculation yields a gain figure of 4.92 dB over a reference half-wavelength dipole in free space. Thus a half-wavelength ground-mounted dipole has 1.91 dB gain over a monopole (4.92 dB − 3.01 dB = *1.91 dB*).

full-wave loops

The full-wavelength loop is used in the shape of a square (the quad loop), a diamond, and a triangle (the delta loop). In this section, the dependence of gain on the loop shape will be calculated. Since I don't know the exact feedpoint resistance for each of

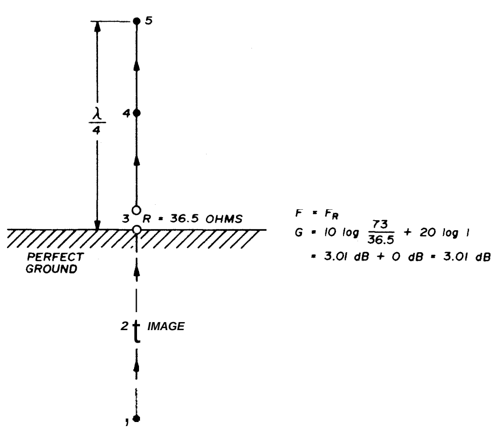


fig. 11. Gain of a monopole over perfectly conducting ground relative to a half-wavelength dipole in free space. The current distribution is the same as that shown in fig. 4.

these forms, a precise analysis is not possible. Some interesting results have been obtained, however, and some suggestions for further work have been identified.

Fig. **12** shows the current distribution on a full-wavelength loop and two special cases. The shorted half-wavelength transmission line will not accept power because $R = 0$, so it is of no interest as an antenna. Fig. **12C** proves that it is consistent for a

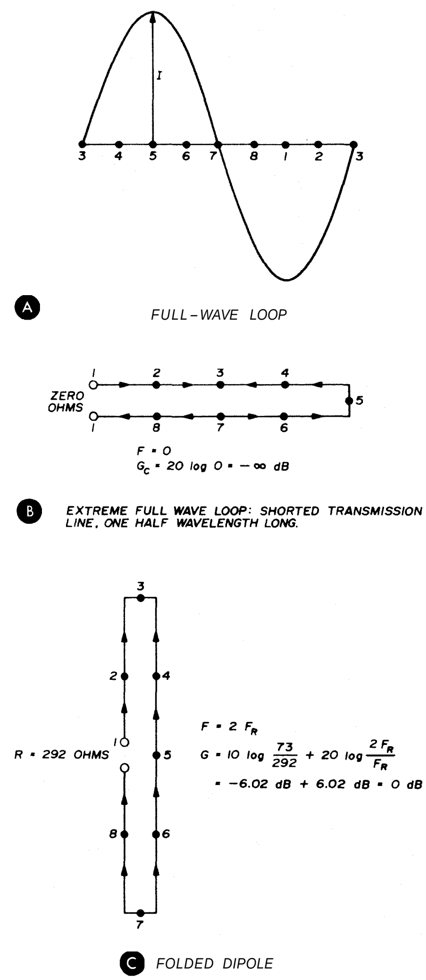
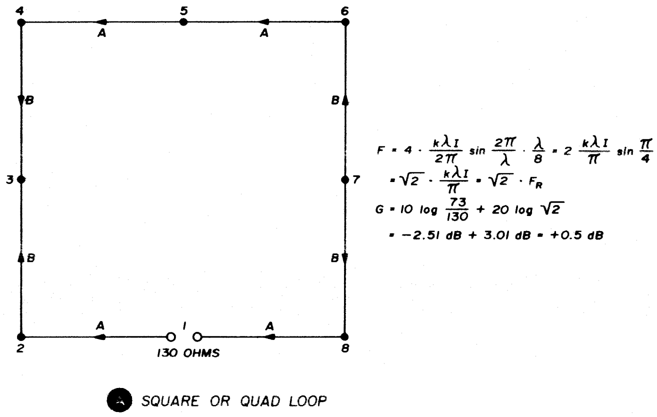
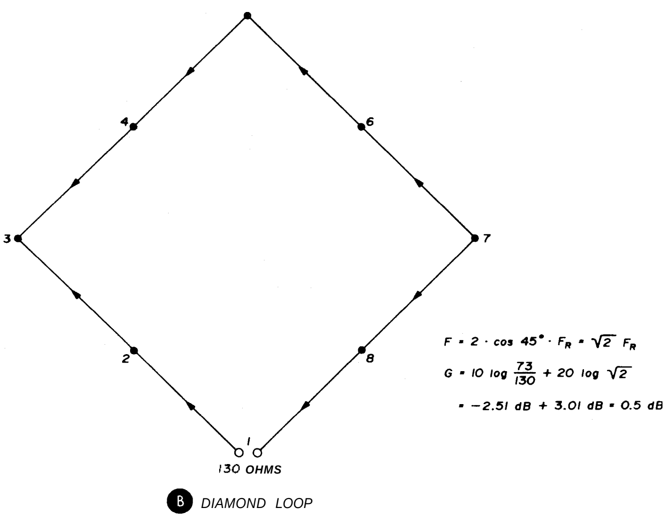


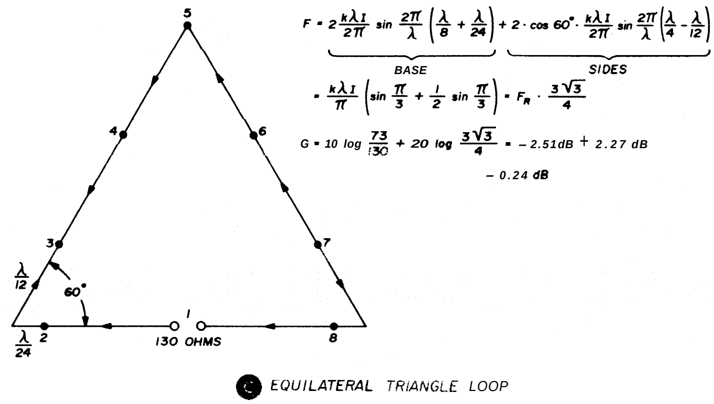
fig. 12. Current distribution of a full-wavelength loop, A, a shorted half-wavelength transmission line, B, and a folded dipole, C. The folded dipole provides the same gain as a reference dipole.

folded dipole to have 0 dB gain and a feedpoint resistance of 292 ohms (4 x 73 ohms).

In the following discussion, the feedpoint resistance of symmetrical loops will be assumed to be 130 ohms. This is consistent with reference 10 which gives a value of 125 ohms for the square loop, and reference 11 which reports that the calculated value for a circular loop is 140 ohms. By symmetrical **I** mean a square, a diamond, an equilateral triangle, or







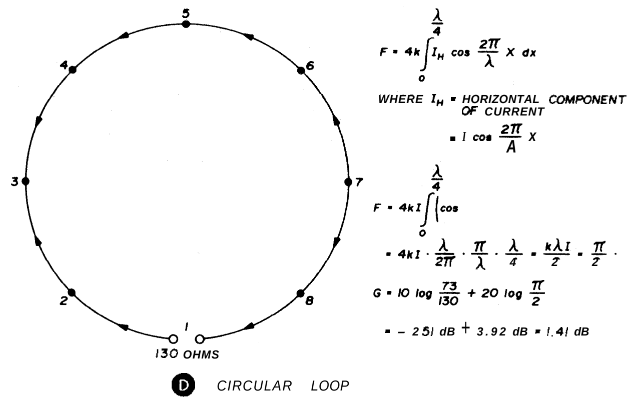


fig. 13. Gain calculations for symmetrical full-wavelength loops. Note that for the square or quad loop, A, the four A segments add in phase, and the four B segments cancel. The diamond loop has the same gain as the square loop. In the circular loop, D, the variable x is measured clockwise around the loop with $x = 0$ at the feedpoint.

a circle. The gains of symmetrical loops are calculated in **fig. 13**.* All of the loops are fed so that polarization is horizontal. Note that movement of the feedpoint to points 3 or 7 will provide vertical polarization. The gain is the same for either polarization; this is obvious for the square, diamond, and circle, and may be easily shown for the equilateral triangle (a calculation similar to that of **fig. 13C** if you assume that the feedpoint resistance is unchanged).

In reference 13 the top-loaded delta loop is introduced as an efficient vertical radiator for use where height is limited. In **fig. 14** gain calculations are shown for a top-loaded delta loop and an isosceles triangle loop (called here the low delta loop) with the same vertical dimension. Again, a feedpoint resistance of 130 ohms is used, but this time the figure comes from measurements made at W1DTV for 80-meter antennas whose bases are 3 meters (10 feet) above ground. The free-space values of feedpoint resistance are unknown.

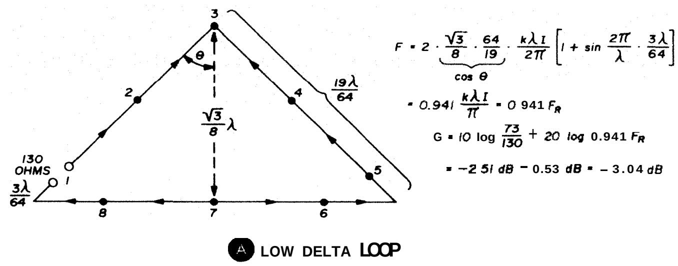
*These gain figures and those in the following table differ from the gain figures given in reference 12 which states that the free-space gain of full-wavelength loops in the shape of squares, diamonds, or circles is approximately 1 dB. The discrepancy may be a result of using incorrect free-space feedpoint resistance values. **Editor.**

It is interesting to compare the results of the full-wavelength loop gain calculations.

shape	gain
Circle	1.4 dB
Square	0.5 dB
Diamond	0.5 dB
Equilateral triangle	0.2 dB
Top loaded delta loop	−0.7 dB
Low delta loop ($\frac{\sqrt{3}}{8}\lambda$ high)	−3.0 dB

All of these results assume the same feedpoint resistance. This is a shaky assumption, and should be examined analytically as was done for the circular loop in reference 11. In any event, the methods given here allow you to revise these results if and when more solid feedpoint resistance data are available.

If the assumed feedpoint resistances are correct, however, there is a significant gain penalty when a low delta loop is used as a vertical radiator. Fortunately, top loading may be used to recover some of this loss. The low gain in a direction perpendicular to



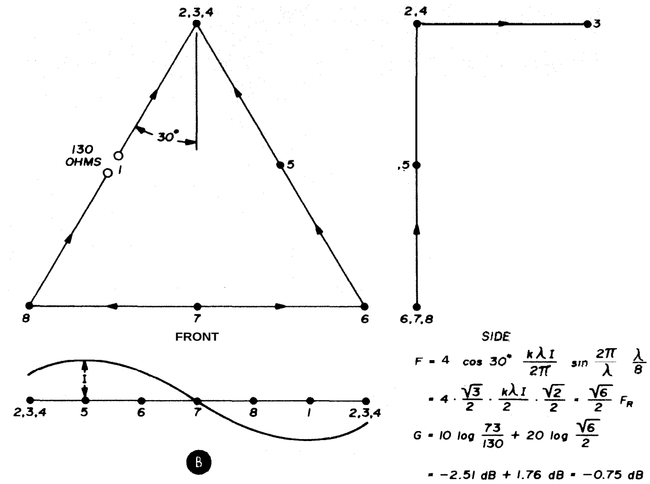


fig. 14. Gain calculation for the low delta loop, A, and the top loaded delta loop, B. The current distribution in the low delta loop is the same as that shown in fig. 12A; note that radiation from the base is cancelled.

the plane of the low delta loop is probably indicative of a pattern somewhat different than that of the more symmetrical configurations. The current distribution suggests more high-angle radiation than the other loops, also.

summary

The results of the gain calculations have been summarized in **table** 3. Also shown is the assumed feedpoint resistance and the size of the antenna in wavelengths. One observation I have made as I became interested in this subject is that there are many conflicting published gain values for various antenna configurations. In particular, the values for full-wavelength loops and the monopole vary considerably. In most cases, however, the method for arriving at a particular published value of gain has not been disclosed. In this paper, a straightforward approach for calculating gain has been presented and applied to a number of examples. The intention is not to present the final word regarding antenna gain figures, but to provide insight and a basis for future work. This has been done by splitting the gain calculation into several parts: the effect of feedpoint resistance, the effect of current distribution, and the effect of the number of driven radiators. Although the approach is not applicable for all antennas, it does apply to enough popularly used antennas to be worthwhile.

It has been pointed out in the past how valuable the knowledge of feedpoint impedance can be for tuning and feeding an antenna; the emphasis here has been to point out another reason for measuring this quantity — the calculation of gain.

table 3. Summary of antenna gains as compared to a half-wavelength dipole in free space.

type	assumed feedpoint (ohms)	size	gain (dB)
Broadside (0.75λ)	51*	0.75λ x 0.5λ	4.6
Broadside (0.5λ)	59*	0.5λ x 0.5λ	3.9
Collinear (0.5λ)	70*	1.5λ x 0λ	3.2
Collinear (0.25λ)	76"	1.25λ x 0λ	2.8
Collinear (0.125λ)	87"	1.125λ x 0λ	2.2
Collinear (0λ)	94*	1.0λ x 0λ	1.9
Circular loop	130	0.32λ x 0.32λ	1.4
Broadside (0.25λ)	116*	0.5λ x 0.25λ	1.0
Square	130	0.25λ x 0.25λ	0.5
Diamond	130	0.35λ x 0.35λ	0.5
Broadside (0.125λ)	136"	0.5λ x 0.125λ	0.3
Half-wavelength dipole	73	0.5λ x 0λ	0
Folded half-wave dipole	292	0.5λ x 0λ	0
Quarter-wave dipole (inductively loaded)	26	0.25λ x 0λ	−0.1
Equilateral triangle loop	130	0.33λ x 0.29λ	−0.2
Inverted V ($\theta = 90°$)	40	0.35λ x .18λ	0.4
L **antenna** ($\theta = 90°$)	40	0.25λ x 0.25λ	0.4
Top-loaded delta loop	130	0.25λ x 0.22λ	−0.7
Low delta loop	130	0.41λ x 0.22λ	−3.0

"Each dipole; spacing shown in parentheses.

references

1. R. King and C. Harrison. *Antennas and Waves,* The MIT Press, Cambridge, Massachusetts, 1969. page 153.
2. *The ARRL Antenna Book,* 13th edition, ARRL, Newington, Connecticut. 1974, pages 312-316.
3. J. Sevick, "Simple RF Bridges," *QST,* April, 1975, page 11.
4. A. Doting and R. Hubbs, "Improvements to the RX Noise Bridge," *ham radio,* February, 1977, page 10.
5. R. Hanson, "Efficiency and Matching Tradeoffs for Inductively Loaded Short Antennas," *IEEE Transactions on Communications,* April, 1975, page 430.
6 G. Hall, "Off-Center Loaded Dipole Antennas," *OST,* September, 1974, page 28
7 D. Covington, "Radiation Resistance of Inverted V Antennas." *OST,* October, 1968, page 36.
8. S. Schelkunoff and H. Friis, *Antennas, Theory and Practice,* John Wiley, New York, 1952. page 162.
9. J Kraus, *Antennas.* McGraw-Hill, New York, 1950, pages 267, 283.
10. W. Orr, *Cubical Quad Antennas,* 2nd edition, Radio Publications, Wilton, Connecticut, 1970, page 15.
11. A. Richtscheid, "Calculations of the Radiation Resistance of Loop Antennas with Sinusoidal Current Distribution," *IEEE Transactions on Antennas and Propagation,* November, 1976, page 889.
12. "Loops versus Dipoles — Where Will It All End?" (Technical Correspondence], *QST,* April, 1977, page 51
13. F. Witt, "Top-Loaded Delta Loop," manuscript to be published by *ham radio* later this year.

ham radio

the ham notebook

Heath HD-1982 Micoder for low-impedance operation

The new Heathkit HD-1982 microphone with installed *Touch-Tone* pad provides a very convenient way to connect a pad into a transceiver which has no auxiliary audio input jack. There is a problem, however, in using the unit with transceivers such as the Drake TR-22C which is designed for low-impedance microphones. The HD-1982 is designed to operate into a load of 10k-ohms or higher.

Thinking that the mismatch might not be too serious, I tried the HD-1982, as designed, with the TR-22C. On-the-air checks were made with four stations. All reported muffled voice quality which cleared up upon changing to the Drake microphone. Some corrective measures were clearly indicated.

An emitter follower circuit was designed and connected into the circuit as shown in **fig. 1.** The input impedance of the circuit consists of the parallel combination of the two 68k resistors and the transistor H_{fe} X (2000), or about 30-k ohms. The output impedance is approximately 2000 ohms. The emitter follower was mounted on a small piece of phenolic board approximately l-inch (2cm) square. The circuit board fits easily in the top of the microphone case between the two mounting posts. The length and routing of the four leads connected into the existing circuitry is not critical. Because the circuit had to be mounted on a board of limited dimensions, the size of the coupling capacitors is critical; I used Caelectro part numbers A1-302 and A1-306. These capacitors are about the size of the head of a match and fit nicely in the available space.

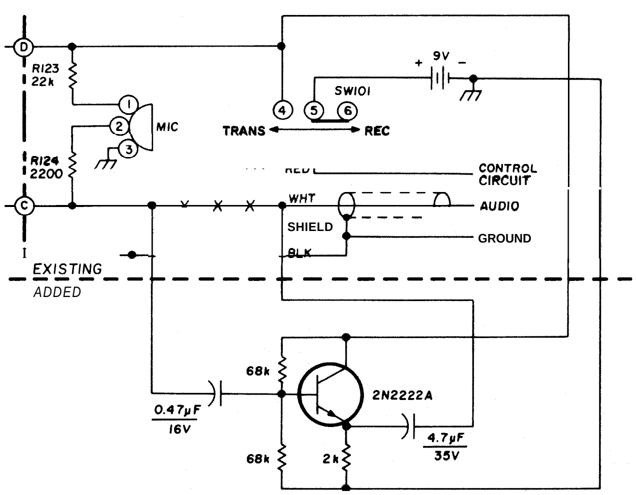


fig. 1. Schematic diagram of the emitter follower used to interface between the HD-1982 Micoder and a low-impedance microphone input.

The modification corrected the problem. On-the-air checks have indicated no essential difference between the performance of the Drake microphone and the modified HD-1982. *Touch-Tone* operation has been very successful with reports of good quality audio.

Wesley Johnson

cleaning teleprinters

My plans to rewire a surplus model 19 Teletype machine and interconnect it to the model 14 TD were hindered by the dirt and oil build-up on the machines. Having an unusually neat and clean shack, it was imperative they be cleaned!

After a few fruitless efforts, the idea to use a commercial degreaser was conceived. A quart can of *Gunk* all-purpose degreaser was sprayed all over the internals (after removing only the motor) using a spray bottle such as a *Fantastik* cleaner applicator. After allowing approximately 20 minutes for the degreaser to work, the machine was rinsed thoroughly using the fine spray of the garden hose hooked to the hot water spigot. Extremely greasy areas were then given a second application of *Gunk,* a light scrubbing with a paint-brush, and more vigorous application of hot-water. Drying can be expedited by setting the unit in the sunshine and/or using your wife's electric hair styler or vacuum cleaner with the hose attached to the exhaust.

After complete drying, the unit should be lubricated to prevent rust. The model 14 TD was also cleaned in this manner. The motors, however, were cleaned with heavy shop-rags and solvent. I have since cleaned several motors using this technique (with *Gunk* and hot water) so I suspect the TTY motors could be cleaned while on their mountings.

As can be imagined, this is a very messy operation. The best place to do it is outside on the gravel driveway. The results, however, are nothing short of fantastic! The aluminum frame sparkles like new; the levers and gears look like the day they were assembled.

It may only be my imagination, but the model 19 seems to run quieter and better since cleaning. It's definitely easier to work on.

F. Neil Urban, W8CD

remote rf current readout

The amount of current flowing in an antenna or a feedline is often more indicative of efficient system operation than vswr. Moreover, having such a current readout conveniently located at the operating position greatly simplifies transmitter adjustment and rapid frequency changes.

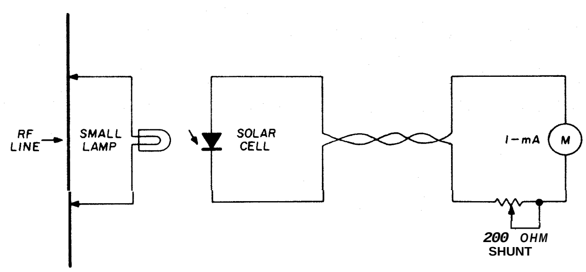


fig. 2. Simple circuit for remote readout of rf current on your transmission line.

The rf current readout described is based on the use of a simple optically-coupled isolator, as illustrated in **fig.** 2. A suitable pilot lamp is illuminated by a small sample of rf and energizes an inexpensive solar cell; the dc current generated by the cell is a measure of relative rf power; and may be routed to a low-current meter located at any convenient point.

A sensitive, low-current pilot lamp is desirable to cause minimum disturbance to normal rf circuit conditions. The Number 48 or 49, 60 mA lamp is suitable for use with transmitters above l-watt output. The solar cell may be an International Rectifier B2M or any similar device. A meter, reading 1 milliampere or less is suitable. A variable current-limiting control, although not absolutely necessary, will add convenience to the system.

The solar cell and lamp may be taped together, using dark tape to prevent light from other sources reaching the cell. The lamp filament should parallel the solar cell surface for maximum sensitivity.

The lamp may be connected directly in series with a conductor which carries very low rf power. For higher current, the lamp is connected across a suitable portion of the line, which then serves as a shunt to limit lamp current to a safe value. Initially, the lamp shunt should have low resistance, and be gradually increased until only useful brilliance is obtained at maximum rf power output.

The rf readout system provides complete isolation between rf and dc circuits, allowing efficient placement of the rf pickup and convenient location of the relative rf-reading meter. During initial adjustment of a new antenna system, several inexpensive pickups may be temporarily installed to monitor rf current in various components.

Gene Brizendine. W4ATE

multiplexed counter displays

We have received many questions from readers concerning the digital display for the counter shown in the article on page 22 of February, 1978, issue of *ham radio.* The author made two references to the type of display used with the counter. In one he stated that, "the multiplexer in the 7208 energizes each LED in sequence." This means that the lines coming from the 7208 (pin 5 for example) are used to turn on (enable) the appropriate digit.

Segment information is also obtained from the 7208 IC. In this case, the same information goes to all the LEDs. Pin 28 of the 7208, would go to the a segment of each LED, and pin 17 to the b segments, etc. During operation, the segment information appears at the same time as the digit enable line. For additional information, refer to the article by John Bordelon, K4JIU, on page 30 of the same issue. He uses the same technique in another version of the same counter.

At the present time, the ICs can be obtained from at least three sources: Circuit Specialists, James Electronics, and Poly-Paks.

Charles Carroll, K1XX

emergency quad antenna repairs

The worst enemy of the cubical quad antenna is the kind of winter storm that simultaneously subjects it to ice and wind loading. Here in the Chicago area we don't get that kind of storm too often, but when we do most quad owners seem to end up with some antenna damage. A quad antenna with a broken spreader, besides being useless as an antenna, is also extremely vulnerable to further damage if not promptly repaired.

A few winters back, I arose one morning to find three broken spreaders on my two-element quad. Not having any replacement spreaders on hand, I nevertheless managed to have the antenna fully operational and structurally sound again within a few hours. I went to the hardware store and purchased a length of 19 x 19 mm x 3 mm (3/4 x 3/4 inch x 1/8 inch) thick aluminum angle stock. I cut it into 30 cm (12 inch) lengths and used two pieces as splints to repair each break, one on each side of the spreader. The splints were secured by wrapping them tightly at each end with no. 14 (1.6mm) solid wire, and a layer of tape was added over the wire for good measure. The antenna was used in this condition for several months with no apparent effect on its performance due to the aluminum splints.

John E. Becker, K9MM

$1.50

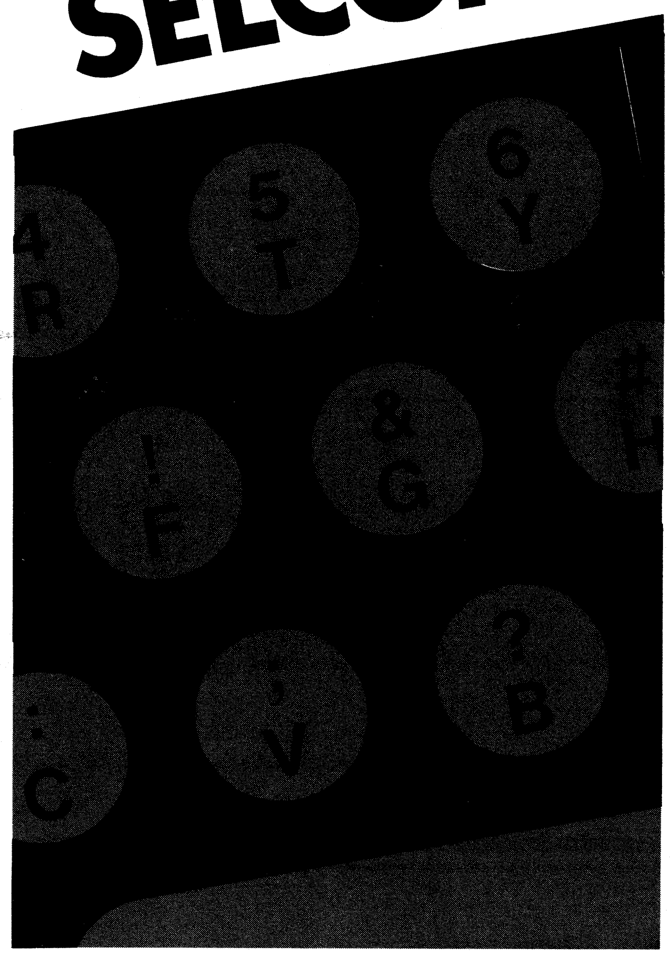
5.75

ham radio

magazine

JUNE 1978

RTTY SELCOM



ham radio magazine

JUNE 1978

volume 11, number 6

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

U.S. and Canada: one year, $12.00
two years, $22.00
three years, $30.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

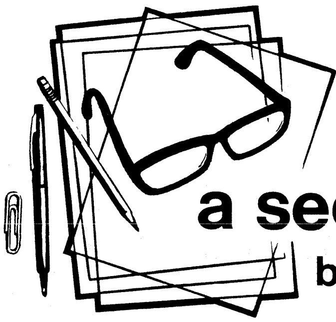
Foreign subscription agents are
listed on page 117

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

A group of atomic physicists at Western Washington University has predicted that sometime this year the first message through the earth, rather than around it by way of the ionosphere, will be transmitted along a beam of neutrino particles from a particle accelerator. The neutrino is one of the fundamental subatomic particles, but one of the more elusive ones — Wolfgang Pauli first proposed its existence on theoretical grounds in 1930; Enrico Fermi christened the new particle the neutrino (for "little neutral one"), but it wasn't until 26 years later, in 1956, that it was first detected by scientists.

The interaction of neutrinos with ordinary matter is so weak that, according to classical theory, a neutrino could pass through a block of lead that stretched from here to the nearest star without disturbing any of the atoms in its path. Since the neutrino carries no charge and has no mass (or nearly no mass, scientists aren't sure), it evaded traditional particle detection methods simply by passing through them without affecting them in any way. Billions of neutrinos from the sun pass through your body every second, day and night, but it's estimated that a neutrino interacts with one of the atoms in your body only about once every ten years. It's no wonder it took 26 years to detect the neutrino's presence!

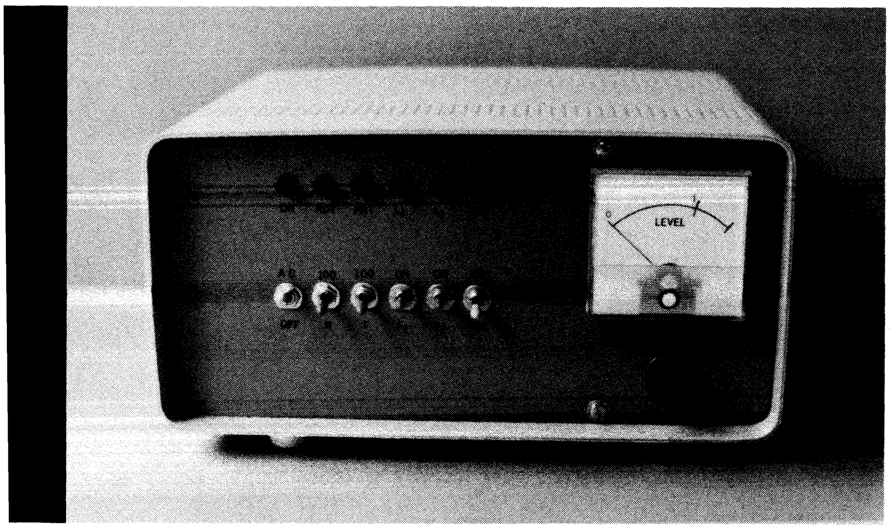
While the average neutrino is capable of passing through most of the matter of the universe without slowing down or losing any of its energy, it's been found that neutrinos fired in eight-second bursts from a high-energy particle accelerator occasionally collide with other particles, at the rate of about one collision for every 17 tons of matter that the beam penetrates. Although neutrinos cannot be detected directly, the particle debris, light, and noise generated by their collisions can be. When a beam of neutrinos is passed through a large volume of water, all along its path some of the collision products emit a forward cone of Cerenkov photons which can be detected by a light collector-phototube system. Prototype experiments using Cerenkov detectors to intercept cosmic neutrinos 1000 meters below the ocean's surface have already been carried out.

In experiments at the 400-billion-electron-volt proton accelerator of the Fermi National Accelerator Laboratory in Illinois, a 20-microsecond pulse of protons is directed into a bar of aluminum — the resulting atomic collisions produce about 10-billion neutrinos per pulse. The beam of neutrinos generates about one reaction per pulse in a bubble chamber containing 25 tons of liquid neon one kilometer away. With a grid of Cerenkov detectors in a large body of water it's predicted that a greater number of reactions per pulse would be detected. If information could be encoded into pulses of the neutrino beam, theoretically a message could be received and decoded virtually any distance away.

In the experiment suggested by the group in Washington, a pulsed neutrino beam from the Fermi Lab accelerator would be directed downward at an angle of about 12 degrees so the beam would pass through the earth and emerge in Puget Sound, nearly 3000 kilometers away. The detector-target would consist of the approximately one-million tons of water in Puget Sound where showers of particles would be recorded with each neutrino collision. The tiny flashes of light from the Cerenkov photons would be recorded and translated into the original message. Funding for the experiment is being considered by a number of government agencies, including the Navy, which is interested in applications of the technique for deep-water communications with nuclear submarines.

An analysis by the Naval Research Laboratory has shown that, if the energy level of the Fermi accelerator was increased to 1000-billion electron volts, with improved beam focusing, the neutrino event rates could be increased by a factor of 10. By using synchronous detection techniques and Cerenkov photo detectors 3000 meters below the surface of the ocean (where they're not bothered by ambient light), it is expected that one 15-bit message per pulse could be transmitted with very low error rates. With one neutrino pulse every 8 seconds, this represents a message rate of 6750 bits per hour. Compared with other methods of communications, this is slow, but unlike radio communications, neutrino beams can't be blocked, they're not affected by solar storms nor dependent on the ionosphere, and they travel great distances with no loss of power.

Jim Fisk, **W1HR**
editor-in-chief



RTTY SELCOM

Discussion of the RTTY SELCOM — an advanced TTL design, providing selective character recognition

One of the first applications of digital logic to RTTY was the RTL SELCAL described by Lamb.[1] Capable of recognizing a single sequence of four characters, it proved very laborious and costly, since there were over 300 wired connections to be made in the basic unit. This concept, however, was expanded and translated into TTL form in the TTL SELCAL described by Branscome.[2] Although this unit was constructionally simpler, even with expanded capabilities, it still did not overcome the cumbersome decoding process, or provide for easy expandability.

Shortly after the TTL SELCAL was introduced, the CATC group[3] tackled the problem of sequential character recognition, hoping to overcome the two main problems presented by the TTL SELCAL. By 1972 the objectives had been met, and circuit boards were fabricated for what was known as SELCOM I. This unit dramatically expanded the flexibility to decode all 32 Baudot characters and could recognize nearly unlimited strings of characters. Many of the sequences were to be used as selective station control commands, hence the name.

Probably the most powerful discrete logic sequential decoder ever developed, the SELCOM was also far easier to program than the earlier SELCALS since only one connection, instead of five, was required per character. This ease in decoding had been achieved by the use of a 1-of-32 decoder. In this manner, the Baudot character set was decoded, providing access to characters, rather than the bits as in the SELCAL versions. And, to provide expandability, a 32-character bus was connected to every sequential decoder board. This bus could be expanded as desired since each line of the bus was capable of handling up to 500 TTL loads.

In May, 1973, the discrete bus drivers were replaced with TTL buffers, lowering the drive to 30 instead of 500 loads. Even with this change, SELCOM II still retained the versatile bus structure of the original version.

Version III of the SELCOM incorporated a MOS UAR/T. This chip, a natural for the SELCOM, had already been in use for several years by various computer manufacturers, but the single quantity price was prohibitively high for amateur work until about December, 1973. Offering significant simplification, the UAR/T provided all functions except those of the clock and character decoders, and in addition, offered new functions not available in earlier versions of the SELCAL or SELCOM. They included regeneration of the received RTTY signal, speed conversion, and the ability to handle any code of 5 through 8 bits with only a simple jumper change.

By Robert C. Clark, K9HVW, Archie Lamb, WB4KUR, and Fred R. Scalf, K4EID. Mr. Clark **may** be reached at 930 Chestwood Avenue, Tallahassee, Florida 32303.

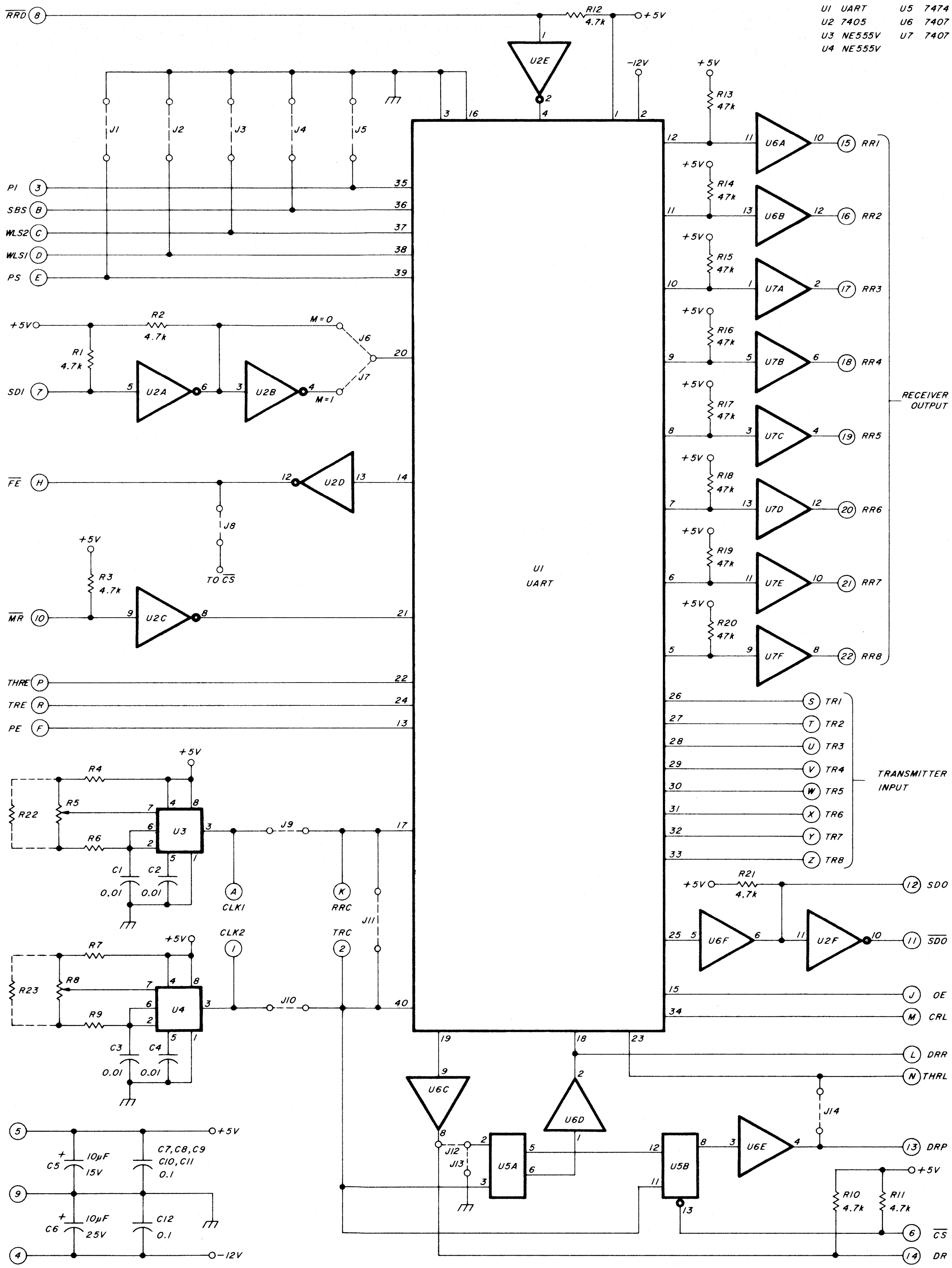


fig. 1. Schematic diagram of the **DU-200** Universal UAR/T module. The jumper placement is explained in the text and also table 1. The UAR/T is available from either Texas **Instruments** or General Instruments. The buffers for the receiver output and flag lines can be eliminated if the lines are used for feeding only **one TTL** load.

In early 1974, an attempt was made to eliminate the mechanical problems presented by the double-width cards used in earlier SELCOMs. To do this, the versatile bus structure was abandoned in favor of a functional module approach. These modules provided a versatility not possible in the earlier versions of the SELCOM. With the change to single-width cards,

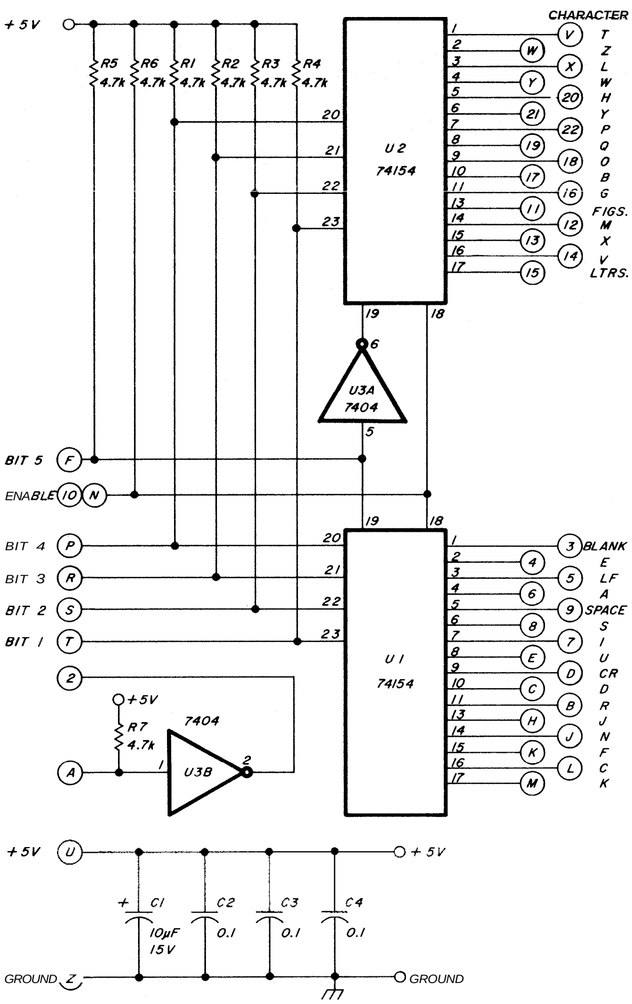


fig. 2. The DU-210 character decoder uses two 1-of-16 decoders as a 1-of-32 decoder. Both 74154s receive bits 1 through 4, while bit 5 is used to select the appropriate 74154. The single inverter is provided if more than one DU-210 decoder is used.

the modules of SELCOM IV are now the same physical size as the DT-500[4] and DT-600 boards.

The group of modules to be described form SELCOM IV, or individual modules may be used in other applications. The modules include:

1. DU-200 Universal UAR/T
2. DU-210 Expandable 1-of-32 Character decoder
3. DU-220 Sequential Decoder
4. DU-300 Mini-SELCOM

SELCOM IV may be used with 5, 6,7 or 8 bit codes at speeds up to 9600 bits per second. The following description is tor the five-bit Baudot code used for amateur RTTY at 45.45 and 74.2 baud, but the unit is designed for expansion to the full 64-character ASCII code group.

SELCOM features

DU-200 Universal UAR/T. The DU-200 **(fig. 1)** is not only the heart of the SELCOM system, but also provides a powerful functional module for many other applications as well. It may be used for teleprinter signal regeneration, speed conversion, serial-to-parallel data conversion, parallel-to-serial data conversion, cede conversion (Baudot to ASCII, ASCII to Baudot, Baudot to Morse, etc.), and many other ways. All the features of the UAR/T have been made available in the DU-200, either through hard-wired jumpers or through external control. In this way, the same board may be configured as a Baudot or ASCII regenerator, an interface between a serial RTTY station and a parallel I/O port of a microprocessor, a SELCOM, or a wide variety of other applications.

The DU-200 consists of the UAR/T, interface (buffering), control, and clock functions. The UAR/T IC (U1) functions as two nearly independent circuits, a digital receiver and a digital transmitter. The receiver accepts serial data in a particular format (selected by the user), checks for format errors (parity error, missing stop bit, etc.), and outputs the data in a parallel form. The transmitter accepts parallel data, adds start, stop, and parity bits, and sends the data in a serial format at the data rate selected. If the parallel output of the receiver is connected to the parallel input of the transmitter, the unit functions as a regenerator. The two sections may be used independently, but the data format for both sections must be the same. That is, both the receiver and transmitter must operate with the same code, parity, and so on. Under certain conditions they may operate at different speeds.

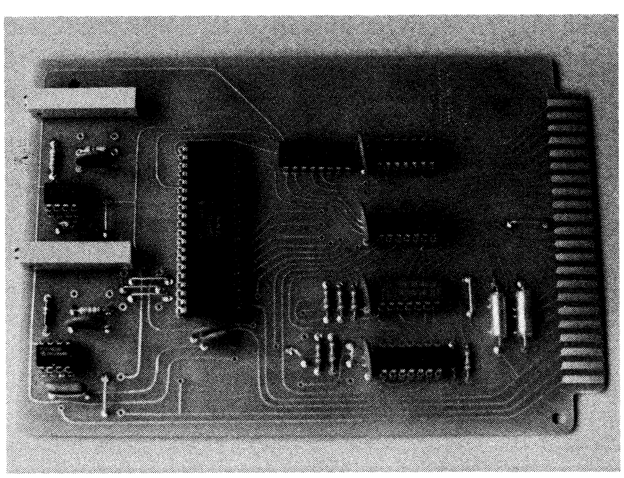
The UAR/T is capable of accepting data with up to 43 per cent distortion (more in some cases) and re-sending it with less than 1 per cent distortion. Typical teleprinters are capable of accepting less than 30 per cent distortion, while many keyboards and transmitter distributors generate signals with large amounts of distortion. As machines age, their ability to accept distortion is diminished and their ability to produce it is increased. The use of the DU-200 as a regenerative repeater offers improvements to all mechanical teleprinters, keyboards, and transmitter distributors.

Under marginal conditions several undesirable situations may exist with the mechanical teleprinter. Typically, high-frequency propagation phenomena tend to add distortion to that which already exists on the transmitted signal. For this reason, it is highly desirable that the transmitted signal have a minimum of distortion. If the DU-200 is used to process the transmitted signal, then this criteria will be met. Even if the transmitted signal is perfect, distortion will be added by the time the signal reaches the receiver.

Another problem that exists with a mechanical teleprinter is that a short noise pulse may be read as a *start.* When this happens, a clutch is released, beginning the sequence of events which decode and print a character. The teleprinter shaft must complete one full revolution before it can recover from this premature start. If the real start bit is received during this revolution, the printer will not be able to get back in synchronization with the sending station. In such a situation, the printer may print *garbage* for several characters. Most brands of the UARIT though, after receiving what appears to be a start bit, recheck to determine that the start bit is still at the appropriate level in the middle of the bit. If the start bit is not valid, then the UAR/T is immediately reset. Thus, the probability of the receiving station staying in synchronization with the sending station is greatly improved.

Another undesirable condition exists when the received signal drops below the noise level and garble is printed. If the signal is not capable of providing the necessary information for character recognition, then it is quite likely that the appropriate level will not be maintained during the stop bit. This is termed a Framing Error and the UAR/T provides a flag to indicate this error. The flag may be used to suppress the transfer of the character to the transmitter section. If this feature is selected, the mutilated character will not be printed. A similar feature is available for parity errors on received characters.

The UAR/T will respond only to those characters that appear to be valid RTTY. It will not respond to a steady space (a single *blank* character will be transferred to the transmitter section unless the Framing Error flag has been used to suppress the transfer)



Photograph of the DU-200 Universal UAR/T board.

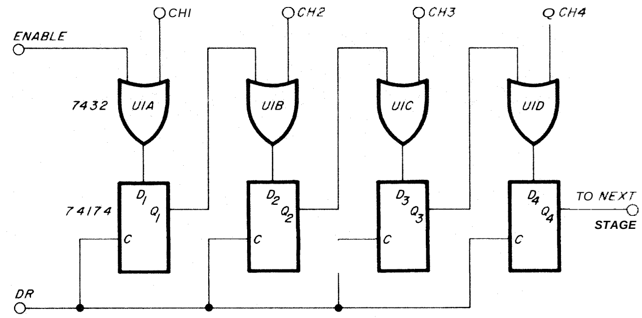


fig. 3. Diagram of the DU-220 sequential decoder. Several similar sections are on each board. Each channel input is connected to the desired character output from the DU-210 decoder.

while most noise and CW will transfer fewer characters to the printer than if the UARIT were not present. With the DU-200 on-line, the appearance of the page is dramatically improved, with any demodulator.

If the clock applied to the transmitter and receiver sections of the UARIT are set for different rates, the UARIT will function as a speed converter. For example, if a 100 wpm (74.2 baud) printer is used, it's possible to receive any speed up to 100 wpm without expensive and noisy gear shifts. By providing a buffer memory between the receiver and transmitter sections, the DU-200 can function as a down converter. Of course, if the size of the buffer is finite, then the UARIT receiver will deliver characters to the buffer faster than the transmitter section clears them, causing the buffer to overflow. In the case of overflow, the UARIT provides an Overrun Error flag which may be used to signal an external device to withhold further characters.

DU-210 Character Decoder. The DU-210 (fig. **2**) recognizes which one of 32 possible characters has actually been received by the UAR/T in the DU-200. Several DU-210 boards may be used to recognize characters from larger character sets. Two DU-210 boards may be used to recognize the 64 characters of the ASCII-6 subset and four may be used to recognize the 128 characters of the full ASCII set. In fact, the DU-210 may be used in many applications where one particular binary code must be recognized.

Sequential Decoder. The DU-220 (fig. **3**) works with the DU-200 and DU-210 to recognize sequences of characters. It might be wired to recognize the station call, setting a latch when the call is received. This latch could be used to prevent the station printer from operating until the call was received. In fact, the DU-220 is capable of detecting a number of sequences, each of which may control some event such as:

1. Turn on reperforator.
2. Turn off reperforator.

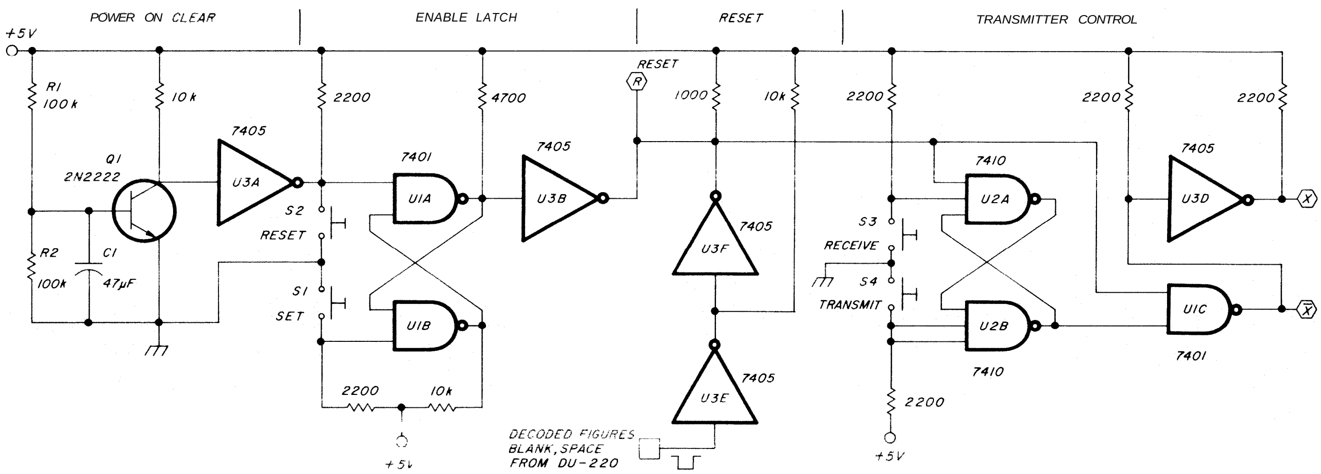


fig. 4. Schematic diagram of the power-up clear and latch circuitry. By using open-collector inverters, the reset line can be fed from several diferent sources. This circuit ensures that the transmitter control will come up in the transmitter off condition.

3. Start CW identification.

4. Turn transmitter on.

5. Turn transmitter off.

If necessary, these control sequences may be configured so that they will be recognized only when the source is the local keyboard.

The DU-220 is quite versatile in that it may be used to detect a sequence of events independent from the SELCOM system. For instance, the DU-220 might be used to recognize a sequence of digits from a *Touch-Tone** decoder. Certain sequences could be prevented from reaching the telephone lines, with others being used to control repeater functions. In addition, the DU-220 could also be used to recognize a sequence of switch closures in an electronic lock.

circuit description

DU-200 UAR/T Board. Serial data for the UAR/T is first applied to pin 5 of U2A. The input to pin 20 (Serial Data Input) of U1 is jumpered to either pin 4 or pin 6 of U2 depending on the sense of the data (mark low or mark high). The UAR/T is programmed, by either hard wiring or external devices, as shown in table **1.**

table **1. UAR/T** programming information.

function	format	conditions (UART pins)
Code Length	5 bit	pin 37 low, pin 38 low
	6 bit	pin 37 low, pin 38 high
	7 bit	pin 37 high, pin 38 low
	8 bit	pin 37 high, pin 38 high
Parity Selection	Parity	pin 35 low
	No Parity	pin 35 high
Odd/Even Parity	Odd Parity	pin 39 low
	Even Parity	pin 39 high
Stop Bit Select	One Stop Bit	pin 36 low
	Two Stop Bits	pin 36 high (for some manufacturers, a 1.5 bit stop is provided for 5 bit codes).

The speed of the received data is selected by an external clock set to sixteen times the baud rate. For 60 wpm, the clock would be set to $45.45\ Hz \times 16 = 727.27\ Hz$. On the first mark to space transition, an internal counter is reset and allowed to count clock pulses. Each brand of UAR/T has some provisions for verifying that the mark to space transition was a valid start bit. If the start is verified, the counter continues, in turn controlling a serial-shift register, so that each received bit is stored in the shift register. At the time of the expected stop bit, another check is performed. If the stop (a mark) is not present at the required time, the Framing Error (FE) flag is raised to indicate an invalid condition. In the same manner, if a parity check has been requested, and the proper parity is not verified, the Parity Error (PE) flag is raised. The Overrun Error (OE) flag is available to indicate that one character has not been removed from the receiver holding register before the next character took its place. These flags may be used to control indicators, keep the character from being printed, or control error-correction schemes.

When a complete character has been received and transferred to the receiver-holding register (only complete characters appear at the receiver holding register output), pin 19 of the UAR/T (Data Ready) goes high to indicate that the character is available in parallel form on pins 5 through 12 of U1. U5 is a two-stage shift register clocked by the receiver clock. The high from pin 19 of U1 (DR) is transferred through U5 after the next two successive clock pulses. The out-

*Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.

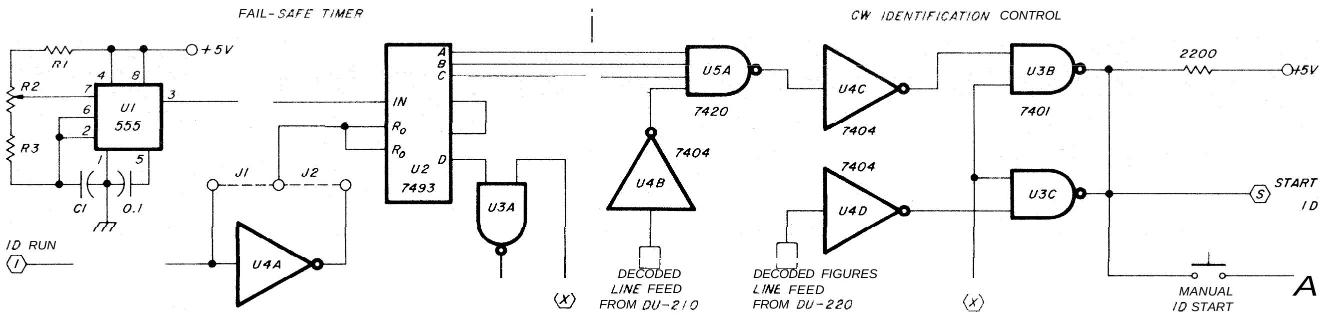


fig. 5. By using this fail-safe timer, the transmitter will be turned off if an identification is not sent every 10 minutes. As with all other diagrams, the power supply connections have not been shown. If the return line from the identification unit is low when the ID is running, J1 should be connected; if high, J2 is connected.

put of the first stage (U5A pin 6) is applied to the Data Ready Reset ($\overline{DRR}$) of U1, resetting the DR line. In addition, the $\overline{Q}$ output of the second stage feeds the Transmitter Holding Register Load (THRL) line of U1, loading the parallel data from the receiver into the transmitter holding register. This action of U5 guarantees that DR stays high for exactly one clock period. Some brands of UAR/T require a rising edge for THRL, while others require a falling edge. The DU-200 provides both a rising and a falling edge, and hence will function with most of the UAR/Ts on the market. The direct clear for the second stage (U5 pin 13) may be used to prevent the character from being loaded into the transmitter holding register. Jumper J8 allows the Framing Error flag to suppress characters with a missing stop pulse. A similar procedure may be used for parity errors and overrun errors.

The UAR/T handles steady spacing in an interesting manner. If J8 is omitted, then a steady space will cause a single blank character to be transferred to the transmitter. No other characters will be transferred to the transmitter until the data line returns to mark and a valid start pulse is detected. If jumper J8 is installed, then no characters will be transferred to the transmitter. Hence, it is not possible for the printer to run open. This action is superior to the anti-space offered on the ST-6 and DT-600.

The character transferred to the transmitter holding register will in turn be transferred to the transmitter register when empty. Notice that the UAR/T may be simultaneously processing 3 characters, sending one character, holding a second character, and receiving a third character. The status of the transmitter registers is indicated by pins 22 (Transmitter Holding Register Empty) and 24 (Transmitter Register Empty) of U1. When a character reaches the transmitter register it is clocked out in serial form (at pin 25 of U1, Serial Data Out) according to the format previously selected, at a speed determined by the transmitter-register clock. If this clock is the same one used for the receiver, then the unit operates as a regenerative repeater. If the clocks are of different frequencies, then the DU-200 operates as a speed converter. If the receiver speed is higher than the transmitter speed, the characters may arrive at the transmitter holding register faster than they may be accepted and a buffer memory must be provided to avoid overrun.

Two clocks are installed on the DU-200 board. In addition, jumpers are provided so that one clock may be used to operate both the receiver and transmitter, or a separate clock may be provided for each. It is also possible to supply clock signals from an external source. A crystal-controlled clock, supplying multiple baud rates, has been designed as part of the CATC line. The 555 IC (U3 and U4) has proven to be adequate as a clock, as long as the ambient temperature is relatively stable. Wide frequency excursions can be expected with wide temperature variations. Most of this frequency shift can be attributed to the thermal characteristics of the resistors and the capacitor which form the RC timing portion of the oscillator. Choice of components, to minimize this shift, will improve the drift characteristics of the oscillator. A polystyrene capacitor is recommended. Also, metal-film resistors will show a significant improvement over the carbon composition types. The configuration of the timing resistors (R4, R5, and R6 for U3) was chosen for stability and is superior to most shown in other articles.

Oscillator frequency is the only adjustment required for the DU-200. The frequency should be set in accordance with **table 2,** while measuring at pin 3 of the 555.

Since the UAR/T is only capable of sinking one TTL load, buffering has been provided on the output and flag lines. If only one load is to be connected to any

table 2. Oscillator frequency of the *555* timer.

speed	baud rate	clock frequency
60 wpm	45.45	727.273 Hz
European	50.00	800.00 Hz
75 wpm	56.83	909.280 Hz
100 wpm	74.18	1186.880 Hz

UAR/T output, and none of the flags or control signals are to be used externally, then U6 and U7 may be eliminated and a jumper used to complete the circuit.

The UAR/T is available from a number of manufacturers. The Texas Instruments TMS-6011NC, General Instrument AY-5-1013, and the Western Digital TR-1602A have all been tested in the DU-200. Difficulties have been experienced with the TR-1602A when the received data did not have a stop bit. Therefore, only the TMS-6011NC and AY-5-1013 are recommended.

The DU-210. The buffered outputs of the UAR/T receiver-holding register (RR1 through RR8 on the DU-200 board) are connected to the one-of-32 decoder on the DU-210 board. Two 74154 (U1 and U2) decoders are used. The first four bits of the received character are used to address the two decoders, while the fifth bit is used to select one of the decoders. When a decoder is selected (pins 18 and 19 low) only one of the output pins will be low. Hence, if the enable line of the DU-210 is held low, then only one of the 32 output lines will be low for any five-bit binary code on the input lines. The characters of the Baudot set have been shown on the output lines in fig. 2.

Because of the nature of the UAR/T, it is permissible to leave the DU-210 enabled at all times, since only complete characters are available at the output lines of the receiver holding register. There is no need to worry about glitches at the output of the DU-210 as one character replaces another in the DU-200.

Several DU-210 boards may be used for a character set larger than the 32 characters in the Baudot set. In such a case, only one enable input at a time may be allowed to go low. The extra section of U3 has been provided to facilitate such connections.

The DU-220 Sequential Decoder. As shown in fig. 3, selected character lines from the DU-210 are connected to the inputs of the DU-220. Suppose that the enable line of the DU-220 is low and that characters A, B, C, and D from the DU-210 have been connected to CHI, CH2, CH3, and CH4 respectively of the DU-220. When the DU-200 receives an A, it is recognized by the DU-210, and the CH1 line of the DU-220 pulled low. Only when both inputs of U1A are pulled low does the output go low. This low is presented to the D input of U2A. Shortly after the character is recognized, the DU-200 DR line goes high, indicating that a character has been received and is stable in the receiver holding register. The leading edge of the DR line pulse clocks the D level through to the Q output of the flip flop. Since the other character lines are not low, all other flip flops in the chain will be reset producing a 1 on the Q output. The high from the output of the first D flip flop is used to enable U1B. If the next character is a B, then the other input of U1B is pulled low (the output of U1A will not change until the next DR pulse is received). This output is applied to the D input of U2B and is transferred to the output when the DR line goes high. If the next two characters are C and D, this sequence is continued through to U2D. The low on the output of U2D indicates that A, B, C, and D have been received in that order. If at any time a different character is received, or the characters are not in the right order, the sequential decoder is reset. It will respond *only* to the right characters in the right order. The output of this sequential decoder may then be used to control a number of station control functions.

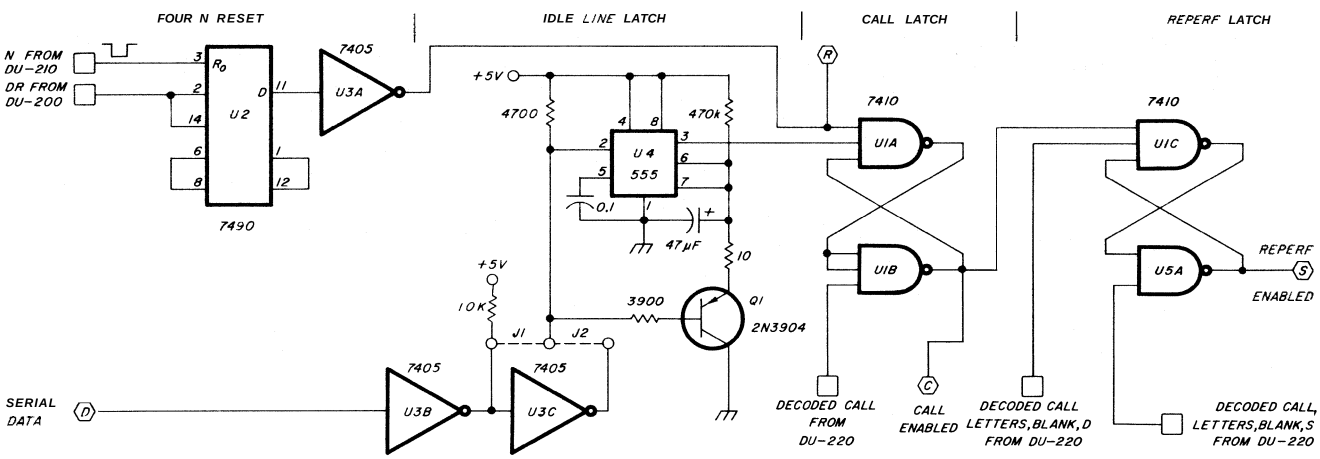


fig. 6. The printer-control portion prevents the printer from operating until the correct sequence of letters is received and the call latch is set. This latch is reset by a four N reset, the idle line reset, or by the normal reset line. The serial data for the idle line reset can be either high or low, with the appropriate jumper connected. For a mark low, use J1 and for mark high, use J2.

station control

Power-up Circuitry. When digital equipment is first turned on, latches and flip flops come up in random states. The portion of the station control logic shown in fig. 4 is a power-up reset which resets all control functions when the power is first turned on. It also guarantees that automatic features are enabled only when desired by the operator.

The latch formed by U1A and U1B is used to enable any automatic functions. The operator may set or reset this latch with S1 and S2. Q1 and U3A form the power-on clear portion of the circuit. As the five-volt

the input of U2, a 7493 four-stage binary counter. When the 7493 has counted eight clock pulses (10 minutes) the D output goes high. This high is inverted by U3A and the resulting low is used to pull the reset line down, resetting the transmit/receive latch.

The counter may be reset to zero at any time by starting the CW identification device. A line from the identification device resets the 7493 when the ID

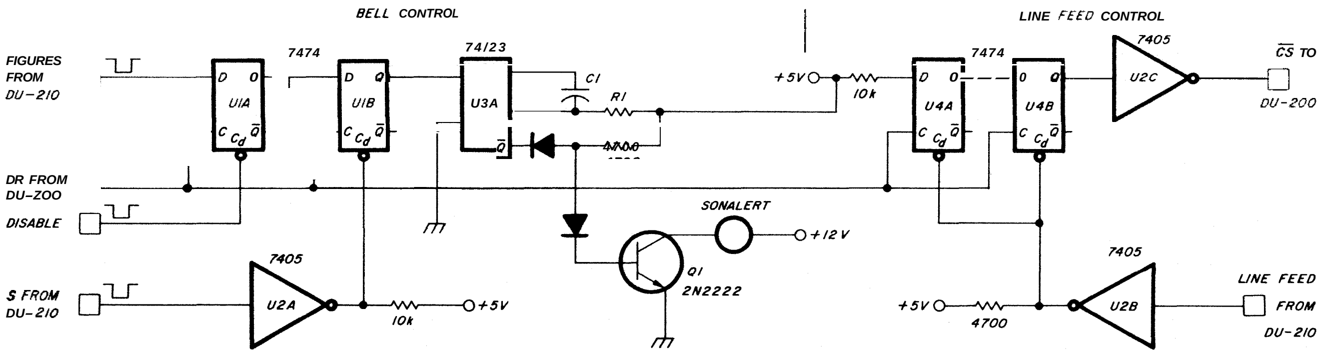


fig. 7. Bell and line-feed control. This circuit prevents the excessive ringing of the bell and the excessive line feeds.

supply starts up from zero, the base of Q1 is held low by the $47\mu F$ capacitor. The capacitor begins to charge slowly through R1, but the voltage on the capacitor will not turn on Q1 for several seconds. By this time the five-volt supply has stabilized and the U1A latch is reset. The R output is low any time the latch is reset, and is used to disable or reset other circuits in the station control. The 7405 open collector hex inverter allows the output of the enable latch to be "QR tied" with other resets, in this case from U3F. U3E and U3F provide an additional reset from the DU-220, resetting all latches.

Fig. 4 also shows the transmit/receive latch. In addition to being set or reset manually by S3 and S4, the latch is reset from the reset line. This means that the transmitter will be turned off by the power-up reset, a manual reset of the enable latch, or by the reset function (figures, blank, space). The provision for a reset function allows the transmitter to be turned off by a code typed on paper tape. Thus, the operator may cut a tape (concluding with the reset sequence) and then look for something more interesting to do than sit and watch the tape play. At the end of the tape, the reset sequence will turn the transmitter off.

Identification and fail-safe timer. The fail-safe timer (fig. 5) guarantees that the transmitter does not stay on the air for an unintentional extended period of time (as when the tape tangles and tears). U1 is another 555 which is enabled only when the transmitter is on the air. The timing components have been chosen so that the period of the oscillator is 1.25 minutes. The pulses from the 555 are applied to

starts. Options are provided so that the CW run line may be active low or active high.

If the 7493 reaches a count of seven without being reset, then pins 12, 9, and 8 are all high. When the next line feed is received, U5A is enabled, and if the transmitter is on, U3B provides a low, starting the CW identification device. The keyboard and transmitter distributor must be inhibited though while the CW device is running.

It should also be possible for the operator to start a CW identification earlier than 8 minutes and 45 seconds into the transmission. In my case, the sequence figures, line feed is used to insert a CW identification at any time the transmitter is on. The outputs of U3B and U3C are OR tied to provide both automatic and semi-automatic identification. A manual push button is also provided to start the identification.

Printer control. The call latch (fig. 6) is mainly used to prevent the station printer from operating until the correct sequence is received. When the DU-220 has recognized the four-character sequence (in my case letters H, V, W) the call latch is set. It may be reset in any of the three different ways: by the sequence N, N, N, N, by a lack of activity on the serial data line, and by the enable latch.

The four N counter (U2) counts DR pulses as long as the N line from the DU-210 is low. When the counter reaches four it is forced to a count of nine. The resultant high on pin 11 is inverted and used to reset the call latch. A unique feature of this counter is that it will automatically reset on the first character after the fourth N.

As all stations have not developed the procedure of sending four Ns at the end of each transmission (or they might be sent, but not received) another method of resetting the call latch is desired. The method chosen is to monitor the serial data line for mark/space transitions. If no transitions are detected for thirty seconds, the call latch will be reset by the action of U4. This 555 is configurated as a monostable with a period of thirty seconds. Each time the serial data line goes to the space level, the 47 μF capacitor is discharged through Q1 and the sequence begins again. Pin 3 of the 555 goes low after thirty seconds of no transitions and resets the call latch.

Another latch (U1C and U5A) is used to control a reperforator. The sequence letters, H, V, W, letters, blank, S sets the latch and turns the reperforator on (the call and reperforator latches are used to enable and disable the DD-350 selector magnet driver and motor control). The reperf latch is reset by the sequence letters, H, V, W, letters, blank, D or any sequence or event that resets the call latch.

At times it is helpful to control machine functions to save wear and tear both on the operator and machine. One function that is not always valuable is the bell. The operation of the bell on random noise is irritating and the excess use of the bell by some operators is infuriating. Fig. **7** shows one method of controlling the bell function. Initially, the printer bell is disabled and replaced by a Mallory *Sonalert.* U1A and U1B form a 2-bit shift register. The first stage is enabled by a figures function and the count is allowed to continue if the next character is an S. The output of U1B goes high when the sequence figures, S has been received, triggering U3A which controls the time the *Sonalert* is on. The reset input of U1A may be used to inhibit the operation of the bell. One way to use this would be to connect the disable pin to the output of the call latch in **fig. *6.*** In this way, the bell will ring only when the station call has been sent and then the sequence figures, S 'is received. This should eliminate all bells directed at someone else and the repeated ringing of the bell.

In a similar manner, the machine may be prevented from responding to a sequence of line feeds. U4A and U4B count DR pulses as long as the received character is a line feed. When two consecutive line feeds are received, the output of U4B goes high, is inverted, and the character suppression input of the DU-200 is pulled low, preventing the character from being transferred to the printer. As long as line feeds are sent, no characters will reach the printer. When a character other than a line feed is received, the U4A/U4B counter is reset and characters will again reach the printer. If the operator wishes to double space, a sequence of carriage return, line feed, letters, line feed, letters will circumvent the line feed counter and allow two lines to be turned up.

Many other station control functions may be provided for with the SELCOM system. Another possibility would be to establish two frequencies within the same band, one for general calling and the other for third-party traffic. Station **A** may call station B on the general calling frequency and send a code to switch station B to the traffic frequency. Station **A** then sends traffic to station B on the second frequency, resetting station **A** to the calling frequency at the termination of the traffic. The SELCOM can control almost any function that the operator can dream of.

summary

The SELCOM is a powerful digital building block which may be used to implement a wide range of station control applications. In a subsequent article, the DU-300 will be presented. The DU-300 Mini-SELCOM is a single board which provides many of the functions of the DU-200, 210, 220, and the station control board. The DU-300 provides regeneration, speed conversion, call-letter recognition, four N turn-off, printer control, and two other (user defined) functions for station control. Also, the trade-offs between the SELCOM and the Mini-SELCOM will be discussed.

All correspondence should be addressed to Robert Clark at the address indicated at the beginning of this article. All inquiries with a self-addressed stamped envelope will be acknowledged.*

acknowledgements

WØRQI, K4OAH, and W6DNT were very helpful in the design of the early versions of the SELCOM. WA4SNL did the layout for versions II, III, and IV. W8IDJ assisted in preparing tables and parts lists. K4TKU was instrumental in the preparation of high-quality drawings as he has done for other CATC project efforts. WA5IAT was particularly helpful with his work on the control section of the DU-200 and in providing comments on the draft article. We extend our appreciation to each of these who have helped, hoping we have not overlooked anyone.

references

1. Tom Lamb and Bill Malloch, "The SelCal," 73, May, 1968, page 58.
2. Kenneth Branscome, "The TTL SelCal," The RTTY Journal, December. 1971, page 7.
3. Robert C. Clark, K9HVW et al, "DT-600 RTTY Demodulator," ham radio, February, 1976, page 8.
4. Robert C. Clark, K9HVW, et al, "The DT-500 RTTY Demodulator," ham *radio,* March, 1976, page 24.

ham radio

*As with the DT-500, DT-600. DI-70, and DD-350, boards are being produced by Data Technology Associates, Box 431912, Miami, Florida 33143. The DU-200 board is available with construction notes for $12.50 plus $.75 for first-class postage. The format is the same as the DT-600.

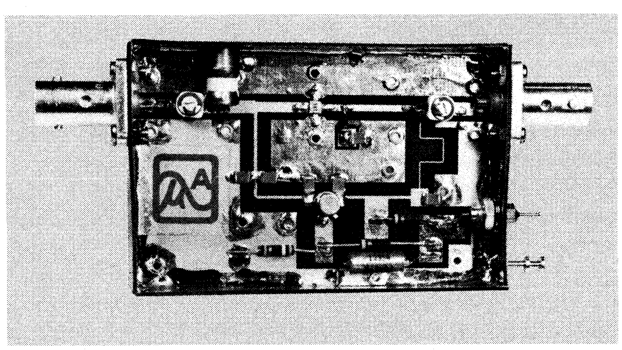
receiving preamplifier
for OSCAR 8 Mode J

Amateurs who have designed and built vhf and uhf transistor circuits in the past are well aware of the fact that high performance often seems to be more art than science; however, the "art" involved in the design of vhf/uhf amplifiers is rapidly giving way to science with the utilization of S parameters and computer optimization. Using these techniques, an engineer can design a multi-stage vhf or uhf amplifier in a few hours with the aid of a computer and be highly confident of the results. The low-noise 435-MHz preamplifier described in this article is a good example of the combination of the manufacturer's transistor S-parameter data, engineering judgment, and computer optimization.

design approach

Virtually all manufacturers of transistors intended for vhf, uhf, and microwave applications now utilize *S* parameters to characterize the performance of their high-frequency transistors. This fact alone attests to the usefulness of this parameter set in the design of high-frequency amplifiers.[1,2] In conjunction with specific noise figure data and bias point considerations, the addition of a computer analysis and optimization program, and sound engineering judgment, you have all the ingredients necessary for a successful design.

This article describes the design of A low—noise 435-MHz receiving preamplifier which is intended for the reception of the downlink communications channel of the OSCAR 8; the preamplifier also provides excellent performance for communications on 432 MHz. Basically, the design approach is developed as outlined below:



The 435-MHz preamp using the Microwave Associates 42141 transistor. The extra chip capacitor, on the collector lead of the bias transistor, is used to ensure a good ac ground. Note the extensive grounding between the two sides of the printed circuit board.

1. Selection of the appropriate transistor.

2. Determination of the terminal impedances (both input and output) required to obtain the specific performance objectives.

3. Synthesis of the appropriate matching networks to present the desired terminal impedances.

4. Computer optimization of network component values.

5. Stability analysis over a broad band of frequencies.

transistor selection

The main criteria in the selection of a transistor for this application is low noise figure coupled with sufficient gain to minimize the second stage contribution to the system noise figure. The intended device should also be completely specified and characterized in terms of S parameters and noise figure data. In the absence of such data, extensive analysis of the amplifier circuit is impossible unless, of course, the designer is willing to perform the transistor evaluation and characterization himself.

The Microwave Associates 42140 series of uhf transistors is ideally suited for this application. The devices are completely specified and characterized over a broad range of frequencies. At optimum bias, or dc operating point which results in minimum noise figure ($V_{CE} = 8$ volts, $I_C = 5$ mA), the MA42141 has the following characteristics:

$$\begin{aligned} S_{11} &= 0.64 \angle -112^\circ \\ S_{21} &= 7.70 \angle 111^\circ \\ S_{12} &= 0.046 \angle 41^\circ \\ S_{22} &= 0.72 \angle -35^\circ \\ \Gamma_{os} &= 0.32 \angle 38^\circ \\ NF_{min} &= 1.80\ dB \end{aligned}$$

where Γ_{os} is the source reflection coefficient required for minimum noise figure.

transformation networks

From this data, the design task is to synthesize matching networks which present Γ_{os} to the input of

By Mark Pride, K1RX, and Kenneth V. Puglia, Microwave Associates, Inc., Burlington, Massachusetts 01803

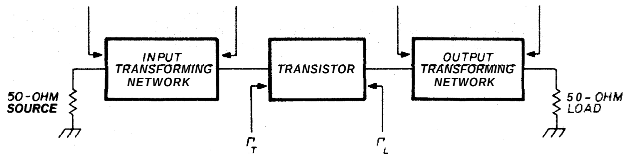


fig. 1. Block diagram of the low-noise preamplifier stage showing the input transformation network, the transistor stage, and the output matching network with applicable reflection coefficients.

the transistor while simultaneously presenting a complex conjugate match at the transistor's output. If low noise figure was not the prime objective, then the design task would be to synthesize input and output networks which would simultaneously provide the complex conjugate impedance to the input and output of the transistor; this would result in maximum gain but not lowest noise figure.

The initial use of the computer is in determining the complex conjugate of the transistor's output impedance. The computer performs this calculation through a program which solves the equation:

$$\Gamma_L = S_{22} + \frac{S_{21}S_{12}\Gamma_{os}}{1 - S_{11}\Gamma_{os}} = 0.35 \angle -54^{\circ}$$

Also of interest is the input impedance of the transistor which may be calculated from:

$$\Gamma_T = S_{11} + \frac{S_{21}S_{12}\Gamma_L{}^*}{1 - S_{22}\Gamma_L{}^*}$$

The asterisk designates the complex conjugate.

The design tasks can be more easily understood by studying **fig. 1** which illustrates, in a block diagram form, a cascade of the input network, the transistor, and the output network. **Fig. 2** is a Smith chart plot showing the locations of the various impedance points. The Smith chart will aid in the synthesis of the transforming networks. (Note that a 150-ohm resistor has been added in shunt across the output of the transistor to provide a margin of stability to the amplifier since the initial analysis in determining the complex conjugate of the output impedance indicated that the transistor was potentially unstable when terminated with these impedances.)

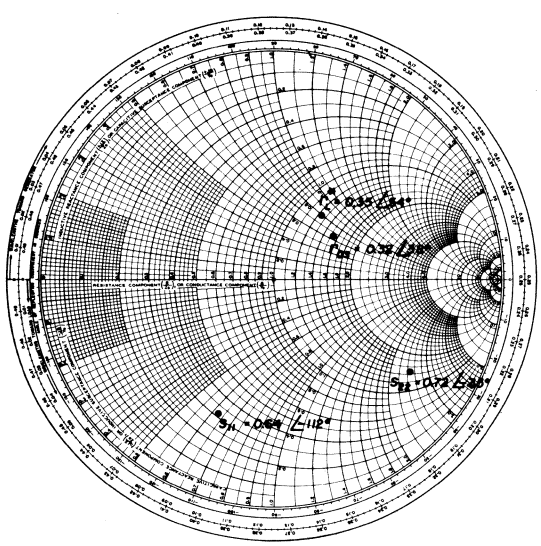


fig. 2. S-parameters for the MA42141 transistor operated at V_{CE} = 8 volts, I_C = 5 mA.

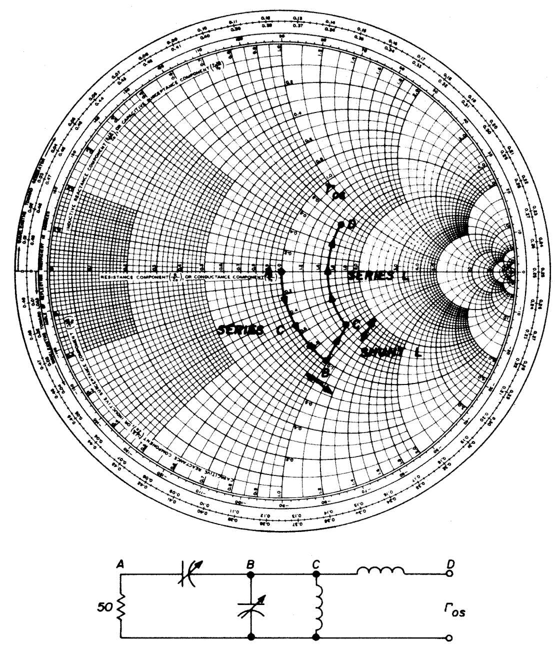


fig. 3. Design of the input impedance transforming network. The Smith chart plot gives preliminary component values which will be optimized with a computer program.

The networks required to transform the 50-ohm source and load to the desired impedances may be designed with the Smith chart. Smith charts are an indispensable tool in the design of impedance transforming networks which use reactive circuit elements and transmission lines. References 3 and 4 provide a clear understanding of Smith charts and their applications.

Figs. 3 and **4** may now be used to determine "ballpark" values for the transformation networks. Variable capacitors have been employed at the input to allow for normal transistor manufacturing variations as well as to extract the absolute minimum noise figure available from the transistor. A single variable capacitor at the output allows you to peak the gain and to minimize the output impedance mismatch.

Variable capacitors are recommended for this application because it's doubtful that either the antenna or the receiver used with the preamplifier will provide the desired 50-ohm impedances; in military and commercial designs where the source and termination impedances are known to be 50 ohms, fixed capacitors are usually installed.

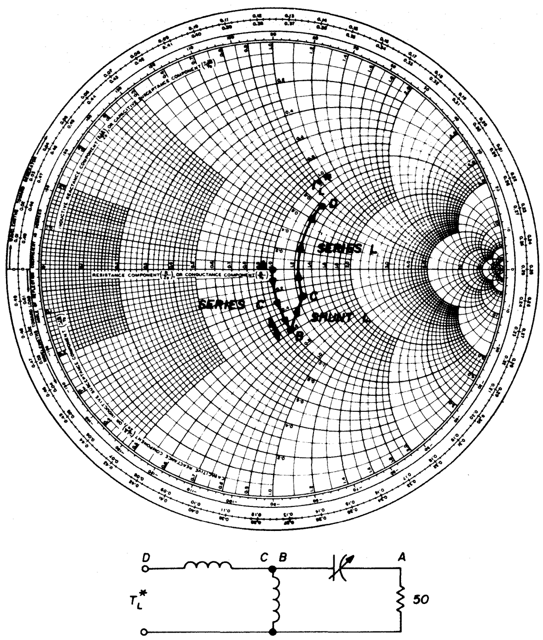


fig. 4. Design of the output matching network with a Smith chart. Component values are optimized with a computer program.

input matching network

At the input of the preamplifier it is necessary to transform the 50-ohm input impedance to the source reflection coefficient required for minimum noise figure (Γ_{os}). This can be accomplished with a T network consisting of a series capacitance, shunt inductance, and series inductance. Beginning at the 50-ohm point at the center of the Smith chart (which has been normalized to 1.0) at point **A,** the series capacitance moves the impedance to $1.0-j1.0$ (point **B**); the shunt inductance rotates this value to 1.5–j0.9 (point **C**); the series inductance transforms this to the source reflection coefficient Γ_{os} at $1.5+j0.66$.

The required reactance values for each of the components in the matching network can be read directly away from the Smith chart. Note that the series capacitance rotates the input impedance from $1.0\pm j0$ at the center of the chart to $1.0-j1.0$ at point B. Therefore, the required capacitive reactance is $-j1.0(50)$ or –j50 ohms; at 435 MHz this is represented by 7.3 pF.

To determine the reactance of the shunt inductor it's necessary to first convert to admittance (lower case letters designate normalized values)

Point B $z=1.0-j1.0$ $y=0.5+j0.5$
Point C $z=1.5-j0.9$ $y=0.5+j0.3$

The desired transformation requires a normalized susceptance of –j0.2; in a 50-ohm system this represents –0.004 Siemens (0.004 mho) or +250 ohms. In this circuit this is provided by 37 nH in parallel with 2.2 pF (a 5 pF variable allows adjustment within the limits indicated by the arrows at point **C** on the Smith chart plot).

The series inductance transforms the impedance of 1.5–j0.9 at point C to $1.5+j0.66$ at Γ_{os}. This requires a normalized reactance of $+j1.56$ or 78 ohms (28.5 nH at 435 MHz).

output matching network

The design procedure for the output matching network is similar to that used for the input network. Working from the 50-ohm load back to the collector of the transistor, the series capacitor transforms the load at **A** ($1.0\pm j0$) to point B (1.0–jO.58); the shunt inductor rotates the impedance to 1.23–j0.35 at point **C.** The series inductor then provides the desired reflection coefficient for the load ($\Gamma_L = 0.35\angle 54^{\circ}$) at point **D** ($1.23+j0.80$).

For the desired transformation the series capacitor must present a reactance of 0.58 x 50 or 29 ohms; at 435 MHz this is provided **by** 12.6 pF. To calculate the value of the shunt inductor, the impedance points are converted to admittance:

Point B $z=1.0-j0.58$ $y=0.75+j0.43$
Point C $z=1.23-j0.35$ $y=0.75+j0.21$

To move from $j0.43$ to j0.21 requires a *negative* susceptance of $-j0.22$ or 4.5 milliSiemens. This is equivalent to 223 ohms of inductive reactance of 82 nH at 435 MHz.

The series inductor required to transform 1.23–j0.35 at point **C** to $1.23+j0.80$ at Γ_L has an inductive reactance of 1.15 x 50 or 57.5 ohms (21 nH at 435 MHz). A preliminary schematic of the amplifier is shown in **fig. 5.**

Note that it is necessary only to determine approximate component values for the matching networks because, in this case, a computer program will be

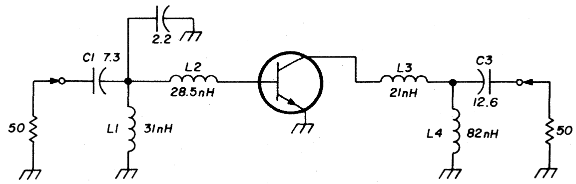


fig. 5. Basic 435-MHz low-noise preamplifier circuit with component values determined with the aid of a Smith chart (figs. 3 and 4).

used to adjust the values for optimum performance. However, values which are close to optimum will result in the usage of less computer time and, hence, lower cost.

Rather than winding inductors which could be lossy and cause stray coupling from unwanted radiation, it is better to use lengths of etched transmission lines for the inductive elements. This can be done providing the line lengths are less than $\lambda/8$ and preferably less than $\lambda/16$. This is more easily seen if you examine the input impedance of a lossless short-circuited transmission line:

$$Z_{in} = +jZ_o \tan \frac{2\pi l}{\lambda}$$

$$= +jZ_o \tan \Theta$$

Where:

Z_{in} = input impedance to the transmission line
Z_o = characteristic impedance of the line
l = length of the transmission line
λ = wavelength
8 = electrical length of the line in degrees

Note that this expression represents a pure reactance which varies almost linearly with the electrical length Θ, provided that Θ is small. Therefore, by varying the characteristic impedance Z_o and the electrical length Θ it's possible to synthesize inductive elements which are very accurate and highly repeatable when printed-circuit techniques are employed.

computer optimization

The next step in the design of the low-noise preamplifier is to select an appropriate computer program to execute the calculations and optimize the component values. The COMPACT* Computer Program is used extensively for this purpose because it has broad capability in terms of network elements and interconnections and is modest in cost when used within certain guidelines.

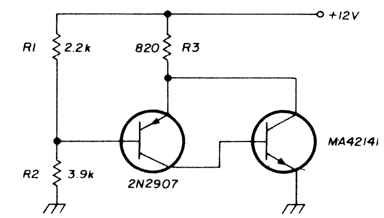


fig. 7. Active bias circuit is used in the low-noise preamplifier to allow direct grounding of the emitter lead. The collector-to-emitter voltage of the MA42141 is determined by voltage divider resistors R1 and R2 (see text).

The information for the computer is written in the form of a data file. Once the data file is written, the computer will vary the network elements and attempt to minimize the error between the desired circuit performance and the actual circuit performance. Specific performance parameters may be weighted so that their attainment carries more importance than other performance parameters. For example, if noise figure is the most important design goal, input impedance match or gain may be sacrificed so that the lowest noise figure may be achieved; the computer will adjust the variable elements in a direction which minimizes noise figure but not necessarily maximizing gain or lowering the input impedance mismatch.

In this case the computer analyzed the circuit and optimized it for operation at one frequency, 435 MHz. Additional or broader optimization could be performed by altering the data file, but this would increase computer time (and cost) because of the larger number of variables. After optimizing the component values, the computer predicted the following preamplifier performance at 435 MHz:

Noise figure	1.81 dB
Power gain	16 dB
Output vswr	1.22:1

The noise figure might be somewhat optimistic since no allowance has been made for circuit losses associated with the variable capacitors and high input vswr.

A complete schematic of the optimized preamplifier circuit is shown in **fig. 6.** The synthesized induc-

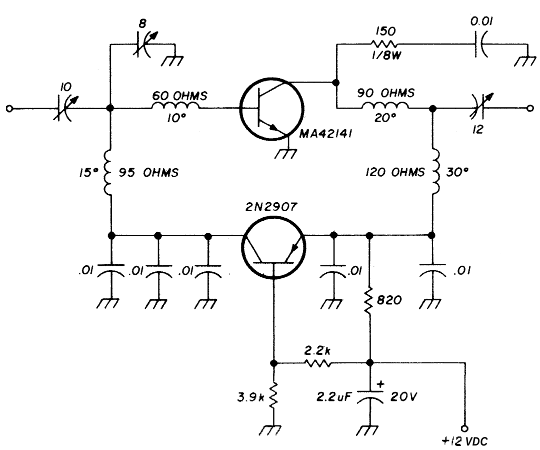


fig. 6. Low-noise 435-MHz preamplifier circuit with computer optimized circuit values. All inductive elements use etched transmission lines of the specified characteristic impedance and electrical length. Fixed values of capacitance are chip capacitors. The 2N2907 is part of the active bias circuit (see fig. 7).

*COMPACT is an acronym for Computer Optimization of Microwave Passive and ACTive circuits. Additional information is available by writing to Compact Engineering, Inc., 1651 Jolly Court, Los Altos, California, 94022.[5,6]

table 1. Comparison of computer predicted performance with measured performance.

	computer predicted	measured performance unit 1	unit 2
Noise figure	1.8 dB	1.9 dB	1.9 dB
Gain	16.1 dB	16.0 dB	16.5 dB
Output vswr	1.22:1	1.15:1	1.12:1

tors are specified in terms of their characteristic impedance and electrical length. **Table 1** shows a comparison between computer predicted performance and the measured performance of two preamplifiers which were built in the lab.

stability considerations

A broadband computer stability analysis reveals that the preamplifier is unconditionally stable over a frequency band from 400 MHz to 2800 MHz. The importance of the stability analysis cannot be over emphasized because, in many applications, the source and load impedances may take on any value outside the particular frequency band of interest. When a high-gain microwave transistor is used, it is most important to assure that the amplifier does not oscillate as a result of various out-of-band source and load impedances

dc bias circuit

To realize the predicted performance when using the manufacturer's transistor data, the designer must mount the transistor in a manner which closely approximates the electrical and mechanical environment under which the manufacturer obtained the data. The introduction of parasitic elements in mounting the transistor can lead to large discrepancies between the predicted and measured performance.

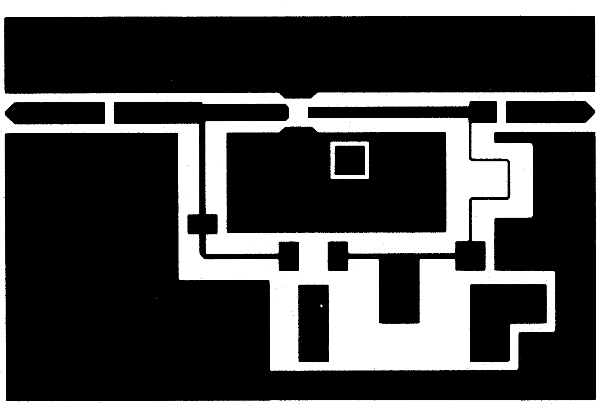


fig. 8. Full-size printed-circuit layout for the low-noise 435-MHz preamplifier. Circuit is etched on 1/16" (1.5 mm) double-clad G-10 fiberglass-epoxy circuit board. Component layout is shown in fig. 9.

The parasitic element which is most sensitive to performance degradation is the emitter lead inductance, which, if not kept to a minimum, will both reduce gain and alter the source impedance required for minimum noise figure. It may also introduce instabilities within the circuit which, under certain conditions, could result in oscillations. For these reasons an active bias circuit has been utilized in this amplifier. The active bias circuit will provide the proper collector-to-emitter voltage and collector current, while allowing direct grounding of the emitter lead to minimize the introduction of any parasitic impedance into the circuit. A schematic of the active bias circuit is shown in fig. **7.**

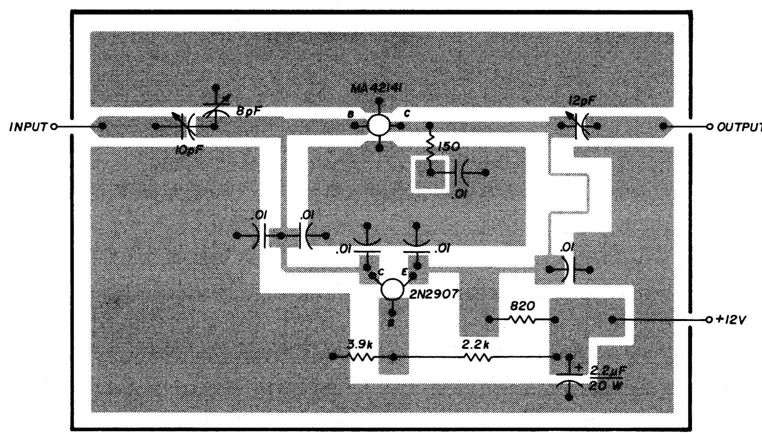


fig. 9. Component layout for the 435-MHz low-noise preamp. For best performance all fixed values of capacitance should be chip capacitors.

The active bias circuit is actually a feedback loop which senses the collector current of the rf transistor and adjusts the base current to hold that collector current fixed. The collector-to-emitter voltage of the rf transistor is held at a fixed potential determined by the voltage divider R1 and R2. The current through resistor R3 becomes the collector current of the rf transistor under the assumption (a good one) that both the 2N2907 and the MA42141 have moderate dc current gain.

references

1. *S-Parameters — Circuit Analysis and Design, Application Note* 95, *Hewlett-Packard,* Palo *Alto, California,* 1968.
2. *S-Parameter Design, Application Note* 154, *Hewlett-Packard,* Palo *Alto, California,* 1972.
3. Philip *H. Smith, Electronic Applications of the Smith Chart in Waveguide, Circuit, and Component Analysis,* McGraw-Hill, *New York,* 1969.
4. *James* R. *Fisk,* W1HR, *"How to Use the Smith Chart," ham radio, March,* 1978, page 92.
5. *COMPACT Version 4.5, Compact Engineering, Los Altos,* California.
6. *Computer Aided Microwave Circuit Design with COMPACT, Application Note* 1, *Compact Engineering, Los Altos, California.*
7. *Transistor Designer's Guide, Microwave Associates, Burlington, Massachusetts.* 1978.

ham radio

single IC
Touch-Tone decoder

A new integrated circuit and high performance active filter are combined into an extremely reliable *Touch-*Tone decoder

A majority of repeater stations today are using *Touch-Tone** circuitry for such functions as autopatch, control, and signaling; individual home and mobile stations make use of encoders and decoders for selective calling and radio control. Dozens of designs have been published for *Touch-Tone* decoders during the last several years. Unfortunately, virtually none of these circuits is reliable.

The use of the 567 phase-lock loop IC, upon which nearly all designs are based, is not recommended unless limited performance is acceptable. Elaborate support circuitry is required, and even then you may not achieve the desired noise immunity or stability. The tradeoffs are many: response time *vs* bandwidth, noise immunity *vs* bandwidth, and economy *vs* reliability.

The 567 PLL chip has several important limitations which must be dealt with regardless of whether you are using the 567, or are a circuit designer developing a totally new decoder IC.

signal consideration

Amplitude Variations. Twisted-pair telephone wires are transmission lines, just as coaxial cables are. Shunt capacitance and series inductance serve to increase attenuation as frequency goes up. This disparity means that high group *Touch-Tone* signals are usually attenuated more than the low group. This variation in amplitude is referred to as "twist." Tone pads attempt to compensate for this by having a stronger output from the high group than the low

Touch-Tone is the registered trademark of the American Telephone and Telegraph Company

By **Larry Nickel, W3QG, 216** Highmeadow Road, Reisterstown, Maryland 21136

group. In addition, when decoding tones from an fm receiver, signal strength and deviation are two more variables. An effective decoding scheme must incorporate an audio ALC system. If the decoder circuitry is sensitive to twist, then separate ALC control could be necessary for the high and low groups.

Bounce And Noise. A *Touch-Tone* source may not present a clean, stable leading edge. Secondly, noise, music, or speech may momentarily produce coincident tones at the proper frequencies, which could be recognized as a legitimate *Touch-Tone* pair. What is required is a delay, of perhaps 40 milliseconds, before a valid tone pair is acknowledged. Then the decoder output must turn on and stay on until the signal disappears, with no bounce. It has been my experience that achieving good noise rejection, fast response without false tripping, and freedom from bounce is not practical with PLL techniques.

Frequency Stability. Tone generation schemes which depend on RC time constants for frequency stability are not reliable, especially at temperature extremes. This is why the Motorola MC14410 and Mostek MK5087 crystal-controlled encoder chips were developed. The same is true of decoding circuitry. Crystal control is a must!

new generation integrated circuits

Within the last two years, several manufacturers have introduced some very sophisticated tone-decoder ICs. The first entry was Rockwell International's Collins CRC8030 dual tone multi-frequency (DTMF — another name for tone control) receiver. It is an MOS decoder chip in a 28-pin dual-in-line package which uses a standard 3.579545 MHz TV color-burst crystal for its reference. This ceramic IC was originally priced at $49 in unit quantities, but a plastic case (CRC8030-3-3) now sells for $42. An ALC system, a filter (which separates high and low group tones), and two voltage comparators are required to complete the decoder. General Instrument also produces a decoder selection designated the AY59800 series. One or more of these chips use a 1 MHz crystal and require dual power supplies.

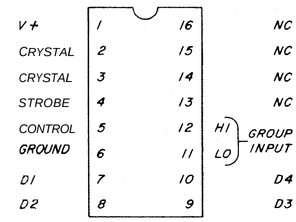
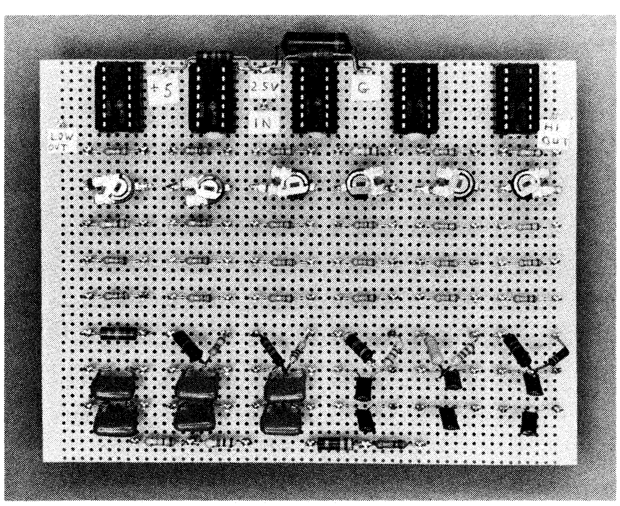


fig. 1. Pinout diagram for the Mostek 5102 Touch-Tone decoder.

table 1. Data output from the MK5102 decoder.

	4-bit binary				dual 2-bit row column			
digit	**D1**	**D2**	**D3**	**D4**	**D1**	**D2**	**D3**	**D4**
1	0	0	0	1	0	1	0	1
2	0	0	1	0	0	1	1	0
3	0	0	1	1	0	1	1	1
4	0	1	0	0	1	0	0	1
5	0	1	0	1	1	0	1	0
6	0	1	1	0	1	0	1	1
7	0	1	1	1	1	1	0	1
8	1	0	0	0	1	1	1	0
9	1	0	0	1	1	1	1	1
0	1	0	1	0	0	0	1	0
	1	0	1	1	0	0	0	1
#	1	1	0	0	0	0	1	1
A	1	1	0	1	0	1	0	0
B	1	1	1	0	1	0	0	0
C	1	1	1	1	1	1	0	0
D	0	0	0	0	0	0	0	0



Layout of the tone separation filter.

The newest entry into the tone-decoder field is the Mostek MK5102. It's a CMOS chip in a 16-pin package, with a typical power dissipation of only 25 mW, at 5 V dc. Available in either a ceramic (MK5102P-5) or plastic case (MK5102N-5), unit quantity prices start at $34.50." Even though this is a large sum of money to pay for an IC, at this time it is the *only* way to build a complete, top-quality decoder without spending more than $60.

A pinout diagram for the MK5102 is shown in **fig. 1.** Compared to the 28-pin CR8030, this is sheer simplicity. The 5 V dc and ground connections, pins 1 and 6 respectively, are self-explanatory. An inexpensive 3.579545-MHz TV-color burst crystal is connected between pins 2 and 3. Pin 5 is used to control the output format of pins 7 through 10 (D1 through D4). This tri-state input line selects a four-bit binary code

"Quality Components, 13628 Neutron Road, Dallas, Texas 75240.

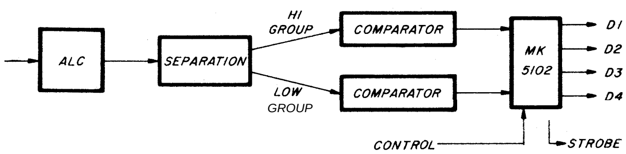


fig. 2. Block diagram of a complete tone decoding system using the new single IC tone decoder.

(input high), a dual two-bit row/column code (input floating), or high-impedance output (input low) for use with bus-structured circuitry.

Pins 7 through 10 are the data out lines. The outputs are CMOS loads when enabled, and open circuited (high impedance) when disabled by the control pin. The output data formats are shown in **table 1.** The two output codes allow the user to obtain either 1-of-16 or 2-of-8 output data by only using a single additional package.

Pin 4, the strobe output, goes high after 40 mS of a valid tone pair, and remains high for a minimum of 10 mS after the input ceases. The output information is valid when the strobe signal goes high and will remain unchanged until the next DTMF digit is detected.

The low- and high-group tones are filtered, separated, and applied to pins 11 and 12, respectively. The MK5102 can detect capacitively-coupled, square-wave signals as small as 1.2 volts pk-pk. The tones are detected, after band splitting, using the digital-counter method. The zero crossings of the incoming tones are counted over a longer period. When a minimum of 40 milliseconds of a valid signal is detected, the proper data is latched into the outputs and the output strobe goes high. When a valid digit is no longer detected, the strobe will return low and the data will remain latched into the outputs. The minimum interdigit time is 35 milliseconds.

A block diagram of a complete *Touch-Tone* decoding system using the MK5102 is shown in **fig. 2.** The ALC reduces any amplitude variations from the signal source.

The low-group tones, the rows on your *Touch-Tone* pad, and the high group tones, the columns, are separated in the tone-separation filter. Its outputs are two sine waves which are squared up in comparators and applied to the MK5102.

active filter

Design Considerations. An active filter for a 1-kHz frequency range is generally of low cost, small, has gain, high-input impedance, low-output impedance, and is easy to design. My intention here is not to make the reader a filter expert but to give a little of the philosophy behind the design of this one. My criteria were:

1. It must be inexpensive.

2. It must use readily available components

3. It must be easily constructed.

4. It must provide adequate out-of-band rejection.

For the bandwidth and Q required, a stagger-tuned circuit is necessary. After examining sample response curves for 2-section filters, I felt that at least 3-sections would be required to achieve adequate out-of-band rejection. One section is tuned to the center frequency, another is tuned to s (the "staggering value") times the center frequency, and

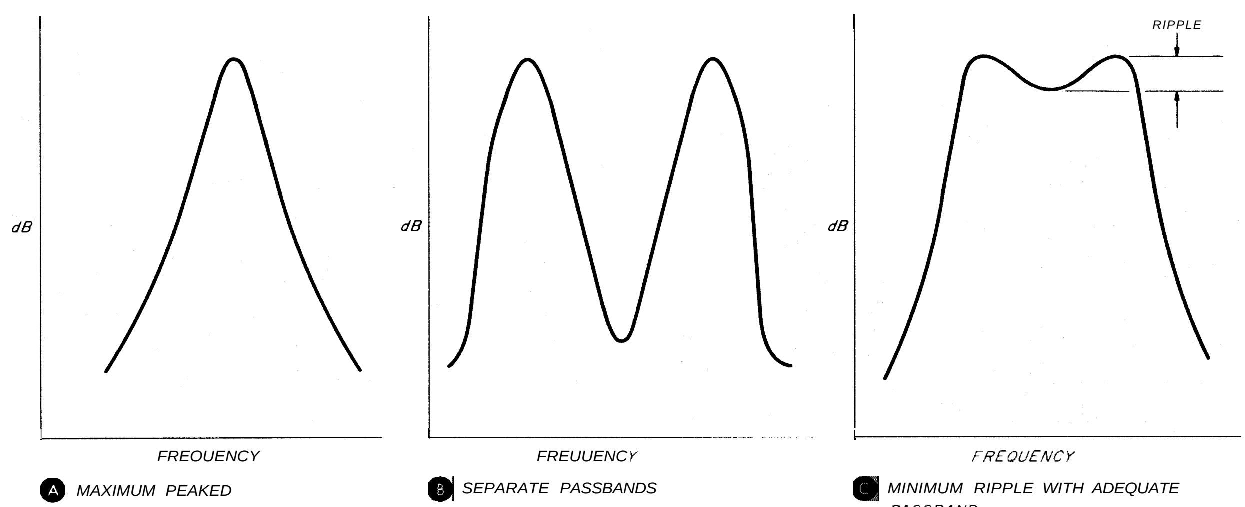


fig. 3. Passband response as a result of varying s in equation 1. A shows a single-peaked response when $s = 1$. For large values of s, B shows the dual-peaked response. In C, a response is obtained that has a minimum passband ripple, yet adequate bandwidth ($s = 1.16$).

the third $1/s$ times the center frequency. For $s=1$, the response has one peak and a very-narrow bandwidth **(fig. 3A)**. For a large value of s, 1.5 for instance, there are two separate distinct passbands **(fig. 3B)**. By selecting the best value of s for two- or three-section filters, a suitable bandwidth with minimum ripple can be achieved, as shown in **fig.** 3C. Each of the three sections is designed for a particular Q. Usually, it is best for the center section to have a lower Qthan the outside sections which have identical higher Qs. This makes for a flatter passband and a steeper slope beyond the passband (if the Q is too high the ripple in the passband will be excessive).

The transfer function for a three-section filter is:

$$\frac{E_{out}}{E_{in}} = 20 \log_{10} (A \bullet B \bullet C) \qquad \textbf{eq. 1}$$

where

$$A = \sqrt{1 + Q^2 \left(\frac{f^2 s^2 - 1}{fs}\right)^2}$$

$$B = \sqrt{1 + Q^2 \left(\frac{f^2/s^2 - 1}{f/s}\right)^2}$$

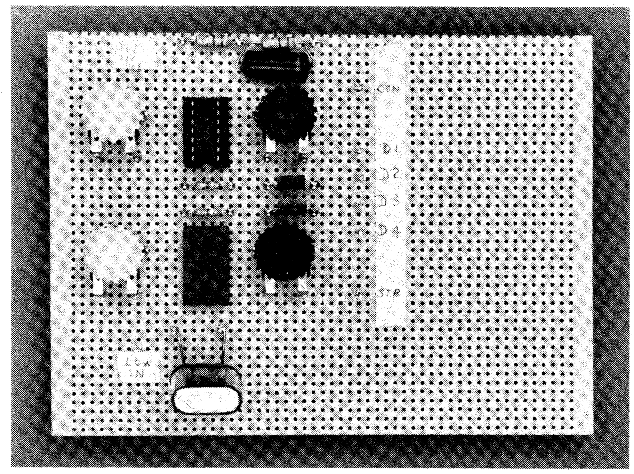
$$C = \sqrt{1 + \left(\frac{Q}{2}\right)^2 \left(\frac{f^2 - 1}{f}\right)^2}$$

s = staggerzng value
f = normalized center frequency, with two outszde sectzons of Q selectivity and one center sectzon of Q/2 selectivity

Various values of s and Q were calculated and plotted. Using $s=1.16$ and $Q=10$ and 20 for the center and outside sections, produced a passband ripple of 3 dB and 32 dB of rejection for the other tone group. With the s and Q values determined, it only remained to select a suitable circuit.

Many designs, using 1, 2, 3, and 4 amplifiers per section, have been published, but the disadvantage of the 1- and 2-op amp versions is that they are not generally suitable for high Q applications, especially with inexpensive ICs which have low gain-bandwidth products. Since low-cost quad op amps are available, and in many pin compatible packages, a 3- or 4-op amp design is preferable; any circuit configuration would be acceptable, especially if one per cent tolerance resistor and capacitors are used, but I did not wish to use precision components, and therefore, expended the additional labor to ensure that standard parts can be used.

Filter Selection. The BIQUAD filter section (see **fig. 4**) was chosen because the center frequency and Q can easily be adjusted by trimming just one resistor for each function. The principle of operation of the



The partially completed decoder board.

BIQUAD is as follows. Since the integral of a sine wave is a cosine wave, or a 90-degree phase shift, U1 is an integrator, giving a 90-degree shift. U2 is an inverter yielding 180-degree shift, for a subtotal of 270 degrees. U3 is another integrator giving an additional 90 degrees, for a grand total of 360 degrees. Positive feedback may then be provided from input to output.

Without the Q setting resistor, the gain and Q of the BIQUAD would be excessively high, causing the circuit to oscillate at the frequency where the phase shift is 360 degrees (the integrator and inverter do not have exactly 90 degrees and 180 degrees shift respectively at more than one frequency). This resistor introduces enough loss so that the Q is controlled and the BIQUAD does not oscillate. The actual frequency response curve for the complete filter is shown in fig. 5 and its schematic is presented in **fig. 6.**

Filter Operation. The tonal input is simultaneously applied to both sides of the filter, passing through the 686 Hz, 809 Hz, and 955 Hz sections of the low-group filter as well as the 1191 Hz, 1404 Hz, and 1657 Hz sections of the high-group filter. Resistive dividers R19/R20 and R41/R42 reduce the outputs of the 686 Hz and 1191 Hz sections so that these sections may be driven farther toward cutoff and saturation without overloading the following stages. Since a single 12-volt supply is used, R39 and R40 divide 12 volts down to 6 volts, establishing the dc bias so that the output may swing equally about 6 V dc. C13 ensures that the 6 V dc bias line is at ac ground.

filter construction

My filter was built using wire-wrap techniques. I used standard, wire-wrap IC sockets with a phenolic

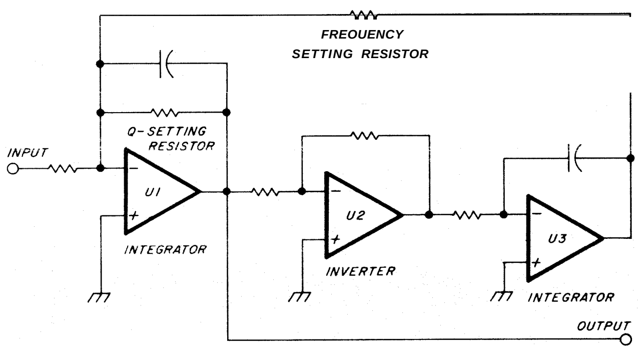


fig. 4. Diagram of the basic BIQUAD active filter. The two integrators produce a 90-degree phase shift and the inverter provides 180 degrees of shift. With proper feedback, this filter will pass a single frequency.

board and Vector wire-wrap terminals to mount the resistors and capacitors. This method has one very attractive advantage (though it is slightly more expensive), it's very fast, especially compared to point-to-point soldered connections.

Drop the Vector pins into place, seating them by pulling from the bottom of the board. Epoxy the sockets into position. The board can be wired using an inexpensive, hand wire-wrap tool. The use of precut and stripped no. AWG 30 (0.25mm) wire further speeds assembly. Use an ohmmeter to check *all* wiring for errors. This is extremely important since it will prevent damaging components and could save considerable time later.

Temporarily install the final 12 resistors using the nominal values shown in the schematic. Miniature potentiometers set to these values are highly recommended. Only six are needed since the low and high group filters can be tuned separately. These pots can be removed later and precision resistors installed.

Make your own precision resistors by wiring two or more resistors in series, or by using the W3QG trimming technique. For instance, to make a precision 11.5k-ohm resistor, file a V notch in a 112-watt, carbon, 10k-ohm resistor until the exact value is attained. A small grinding wheel on a Dremel tool is even better; use a light-stroking motion. And finally, seal the exposed carbon with a dab of epoxy.

tuning and alignment

For checkout, you will need a sine-wave audio source and an oscilloscope. It would be helpful if the sine-wave source does not change in amplitude as it is tuned from one frequency to another, but if necessary its output can be readjusted with the scope.

You should now install the ICs and apply power. Set the generator for approximately 50 mV peak-to-peak. Beware, a larger value may reduce the Q and/or drive the op amps into nonlinearity. Be sure the generator does not cause a dc-bias problem with the filter; you may want to include a 0.01 μF capacitor in series with the filter input.

Tune the generator from approximately 600 Hz to 1200 Hz with the scope connected to the low group output. Notice where the three filter sections are peaking. Don't expect the peaks to be the same amplitude since the Q has not been tuned yet. If you have chosen to use pots, tune the center frequency of each section to the correct value, using R6, R12, and R18.

If you are using individual resistors, divide the actual center frequency of each section by the desired frequency for that section. Square this fraction and multiply it times the existing resistor value. This will give you the approximate value for a resistor which will put you very close to the desired frequency. Set the center frequency for the three high group sections in a similar manner by tuning the generator from approximately 1000 Hz to 2000 Hz.

To tune the Q, each BIQUAD could be driven and monitored separately and adjusted to the desired Q. An easier method is as follows. The Q of the first low group section is 20; the frequency is 686 Hz; the bandwidth is 686120 or 34 Hz. Hook the oscilloscope to U3 pin 7. Adjust the generator to the center frequency and note the amplitude. Find the two frequencies where the output is 3 dB (0.707 times) less than at the center. Is the bandwidth more or less than 34 Hz? Adjust R2 upward to increase Q or down to decrease Q. Once this value equals 20, connect the scope to the output of the low group filter.

Set the generator for the center frequency of the 686 Hz section. If, for instance, the output is 500 mV peak-to-peak, tune the generator to the center of the 809-Hz section and adjust R8 for 500 mV, then tune the generator to the center of the 955-Hz section and adjust R14 for 500 mV.

The high-group filter may now be tuned in a similar manner at its respective frequencies, first adjusting the 1191-Hz section, via R22 for a Q of 20 and a 60-Hz bandwidth with your scope on U3 pin 8. Then,

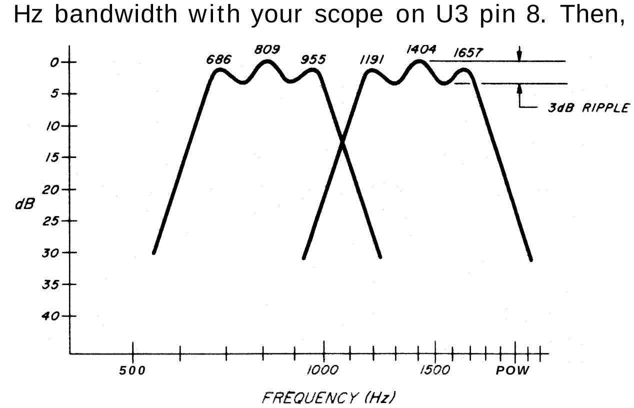


fig. 5. Filter response for the high and low tone separation filter.

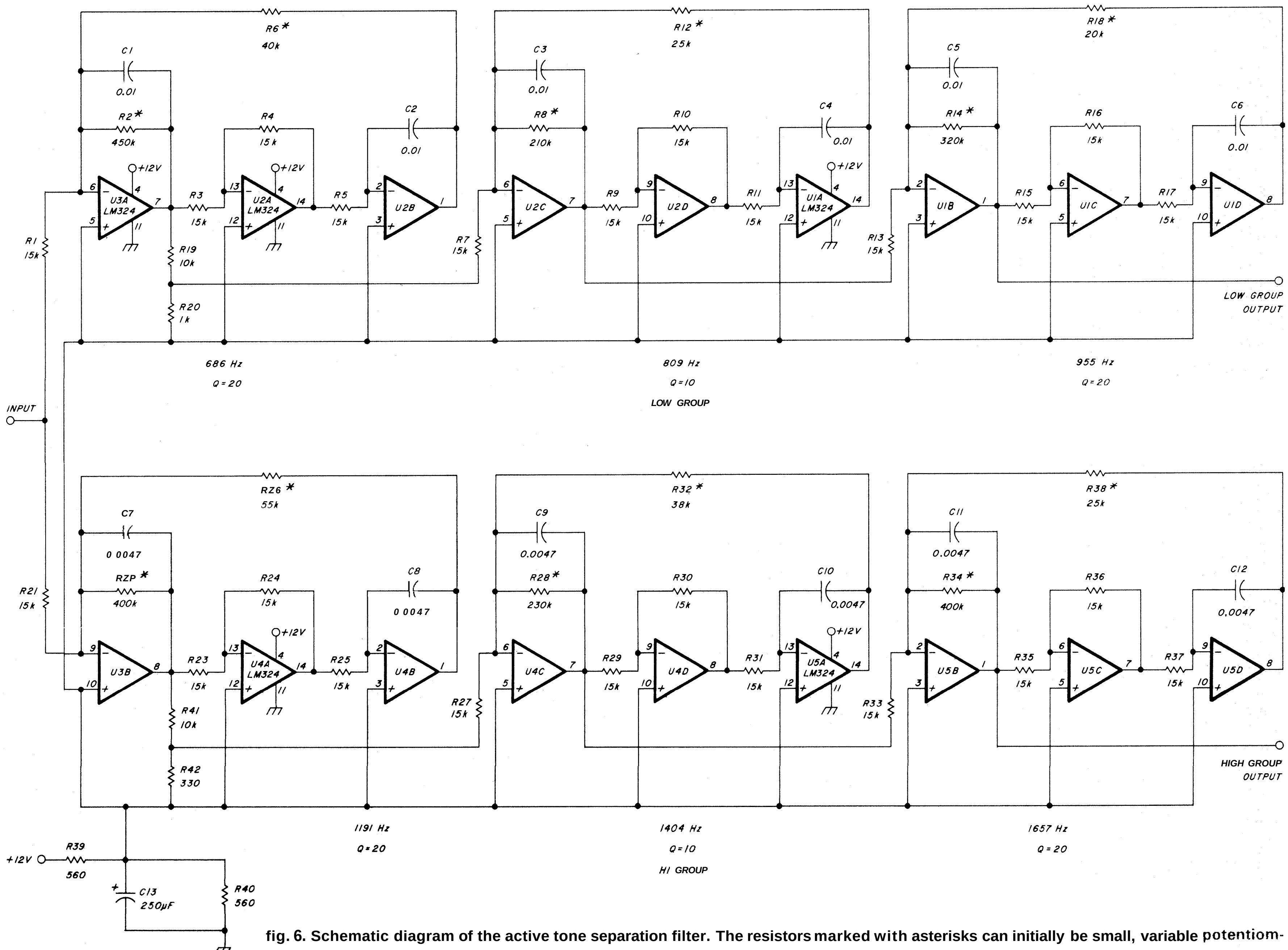


fig. 6. Schematic diagram of the active tone separation filter. The resistors marked with asterisks can initially be small, variable potentiometers set to these values. Their final value, however, will be determined during alignment. All capacitors must be mylar or polystyrene

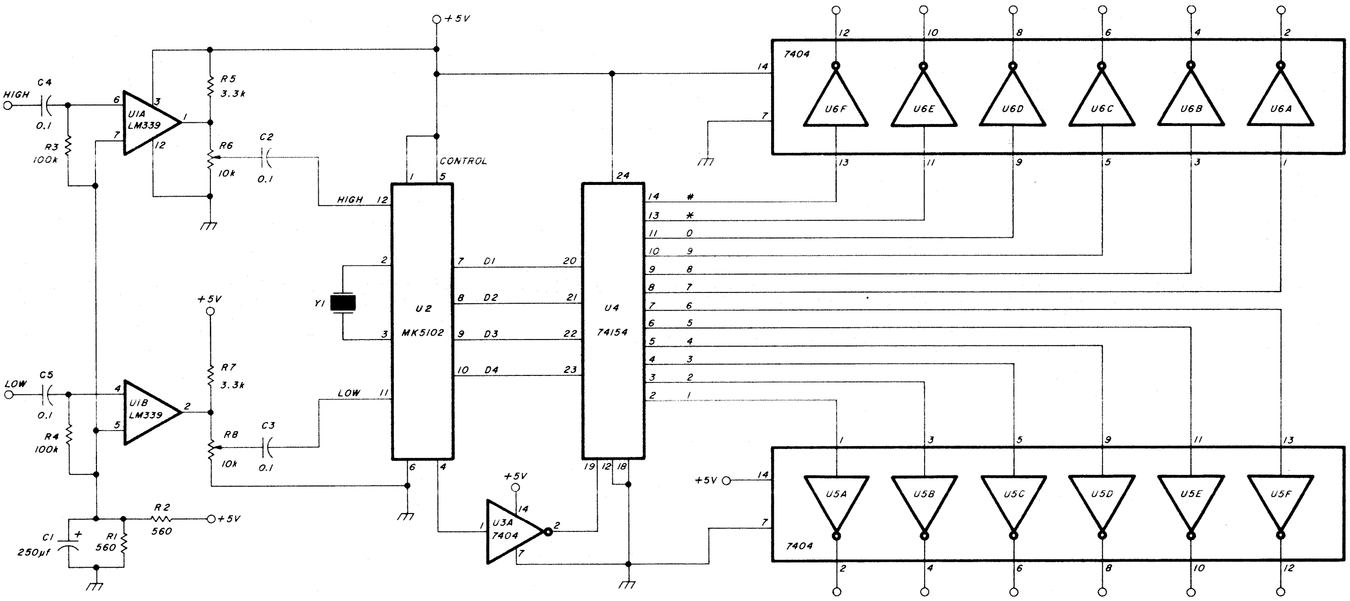


fig. 7. Schematic diagram of the decoder section. The data from the 74154 is active low, and can be inverted (by U5 and U6), depending upon the needs of your system.

with your scope on pin 1 of U5, adjust R28 and R34 at 1404 Hz and 1657 Hz respectively as you did for the low-group sections. Finally, check to ensure that 1209 Hz is really 30 dB down from 941 Hz in the low-group filter, and vice versa.

A voltmeter may be substituted for an oscilloscope for every filter test, but it will not allow you to see nonlinearities, oscillations, and hum. Also, a *Touch-Tone* pad can be used as a frequency standard to calibrate the generator by setting up a Lissajous pattern. On most pads, pressing two row or column buttons will produce only the sine wave for that row or column.

If your filter oscillates, the problem may be:

1. The Q of one or more stages is too high.
2. The power supply is not adequately bypassed.
3. The ground circuit is not adequate.
4. Circuit layout causes unwanted feedback.

The filters I have constructed have not had these problems. Should you decide to use a substitute op amp be aware that if it has poor power supply rejection or insufficient phase margin, it could cause oscillation.

comparator and decoder circuitry

Fig. 7 is the schematic for the comparator and decoder. The two sections of the comparator U1 are used to square up the filter outputs. The resistive dividers on the output of the comparators allow adjustment of the drive to the MK5102, although this does not seem critical; you may choose to replace the pots with fixed dividers.

In this circuit, the decoder's data control pin is tied to 5 V dc to constantly enable the proper format. A 74154 separates the twelve subsequent commands. The 74154 outputs are active low, so if you require active high signals include hex inverters U5 and U6. If the digit 1 is received, then pin 2 of the 74154 will stay low just as long as the 1 is being received.

The filter, decoder, and ALC circuitry has been in operation for many hours. It's presently connected to the output of a 2-meter fm receiver tuned to a noisy, busy, repeater channel, and is used every day for selective call and remote control. The circuit does not trigger on false signals. In fact, it has proven to be stable and reliable, sufficiently so that it has permanently replaced an earlier sophisticated 567 PLL decoder system which was in use here for over a year and a half.

Incidentally, there is a minor limitation of which you should be aware. Because of the ALC used with this system, a signal with somewhat low deviation will not be a problem. However, a signal with excessive deviation will be wider than your receiver's passband and will not be decoded. This is not a fault of the decoder, but must be corrected at the transmitter.

Soon the MK5102, with its tri-state outputs, will be connected to a MOS-Technology 6502 microprocessor system so that more complex functions may be performed. I welcome and solicit any and all comments and improvements to this *Touch-Tone* decoder.

ham radio

antenna guys

and structural solutions

Advice on choosing guying materials and installing them for safety and long life

Amateurs with free-standing antenna towers are fortunate indeed: no need to worry about guy wires, anchors, and supports. But if your antenna tower must be guyed, how do you ensure that your guying system will withstand the forces of high winds? What about your soil conditions? Can your soil hold guy anchors?

This article gives some guidelines on how to handle the problem of guying antenna towers erected on various types of soil (from hard rock through loose sand and gravel). Also included are some tips on guying materials and how to use them. The author has had some 30 years of experience in the engineering and construction of antenna systems in locations where wind velocity often exceeDS 185 km/hr (100 knots).

guy-anchor placement

Starting at ground level and going up, the location of each guy anchor is our first consideration. Ideally, guy anchors should be placed the same distance from the tower or mast base as the guy attachment to the structure, a 1:1 ratio. A minimum of three guys, spaced 120 degrees apart are considered while four guys spaced 90 degrees apart is most desirable. When real estate is not available to maintain the desired 1:1 ratio, consideration should be given to an acceptable ratio of 6:4.

anchors

Table 1 shows how we may expect our anchors to accept the strain when installed in various soils. Considering the load that an average guy will put on the

table 1. Classification of soils. The soil classification is used for determining the type of holding anchor guy wires (Courtesy Graybar Electric).

soil class	soil description
1	hard rock, solid
2	shale or sandstone, solid or layered
3	hard, dry, hardpan
4	crumbly, damp
5	firm, moist
6	plastic, wet
7	loose dry sand, gravel

anchor and rod, the stamped 152-mm (6-inch) anchor with a 13 mm (1/2 inch) by 1.5-meter (5 foot) rod will perform satisfactorily in most locations **(fig. 1)**.

In my case anchor holes were drilled with a power-driven post-hole digger leased from a local equipment rental agency. Angle the anchor hole so that guy tension is in a straight line with the anchor rod. The holding ability of this type of construction is shown in **table 2.** Before placing the anchor and rod, dress the rod threads to prevent the nut from backing off.

These units will survive for many years if the soil composition is not corrosive. A test with litmus

table 2. Holding power of a 152 rnm (6 inch) cone anchor in several soil types (courtesy Graybar Electric).

soil classification (table 1)	3	4	5
pounds (kg) ultimate	10,000(4540)	8,000(3632)	6,000(2724)

By Marchal H. Caldwell, Sr., W6RTK, 4620 Greenholme Drive, No. 4, Sacramento, California 95842

paper will reveal the presence of any contaminants. Coat the rod and anchor with thinned roofing mastic before back filling.

guy wire material

Consider the actual guy wire. There are probably more different guy wire types than vacuum tubes, and most will survive for a long time. Since an antenna and its supporting structure represent a considerable investment, the proper selection and use of guying materials is a must. **Table** 3 lists some of the available solid and stranded guy wire.

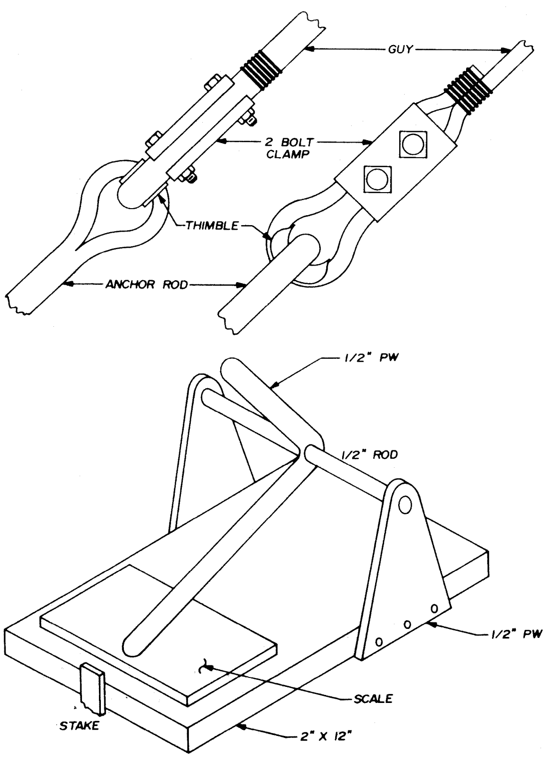


fig. 1. Mechanical details for securing anchor hardware on guy wires. Lower sketch shows a method for adjusting **guy**-wire tension. An ideal ratio of **1:1** is assumed for the distance of guy-wire attachment to structure (that is, guy locations should be the same distance from tower base as the guy attachment to the structure). A compromise ratio of 6:4 is acceptable.

attaching the guys

Guy-wire attachment to the anchor rod eye is usually by a clevis-and-eye turnbuckle, with the clevis end attached to the rod and the guy strand attached to the eye by wire rope clamps. At least ***two*** clamps must be used spaced six times the diameter of the wire. The wire dead-end should be served around the wire adjacent to the outer clamp. Turnbuckles should be safety-wired. A thimble should be used under the wire and inside the eye **(fig. 2**).

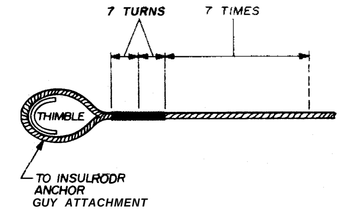


fig. 2. Method for securing terminal hardware to a guy. Note how guy wire should be served (7 turns, 7 times).

Towers and masts are sometimes provided with a guy attachment bracket. Terminate the guy wire in the same manner as used at the turnbuckle **(fig.** 3). When no guy bracket is provided, wrap the guy around the tower leg above a cross brace.

Proper guy tension has always presented a problem. The only sure and safe method is to use a strand dynamometer, which is expensive and normally not available. However, most amateurs use the eyeballing method – keeping too much tension out and not too much sag.

strain insulators

Some antenna structures will require the guy wires to be broken with strain insulators while others won't need this treatment. Fig. **4** shows how to connect strain insulators in the guy wires. Note the recommended method of serving the loose ends of the wires, which is extremely important. At amateur frequencies, the most desirable maximum distance between guy-wire insulators is 3 meters (10 feet). Your

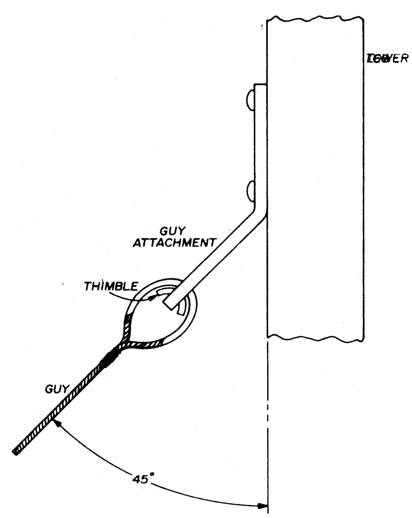


fig. 3. Connecting terminal hardware to a mast or tower, showing recommended geometry for guy placement. An angle of 45 degrees is ideal, but compromises are acceptable depending on your local wind conditions and available real estate.

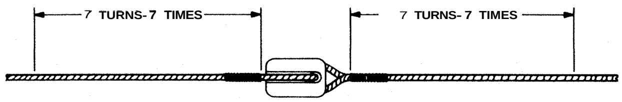


fig. 4. How to connect strain insulators to your guys. Serving the loose ends of the guys follows the rule: 7 turns, 7 times.

pocketbook will be the deciding factor since the insulators and guy wire clamps could cost about $5 at each point.

Reasonably corrosion-proof materials should be used in your antenna construction. Hot-galvanized materials are among the best. An excellent source of supplies is your local electrical contractor. Exploration of surplus and salvage agencies often produces items at a considerable saving — but again, don't compromise on quality!

Personal safety during construction is a most important consideration, which must be practiced and observed. Have an adequate supply of strong hands and backs available when erecting any type of supporting structure. Ground-crew members should be equipped with hard hats, safety shoes, work gloves, and a knowledge of your construction plans. Climbers should be similarily equipped and have a good safety-belt.

If you're considering constant experimental work at the top of your structure, fall-safe units are available, which will prevent a disastrous fall. (This stuff is expensive, but so is a hospital bed!)

le 3. Guy-wire strength in terms of size, breaking strength, and maxim load (courtesy Graybar Electric).

	size, mm(AWG)	breaking strength, kg (lb)	maximum load, kg (lb)
/anized telephone	2.6 (10)	293 (645)	136 (300)
elegraph	2.1 (12)	193 (425)	91 (200)
	1.6 (14)	112 (247)	52 (115)
	diameter, mm (in)	**breaking strength, kg (lb)**	**maximum load, kg (lb)**
en wires twisted	5.0 (3/16)	522 (1150)	250 (550)
1 strand (common	6.5 (1/4)	863 (1900)	409 (900)
de)	8.0 (5/16)	660 (3200)	681 (1500)
en wires twisted into	5.0 (3/16)	999 (2200)	477 (1050)
rand (utility grade)	6.5 (1/4)	1816 (4000)	908 (2000)
Western Union, AT&T	8.0 (5/16)	2724 (6000)	1362 (3000)

Finally, but by no means of less importance, remember that erecting a supporting structure near high-voltage power lines could cause you to miss out on the Quarter Century Wireless Club.

ham radio

vfo design

using characteristic curves

Graphical aid to help you choose vfo component values on the basis of working-frequency range and scale linearization

This article will help you design a vfo using what I call the "universal characteristics" of the Colpitts oscillator. The hard work has already been done with the aid of a computer. The result is a family of parametric curves that allow you to choose frequency determining values of critical components precisely and without guesswork.

background

Several months ago I started the project with an ssb transceiver using ICs, following the new and well-known concept of concentrating all common circuits that don't depend on reception-transmission frequency in a single printed circuit. An external vfo and a tuned amplifier in the input-output circuit defined the transceiver operating frequency. Several IC manufacturers have developed exclusive ICs for the common parts of the transceiver mentioned above by standardization and by reducing the number of discrete components.

the external vfo

The first problem I met was designing the external vfo. I needed a vfo in the 5 to 5.5-MHz frequency range with good stability and linearity. Because of the chosen frequency and the oscillator characteristics, I chose the Colpitts oscillator. I first tried to find articles that already described vfos working in my desired frequency range. I found a couple of circuits, and despite the different component values shown, I started construction following the scheme in fig. 1. First of all I used a capacitor, $C_v = 80\ pF$ with an inductance of approximately $L = 5\ \mu H$, using different values for C1 and C2. I immediately realized the dif-

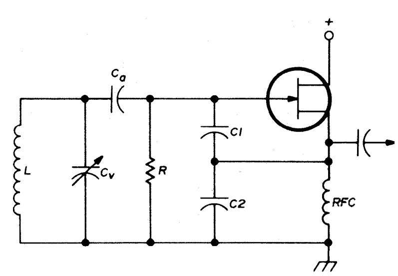


fig. 1. Typical vfo Colpitts oscillator circuit used in the text example. Components L, C_v, C_a define the oscillation frequency, for which parametric curves have been derived and are shown in fig. 3.

By Maurizio Gramigni, I2BVZ, 1621 16th Avenue NW, Rochester, Minnesota 55901

ficulties ahead; in fact, when turning C_V 180 degrees, the frequency range was around 1.6 MHz, which was not the desired range. After this failure I decided to face the problem from a technical point of view, starting from these assumptions:

1. Determine frequency stability *vs* temperature variation.

2. Calculate circuit-component values on the basis of the chosen working frequency range and scale linearization.*

Regarding point **1,** the use of an fet is highly recommended compared with bipolar transistors or vacuum tubes. The advantages are listed below:

1. An fet, with its high input impedance, can amplify signals with very low current level. As a result the power dissipation will be less than with bipolar transistors and vacuum tubes.

2. As a consequence of point **1** above, the heat dissipated and transferred to the oscillator components is lower, resulting in better thermal stability.

3. Since the power used is very low and the fet mass is very small and compact, the thermal equilibrium will be reached in a very short time — 30-40 seconds.

Using components such as silver mica capacitors helped to provide an oscillator with a very good frequency stability.

Calculating component values as a function of frequency range and scale linearization is the subject of this article. Fig. **1** shows our vfo circuit. Let's say, first of all, that L, C_v, C_a define oscillation frequency,

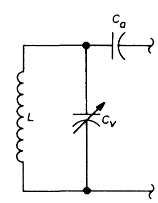


fig. 2. At the resonant frequency the oscillator impedance will equal zero. The network L, C_v, C_a formed the basis of the FORTRAN **program, which resulted in the universal characteristic curves for determining vfo values as shown in fig. 3.**

while $C1$ and $C2$ are the positive feedback network, which provides self starting and maintains oscillation. Distortion due to the nonlinear elements is limited by the proper choice of resistor R.

*Scale linearization means that any variation of C_v capacitance, ΔC_v, always corresponds to the same frequency variation, Δf, in the entire oscillator range; that is, the linearity ratio, $\Delta f / \Delta C$, remains constant in the chosen range.

Generally it's easy to design an oscillator working at one predetermined frequency. In fact, this occurs when the capacitive reactance of the oscillator circuit equals its inductive reactance. If a variable frequency oscillator is needed, however, a different approach to the problem is required. In this case, once the frequency range is chosen we need to know L, C_v, and C_a values.

universal characteristics

To reach our goal we start by calculating the transfer function of the complete oscillatory circuit. Then if we indicate

$$\omega = 2\pi f$$

$$C = \frac{C_1 \bullet C_2}{C_1 + C_2}$$

$$X_L = j\omega l$$

$$X_v = -j\frac{1}{\omega C_v}$$

$$X_a = -j\frac{1}{\omega C_a}$$

$$X_C = -j\frac{1}{\omega C}$$

the final equation is

$$\vec{Z} = \frac{\left(\frac{X_L \bullet X_v}{X_L + X_v} + X_a\right) \bullet \left(\frac{R \bullet X_C}{R + X_C}\right)}{\left(\frac{X_L \bullet X_v}{X_L + X_v} + X_a\right) + \frac{R \bullet X_C}{R + X_C}}$$

Because of the quantity of calculations I used a computer to solve for $\vec{Z}$.

As mentioned previously the real oscillator circuit consists of L, C_v, C_a components (see fig. **2**). At the resonant frequency, its impedance will equal zero, therefore our complex impedance, $\vec{Z}$, must also equal zero. The program, in FORTRAN language, was based on the observation above. In fact, the computer printout directly supplied C_v values for that particular frequency value where $\vec{Z}$ equals zero.

parametric curves

Starting from frequency values of 2.5 MHz, and for fixed values of L, C, C_a, and R, we obtained from the computer the C_v values for $\vec{Z} = 0$. The results are summarized in the curves of fig. 3.

Fig. **3A** shows the universal characteristics family for $L = 7.5 \mu H$. C_a values vary from 150 to 400 pF in 50-pF

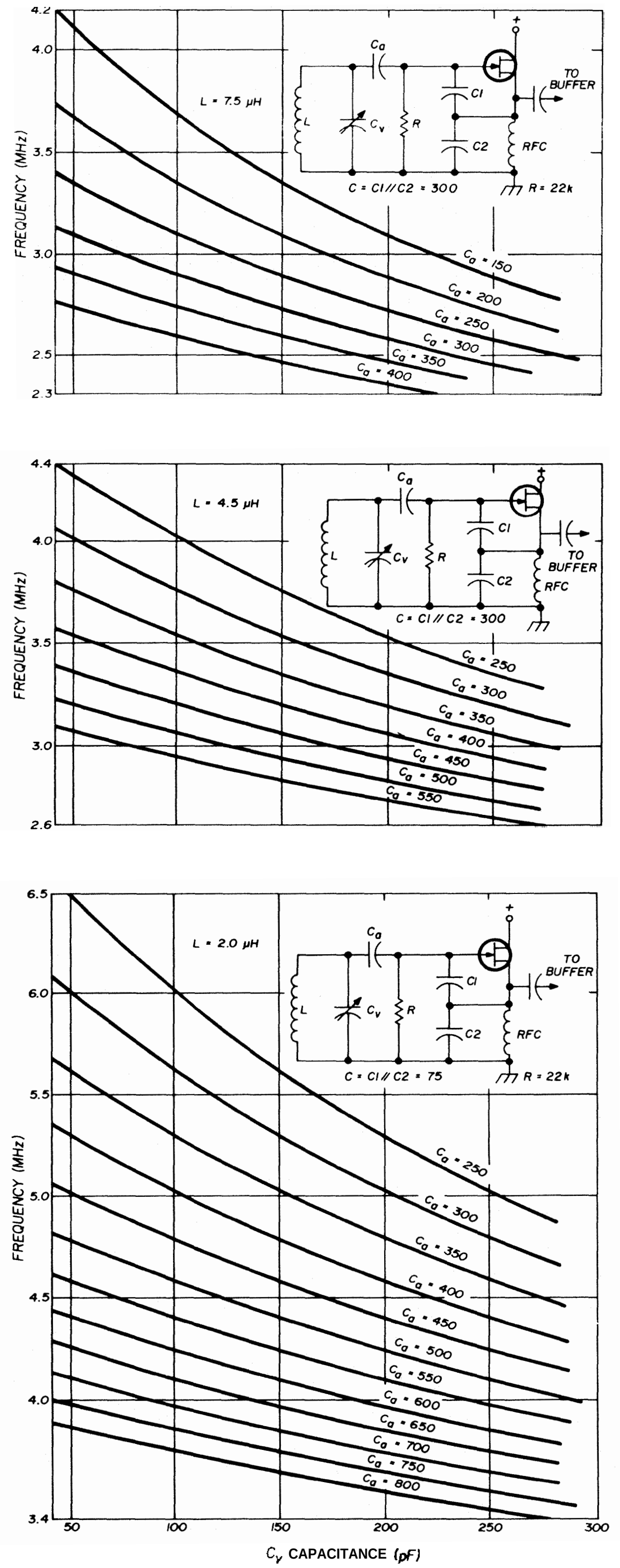


fig. 3. A family of parametric curves for determining component values for the Colpitts vfo in fig. 1. Operating frequency as a function of variable capacitance, C,,, is shown as a function of oscillator frequency with values of C_a as a parameter. The three sets of curves are for various values of inductance. L.

steps. **Fig. 3B** shows the universai characteristics for $L = 4.5\ \mu H$. Seven curves for C_a values between 250 and 550 pF have been calculated.

Fig. 3C refers to the universal characteristics for $L = 2\ \mu H$. These curves show that the effect of decreasing L causes an extension of the frequency range of interest by universal characteristics. For the characteristics of **figs. 3A** and **3B** it was not possible to calculate other C_a values because of the unstable conditions I found. Therefore, only 12 curves for C_a, from 250 to 800 pF, have been calculated.

how to use the universal characteristics

I call these characteristics "universal" because they are valid for all vfos (Colpitts) as shown in **fig. 1.** With these characteristic features you can determine directly the values of all vfo components, as shown in the following example.

Assume you want a variable oscillator covering 3.5-3.9 MHz. Choose two curves, the first from **fig. 3B** with $C_a = 300\ pF$ and the second one from **fig. 3C** with $C_a = 800\ pF$. In the first case, the linearity ratio value is $\Delta f/\Delta C = 4.4\ \frac{kHz}{pF}$; while in the second case it is $\Delta f/\Delta C_v = 2.1\ \frac{kHz}{pF}$. To obtain C,, (which in this case is 90 pF instead of 190 pF as in the second case) the curve of **fig. 3B** could be used. From this figure it's possible to get all the other values, which are:

$$C_a = 300\ pF \qquad C_v = 90\ pF$$
$$L = 4.5\ pF \qquad R = 22\ K$$
$$C1 = C2 = 600\ pF$$

Usually $C1$ and $C2$ have the same values to guarantee self-starting oscillations.

in conclusion

Once the working frequency range is chosen, it's possible to satisfy, through the universal characteristics of **fig. 3,** the following points:

1. Dial linearization; that is, keeping the $\Delta f/\Delta C_v$ ratio constant within the chosen range.

2. Definition of minimum and maximum C_v values.

In any case, for about 200 kHz of frequency range, the linearity error is very low for each curve of the families. On the contrary, for a wider frequency range, the choice of the curve is much more limited and depends on the desired $\Delta f/\Delta C_v$ ratio; that is the derivative value of the curve. Generally, for the $\Delta f/\Delta C_v$ ratio, it's advisable not to select very high values, around 4 kHz/pF.

ham radio

rf chokes

their performance above and below resonance

An investigation into the properties of rf chokes over wide frequency ranges

For years I've heard various electronics engineers quote a simple rule-of-thumb about rf chokes: "Never use an rf choke above its self-resonant frequency because it's capacitive!" The inference is, I think, that above its self-resonant frequency, the rf choke will have a very low reactance, which is the opposite effect usually desired from an rf choke. After all, capacitors are used for bypass and coupling functions, and an rf choke *does* show capacitive reactance above its self-resonant frequency.

This rule-of-thumb is a very safe and conservative way to apply rf chokes in circuit design, I suppose, but sometimes it makes it difficult to get the impedance you want over a desired frequency range; in general, the larger the inductance of an rf choke, the lower will be its self-resonant frequency.

A frequent application of rf chokes by amateurs is the use of a 2.5 millihenry choke in a circuit which operates up to 30 MHz. Such use is in direct conflict with the stated rule-of-thumb, because the self-resonant frequency of these chokes is usually on the order of 3 MHz! In the paragraphs which follow, I have attempted to resolve this conflict by mathematical analysis.

mathematical analysis of the rf choke

To evaluate the performance of rf chokes at different frequencies, I felt it necessary to develop a general mathematical expression for impedance as a function of self-resonant frequency. Starting with the mythical ideal rf choke, we have a pure inductor as shown in **fig. 1.** The equation for its impedance in ohms is

$$Z = j2\pi fL \qquad (1)$$

where Z is the impedance, f is frequency in Hertz, and L is inductance in henries. The j is simply a notation which indicates that the phase angle of the impedance is $+90°$ with respect to a pure resistance. To simplify the following equations a little, I will substitute ω for $2\pi f$

$$\omega = 2\pi f \qquad (2)$$

where ω is frequency in radians-per-second. You can always get back to frequency in Hertz by a rearrangement of **eq. 2.**

$$f = \frac{\omega}{2\pi} \qquad (3)$$

The solid line on the graph of **fig. 2** shows how the impedance of a pure inductor, having an inductance of 1 henry, rises linearly with increasing frequency, according to **eq. 1.** The linear relationship will continue to any frequency.

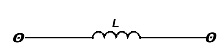


fig. 1. The ideal rf choke is a pure inductance.

Practical rf chokes, however, are not pure inductors; stray capacitance between turns of the coil is effectively in parallel with the ideal inductor. For the purpose of this discussion all of the stray capacitance may be lumped into one equivalent parallel capacitor, as shown in **fig.** 3. The **rf** choke now has a self-resonant frequency whose value is determined by the values of L and C. The equation for the impedance of this parallel circuit is

$$Z = \frac{(j\omega L)\frac{1}{-j\omega C}}{j\omega L + \frac{1}{-j\omega C}} = \frac{j\omega L}{1 - \omega^2 LC} \qquad (4)$$

The self-resonant frequency of the circuit will be

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248

designated w_o, and its value is

$$\omega_o = \frac{1}{\sqrt{LC}} \quad (5)$$

If ω_o is substituted for ω in **eq. 4,** we obtain the value of the choke's impedance at the self-resonant frequency.

$$Z_o = \frac{j\omega_o L}{1 - \omega_o^2 LC} = \frac{j\omega_o L}{1 - \left(\frac{1}{LC}\right) LC} = \frac{j\omega_o L}{1 - 1}$$

The denominator in **eq. 6** is equal to zero, so the impedance, Z_o, must be infinitely large (remember, we don't have any losses in our mathematical model, yet).

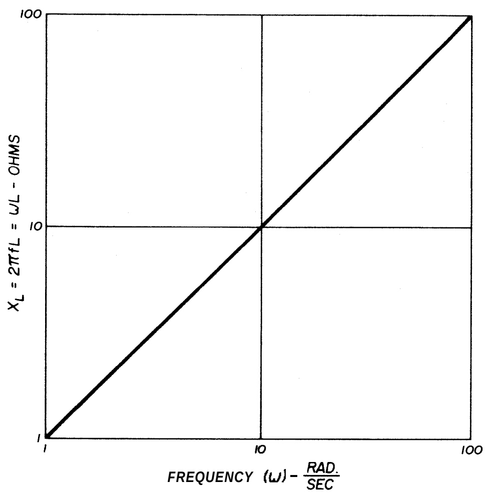


fig. 2. Reactance of a pure 1 henry inductor vs frequency (no stray capacitance or loss resistance).

To solve for the choke's impedance at one octave below and above the self-resonant frequency, substitute $\frac{w_o}{2}$ and $2\omega_o$, respectively, for w in **eq. 4**

$$Z_{\frac{\omega_o}{2}} = \frac{j\left(\frac{\omega_o}{2}\right) L}{1 - \frac{\omega_o^2}{4} LC} = \frac{j\left(\frac{\omega_o}{2}\right) L}{1 - \frac{LC}{4LC}}$$

$$= \frac{2}{3}\,(j\omega_o L) \quad (7)$$

$$Z_{2\omega_o} = \frac{j(2\omega_o) L}{1 - 4\omega_o^2 LC}$$

$$= \frac{j(2\omega_o) L}{1 - \frac{4LC}{LC}} = -\frac{2}{3}\,(j\omega_o L)$$

Let's closely examine these two results. Notice that the quantity $(j\omega_o L)$ in each answer is the reactance a pure inductor would have at ω_o, the frequency at which the practical choke is self-resonant. Notice too that the magnitudes of the two answers are equal; the choke has as much impedance one octave above the self-resonant frequency as it does one octave below. The minus sign of the impedance one octave above self-resonance indicates the impedance has a phase angle of $-90°$ with respect to a pure resistance, and this means it is a capacitive reactance.

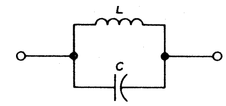


fig. 3. Practical rf choke has stray capacitance, creating a resonant circuit.

To explore this a little further, let's find the choke's impedance one decade below and above self-resonance

$$Z_{\frac{\omega_o}{10}} = \frac{j\frac{\omega_o}{10} L}{1 - \frac{\omega_o^2}{100} LC} = \frac{j\frac{\omega_o}{10} L}{1 - \frac{LC}{100LC}} = \frac{10}{99}\,(j\omega_o L) \quad (9)$$

$$Z_{10\omega_o} = \frac{j\,(10\omega_o)L}{1 - 100\omega_o^2 LC}$$

$$= \frac{j(10\omega_o)L}{1 - \frac{100LC}{LC}} = -\frac{10}{99}\,(j\omega_o L)$$

Again, both answers have the same magnitude, and the impedance is inductive below self-resonance and capacitive above self-resonance. I have solved for the choke's impedance at several more frequencies and plotted the curve shown in **fig. 4.** Although the choke's impedance is capacitive above self-resonance, it doesn't fall off any faster than it does below self-resonance.

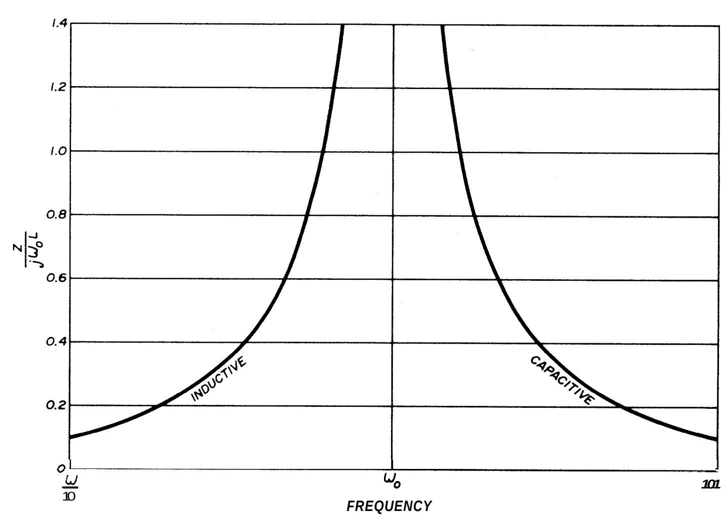


fig. 4. Relative frequency response of a lossless, but self-resonant rf choke. The frequency of resonance is at ω_o.

We must add an equivalent parallel resistance, R_p, as shown in **fig.** 5, to make the mathematical choke completely realistic. Another way to show this is pictured in **fig. 6,** where R_p is connected in parallel with the impedance, Z, solved for above. R_p accounts for losses in the choke; at the self-resonant frequency, the Impedance of the choke will be equal to R_p.

The value of R_p does not stay constant as frequency changes because losses tend to increase at higher frequencies. A major contributor to this characteristic is the skin effect in the wire of the coil, which causes R_p to decrease as a function of the square-root of frequency. Solving for the exact impedance of the rf choke with R_p present is a little tedious and will not be dealt with here. It is assumed, however, that R_p will be large enough so that the calculations above may be taken as rough approximations at fre-

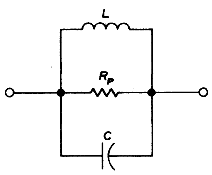


fig. 5. Equivalent parallel resistance, R_p, accounts for losses in rf chokes.

quencies at least one octave away from self-resonance. Thus the curve of **fig. 4** is assumed to be roughly correct outside the frequency range from $0.5\omega_o$ to 20,. An example may help to verify this.

example

Using manufacturer's specifications for a typical 2.5 mH rf choke, let's examine its performance using the relationships developed above. Inductance is specified as 2.5 mH at 250 kHz with a Q of 55. The choke's reactance and R_p at 250 kHz are

$$X = 2\pi(.25x10^6)(2.5x10^{-3}) = 3927 \text{ ohms} \quad \textbf{(11)}$$

$$R_p = QX = 55 \times 3927 = 215{,}985 \text{ ohms} \quad \textbf{(12)}$$

Self-resonant frequency is given as 2.5 MHz. Therefore,

$$j\omega_o L = j(2\pi f_o) L = j(2\pi \times 2.5 \times 10^6)(2.5 \times 10^{-3})$$
$$= j39{,}270 \text{ ohms} \quad \textbf{(13)}$$

Assuming R_p has decreased by the square-root of frequency change, we can estimate its value at 2.5 MHz as

$$R_p = \frac{215{,}985}{\sqrt{\frac{2.5 \text{ MHz}}{.25 \text{ MHz}}}} = \frac{215{,}985}{\sqrt{10}} = 68{,}300 \text{ ohms} \quad \textbf{(14)}$$

Dividing R_p by $j\omega_o L$, we find the choke's Q at 2.5 MHz is about 1.74, a very low value. To satisfy our curiosity, let's see what the value of the equivalent parallel capacitor is. The capacitor's reactance must

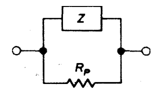


fig. 6. Another way of schematically representing an rf choke. Z is the lossless impedance calculated in the text.

equal the inductor's reactance at resonance, so

$$C = \frac{1}{2\pi(2.5 \times 10^6)(39{,}270)} = 1.6 \text{ pF} \quad \textbf{(15)}$$

Now let's see what sort of impedance the choke has at 25 MHz, one decade above its self-resonant frequency. Substituting the results of **eq. 13** into **eq. 10**

$$Z_{10\omega_o} = -\frac{10}{99}(j\omega_o L) = -j\frac{10}{99}(39{,}270) = -j3967 \text{ ohms}$$

The value of R_p at 25 MHz is estimated to be

$$\frac{215{,}985}{\sqrt{\frac{25 \text{ MHz}}{.25 \text{ MHz}}}} = \frac{215{,}985}{\sqrt{100}} = 21{,}598.5 \text{ ohms} \quad \textbf{(17)}$$

so it won't cause too large an error in the result of **eq. 16.** Notice that the impedance at 25 MHz, one decade above self-resonance, is very nearly the same as the impedance at 250 kHz, one decade below self-resonance (see **eq. 11**).

One final calculation of interest is the value of capacitance the choke represents at 25 MHz

$$C = \frac{1}{2\pi(25 \times 10^6)(3967)} = 1.6 \text{ pF} \quad \textbf{(18)}$$

Well, how about that? The same value as at self-resonance. Obviously, the equivalent parallel capacitance of the choke is totally dominating the value of reactance at 25 MHz. But then, it should.

conclusion

After going through the exercise above, it is evident that an rf choke behaves like any parallel tuned circuit. But since an rf choke has a self-resonant frequency, we knew it was just a parallel tuned circuit all along, didn't we? If we can use an rf choke below its self-resonant frequency, I see no reason not to use it above its self-resonant frequency, so long as it will provide the required impedance. An rf choke does look like a capacitor above self-resonance, but as we saw with the 2.5 mH choke, it can be a very small capacitor.

ham radio

techniques for preventing rf leakage from your transmitter

Lowering unwanted rf leakage from your transmitter reduces radio-frequency interference

We are all aware of the increasing incidence of RFI problems caused by solid-state home entertainment products which are not designed to function normally in the presence of a strong RF field. These kinds of problems can be cured only by adding filtering and/or shielding to the affected device.

At the same time it behooves us all not to forget to be sure that our own transmitters are clean. This means using a good lowpass filter and making sure that the only path out of the transmitter for rf energy is through that lowpass filter. If you have any RFI problems, this is the first thing that should be checked.

Run your transmitter into a shielded dummy load. If the RFI is still present, you have rf leakage which should be eliminated. If it is harmonic interference, no lowpass filter will do any good if the harmonics are leaking out ahead of it. If it is overload interference, it may be possible to eliminate it by curing the leakage, depending on the amount of leakage to begin with and the relative distances from the equipment and the antenna to the device suffering the interference.

Three areas need to be addressed to solve rf leakage problems: filtering of power, audio, and control leads; shielding effectiveness of coaxial cable; and leakage at joints in shielded enclosures. Techniques

By John E. Becker, K9MM, 201 East Marion Street, Prospect Heights, Illinois 60070

for filtering of power, audio, and control leads has been thoroughly discussed in many amateur publications so that topic will not be addressed here. Instead, I will concentrate on the other two areas.

There is considerable difference in the shielding efficiencies of different types of coaxial cable. The numbers to be quoted are from manufacturers' literature, and while different manufacturers do not necessarily agree on the shielding efficiency of a particular type of cable, there is no doubt about the significant difference in the shielding efficiency afforded by the different types of cable construction to be discussed, and the numbers should be considered with that in mind. The most commonly used type of coax, that having a single shield braid of stranded copper wire, has a shielding efficiency of −48 dB at 100 MHz for a 30 cm (1 foot) length of cable. That is, the total power radiated by that 30 cm of cable is 48 dB down from the signal level in the cable. This assumes 95 per cent shield coverage. This shielding efficiency increases only slightly with decreasing frequency, to −52 dB at 10 MHz according to data published by the Times Wire and Cable Company.

The shielding efficiency is significantly reduced by lower shield coverage, and 85-90 per cent is more typical. This means that if you have 30 cm of coax between your transmitter and lowpass filter, harmonics radiating from that piece of coax are 48 dB down from their level at the transmitter output. Under these conditions, a lowpass filter with 100 dB harmonic attenuation is no better than one with 48 dB attenuation as far as nearby interference problems are concerned. The filter will keep the harmonics from getting to the antenna to be radiated for long distances, but the TVI problem next door is just as likely to be due to the radiation from the cable.

While we're still on the subject of single shielded coax, be especially wary of bargain priced coax that is not made to military specifications. Some manufacturers have drastically reduced shield coverage in the last few years to keep costs down.

There are more sophisticated types of coaxial cable with much better shielding efficiencies. The most often encountered construction is double shielded coax. This type of cable uses two shield layers with no dielectric between them. Shielding efficiency is typically −87 dB at 100 MHz for a 30 cm (1 foot) length, and since this is a premium cable designed for excellent shielding you can be sure that the shield coverage is as great as possible.

A still better type of construction for shielding efficiency is triaxial cable. This cable has two shield layers with a dielectric layer between them. Shielding efficiency is further improved to −97 dB at 100 MHz for a 30 cm length.

The ultimate in shielding efficiency is the solid jacketed type of cable. Radiation from this type of cable is below measurable limits, and system shielding efficiency is limited only by leakage at a connector interface.

The use of one of these improved types of cable for all station interconnections up to the lowpass filter will ensure that you are getting all the harmonic protection the filter is capable of. Double-shielded substitutes for RG-8A/U or RG-213/U are RG-9B/U or RG-214/U. The double-shielded substitute for RG-58C/U is RG-55A/U or RG-223/U. A triaxial substitute for RG-8A/U or RG-213/U is Times TRF-8. A triaxial substitute for RG-58C/U is Times TRF-58 or Essex 21-204.

Whatever kind of cable is used, be sure the connectors are properly installed. An improperly attached shield will seriously degrade shielding effectiveness. With double-shielded and triaxial cables, both shields should be attached to the connector shell at both ends of each cable run. Assembly instructions for UHF, N, and BNC connectors can be found in both the ARRL *Handbook* and *Antenna Book.*

For sealing joints of shielded enclosures, the 3M company has two products that are hard to beat for effectiveness and ease of application. They are copper foil tapes with conductive adhesives, especially designed for RFI shielding. Type X-1181 is a smooth tape, and type X-1245 is embossed with a grid pattern for more reliable electrical contact with the surface on which it is applied. Both are available in widths from 6 to 25 mm (1/4 to 1 inch). Shielding efficiency of up to −65 dB is claimed by 3M, depending on the type of metal the tape is applied to. If desired, it is possible to solder to the tape without damaging it.

I have applied this tape to the edges of all shield enclosures in my equipment, to the seams of lowpass filter boxes, and over the edges of coax connectors. The same piece has been removed and re-applied on my linear amplifier cover shield several times with no apparent loss of adhesion. If the metal surface is clean to begin with, the adhesion is excellent. This type of tape may also be spirally wound over a length of conventional single shielded coax to obtain improved shielding effieciency. Each turn should overlap the previous turn by half the width of the tape, and the tape should extend over the shell of the connector at each end.

ham radio

simple and effective vertical antenna for portable communications

How to build
a simple
vertical antenna mount
that can be
set up quickly
for emergency
radio communications

One of the important aspects of amateur radio is the provision for efficient and timely communications in emergency or disaster situations. Having been a ham for several years I have found it interesting that, in the various ham-oriented magazines, there is substantial coverage given for proper procedures to follow during disaster or emergency communications but little mention or suggestion for effective and simple antenna systems. The following presents an idea which fulfills both the "simple" as well as "effective" aspects of an emergency antenna set-up and is easy and inexpensive to make. The secret to success is in the use of a vertical antenna (for the bands you intend to operate) with ground radials, and a homebrew base section which mounts in the spare tire from your vehicle (see **fig. 1**).

construction

The materials required to build the base section are a short length of mast or pipe and a front wheel hub that will bolt into your vehicle's spare tire (the wheel hub can be obtained from a local junk yard for a few dollars). You will also need a set of lug nuts that will fit the bolts on the hub.

The outside diameter of the mast should be as close to the inside diameter of the outside grease-seal surface of the hub as possible to make it easier to position the mast perpendicular to the hub. The mast must be welded to the hub.

Clean the hub of all oil or grease. Place it on a hard surface (that will not burn) with the lug-bolts facing up, stand the mast in the hub, and weld the mast and hub together as pictured in **fig.** 2. Be sure to position the mast perpendicular to the hub so the antenna will be straight and not tilted to one side or the other.

field use

To use the antenna set-up for emergency or portable operation (such as Field Day or a camping trip)

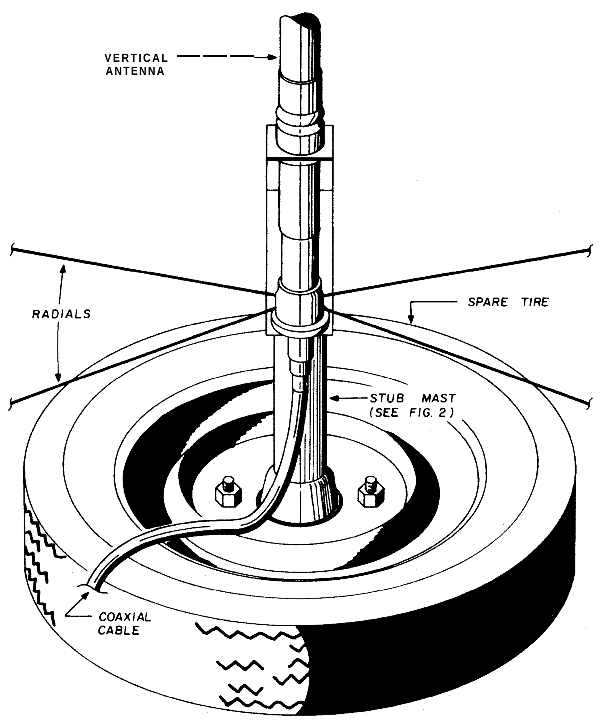


fig. 1. Portable vertical antenna mount consists of a junk-yard hub bolted to the spare tire from your car. In high wind conditions you may want to weight the tire down with rocks, but in most cases this isn't necessary. Construction of the hub section is shown in fig. 2.

all you need to do is remove the spare tire from your automobile, bolt in the base-section, lay the spare on the ground so the mast points up, and mount the antenna on the mast. If the vertical leans to one side, remove some dirt from under the high side of the tire or place some dirt, rocks, or a board under the low side to correct it. If you wish, you can place several

By John S. Jolly, WA7NWL, 1840 North 64th Lane, Phoenix, Arizona 85035

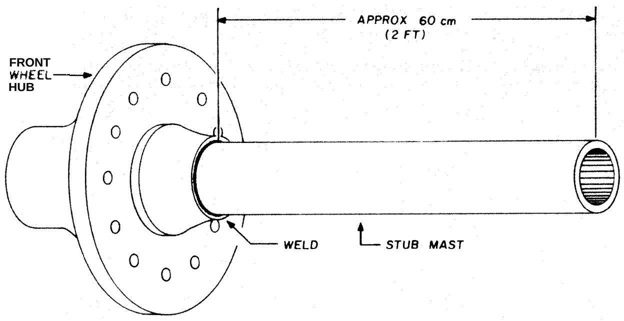


fig. 2. Stub mast for the portable antenna base consists of a front hub from the junk yard which fits your car, and a short steel mast which is welded to the hub. The outside diameter of the short mast should be very close to the inside diameter of the hub's grease-seal surface so the mast is perpendicular to the stub.

rocks on top of the spare for additional weight and stability (under normal conditions this should not be necessary). Attach the ground radials and coax, and you are ready to go on the air.

conclusion

It is the intent of this article to describe a relatively simple and inexpensive method for an average amateur to have an antenna set-up which is portable and effective for use when conditions necessitate it. In addition to emergency and portable use, this set-up could be used by those hams living in apartments or townhouses which forbid permanent fixture on their roofs.

ham radio

cleanup tips for amateur equipment

Being in a position that would be considered enviable among some amateur radio operators (this was written during my fifth year in the R.L. Drake Company Customer Service Department; I now work for ETO), I've been able to pick up on some tips and hints on routine maintenance for amateur equipment. Even though I work for one of the manufacturers, don't wrinkle your nose and flip the page in favor of the ads on the latest offerings from your local dealer. What I'm going to talk about is directly applicable to any amateur equipment.

The item I want to deal with is cleaning the exterior of your gear. The benefits here are numerous; you can operate without having an air-sick bag handy, most technicians would rather spend an hour giving the gear a little extra tweaking instead of cleaning it, and it keeps the wife from nagging you about operating a home for wayward cockroaches.

For those of you who smoke, the gear really gets coated with tobacco stains in a hurry. The primary reason for this fact is that most of us get so wrapped up in a hot contest that we wind up with two cigarettes between the lips and three burning in the ashtray without realizing it. In time, the panel markings get dim and bturred and you find yourself shading the vfo with one hand while squinting your eyes trying to read the numbers and keying with the other.

There are only two cures for this condition that I know of. One is to stop smoking (if you have the willpower — I don't), the other is a couple of products called *Sani-Wax* and *Fantastik* along with about two hours work. Even if you decide to quit smoking, you still have the gear to clean up or trade in, and clean equipment almost always brings a few extra bucks when you trade.

Sani-Wax is a white liquid that comes in a handy-dandy plastic quart container and retails for a dollar ninety-nine in my neck of the woods. The things it does for metal panels and cabinet borders is magnificent. It's the only product I've used that will cut through smoke, grease, oil, and coffee stains while leaving a nice shine behind. The best part is that it goes on easy and wipes off easier.

Fantastik is a spray cleaner from Texize Chemicals Company that will totally fascinate you with what it does for meters, vfo dials and windows, and especially knobs. It knocks the skin oils and accumulated dirt out of the little grooves in those plastic knobs like crazy and it doesn't scratch or discolor them. An old toothbrush makes the job easier and a lot faster. Just make sure you get it back in the rack before your mother-in-law shows up for her weekly overnight visit.

A word of caution before you start. It's a good idea to remove all knobs before beginning and to mark down what position your knobs are in BEFORE removing them. A lot of the controls, particularly bfo, audio/rf gain, passband tuning, and veniers, don't have flatted or keyed shafts to realign the pointers.

If you can't find *Fantastik* at your local supermarket or electronics surplus store, try writing to the Consumer Relations Department, Texize Chemicals Company — Division of Morton-Norwich Products, Inc., Greenville, South Carolina 29602. They can put you hot on the heels of a local distributor. The people at *Sani-Wax* can be contacted by writing to Roger Solem, Market Mechanics, 1900 East Randoll Mill Road, Suite 106, Arlington, Texas 76011.

By using *Sani-Wax* on all the metal parts, and *Fantastik* on all the plastic parts, your amateur equipment will shine like new with a minimum of work.

William D. Fisher

pi network design

for high-frequency power amplifiers

A complete discussion of pi and pi-L networks with computer derived component values for a wide range of operating conditions

The design of rf power amplifiers has always fascinated the typical radio amateur, and it remains one of the few fields in which a person of modest technical capability can still actively participate. Although the number of home-built transmitters has steadily diminished as more commercial companies have entered the market, many amateurs still like to design and build their own final amplifier. The information contained in this article should greatly assist those so inclined. Many interesting comparisons will be presented between amplifiers running at different power levels as well as pertinent computer-derived data for the proper selection of component values.

With single sideband and its legal 2-kW PEP maximum input power, certain problems crop up which many amtateurs overlook or are unable to handle. This is because the operator wants to run the amplifier at one power level for ssb and another for CW. The problems are compounded when the operator also wants to run RTTY, which is 100 per cent key-down continuous-carrier operation.

There is also a growing tendency to build power amplifiers with higher plate voltages than were common a few years ago. Part of this trend is due to the fact that the newer power tubes provide maximum performance at high plate voltages. Many of the pi-network design charts previously published have not been extended to include these higher operating voltages.

pi networks

The pi network is so named because of its resemblance to the Greek letter pi as shown in fig. **1.** The same network in its electrical form with input and output impedances is shown in fig. **2.** Since most amateurs use 50-ohm coaxial transmission line, the output load impedance of the pi network is usually 50 ohms.

When the pi network is used in a power amplifier, the circuit looks like that shown in fig. 3. The antenna provides the output load impedance, Z_L, and the power tube provides the input load impedance, Z_p. Since the plate load impedance usually falls into the range from 1200 to 5000 ohms, the pi network transforms the high impedance of the vacuum tube into the 50-ohm antenna load. It performs this job quite efficiently, and with predictable results.

Actually, the pi network is a basic form of a three-pole lowpass filter. With proper care in design it will attenuate the second harmonic by 35 dB or more.[1] This would be for a loaded Q of 12; if the Q is doubled, attenuation is increased by approximately 6 dB.

The pi-L network shown in fig. **4** consists of a standard pi network with an additional inductor. Since the pi-L network is a four-pole lowpass filter, second harmonic attenuation is increased to approximately 50 dB. This is particularly important if you want maximum suppression of TVI.

In addition to increased harmonic suppression, the pi-L network offers greater bandwidth for a given variation in operating Q, requires less output capacitance, and is able to operate efficiently with lower Q at very high plate load impedances. These advantages will become more apparent later.

By Irvin M. **Hoff, W6FFC** (reprinted from the September, 1972, issue of *ham radio)*

The dc plate resistance of a vacuum tube, at a given input power level, can be calculated with Ohm's law: $\boldsymbol{R} = E/I$, where E is the dc plate voltage and I is the dc plate current. However, since we are dealing with an ac circuit, this is of little value. What we need to know is the plate load *impedance.* This is

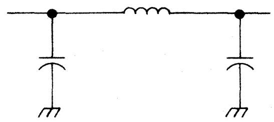


fig. 1. The pi network is so named because of its basic resemblance to the Greek letter π.

given approximately by the following equation which has been derived from the complex functions of a vacuum tube operating in class B.

$$Z_p \approx \frac{E}{1.57I} \qquad \textbf{(1)}$$

where Z_p is the plate load impedance, I is the indicated plate current, and E is the dc plate voltage.

When the vacuum tube is operated in class C, as for CW, the plate load impedance is approximated by

$$Z_p \approx \frac{E}{2I} \qquad \textbf{(2)}$$

If you are using a linear amplifier that runs with very high idling current, and approaches class A, the following approximation for plate load impedance would be more appropriate.

$$Z_p \approx \frac{E}{1.3I} \qquad \textbf{(3)}$$

Zero-bias grounded-grid linears are usually thought of as being class B, but there is no hard and fast rule in this regard. A number of articles has been written on this subject, and you are likely to have already formed some opinions of your own.

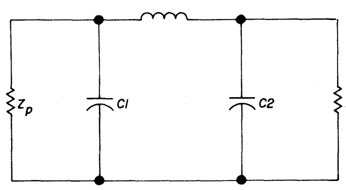


fig. 2. Basic pi network showing the input and output load impedance. The input load impedance in transmitters is the plate load impedance; output load impedance is usually 50 ohms in amateur stations.

Consider the case of a class-B rf power amplifier with a 2100-volt plate power supply and indicated plate current of 476 mA (1kW input). As calculated from **eq.** 2, the plate load impedance is 2800 ohms:

$$Z_p - \frac{2100}{1.57 \bullet 0.476} = 2800 \text{ ohms}$$

Typical plate load impedances for various power levels and different operating voltages and currents are shown in **table 1.** It can be seen from this data that the plate load impedance rises to very high levels when the plate voltage is increased above 4000 volts. More amateurs than might be expected use 4000 to 6000 volt power supplies, and many of the associated problems have not been adequately discussed in the past.

circuit Q

The letter Q stands for quality factor, and is used to describe, in simple numerical terms, the efficiency and performance of capacitors and inductors. Actually, there are two types of Q — *loaded Q* and *unloaded* Q. The unloaded Q is the inherent quality factor of the component itself; loaded Q is the quality factor of the component when it is used (and loaded down) by the circuit.

The unloaded Q of a component is given by

$$Q_u = Q\,(unloaded) = \frac{X}{\;} \qquad \textbf{(4)}$$

where X is reactance and r is ac resistance. The unloaded Q of a high-quality capacitor might be 1000 or more, and a silver-plated inductor might have an unloaded Q of more than 500.

The loaded Q of a pi network is usually on the order of 10 to 20 for maximum harmonic attenuation, and is given by:

$$Q_o = Q\,(loaded) = \sqrt{\frac{Z_p - Z_L}{Z_L}} \qquad \textbf{(5)}$$

where Z_p is the input impedance to the network, and Z_L is the output impedance.

When designing pi networks a value of loaded Q is chosen on the basis of harmonic attenuation, and is used in the design equations to determine the inductance and capacitance values for a given operating frequency.

L-networks

A typical step-down L-network is shown in **fig.** 5. This network is used to transform its input impedance to a lower output impedance. The Q of this circuit is entirely dependent upon the ratio of the input and output impedances as given in **eq.** 5.

For example, if the input impedance to an L-network is 2500 ohms, and the output impedance is 50 ohms, the loaded Q of the network is 7:

$$Q_o = Q\,(loaded) = \sqrt{\frac{2500 - 50}{50}} = \sqrt{49} = 7$$

However, a loaded Q of 7 is much too low for good harmonic suppression. To determine the L-network input impedance required to provide a desired value of loaded Q, **eq.** 5 is rearranged as shown below:

$$Z_p = Z_L(Q_o^2 + 1) \qquad \textbf{(6)}$$

For example, with an output load impedance of 50 ohms, and a desired loaded Q of 12 (for good harmonic suppression), the required input impedance is 7250 ohms. This is very restrictive and does not allow the designer sufficient latitude. So, although the L network is extremely efficient (98 per cent, typical), a pi network is usually used in transmitter output circuits.

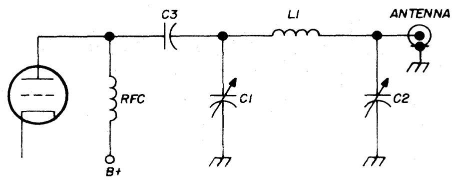


fig. 3. Pi network used in the output of an rf power amplifier is coupled to the power tube through a dc blocking capacitor (C3). C1 is the tuning capacitor, C2 is the loading capacitor, and L1 is the tank inductor.

pi network analysis

You can think of the pi network as being two L networks in tandem as shown in **fig. 6.** The first L network is a step-down type while the second L network is reversed for impedance step up. As an example, consider the case where the input impedance to the dissected pi network in **fig. 6** is 2900 ohms. With a Q of 12, the first L network would step the input impedance down to 20 ohms. This is often called the *virtual* impedance.

$$Z_L = \frac{Z_p}{Q_o^2 + 1} = \frac{2900}{12^2 + 1} = \frac{2900}{145} = ohms \qquad \textbf{(7)}$$

The second L network would then be designed to raise this virtual impedance of 20 ohms to 50 ohms to match the antenna. The Q of the second section would be quite low, on the order of 1.5.

As the input impedance is increased with Q held constant, the virtual impedance increases, and when the virtual impedance is equal to the desired output impedance, the pi network reverts to an L network. For example, with a plate load impedance of 7250 ohms and a Q of 12, the virtual impedance is 50 ohms. This is the maximum possible impedance transformation for a Q of 12 and an output impedance of 50 ohms.

Normally, about 70 per cent of the maximum possible impedance transformation is used in a practical circuit. For a Q of 12 and an antenna load of 50 ohms, this would represent a plate load resistance of 5075 ohms. If the plate load resistance in an rf power amplifier is higher than 5075 ohms, a Q of more than 12 is required to retain the same level of harmonic suppression. This problem is circumvented by the use of the pi-L network, as discussed below.

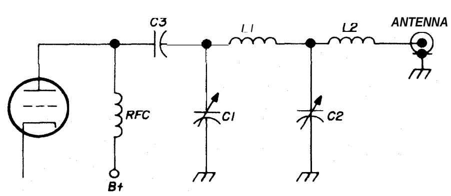


fig. 4. The pi-L network requires an additional inductor and provides increased second harmonic attenuation.

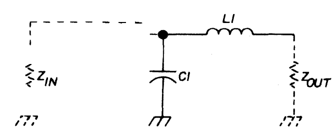
pi-L network design

Another L network may be added to the pi network as shown in **fig. 7** for additional harmonic attenuation. In actual practice C2 and C3 are combined into one capacitor so the circuit used in the transmitter is like that shown in **fig. 4.**

In the pi-L network, the input pi section transforms the plate load impedance to some lower figure, such as 300 ohms; this is often called the *image* impedance. The final L network transforms the image impedance down to 50 ohms to match the antenna.

From **eq. 6** it can be seen that with an image impedance of 300 ohms and a Q of 12, the pi network has a maximum transformation of 43500 ohms.

fig. 5. Typical step-down L network is highly efficient but very restrictive as far as acceptable Q is concerned.



Using 70 per cent of the maximum possible transformation as a practical maximum, as noted before, results in a maximum practical input impedance of 30500 ohms with a Q of 12. This is far in excess of what you will ever need in a power amplifier designed for amateur service.

The image impedance usually falls in the range between 200 and 400 ohms. It is selected for good harmonic attenuation, as well as balance in the T section of the pi-L network, and reasonable component values for the capacitors and inductors. If the image impedance is too high, the tuning capacitor (C1) will be too small on 10 and 15 meters, and the two inductors will be very large. Large inductors, of course, increase circulating currents which result in higher losses due to heat.

Q vs frequency

The loaded Q of a pi network (or any tank circuit, for that matter) is equal to its parallel-resonant impedance divided by either the inductive or capacitive reactance of the network.

$$Q_o = \frac{Z_p}{X} \qquad \textbf{(8)}$$

The reactance of any inductor is directly proportional to frequency, increasing as the frequency increases. Therefore, from **eq. 8** it can be seen that if a particular inductor is used, loaded *Q* will vary inversely with frequency. As the frequency is lowered, for example, *Q* is raised a proportionate amount. With this in mind, it is easy to determine the *Q* for a given network on a different frequency from the following formula:

$$\frac{f1}{f2} \cong \frac{Q2}{Q1} \quad (9)$$

Where $f1$ and $Q1$ are the frequency and *Q* at one frequency, and $f2$ and $Q2$ are at the second, different frequency.

For example, if an 80-meter pi network has a *Q* of 12 at 4.0 MHz, what is the *Q* at 3.5 MHz?

$$\frac{4.0}{3.5} = \frac{Q2}{12} \quad Q2 = 13.7$$

Although the actual loaded Q is somewhat dependent upon the value of plate load impedance used in the circuit, this approximation is accurate within 1 per cent. In the above example, with a plate load impedance of 3000 ohms, $Q2$ would actually be 13.84.

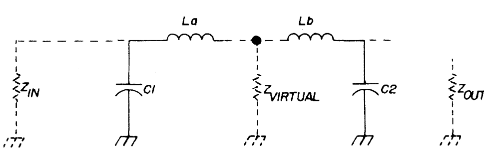


fig. 6. Pi network is basically two L networks in tandem.

Since the *Q* of the network goes up as the frequency goes down, it's a good idea to design the pi network for the highest frequency that is to be used.2 With this approach, when the same inductor is used at lower frequencies within an amateur band, *Q* increases somewhat, improving harmonic attenuation.

Table 2 shows how *Q* varies as a pi network is retuned to a different frequency (same inductor). **Table 2A** shows a pi network designed for 4.0 MHz which is retuned to 3.5 MHz; *Q* increases from **12** at 4.0 MHz to **13.8** at 3.5 MHz. The values of the tuning and loading capacitors are shown for comparison.

Table 2B shows the case where a pi network is designed for 3.5 MHz with a Q of **12** and retuned to 4.0 MHz (same inductor). The *Q* drops to 10.4, well below the selected minimum of 12.

network efficiency

As the loaded *Q* of a network is increased, efficiency goes down because of higher circulating currents and higher losses in the components. Approximate efficiency is given by

$$efficiency = 100\left(1 - \frac{Q_o}{Q_u}\right) \quad (10)$$

where Q_o is the loaded circuit *Q* and Q_u is the unloaded component *Q*. The graph in **fig. 8** shows that efficiency is a linear function of loaded *Q*. For minimum loss, the loaded Q should be as low as convenient, while still providing adequate harmonic attenuation. This figure has arbitrarily been chosen as 12.

When the pi network is designed, the minimum *Q* of **12** can only be obtained at the upper frequency of

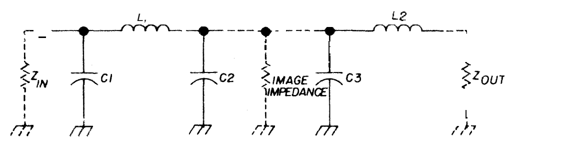


fig. 7. In the pi-L network a second L network is added to the basic pi network. Capacitors C2 and C3 are combined into one capacitor in a practical circuit, as shown in fig. 4.

each amateur band, and then only at the maximum input power level. For other frequencies or lower input powers, the loaded *Q* is higher than 12.

pi network design

Usually, when you are trying to design a pi network for your transmitter or linear amplifier, you must refer to graphs shown in reference books such as the ARRL "Radio Amateur's Handbook."[3] These graphs are often somewhat confusing because you must first determine the plate load impedance, select a value of *Q* and then find the reactance of each of the components. Then you must locate yet another graph to convert these reactance values into actual values of inductance and capacitance.

Few of these charts and graphs are extended above plate load impedances of 5000 ohms, and most give vague reference to the fact that if the plate load impedance is greater than about 5000 ohms, the *Q* should be increased.

The charts in **table** 3 and **table 4** are computer derived and offer all the required information to build

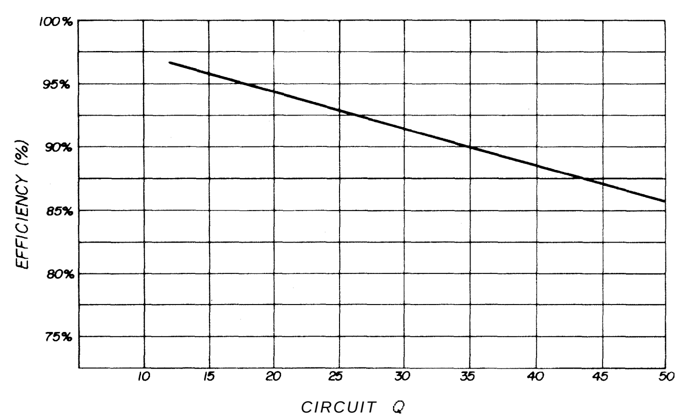


fig. 8. Efficiency of a network is inversely proportional to the loaded *Q* of the network.

table 1. Plate load impedances for different input power levels and different operating voltages and currents.

input power (W)	volts	mA	plate impedance (ohms)
1000	2000	500	2546
2000	2000	1000	1273
2500	2000	1250	1019
1000	2500	400	3979
2000	2500	800	1989
2500	2500	1000	1592
1000	2800	357	4991
2000	2800	714	2496
2500	2800	893	1996
1000	3300	303	6933
2000	3300	606	3466
2500	3300	758	2773
1000	4000	250	10186
2000	4000	500	5093
2500	4000	625	4074
1000	5000	200	15915
2000	5000	400	7958
2500	5000	500	6366
1000	6000	167	22918
2000	6000	333	11459
2500	6000	417	9167

a practical pi network. The pi networks in table 3 for plate load impedances above 5000 ohms have increasing Q so that the transformation ratio never exceeds the 70 per cent maximum. In addition, the inductors chosen for each of the designs are calculated for the highest frequency in the band.

The Q of the network at the highest frequency is 12 except when the plate load impedance is greater than 5075 ohms. The chart shows the capacitance values required to resonate the network to the lowest frequency in the band (maximum capacitance), as well as the operating Q at that frequency. In table **4,** the image impedance (R3) at the lower frequency is also given.

You will notice that the Q of the pi-L network does not go up as fast when frequency is lowered as it does with the pi network. Also, the Q remains the same for the pi-L for higher plate loads.

In both table 3 and table **4** the values for ten meters are for 29.7 MHz, the highest frequency in the band. This is because you need to know minimum capacitance values to reach this frequency in a five-band transmitter.

effect of swr

A standing-wave ratio of 4:1 will affect the capacitance required at C1 by ± 10 per cent and at C2 by ± 25 per cent. If the swr is caused by capacitive reactance, the tuning and loading capacitors are on the smaller side, and if the swr is the result of inductive reactance, the loading and tuning capacitors must be larger. Keep this in mind when selecting component values for a transmitter so you will be able to compensate for an antenna that is not exactly matched to your transmission line.

ssb and CW operation

Table **1** shows that for a given plate supply voltage, the plate load impedance is inversely proportional to the input power level. That is, 1000 watts at 2800 volts represents 5000 ohms, while 2 kW at the same supply voltage is 2500 ohms. Pi network values for a Q of 12 for each of these impedances are shown below (capacitace in pF, inductance in μH):

f	R1	C1	L1	C2	Q
4.0MHz	2500	191	9.18	1097	12.0
4.0MHz	5000	95	17.38	534	12.0

Note that the required inductance values are quite different.

As an example of what you may expect under actual operating conditions, consider the 2-kW design above (9.18 μH inductor). At 3.5 MHz with 2 kW input, C1 is 252 pF, C2 is 1536 pF, and Q is 13.8; not too bad. However, if the input power is reduced to 1000 watts at 3.5 MHz, C1 is 246 pF, C2 is 2287 pF, and Q reaches 27.0, increasing the circulating currents and heat losses in the network.

These figures point out the problems you can run into when you use the same operating voltage and same inductor at different power levels. Fortunately, there are several things which the designer can do to minimize these problems.

variable inductors

There are various variable rotary inductors available on the market which allow the operator to select the proper inductance for 1000 watts CW at the bottom of a band as well as 2000 watts PEP ssb at the top. When compared to fixed inductors, these variable units are fairly expensive, and require a turns counter, further increasing cost. However, they are available in various inductance and current-carrying abilities, so they encompass practically any design requirement. Also, using a variable inductor eliminates the need for a bandswitch.

bandswitch

The primary purpose of the bandswitch is to change the tap on the tank-circuit inductor to one better suited for the band in use. However, there are several other important functions for the bandswitch:

1. Used to switch input networks to match the 50-ohm output of the exciter.

2. Changes taps on the second inductor in a pi-L network.

3. Sometimes used to switch in additional tuning/loading capacitance on the 80-meter band so smaller variable capacitors may be used in the circuit.

Since you may wish to use a bandswitch in your power amplifier because of these additional uses, the variable inductor may lose some of its appeal.

In a novel approach to this problem a ten-position bandswitch has been used in one design to select different amounts of inductance for the CW and ssb ends of the band.4 However, the additional switch leads and the large number of inductor taps make this approach seem impractical for the typical home builder.

Table 5 compares the performance of 1000- and 2000-watt transmitters, as well as a 2000-watt transmitter run at 1000 watts input. In the latter case some additional losses are evident, but they're hardly large enough to cause much excitement. The same comparison shows that the 2-kW transmitter with a Q of 12 at 4 MHz, has a Q of 16.2 at 3.0 MHz. However, considerably more capacitance is required at C1 and C2. The pi-L network will alleviate this problem to some extent, as the Q of the pi-L does not increase as rapidly as frequency is lowered as it does with the straight pi network.

Since the 80-meter band is proportionally wider than any other high-frequency amateur band, there is some merit in using an extra bandswitch position for 80 meters.5 While I have shown previously that this is not required, it would be beneficial because you could select the 75-meter inductor for 4 MHz and 2-kW input, with the 80-meter inductor chosen for 3.7 MHz and 1-kW input.

The primary advantage in such an arrangement would be the ability to add a second input network to match the exciter. Since the input networks have low Q (typically 2 to 3), they are quite broadband and are usually set to a frequency in the middle of the band. However, it would be literally impossible for the same input network to work equally well on both 3.5 and 4.0 MHz, so it would be desirable to have one for each end of the band. From a practical standpoint, this might not be necessary because most operators have ample drive on CW if they are able to push the final to 2000 watts PEP on ssb.

table 2. Variations in Q as the resonant frequency of the pi network is changed (same inductor).

	frequency	plate impedance (ohms)	load impedance (ohms)	C1 (pF)	L1 (μH)	C2 (pF)	Q
A. Decreasing	4.0 MHz	2500	50	191	9.18	1097	12.0
frequency	3.5 MHz	2500	50	252	9.18	1536	13.8
B. Increasing	3.5 MHz	2500	50	218	10.49	1254	12.0
frequency	4.0 MHz	2500	50	165	10.49	863	10.4

power supply voltage

Since, as I just mentioned, most exciters have more than ample drive for 1000-watts input on CW if they are capable of driving the final to 2000-watts PEP on ssb, it's desirable to include some sort of automatic swamping so the exciter can be run in a normal manner for both ssb and CW. Lowering the plate supply voltage on the final-amplifier tubes decreases the plate load impedance required for a given input power level, therefore requiring more drive to reach this input power level,

For example, if it takes 70 watts drive with 3000 volts on the plate to reach 2000-watts input, then, depending upon the tubes used, it would take 70 to 80 watts drive to reach 1000-watts input with a substantially lower plate supply voltage. At the same time, the voltage-current relationship has changed, lowering the plate load impedance to something much closer to that which would give a Q of 12 with the same inductor.

Also, the plate voltage must be lowered to retain the same Q with the same inductor at the same operating frequency. This voltage reduction can be determined from

$$E2 = 0.71\,(E1) \tag{11}$$

where $E1$ is the original plate voltage for 2000-watts input and $E2$ is the lowered plate voltage for 1000-watts input. For example, a plate supply of 2800 volts for 2000-watts input must be changed to 2000 volts for 1000-watts input at the same operating frequency and circuit Q. Actually, on 3.5 MHz, this would be perhaps 1800 to 1900 volts to provide a Q of 12 at 3.5 MHz (1000 watts input) using a 2-kW transmitter designed for a Q of 12 at 4.0 MHz. However, it is unlikely that you could get 1000-watts input at this plate voltage, even with 100 watts drive on a cathode-driven grounded-grid amplifier.

tuning capacitance

Table 3 and table 4 show that the C1 tuning capacitance becomes quite small on 10 and 15 meters as the plate load impedance is raised. A typical 2000-watt transmitter might use 2800 volts on the plate,

table 3A. Pi network component values for matching a 50-ohm antenna load. Values have been chosen for a *Q* of 12 at the top edge of each amateur band. For plate load impedances greater than 5000 ohms, the *Q* of the network has been adjusted upward to compensate for the maximum transformation ratio, as discussed in the text. R1 is the plate load impedance.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
200	3.5	3147	0.98	6244	50	13.8
200	7.0	1427	0.54	2827	50	12.6
200	14.0	701	0.27	1387	50	12.3
200	21.0	465	0.18	921	50	12.3
200	29.7	322	0.13	636	50	12.0
300	3.5	2098	1.38	5071	50	13.8
300	7.0	952	0.76	2294	50	12.6
300	14.0	467	0.39	1125	50	12.3
300	21.0	310	0.26	747	50	12.3
300	29.7	214	0.19	516	50	12.0
400	3.5	1573	1.76	4368	50	13.8
400	7.0	714	0.97	1973	50	12.6
400	14.0	350	0.49	968	50	12.3
400	21.0	233	0.33	643	50	12.3
400	29.7	161	0.24	444	50	12.0
500	3.5	1259	2.14	3886	50	13.8
500	7.0	571	1.18	1753	50	12.6
500	14.0	280	0.60	859	50	12.3
500	21.0	186	0.40	571	50	12.3
500	29.7	129	0.29	394	50	12.0
600	3.5	1049	2.51	3528	50	13.8
600	7.0	476	1.38	1590	50	12.6
600	14.0	234	0.70	779	50	12.3
600	21.0	155	0.47	517	50	12.3
600	29.7	107	0.34	357	50	12.0
700	3.5	899	2.88	3248	50	13.8
700	7.0	408	1.58	1462	50	12.6
700	14.0	200	0.81	716	50	12.3
700	21.0	133	0.54	475	50	12.3
700	29.7	92	0.39	328	50	12.0
800	3.5	787	3.24	3021	50	13.8
800	7.0	357	1.78	1358	50	12.6
800	14.0	175	0.91	665	50	12.3
800	21.0	116	0.61	442	50	12.3
800	29.7	80	0.44	304	50	12.0
900	3.5	699	3.60	2832	50	13.8
900	7.0	317	1.98	1271	50	12.6
900	14.0	156	1.01	622	50	12.3
900	21.0	103	0.67	413	50	12.3
900	29.7	71	0.49	285	50	12.0
1000	3.5	629	3.96	2671	50	13.8
1000	7.0	285	2.18	1197	50	12.6
1000	14.0	140	1.11	586	50	12.3
1000	21.0	93	0.74	389	50	12.3
1000	29.7	64	0.54	268	50	12.0
1100	3.5	572	4.32	2532	50	13.8
1100	7.0	260	2.37	1133	50	12.6
1100	14.0	127	1.21	555	50	12.3
1100	21.0	85	0.81	368	50	12.3
1100	29.7	58	0.58	253	50	12.0
1200	3.5	524	4.67	2410	50	13.8
1200	7.0	238	2.57	1077	50	12.6
1200	14.0	117	1.31	527	50	12.3
1200	21.0	78	0.87	350	50	12.3
1200	29.7	54	0.63	241	50	12.0
1300	3.5	484	5.03	2302	50	13.8
1300	7.0	220	2.76	1027	50	12.6
1300	14.0	108	1.40	502	50	12.3
1300	21.0	72	0.94	333	50	12.3
1300	29.7	49	0.68	229	50	12.0
1400	3.5	450	5.38	2205	50	13.8
1400	7.0	204	2.95	982	50	12.6
1400	14.0	100	1.50	480	50	12.3
1400	21.0	66	1.01	319	50	12.3
1400	29.7	46	0.73	219	50	12.0
1500	3.5	420	5.73	2117	50	13.8
1500	7.0	190	3.14	942	50	12.6
1500	14.0	93	1.60	460	50	12.3
1500	21.0	62	1.07	305	50	12.3
1500	29.7	43	0.77	210	50	12.0
2000	3.5	315	7.46	1776	50	13.8
2000	7.0	143	4.09	783	50	12.6
2000	14.0	70	2.08	382	50	12.3
2000	21.0	47	1.39	253	50	12.3
2000	29.7	32	1.01	174	50	12.0
2500	3.5	252	9.17	1536	50	13.8
2500	7.0	114	5.03	670	50	12.6
2500	14.0	56	2.56	326	50	12.3
2500	21.0	37	1.71	216	50	12.3
2500	29.7	26	1.24	148	50	12.0
3000	3.5	210	10.86	1352	50	13.8
3000	7.0	95	5.95	583	50	12.6
3000	14.0	47	3.03	283	50	12.3
3000	21.0	31	2.02	187	50	12.3
3000	29.7	21	1.46	128	50	12.0
3500	3.5	180	12.53	1203	50	13.8
3500	7.0	82	6.86	512	50	12.6
3500	14.0	40	3.49	247	50	12.3
3500	21.0	27	2.34	164	50	12.3
3500	29.7	18	1.69	111	50	12.0
4000	3.5	157	14.19	1079	50	13.8
4000	7.0	71	7.77	451	50	12.6
4000	14.0	35	3.95	217	50	12.3
4000	21.0	23	2.64	144	50	12.3
4000	29.7	16	1.91	97	50	12.0
4500	3.5	140	15.84	971	50	13.8
4500	7.0	63	8.66	397	50	12.6
4500	14.0	31	4.40	190	50	12.3
4500	21.0	21	2.95	126	50	12.3
4500	29.7	14	2.13	84	50	12.0
5000	3.5	126	17.48	875	50	13.8
5000	7.0	57	9.55	348	50	12.6
5000	14.0	28	4.85	165	50	12.3
5000	21.0	19	3.25	109	50	12.3
5000	29.7	13	2.34	72	50	12.0
5500	3.5	119	18.41	861	50	14.4
5500	7.0	54	10.05	341	50	13.1
5500	14.0	27	5.11	162	50	12.8
5500	21.0	18	3.42	107	50	12.8
5500	29.7	12	2.47	70	50	12.5
6000	3.5	114	19.18	862	50	15.1
6000	7.0	52	10.48	341	50	13.7
6000	14.0	25	5.33	162	50	13.4
6000	21.0	17	3.56	107	50	13.4
6000	29.7	12	2.57	70	50	13.1
6500	3.5	110	19.92	862	50	15.7
6500	7.0	50	10.89	341	50	14.2
6500	14.0	24	5.54	162	50	14.0
6500	21.0	16	3.70	107	50	13.9
6500	29.7	11	2.67	70	50	13.6
7000	3.5	106	20.63	862	50	16.3
7000	7.0	48	11.28	341	50	14.8
7000	14.0	24	5.74	162	50	14.5
7000	21.0	16	3.84	107	50	14.4
7000	29.7	11	2.77	70	50	14.1
7500	3.5	102	21.31	862	50	16.8
7500	7.0	46	11.66	341	50	15.3
7500	14.0	23	5.93	162	50	15.0
7500	21.0	15	3.97	107	50	14.9
7500	29.7	10	2.86	70	50	14.6
8000	3.5	99	21.98	862	50	17.4
8000	7.0	45	12.02	341	50	15.8
8000	14.0	22	6.11	162	50	15.5
8000	21.0	15	4.09	107	50	15.4
8000	29.7	10	2.95	70	50	15.1
8500	3.5	96	22.62	862	50	17.9
8500	7.0	44	12.38	341	50	16.3
8500	14.0	21	6.29	162	50	16.0
8500	21.0	14	4.21	107	50	15.9
8500	29.7	10	3.04	70	50	15.6
9000	3.5	93	23.24	862	50	18.5
9000	7.0	42	12.72	341	50	16.7
9000	14.0	21	6.47	162	50	16.4
9000	21.0	14	4.33	107	50	16.4
9000	29.7	10	3.12	70	50	16.0
9500	3.5	91	23.85	862	50	19.0
9500	7.0	41	13.06	341	50	17.2
9500	14.0	20	6.64	162	50	16.9
9500	21.0	13	4.44	107	50	16.8
9500	29.7	9	3.21	70	50	16.4
10000	3.5	88	24.44	862	50	19.5
10000	7.0	40	13.38	341	50	17.7
10000	14.0	20	6.81	162	50	17.3
10000	21.0	13	4.55	107	50	17.3
10000	29.7	9	3.29	70	50	16.9
10500	3.5	86	25.02	862	50	19.9
10500	7.0	39	13.70	341	50	18.1
10500	14.0	19	6.97	162	50	17.8
10500	21.0	13	4.66	107	50	17.7
10500	29.7	9	3.37	70	50	17.3
11000	3.5	84	25.58	862	50	20.4
11000	7.0	38	14.01	341	50	18.5
11000	14.0	19	7.13	162	50	18.2
11000	21.0	12	4.77	107	50	18.1
11000	29.7	9	3.44	70	50	17.7
11500	3.5	83	26.13	862	50	20.9
11500	7.0	37	14.32	341	50	18.9
11500	14.0	18	7.28	162	50	18.6
11500	21.0	12	4.87	107	50	18.5
11500	29.7	8	3.52	70	50	18.1
12000	3.5	81	26.67	862	50	21.3
12000	7.0	37	14.61	341	50	19.3
12000	14.0	18	7.43	162	50	19.0
12000	21.0	12	4.97	107	50	18.9
12000	29.7	8	3.59	70	50	18.5

providing a plate load impedance of approximately 2500 ohms. This transmitter would require only 26 pF tuning capacitance to reach the top end of the 10-meter band.

Unfortunately, most modern rf power tubes designed for the 2000-watt level have output capacitances on the order of 10 pF — this leaves about 16 pF for tuning, including stray circuit capacitance.

If you study the various air-variable capacitors available you will find that it is virtually impossible to find a variable capacitor that will provide the necessary spacing for this operating voltage as well as tune the capacitance range needed for both 10 and 80 meters. Also, you must keep in mind that ±10 per cent leeway should be provided to compensate for any swr on the transmission line.

As the plate load impedance increases, the situation becomes even more acute. A 1000-watt transmitter with a plate supply of 2800 volts has a plate load impedance of 5000 ohms — on ten meters this means the tuning capacitor C1 is a total of 13 pF. In this case you would probably have to delete C1 entirely from the circuit and let the capacitance of the power tube supply the necessary tuning capacitance. However, this is not practical.

Fortunately, there are several things you can do to help alleviate this situation. You can use a smaller capacitor and add fixed capacitance on 40 and 80 meters, or use two variable capacitors, switching in the larger one on the lower bands. The vacuum capacitor is another possibility because of its low minimum capacitance, often as low as 3 pF. You can also blunder ahead and use a too-large capacitor, allowing the *Q* to be higher than normal.

table 3B. Pi-network component values for use within the 160-meter amateur band. Values were determined as in table 3A.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.	R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.	R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
1000	1.8	1187	7.94	5019	50	13.4	4750	1.8	250	33.33	1684	50	13.4	8500	1.8	181	45.26	1567	50	17.4
1000	1.9	1062	7.95	4459	50	12.7	4750	1.9	224	33.26	1403	50	12.7	8500	1.9	162	45.21	1290	50	16.4
1000	2.0	955	7.96	3979	50	12.0	4750	2.0	201	33.17	1155	50	12.0	8500	2.0	146	45.14	1042	50	15.6
1250	1.8	949	9.71	4419	50	13.4	5000	1.8	237	34.96	1593	50	13.4	8750	1.8	178	45.89	1567	50	17.6
1250	1.9	849	9.72	3917	50	12.7	5000	1.9	212	34.87	1315	50	12.7	8750	1.9	160	45.84	1290	50	16.7
1250	2.0	764	9.74	3487	50	12.0	5000	2.0	191	34.77	1058	50	12.0	8750	2.0	144	45.77	1042	50	15.8
1500	1.8	791	11.47	3969	50	13.4	5250	1.8	230	36.01	1566	50	13.7	9000	1.8	176	46.51	1567	50	17.9
1500	1.9	708	11.48	3510	50	12.7	5250	1.9	206	35.93	1290	50	12.9	9000	1.9	157	46.47	1290	50	16.9
1500	2.0	637	11.49	3116	50	12.0	5250	2.0	185	35.82	1042	50	12.2	9000	2.0	142	46.40	1042	50	16.0
1750	1.8	678	13.21	3613	50	13.4	5500	1.8	225	36.81	1566	50	14.0	9250	1.8	173	47.13	1567	50	18.1
1750	1.9	607	13.22	3197	50	12.7	5500	1.9	201	36.73	1290	50	13.2	9250	1.9	155	47.08	1290	50	17.1
1750	2.0	546	13.23	2822	50	12.0	5500	2.0	181	36.63	1042	50	12.5	9250	2.0	140	47.02	1042	50	16.2
2000	1.8	593	14.94	3322	50	13.4	5750	1.8	220	37.59	1567	50	14.3	9500	1.8	171	47.73	1567	50	18.4
2000	1.9	531	14.94	2922	50	12.7	5750	1.9	197	37.52	1290	50	13.5	9500	1.9	153	47.69	1290	50	17.4
2000	2.0	477	14.95	2579	50	12.0	5750	2.0	177	37.41	1042	50	12.8	9500	2.0	138	47.63	1042	50	16.4
2250	1.8	527	16.65	3076	50	13.4	6000	1.8	215	38.36	1567	50	14.6	9750	1.8	169	48.33	1567	50	18.6
2250	1.9	472	16.65	2697	50	12.7	6000	1.9	193	38.28	1290	50	13.8	9750	1.9	151	48.29	1290	50	17.6
2250	2.0	424	16.66	2373	50	12.0	6000	2.0	173	38.18	1042	50	13.1	9750	2.0	136	48.23	1042	50	16.7
2500	1.8	475	18.36	2864	50	13.4	6250	1.8	211	39.11	1567	50	14.9	10000	1.8	167	48.92	1568	50	18.9
2500	1.9	425	18.36	2504	50	12.7	6250	1.9	189	39.03	1290	50	14.1	10000	1.9	149	48.88	1290	50	17.8
2500	2.0	382	18.36	2194	50	12.0	6250	2.0	170	38.94	1042	50	13.3	10000	2.0	134	48.82	1042	50	16.9
2750	1.8	432	20.05	2678	50	13.4	6500	1.8	207	39.84	1567	50	15.2	10250	1.8	165	49.50	1568	50	19.1
2750	1.9	386	20.05	2333	50	12.7	6500	1.9	185	39.77	1290	50	14.4	10250	1.9	147	49.46	1290	50	18.0
2750	2.0	347	20.04	2036	50	12.0	6500	2.0	166	39.68	1042	50	13.6	10250	2.0	133	49.41	1042	50	17.1
3000	1.8	396	21.74	2513	50	13.4	6750	1.8	203	40.56	1567	50	15.5	10500	1.8	163	50.07	1568	50	19.3
3000	1.9	354	21.73	2181	50	12.7	6750	1.9	182	40.49	1290	50	14.6	10500	1.9	146	50.04	1290	50	18.3
3000	2.0	318	21.72	1894	50	12.0	6750	2.0	163	40.40	1042	50	13.9	10500	2.0	131	49.98	1042	50	17.3
3250	1.8	365	23.41	2364	50	13.4	7000	1.8	199	41.27	1567	50	15.8	10750	1.8	161	50.64	1568	50	19.6
3250	1.9	327	23.40	2043	50	12.7	7000	1.9	178	41.20	1290	50	14.9	10750	1.9	144	50.61	1290	50	18.5
3250	2.0	294	23.38	1766	50	12.0	7000	2.0	160	41.11	1042	50	14.1	10750	2.0	130	50.56	1042	50	17.5
3500	1.8	339	25.08	2228	50	13.4	7250	1.8	196	41.96	1567	50	16.1	11000	1.8	159	51.20	1568	50	19.8
3500	1.9	303	25.06	1917	50	12.7	7250	1.9	175	41.90	1290	50	15.2	11000	1.9	142	51.17	1290	50	18.7
3500	2.0	273	25.04	1647	50	12.0	7250	2.0	158	41.81	1042	50	14.4	11000	2.0	128	51.12	1042	50	17.7
3750	1.8	316	26.75	2104	50	13.4	7500	1.8	193	42.64	1567	50	16.3	11250	1.8	157	51.76	1568	50	20.0
3750	1.9	283	26.72	1801	50	12.7	7500	1.9	172	42.58	1290	50	15.4	11250	1.9	141	51.73	1290	50	18.9
3750	2.0	255	26.68	1538	50	12.0	7500	2.0	155	42.50	1042	50	14.6	11250	2.0	127	51.68	1042	50	17.9
4000	1.8	297	28.40	1938	50	13.4	7750	1.8	189	43.31	1567	50	16.6	11500	1.8	156	52.30	1568	50	20.2
4000	1.9	265	28.37	1693	50	12.7	7750	1.9	169	43.25	1290	50	15.7	11500	1.9	139	52.28	1290	50	19.1
4000	2.0	239	28.32	1435	50	12.0	7750	2.0	152	43.17	1042	50	14.8	11500	2.0	125	52.23	1042	50	18.1
4250	1.8	279	30.05	1881	50	13.4	8000	1.8	186	43.97	1567	50	16.9	11750	1.8	154	52.85	1568	50	20.5
4250	1.9	250	30.00	1591	50	12.7	8000	1.9	167	43.92	1290	50	15.9	11750	1.9	138	52.82	1290	50	19.3
4250	2.0	225	29.95	1337	50	12.0	8000	2.0	150	43.84	1042	50	15.1	11750	2.0	124	52.78	1042	50	18.3
4500	1.8	264	31.69	1779	50	13.4	8250	1.8	184	44.62	1567	50	17.1	12000	1.8	152	53.38	1568	50	20.7
4500	1.9	236	31.64	1495	50	12.7	8250	1.9	164	44.57	1290	50	16.2	12000	1.9	136	53.36	1290	50	19.5
4500	2.0	212	31.57	1244	50	12.0	8250	2.0	148	44.49	1042	50	15.3	12000	2.0	123	53.32	1042	50	18.5

Oddly enough, each of these different techniques is currently being used in commercial amateur-band power amplifiers. The vacuum variable provides the best answer to this problem, but it is also the most expensive (by a wide margin). However, the vacuum variable has many advantages worth considering if you are more interested in performance than in total cost.

From **table *6*** you can see that the *Q* on ten meters goes up quite rapidly if too much capacitance is used at C1. One currently available commercial amplifier uses 2800 volts at 2-kW input (plate impedance, 2500 ohms). For ten meters this calls for an input capacitor of about 15 pF, after the output capacitance of the tubes has been subtracted. However, this amplifier uses two 20-150 pF capacitors in parallel which are tuned in tandem with a geared arrangement. Thus, their minimum capacitance is about 40 pF, plus 10 pF added by the power tubes, providing a minimum input capacitance of more than 50 pF without any allowance for strays.

Table *6* shows that this gives a minimun *Q* of 24.0 at the top end of the ten-meter band (around 25.5 at the bottom end). If the amplifier were used at 1000-watts input, the *Q* would be nearly 48 at the top band edge and over 50 at the bottom!

This amplifier would obviously lose substantial power output in the form of heat in the tank inductor, and proper tuning would be very critical. It would also have to be retuned more often as frequency was changed.

This design is what I call the ***blunder-ahead*** method. In my mind, it would have been relatively simple for the manufacturer to use only one of the two tuning capacitors on 10, 15, and 20 meters, switching in the second tuning capacitor on 40 and 80.

Another manufacturer does precisely this. He uses a dual-section capacitor – half is used for the three upper bands and the other half is added in parallel on 40 and 80 meters. This provides normal *Q* for 2000-watts input on 10 meters. It still gives *Q* in excess of 20 with 1000-watts input, but that's really not too bad. This tuning system gives more than twice the ***vernier*** of the other system since the maximum capacitance on 20 meters, for example, is 120 pF. On the previous amplifier there is 300 pF available, even on 20 meters. The unit with the lower capacitance is far easier to tune on the upper three bands.

table 4A. Pi-L network component values for matching a 50-ohm antenna load. Values have been chosen for a *Q* of 12 at the top edge of each amateur band. The image impedance (R3) has been chosen to provide a balanced transformation in the T section of the pi-L network. R1 is the plate load impedance.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	L2 UH	R2 OHMS	R3 OHMS	'Q' QUAL
500	3.5	1219	2.86	2118	4.45	241	50	13.4
500	7.0	565	1.59	927	2.44	280	50	12.4
500	14.0	277	0.81	451	1.24	288	50	12.2
500	21.0	185	0.54	300	0.83	288	50	12.2
500	29.7	129	0.39	206	0.60	300	50	12.0
600	3.5	1015	3.30	1963	4.45	241	50	13.4
600	7.0	471	1.83	861	2.44	280	50	12.4
600	14.0	231	0.94	419	1.24	288	50	12.2
600	21.0	154	0.63	279	0.83	288	50	12.2
600	29.7	107	0.45	191	0.60	300	50	12.0
700	3.5	870	3.74	1843	4.45	241	50	13.4
700	7.0	403	2.07	809	2.44	280	50	12.4
700	14.0	198	1.06	394	1.24	288	50	12.2
700	21.0	132	0.71	262	0.83	288	50	12.2
700	29.7	92	0.51	180	0.60	300	50	12.0
800	3.5	762	4.17	1746	4.45	241	50	13.4
800	7.0	353	2.31	767	2.44	280	50	12.4
800	14.0	173	1.18	373	1.24	288	50	12.2
800	21.0	116	0.79	249	0.83	288	50	12.2
800	29.7	80	0.57	170	0.60	300	50	12.0
900	3.5	677	4.60	1666	4.45	241	50	13.4
900	7.0	314	2.54	733	2.44	280	50	12.4
900	14.0	154	1.30	357	1.24	288	50	12.2
900	21.0	103	0.87	238	0.83	288	50	12.2
900	29.7	71	0.63	163	0.60	300	50	12.0
1000	3.5	609	5.02	1598	4.45	241	50	13.4
1000	7.0	282	2.77	703	2.44	280	50	12.4
1000	14.0	139	1.42	342	1.24	288	50	12.2
1000	21.0	92	0.95	228	0.83	288	50	12.2
1000	29.7	64	0.68	156	0.60	300	50	12.0
1100	3.5	554	5.43	1539	4.45	241	50	13.4
1100	7.0	257	3.00	678	2.44	280	50	12.4
1100	14.0	125	1.53	330	1.24	288	50	12.2
1100	21.0	84	1.02	220	0.83	288	50	12.2
1100	29.7	58	0.74	151	0.60	300	50	12.0
1200	3.5	508	5.85	1488	4.45	241	50	13.4
1200	7.0	235	3.23	656	2.44	280	50	12.4
1200	14.0	116	1.65	320	1.24	288	50	12.2
1200	21.0	77	1.10	213	0.83	288	50	12.2
1200	29.7	54	0.80	146	0.60	300	50	12.0
1300	3.5	469	6.25	1443	4.45	241	50	13.4
1300	7.0	217	3.45	637	2.44	280	50	12.4
1300	14.0	107	1.76	310	1.24	288	50	12.2
1300	21.0	71	1.18	207	0.83	288	50	12.2
1300	29.7	49	0.85	142	0.60	300	50	12.0
1400	3.5	435	6.66	1403	4.45	241	50	13.4
1400	7.0	202	3.67	619	2.44	280	50	12.4
1400	14.0	99	1.88	302	1.24	288	50	12.2
1400	21.0	66	1.25	201	0.83	288	50	12.2
1400	29.7	46	0.90	138	0.60	300	50	12.0
1500	3.5	406	7.06	1367	4.45	241	50	13.4
1500	7.0	188	3.89	604	2.44	280	50	12.4
1500	14.0	92	1.99	294	1.24	288	50	12.2
1500	21.0	62	1.33	196	0.83	288	50	12.2
1500	29.7	43	0.96	134	0.60	300	50	12.0
2000	3.5	305	9.05	1228	4.45	241	50	13.4
2000	7.0	141	4.97	544	2.44	280	50	12.4
2000	14.0	69	2.54	265	1.24	288	50	12.2
2000	21.0	46	1.69	177	0.83	288	50	12.2
2000	29.7	32	1.22	121	0.60	300	50	12.0
2500	3.5	244	10.99	1132	4.45	241	50	13.4
2500	7.0	113	6.03	503	2.44	280	50	12.4
2500	14.0	55	3.08	245	1.24	288	50	12.2
2500	21.0	37	2.05	164	0.83	288	50	12.2
2500	29.7	26	1.48	112	0.60	300	50	12.0
3000	3.5	203	12.90	1062	4.45	241	50	13.4
3000	7.0	94	7.07	472	2.44	280	50	12.4
3000	14.0	46	3.61	231	1.24	288	50	12.2
3000	21.0	31	2.41	154	0.83	288	50	12.2
3000	29.7	21	1.74	106	0.60	300	50	12.0
3500	3.5	174	14.79	1006	4.45	241	50	13.4
3500	7.0	81	8.10	449	2.44	280	50	12.4
3500	14.0	40	4.13	219	1.24	288	50	12.2
3500	21.0	26	2.76	146	0.83	288	50	12.2
3500	29.7	18	1.99	100	0.60	300	50	12.0
4000	3.5	152	16.67	961	4.45	241	50	13.4
4000	7.0	71	9.12	429	2.44	280	50	12.4
4000	14.0	35	4.65	210	1.24	288	50	12.2
4000	21.0	23	3.10	140	0.83	288	50	12.2
4000	29.7	16	2.24	96	0.60	300	50	12.0

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	L2 UH	R2 OHMS	R3 OHMS	'Q' QUAL
4500	3.5	135	18.53	924	4.45	241	50	13.4
4500	7.0	63	10.12	413	2.44	280	50	12.4
4500	14.0	31	5.17	202	1.24	288	50	12.2
4500	21.0	21	3.44	135	0.83	288	50	12.2
4500	29.7	14	2.49	93	0.60	300	50	12.0
5000	3.5	122	20.37	892	4.45	241	50	13.4
5000	7.0	56	11.13	399	2.44	280	50	12.4
5000	14.0	28	5.68	195	1.24	288	50	12.2
5000	21.0	18	3.78	130	0.83	288	50	12.2
5000	29.7	13	2.73	90	0.60	300	50	12.0
5500	3.5	111	22.21	864	4.45	241	50	13.4
5500	7.0	51	12.12	388	2.44	280	50	12.4
5500	14.0	25	6.18	189	1.24	288	50	12.2
5500	21.0	17	4.12	126	0.83	288	50	12.2
5500	29.7	12	2.97	87	0.60	300	50	12.0
6000	3.5	102	24.03	840	4.45	241	50	13.4
6000	7.0	47	13.11	377	2.44	280	50	12.4
6000	14.0	23	6.69	184	1.24	288	50	12.2
6000	21.0	15	4.46	123	0.83	288	50	12.2
6000	29.7	11	3.22	85	0.60	300	50	12.0
6500	3.5	94	25.85	819	4.45	241	50	13.4
6500	7.0	43	14.09	368	2.44	280	50	12.4
6500	14.0	21	7.19	180	1.24	288	50	12.2
6500	21.0	14	4.79	120	0.83	288	50	12.2
6500	19.7	10	3.46	83	0.60	300	50	12.0
7000	3.5	87	27.65	799	4.45	241	50	13.4
7000	7.0	40	15.07	360	2.44	280	50	12.4
7000	14.0	20	7.69	176	1.24	288	50	12.2
7000	21.0	13	5.12	117	0.83	288	50	12.2
7000	29.7	9	3.70	81	0.60	300	50	12.0
7500	3.5	81	29.45	782	4.45	241	50	13.4
7500	7.0	38	16.05	352	2.44	280	50	12.4
7500	14.0	18	8.18	172	1.24	288	50	12.2
7500	21.0	12	5.45	115	0.83	288	50	12.2
7500	29.7	9	3.93	79	0.60	300	50	12.0
8000	3.5	76	31.25	766	4.45	241	50	13.4
8000	7.0	35	17.02	345	2.44	280	50	12.4
8000	14.0	17	8.68	169	1.24	288	50	12.2
8000	21.0	12	5.78	113	0.83	288	50	12.2
8000	29.7	8	4.17	78	0.60	300	50	12.0
8500	3.5	72	33.03	751	4.45	241	50	13.4
8500	7.0	33	17.99	339	2.44	280	50	12.4
8500	14.0	16	9.17	166	1.24	288	50	12.2
8500	21.0	11	6.11	111	0.83	288	50	12.2
8500	29.7	8	4.41	76	0.60	300	50	12.0
9000	3.5	68	34.82	738	4.45	241	50	13.4
9000	7.0	31	18.95	333	2.44	280	50	12.4
9000	14.0	15	9.66	163	1.24	288	50	12.2
9000	21.0	10	6.44	109	0.83	288	50	12.2
9000	29.7	7	4.64	74	0.60	300	50	12.0
9500	3.5	64	36.59	725	4.45	241	50	13.4
9500	7.0	30	19.91	328	2.44	280	50	12.4
9500	14.0	15	10.15	160	1.24	288	50	12.2
9500	21.0	10	6.77	107	0.83	288	50	12.2
9500	29.7	7	4.88	74	0.60	300	50	12.0
10000	3.5	61	38.36	714	4.45	241	50	13.4
10000	7.0	28	20.87	323	2.44	280	50	12.4
10000	14.0	14	10.64	158	1.24	288	50	12.2
10000	21.0	9	7.09	105	0.83	288	50	12.2
10000	29.7	6	5.11	73	0.60	300	50	12.0
10500	3.5	58	40.13	703	4.45	241	50	13.4
10500	7.0	27	21.82	318	2.44	280	50	12.4
10500	14.0	13	11.12	156	1.24	288	50	12.2
10500	21.0	9	7.42	104	0.83	288	50	12.2
10500	29.7	6	5.34	72	0.60	300	50	12.0
11000	3.5	55	41.89	693	4.45	241	50	13.4
11000	7.0	26	22.78	314	2.44	280	50	12.4
11000	14.0	13	11.61	154	1.24	288	50	12.2
11000	21.0	8	7.74	102	0.83	288	50	12.2
11000	29.7	6	5.58	71	0.60	300	50	12.0
11500	3.5	53	43.65	683	4.45	241	50	13.4
11500	7.0	25	23.73	310	2.44	280	50	12.4
11500	14.0	12	12.09	152	1.24	288	50	12.2
11500	21.0	8	8.06	101	0.83	288	50	12.2
11500	29.7	6	5.81	70	0.60	300	50	12.0
12000	3.5	51	45.40	674	4.45	241	50	13.4
12000	7.0	24	24.67	306	2.44	280	50	12.4
12000	14.0	12	12.57	150	1.24	288	50	12.2
12000	21.0	8	8.38	100	0.83	288	50	12.2
12000	29.7	5	6.04	69	0.60	300	50	12.0

table 4B. Pi-L network component values for the 160-meter amateur band. Values were determined as in table 4A.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	L2 UH	R2 OHMS	R3 OHMS	'Q' QUAL
1000	1.8	1155	10.08	2985	8.90	252	50	13.1
1000	1.9	1047	10.12	2626	8.90	275	50	12.5
1000	2.0	955	10.16	2322	8.90	300	50	12.0
1250	1.8	924	12.14	2739	8.90	252	50	13.1
1250	1.9	838	12.18	2412	8.90	275	50	12.5
1250	2.0	764	12.22	2135	8.90	300	50	12.0
1500	1.8	770	14.17	2556	8.90	252	50	13.1
1500	1.9	698	14.20	2253	8.90	275	50	12.5
1500	2.0	637	14.23	1997	8.90	300	50	12.0
1750	1.8	660	16.17	2413	8.90	252	50	13.1
1750	1.9	590	16.19	2129	8.90	275	50	12.5
1750	2.0	546	16.22	1889	8.90	300	50	12.0
2000	1.8	578	18.14	2298	8.90	252	50	13.1
2000	1.9	524	18.15	2029	8.90	275	50	12.5
2000	2.0	477	18.17	1801	8.90	300	50	12.0
2250	1.8	514	20.09	2203	8.90	252	50	13.1
2250	1.9	466	20.10	1946	8.90	275	50	12.5
2250	2.0	424	20.11	1729	8.90	300	50	12.0
2500	1.8	462	22.02	2121	8.90	252	50	13.1
2500	1.9	419	22.02	1876	8.90	275	50	12.5
2500	2.0	382	22.02	1667	8.90	300	50	12.0
2750	1.8	420	23.94	2051	8.90	252	50	13.1
2750	1.9	381	23.93	1815	8.90	275	50	12.5
2750	2.0	347	23.92	1614	8.90	300	50	12.0
3000	1.8	385	25.85	1990	8.90	252	50	13.1
3000	1.9	349	25.83	1762	8.90	275	50	12.5
3000	2.0	318	25.81	1568	8.90	300	50	12.0
3250	1.8	356	27.74	1936	8.90	252	50	13.1
3250	1.9	322	27.71	1715	8.90	275	50	12.5
3250	2.0	294	27.68	1527	8.90	300	50	12.0
3500	1.8	330	29.62	1887	8.90	252	50	13.1
3500	1.9	299	29.58	1673	8.90	275	50	12.5
3500	2.0	273	29.54	1490	8.90	300	50	12.0
3750	1.8	308	31.50	1844	8.90	252	50	13.1
3750	1.9	279	31.45	1635	8.90	275	50	12.5
3750	2.0	255	31.40	1457	8.90	300	50	12.0
4000	1.8	289	33.37	1804	8.90	252	50	13.1
4000	1.9	262	33.30	1600	8.90	275	50	12.5
4000	2.0	239	33.24	1427	8.90	300	50	12.0
4250	1.8	272	35.23	1768	8.90	252	50	13.1
4250	1.9	246	35.15	1569	8.90	275	50	12.5
4250	2.0	225	35.08	1399	8.90	300	50	12.0
4500	1.8	257	37.08	1735	8.90	252	50	13.1
4500	1.9	233	36.99	1540	8.90	275	50	12.5
4500	2.0	212	36.91	1374	8.90	300	50	12.0
4750	1.8	243	38.92	1704	8.90	252	50	13.1
4750	1.9	221	38.83	1513	8.90	275	50	12.5
4750	2.0	201	38.73	1351	8.90	300	50	12.0
5000	1.8	231	40.76	1675	8.90	252	50	13.1
5000	1.9	209	40.65	1489	8.90	275	50	12.5
5000	2.0	191	40.54	1329	8.90	300	50	12.0
5250	1.8	220	42.60	1649	8.90	252	50	13.1
5250	1.9	200	42.47	1466	8.90	275	50	12.5
5250	2.0	182	42.35	1309	8.90	300	50	12.0
5500	1.8	210	44.43	1624	8.90	252	50	13.1
5500	1.9	190	44.29	1444	8.90	275	50	12.5
5500	2.0	174	44.16	1290	8.90	300	50	12.0
5750	1.8	201	46.25	1601	8.90	252	50	13.1
5750	1.9	182	46.10	1424	8.90	275	50	12.5
5750	2.0	166	45.95	1273	8.90	300	50	12.0
6000	1.8	193	48.07	1579	8.90	252	50	13.1
6000	1.9	175	47.91	1405	8.90	275	50	12.5
6000	2.0	159	47.75	1256	8.90	300	50	12.0
6250	1.8	185	49.88	1559	8.90	252	50	13.1
6250	1.9	168	49.71	1387	8.90	275	50	12.5
6250	2.0	153	49.53	1241	8.90	300	50	12.0
6500	1.8	178	51.69	1539	8.90	252	50	13.1
6500	1.9	161	51.50	1370	8.90	275	50	12.5
6500	2.0	147	51.32	1226	8.90	300	50	12.0
6750	1.8	171	53.50	1521	8.90	252	50	13.1
6750	1.9	155	53.30	1354	8.90	275	50	12.5
6750	2.0	141	53.10	1212	8.90	300	50	12.0
7000	1.8	165	55.30	1503	8.90	252	50	13.1
7000	1.9	150	55.09	1339	8.90	275	50	12.5
7000	2.0	136	54.87	1199	8.90	300	50	12.0
7250	1.8	159	57.10	1487	8.90	252	50	13.1
7250	1.9	144	56.87	1325	8.90	275	50	12.5
7250	2.0	132	56.64	1186	8.90	300	50	12.0
7500	1.8	154	58.89	1471	8.90	252	50	13.1
7500	1.9	140	58.65	1311	8.90	275	50	12.5
7500	2.0	127	58.41	1174	8.90	300	50	12.0
7750	1.8	149	60.69	1456	8.90	252	50	13.1
7750	1.9	135	60.43	1298	8.90	275	50	12.5
7750	2.0	123	60.17	1163	8.90	300	50	12.0
8000	1.8	144	62.47	1442	8.90	252	50	13.1
8000	1.9	131	62.20	1285	8.90	275	50	12.5
8000	2.0	119	61.93	1152	8.90	300	50	12.0
8250	1.8	140	64.26	1428	8.90	252	50	13.1
8250	1.9	127	63.97	1273	8.90	275	50	12.5
8250	2.0	116	63.69	1141	8.90	300	50	12.0
8500	1.8	136	66.04	1414	8.90	252	50	13.1
8500	1.9	123	65.74	1262	8.90	275	50	12.5
8500	2.0	112	65.44	1131	8.90	300	50	12.0
8750	1.8	132	67.82	1402	8.90	252	50	13.1
8750	1.9	120	67.50	1251	8.90	275	50	12.5
8750	2.0	109	67.19	1122	8.90	300	50	12.0
9000	1.8	128	69.59	1390	8.90	252	50	13.1
9000	1.9	116	69.27	1240	8.90	275	50	12.5
9000	2.0	106	68.94	1112	8.90	300	50	12.0
9250	1.8	125	71.37	1378	8.90	252	50	13.1
9250	1.9	113	71.02	1230	8.90	275	50	12.5
9250	2.0	103	70.69	1104	8.90	300	50	12.0
9500	1.8	122	73.14	1366	8.90	252	50	13.1
9500	1.9	110	72.78	1220	8.90	275	50	12.5
9500	2.0	101	72.43	1095	8.90	300	50	12.0
9750	1.8	119	74.91	1355	8.90	252	50	13.1
9750	1.9	107	74.53	1211	8.90	275	50	12.5
9750	2.0	98	74.17	1087	8.90	300	50	12.0
10000	1.8	116	76.67	1345	8.90	252	50	13.1
10000	1.9	105	76.28	1202	8.90	275	50	12.5
10000	2.0	95	75.90	1079	8.90	300	50	12.0
10250	1.8	113	78.43	1335	8.90	252	50	13.1
10250	1.9	102	78.03	1193	8.90	275	50	12.5
10250	2.0	93	77.64	1071	8.90	300	50	12.0
10500	1.8	110	80.19	1325	8.90	252	50	13.1
10500	1.9	100	79.78	1184	8.90	275	50	12.5
10500	2.0	91	79.37	1063	8.90	300	50	12.0
10750	1.8	107	81.95	1315	8.90	252	50	13.1
10750	1.9	97	81.52	1176	8.90	275	50	12.5
10750	2.0	89	81.09	1056	8.90	300	50	12.0
11000	1.8	105	83.71	1306	8.90	252	50	13.1
11000	1.9	95	83.26	1168	8.90	275	50	12.5
11000	2.0	87	82.82	1049	8.90	300	50	12.0
11250	1.8	103	85.46	1297	8.90	252	50	13.1
11250	1.9	93	85.00	1160	8.90	275	50	12.5
11250	2.0	85	84.54	1042	8.90	300	50	12.0
11500	1.8	100	87.21	1288	8.90	252	50	13.1
11500	1.9	91	86.73	1152	8.90	275	50	12.5
11500	2.0	83	86.26	1036	8.90	300	50	12.0
11750	1.8	98	88.96	1280	8.90	252	50	13.1
11750	1.9	89	88.47	1145	8.90	275	50	12.5
11750	2.0	81	87.98	1029	8.90	300	50	12.0
12000	1.8	96	90.71	1271	8.90	252	50	13.1
12000	1.9	87	90.20	1138	8.90	275	50	12.5
12000	2.0	80	89.70	1023	8.90	300	50	12.0

band	turns	approximate inductance
80	11.000	4.43 μH
40	7.125	2.43 μH
20	4.5000	1.23 μH
15	3.500	0.83 μH
10	2.875	0.60 μH

A suitable coil for the L-network inductor consists of two inches (5cm) of Air-Dux 1606T [6 turns-per-inch, (2.5cm), no. 14 wire, 2″ (2.5cm) diameter]. This inductor should be placed at right angles to the main pi inductor to avoid mutual inductance. In the following chart, 7.125 would be slightly over 7 full turns, 2.875 is slightly less than 3 full turns.

table **5.** Comparisons between a **1-kW** transmitter, a **2-kW** transmitter, **and** a **2-kW** transmitter operated at **1-kW** input. **A** frequency of **4 MHz** was used, but other frequencies from 3 to 30 MHz should produce comparable results. (These calculations are computer derived for comparative **purposes** and only approximate actual operating conditions.)

	1 kW	2 kW	1 kW on 2-kW transmitter	
Plate voltage	2800	2800	2800	volts
Plate current	357	714	357	mA
Plate load impedance	5000	2500	5000	ohms
Power input	1000	2000	1000	watts
Tube output (typical)	700	1400	700	watts
Power at antenna	672	1343	647	watts
Transmitter efficiency	67.2	67.1	64.7	per cent
Network efficiency	96.0	95.9	92.5	per cent
Lost in L1 (heat)	27.9	56.9	52.6	watts
Circuit Q	12	12	23.6	
Inductor Q (typical)	350	350	350	
Frequency	4.0	4.0	4.0	MHz
Antenna load	50	50	50	ohms
C1 tuning capacitor	95.5	191.0	187.8	pF
L1 inductor	17.38	9.18	9.18	μH
C2 loading capacitor	533.8	1096.9	1703.0	pF
C1 reactance	416.7	208.3	211.9	ohms
L1 reactance	436.9	230.7	230.7	ohms
C2 reactance	74.5	36.3	23.4	ohms
Current in C1	4.49	8.98	8.83	amps
Current in L1	4.73	9.29	8.93	amps
Current in C2	2.46	7.14	7.70	amps
Voltage across C1	2645.8	2645.8	2645.8	peak volts
Voltage across C2	259.2	366.5	254.4	peak volts
Voltage on antenna	183.3	259.1	179.9	rms volts
Current in antenna	3.67	5.18	3.60	amps

One other circuit trick which can be used quite successfully is to use a dual-section variable, placing the two sections in series rather than parallel. This reduces the minimum capacitance to 10 pF or less.

broadband power amplifier

Many operators need special frequencies outside the five amateur bands for MARS or other purposes, and need a power amplifier which can be tuned up at any frequency in the range from 3.0 to 30 MHz. Table **7** shows a pi-network design that gives continuous frequency coverage in five switch positions. A pi-L network for similar use is shown in table **8.** The pi-L is more broadband for a given *Q* variation, and requires substantially less output capacitance. Both designs are for 2000-watts input with a 2800-volt plate supply, or 1000-watts input at 2000 volts.

component ratings

To determine the peak voltage across C1 you can use the maximum dc plate voltage. This is not precisely correct, but it's close enough. Normally you would increase the voltage by at least 30 per cent when selecting a capacitor to prevent arcing if the tank circuit is not prefectly resonated, and to allow for some oxidation if you use an air variable.

There are several ways to determine peak voltage. If the power output is known at this point you can use eq. **12** to determine peak voltage:

$$E_{pk} = \sqrt{2PZ} \qquad \textbf{(12)}$$

where E_{pk} is the peak rf voltage, P, is output power, and Z is plate load impedance. For example, in a 1-kW transmitter with 2800 volts on the plate, the peak voltage across C1 and L1 is 2646 volts. (The power output of class-B stages may be estimated at 70 per cent of the input power as this gives some margin of protection and is suitable for this purpose).

The peak voltage across C2 can also be figured in a similar manner, except that Z in eq. 12 is the antenna load impedance. Power output may be estimated at 65 per cent of the input. For example, if the output power is 650 watts (for a 1-kW amplifier), and the antenna load is 50 ohms, this represents approximately 254 peak volts across C2. Thus, for a 1000-watt transmitter, a 350-volt, 365-pF broadcast receiver type capacitor could be used successfully. For 2000 watts input at 2800 volts, the peak voltage across C2 would be 367 volts, and the broadcast-tuning capacitor would be too marginal.

In the pi-L network the image impedance must be used when calculating the peak voltage across capacitor C2, and the voltage rating must be substantially higher than for the same capacitor in the pi network. For example, in a 1-kW transmitter, the peak voltage across C2 is about 635 volts; for a 2000-watt amplifier the peak voltage is about 895 volts.

The peak voltage across C1 has already been determined, but to find the current through C1, rms voltage is more useful. This can be found from eq. 13:

$$E_{rms} = \sqrt{PZ} \qquad \textbf{(13)}$$

where E_{rms} is the rms voltage, P is the output power, and Z is the plate load impedance. In the previous example of the 1000-watt transmitter with a 2800-volt plate supply, the rms voltage across C1 is nearly 1870 volts.

To calculate the current through C1 you must first determine the reactance of C1 (**eq.** 14) and calculate its impedance (**eq.** 15). The current is found from eq. 16:

$$X_c = \frac{Z_p}{Q} \qquad \textbf{(14)}$$

$$Z_{C1} = \sqrt{R^2 + X_C^2} \qquad \textbf{(15)}$$

$$I = \frac{E_{rms}}{Z_{C1}} \qquad \textbf{(16)}$$

However, since the resistance of a high-quality air-variable capacitor is very small, less than 1 ohm, for all practical purposes the impedance of the capacitor is equal to its reactance. Therefore, the current can be found from

$$I = \frac{E_{rms}}{X_{C1}} \tag{17}$$

As you can see in **table 5,** the current through C1 is much higher than you might think, with nearly 4.5 amperes flowing through C1 in the 1000-watt transmitter with 2800 volts on the plate. Most air variables and vacuum capacitors can handle this current easily, but you must be careful when selecting fixed capacitors to pad the variables. Transmitting-type capacitors with high Q and good current-carrying capability are required (such as the Centralab 850 series).

The current through C2 can also be determined with **eq. 17.** However, when calculating the rms voltage across C2 the antenna load impedance must be used in **eq. 13.** Again, there is substantial current flowing through C2 — nearly 2.5 amperes in the 1000-watt transmitter.

For all practical purposes, the current through inductor L1 is equal to that through C1. It is actually a little higher, and the following formula is reasonably correct for class B:

$$I_{cc} = 1.05\, Q_o I_p \tag{18}$$

where I_{cc} is the circulating current, Q_o is loaded circuit O, and I_p is the indicated plate current. **Eq. 18** is a close approximation that compares favorably with answers derived from using complex vector analysis of reactive components used in rf circuits at resonance.

inductor power loss

To determine heat losses in the inductor, it is necessary to know the rf resistance of the inductor.

table 6. Large value at C1 and smaller inductor cause the *Q* on ten meters to rise very rapidly, especially when running the transmitter at a lower power input which requires 5000 ohms plate load impedance.

f (MHz)	R1	C1 (pF)	L1 (μH)	C2 (pF)	Q
29.7	2500	26	1.24	148	12.0
29.7	2500	32	1.00	210	15.0
29.7	2500	39	0.84	251	18.0
29.7	2500	45	0.72	300	21.0
29.7	2500	51	0.63	348	24.0
29.7	5000	26	1.24	234	24.0
29.7	5000	32	0.98	303	30.0
29.7	5000	45	0.70	437	42.0
29.7	5000	51	0.61	503	48.0

table 7. Pi-network component values for a broadband 3-30 MHz rf power amplifier matching a 50-ohm antenna load. This is accomplished in five bands: 3.0-5.0 MHz, 5.0-8.5 MHz, 8.5-14.4 MHz, 13.5-22.0 MHz and 20.0-30.0 MHz. The Q is set for a minimum of 12 at the top of each band. The 2500-ohm plate load impedance corresponds to a grounded-grid amplifier running 2000 watts at 2800 volts, or a 1000-watt amplifier with 2000 volts on the plate.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R? OHMS	'Q' QUAL.
2500	3.q	433	7.34	2878	50	20.4
2500	3.5	317	7.34	2053	5n	17.4
2500	4.q	242	7.34	1517	50	15.2
2500	5.0	153	7.34	875	50	12.0
2500	5.0	265	4.32	1764	50	20.8
2500	7.0	134	4.32	834	50	14.7
2500	7.3	123	4.32	755	50	14.1
2500	8.5	90	4.32	516	50	12.0
2500	8.5	155	2.55	1034	50	20.8
2500	14.0	56	2.55	327	50	12.4
2500	14.35	54	2.55	308	50	12.1
2500	14.4	53	2.55	305	50	12.0
2500	13.5	94	1.67	621	50	19.9
2500	21.0	38	1.67	225	50	12.6
2500	21.45	37	1.67	212	50	12.3
2500	22.0	35	1.67	199	50	12.0
2500	20.0	59	1.22	383	50	18.4
2500	28.0	30	1.22	176	50	13.0
2500	29.7	26	1.22	151	50	12.2
2500	30.0	25	1.22	146	50	12.0

Then you can use **eq. 19** to find power loss.

$$P = I^2 r \tag{19}$$

where I is the circulating current and r is the rf resistance.

To minimize these losses, the inductor should be silver plated, as should all leads to the bandswitch. Power losses on the order of 30 to 100 watts are not unusual, even with low standing-wave ratios. The use of tubing is encouraged, particularly on the higher frequencies to provide better unloaded O.

It may come as a surprise to find that the conductivity of silver is only slightly superior to that of copper. In fact, a silver-plated coil is little more efficient than a new tank coil made of copper. Copper, however, oxidizes, and the outer rf-current-carrying layer becomes less effective. On the other hand, silver develops a form of silver sulfide on its outer surface which barely affects its conductivity. Over a period of years the silver-plated coil will retain most of its original conductivity.

safety

An rf choke should be used at the antenna output of any pi or pi-L network. This choke should be large enough to blow the overload relay (or fuse) in the high-voltage power supply if the dc blocking capacitor should short out. This is the only backup protection you have to keep high dc voltage off the pi-network components if the blocking capacitor

table 8. Pi-L network component values for a broadband 3-30 MHz rf power amplifier matching a 50-ohm antenna load. This is accomplished in five bands: 3.0-5.0 MHz, 5.0-8.5 MHz, 8.5-14.4 MHz, 13.5-22.0 MHz and 20.0-30.0 MHz. The Q is set for a minimum of 12 at the top of each band. The 2500-ohm plate load impedance corresponds to a grounded-grid amplifier running 2000 watts at 2800 volts, or a 1000-watt amplifier with 2000 volts on the plate.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	L2 UH	R2 OHMS	R3 OHMS	'Q' QUAL
2500	3.0	388	9.00	2004	3.90	158	50	18.3
2500	3.5	292	9.00	1409	3.90	197	50	16.0
2500	4.0	228	9.00	1042	3.90	242	50	14.4
2500	5.0	153	9.00	622	3.90	350	50	12.0
2500	5.0	237	5.29	1231	2.29	154	50	18.6
2500	7.0	127	5.29	572	2.29	253	40	14.0
2500	7.3	118	5.29	519	2.29	270	50	13.5
2500	8.5	90	5.29	366	2.29	350	50	12.0
2500	8.5	139	3.12	724	1.35	154	50	18.6
2500	14.0	56	3.12	232	1.35	330	50	12.3
2500	14.35	54	3.12	219	1.35	345	50	12.1
2500	14.4	52	3.12	216	1.35	350	50	12.0
2500	13.5	85	2.04	431	0.89	165	50	18.0
2500	21.0	38	2.04	158	0.89	325	50	12.5
2500	21.45	36	2.04	150	0.89	335	50	12.3
2500	22.0	35	2.04	141	0.89	350	50	12.0
2500	20.0	53	1.50	263	0.65	183	50	16.7
2500	28.0	29	1.50	121	0.65	315	50	12.7
2500	29.7	26	1.50	106	0.65	345	50	12.1
2500	30.0	25	1.50	104	0.65	350	50	12.0

A suitable inductor for the L-section of the pi-L network consists of 5 cm (2 inches) of Air-Dux 1606T [6 turns-per-inch (2.5cm), no. 14, 5 cm (2 inches) diameter]. It should be placed at right angles to the main pi inductor to avoid mutual inductance.

frequency	number turns	approximate inductance
3.0-5.0 MHz	10.00	3.90 pH
5.0-8.5 MHz	6.75	2.25 pH
8.5-14.4 MHz	4.75	1.33 μH
12.5-22.0 MHz	3.50	0.83 μH
20.0-30.0 MHz	3.00	0.65 μH

shorts out. This rf choke also keeps any dc component off the antenna.

RTTY and ssb

Many amateurs are interested in RTTY as well as CW and ssb. Since RTTY is essentially 100 per cent key-down, it's quite hard on the various components in the transmitter. On ssb, the typical duty factor is 30 to 50 per cent, depending on how much ALC and other compression you use. Typically, however, the *average* circulating current in the network is perhaps one-third of that for key-down operation.

Table 5 shows that 2000-watts key-down gives comparable circulating currents to that of the same transmitter run at 1000-watts key-down with the same plate voltage and same inductor. This is due to the higher Q that is being used. Because of the lower duty cycle of ssb, running a 2000-watt transmitter key-down at 1000 watts for RTTY is three times as hard on the transmitter as running 2000-watts PEP! This is rather startling, and indicates why some rf power amplifiers should not be used on RTTY, although they are perfectly suitable for ssb at hrgher input power levels.

Conversely, it follows that if a manufacturer guarantees his unit to run indefinitely at 1000-watts key-down RTTY, that same transmitter should last forever at 2000-watts PEP ssb. Some manufacturers hedge if this specific question is posed to them.

Using a 2000-watt rf power amplifier at the 1000-watt level for RTTY or CW poses certain inherent problems regarding heat and efficiency. High plate supply voltages raise the plate load impedance to the point where it may be difficult to get the minimum capacitance required for resonance on 10 and 15 meters.

When building a high-power final amplifier, consideration must be given to selecting components which will handle the voltage and currents encountered in the circuit. The formulas given in this article should make it relatively easy for the builder to predict what these voltages and circulating currents will be before he actually builds the amplifier.

Computer-derived tables provide much data for the builder, and clarify many design points only hinted at in previous articles. I hope that the information presented here will be of benefit to anyone who builds or buys a final rf power am'plifier.

acknowledgements

Many people are interested in pi and pi-L networks, and have been of direct assistance. Providing particular assistance was Bob Sutherland, WGUOV, or EIMAC. I also spent a great deal of time reading articles written by George Grammer, W1DF, former Technical Editor of QST. His work in this field, and his series of three articles in QST[5] represent an outstanding contribution. Bill Craig, WB4FPK, was most helpful, as was Garey Barrell, K4OAH. Bill Carver, KGOLG, also provided stimulating comments.

The Computer Terminal Corporation of San Antonio, Texas, provided over 100 hours of computer terminal time which was invaluable in this project.

references

1. Collins Radio Company, "Fundamentals of Single Sideband," second edition, 1965, page 7-6.
2. George Grammer, W1DF, "Simplified Design of Impedance-Matching Networks, Part II," *QST,* April, 1957, page 32.
3. American Radio Relay League, "Radio Amateur's Handbook," 49th edition, 1971, page 158.
4. Douglas Blakeslee, W1KLK, and Carl Smith, W1ETU, "Some Notes on the Design and Construction of Grounded-Grid Linear Amplifiers," *QST,* December, 1970, page 22.
5. George Grammer, "Simplified Design of Impedance-Matching Networks," Parts I and III, *QST,* March, 1957, page 38, and May, 1957, page 29.

ham radio

If you were one of the fortunate 29,069 hams who worked Clipperton Island in March of 1978, you've worked an Atlas 350-XL transceiver. The 350-XL was selected by the DXpedition logistics team, headed by Don Bostrom, N6IC, because it had all of the necessary features required for the operation contained in one compact package. This included primary and auxiliary VFO's for split frequency operation, digital frequency display with accuracy of ± 50 Hz, VOX for SSB and full break-in for CW, sidetone, more than 200 watts output (twice that of most other transceivers), all solid state design permitting efficient operation from a storage battery if necessary. And above all, rugged design and construction that permits hour after hour of continuous operation without failure.

> ***"The 350-XL is a fine, rugged transceiver. . . even works after a salt water bath. . ." Willy, HB9AHL***

One very important point we want to make clear ... the Clipperton DXpedition was financed by the 16 operators who went there, and by many generous donations from DX clubs, radio clubs, individual hams, and others. Atlas Radio was not a financial sponsor, except to the extent of loaning equipment. Other manufacturers provided similar support.

> ***"As equipment logistics manager, my selection of the Atlas 350-XL proved to be the perfect choice . . ." Don, N6IC***

Needless to say, we at Atlas Radio were very pleased when the team chose the 350-XL as the transceiver for all 3 stations. At that point, how could I (W6QKI) turn down the invitation to join the team, and to share in a tremendous adventure? Did I go along to keep our radios working? Well, truthfully I brought along a box full of spare parts and pieces. Happily I can report that the box could have stayed at home. And there are 15 witnesses who will verify this. Their unanimous and whole hearted endorsement of the 350-XL is most gratifying.

Many of you will be interested in how the 3 stations were organized. Number 1 station was set up in the metal Quonset-type building which the French put up in 1957 during the IGY scientific work conducted on the island. This station worked strictly 20 meters 'round the clock for 7 days, SSB and CW. It included a Dentron MLA-2500 Linear which was used much of the time to break through to Europe and other distant points. The antenna was a Wilson 4 element monobander about 30 feet high. Power was supplied by a 2500 watt Honda gasoline generator. This station ran continuously for 7 days, and made 11,158 contacts! Problems, zero!

> ***"Unbelievable performance and reliability under extremely adverse portable conditions and constant use by DXpedition multi-operators . . ." Hugh, WA4WME***

Incidentally, we took one box ashore which contained 3 fans. They were intended for blowing air on the transceiver heat sinks. The box is still on the island, unopened! Ambient temperature outside was 85 degrees F. Inside the metal building? Up to 95 degrees!

Station Number 2 was located about 200 feet from Number 1, and was set up in a tent. It worked 10 meters daytime, 80 and 160 meters at night. A Dentron MLA-2500 Linear was used, mainly on 80 and 160, some of the time on 10 meters. A 3 element Wilson monobander was used on 10. A doublet was used on 80 meters, later changed to a Delta loop by F6ARC, a KLM vertical with ground radials worked very well on 160 meters. A Dentron MT.3000 antenna tuner was used on 80 and 160. Power was supplied by a Sears 2200 watt generator. This station averaged 21 to 22 hours operation each of the 7 days. Problems? The digital frequency display made signs of acting up. One of the IC's was replaced. A 5 minute job. The rig had been liberally sprayed with salt water on the trip in through the surf, as also was the Dentron linear. Total contacts from station Number 2 were 6401 on 10 meters, 1644 on 80 meters, and 202 on 160 meters.

> ***"Clipperton: The best location for DXers. Atlas 350-XL: The best equipment for hardest DXpedition. Result: One of the best DXpeditions ever. . ." Jack, F5II/FO∅XB***

Station Number 3 was located in a tent about 300 feet (and 5000 crabs) from Number 2. It operated on 15 and 40 meters. Foreign broadcast QRM was very rough on 40, so most operating time was on 15 using a Wilson 3 element monobander. No linear was used at this station because the generator would not provide enough power.

So, if you heard Clipperton on 15 or 40 meters, it was strictly barefoot. A Dentron MT-3000 tuner was used with a KLM vertical on 40 meters. Station Number 3 ran all week on a generator that delivered 155 volts AC when receiving . . . and only 75 to 90 volts during transmit! We were unable to adjust the problem, so simply let it go. Didn't bother the rig. Total contacts on 15 meters numbered 7194, second only to 20 meters! 40 meters netted 2450 contacts.

This report hardly is complete if we don't mention 6 meters and Oscar. N6IC and W6SO were the Oscar specialists. Unfortunately, some equipment difficulty (not Atlas) limited Oscar contacts to only 20. Rather disappointing, but the best we could do, and the guys really tried. 6 meters just never produced an opening. We monitored everyday without ever hearing a signal.

> ***"I cannot say enough about the excellent performance of the Atlas equipment. Under the most trying conditions of operation the gear came through with flying colors. With 16 operators pushing switches and twisting knobs 24 hours a day for 7 days, the equipment never faltered. Truly remarkable. The success of the DXpedition was due in large to the faultless operation of the 350-XL . . ." Hoppy, W6SO***

All in all, we feel the performance record on the HF bands is something to brag about, and hope you'll pardon us for indulging. One final thing to boast about was really unexpected. The ride through the surf back to the ship was quite a ride. Everyone, and everything thoroughly soaked. Much of the gear was full submerged. But all 3 of the 350-XL's worked normally after drying out! Being very low on fresh water we could not afford to wash the gear down. All we could do was dry them out in the sun. Obviously, as soon as we got back we had to wash out the salt and clean the sets up. But, they were used "maritime-mobile" on the trip back to San Diego.

The Clipperton '78 DXpedition was undoubtedly the biggest expedition and adventure of its kind ever put together, and turned out to be a smashing success in all respects. All the gang at Atlas is mighty proud at how well the 350-XL proved itself, truly a great performer; a real classic that will set the pace for years to come.

73 Herb Johnson W6QKI

RADIO INC.
417 Via Del Monte, Oceanside, CA 92054
(714) 433-1983
Special Customer Service Direct Line (714) 433-9591

programming for automated satellite communication

Following the lead of Ball in February ham *radio*, author Milazzo has also derived the equations for tracking Oscar — emphasis in this case has been placed on developing a program suitable for use with the Texas Instruments calculators which use algebraic notation

Of primary importance in satellite communication is the required antenna orientation and the time during which the satellite is available from the ground station. Such information increases the dependability and efficiency of satellite use. Presently, the means of obtaining this data is widely available due to the increased popularity of low-cost programmable calculators and minicomputers.

Previous articles have dealt with the prediction of equatorial crossings.[1,2] These offer few advantages since such information is published monthly and interpolation of this data is easily accomplished. Other articles do offer more useful information but require the use of slide-rule type devices which are imprecise and incapable of being interfaced with station equipment.[3,4]

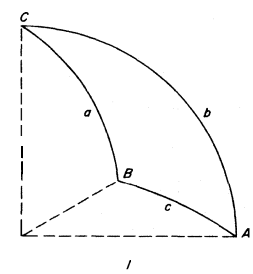


fig. 1. Basic figure used for the derivation of the main equations to track the satellite; the law of cosines for spherical trigonometry applies to this figure.

This article presents a series of equations for deter-

By Charles F. Milazzo, KP4MD, University of Puerto Rico School of Medicine, Apartado 2-2, GPO Box 5067, San Juan, Puerto Rico 00936.

mining the exact position of any earth satellite that approximates a circular orbit, at any given time. The calculations are useful for manually or automatically tracking a satellite, for preparing tables for future reference, or for alerting the operator when a satellite is approaching. This algorithm has been used to prepare programs for the Texas Instruments SR-52 calculator, but can be programmed for other calculators and computers as well.

theory

The theoretical model considers the earth as a stationary sphere with the satellite travelling in a circular orbit moving from east to west. Thus, the coordinates of the ground station remain constant while those of the satellite vary as a function of time. Solving this problem involves the application of spherical trigonometry which implies that the computer must be supplied subroutines for trigonometric functions. Algebraic methods for evaluating these functions are found in various references.[5,6]

The main equations are derived from the law of cosines of spherical trigonometry (see **fig. 1**).

$$\cos a = \cos b \cos c + \sin b \sin c \cos A$$

To calculate the latitude and longitude of the satellite, the angle labelled A is placed at the equator crossing point of a reference orbit, angle B is at the satellite's position at time T, and angle C is at the North Pole. Thus, A is equal to 360° minus the orbital azimuth during equator crossing ($A = 360° - \alpha_{eqx}$), and C is equal to longitudinal displacement of the satellite from the reference crossing ($C = \lambda_s - \lambda_{eqx}$). Side a is the complement of the satellite's latitude ($a = 90° - \psi_s$), side b is a function of the orbital period P_s and of the time elapsed from the reference crossing [$b = 360°(T - T_{eqx})/P_s$], and side c is 90°.

Substituting known values into the equation for the law of cosines results in:

$$\cos(90° - \psi_s) = \cos[360°(T - T_{eqx})/P_s] \cos 90° + \sin[360°(T - T_{eqx})/P_s] \sin 90° \cos(360° - \alpha_{eqx})$$

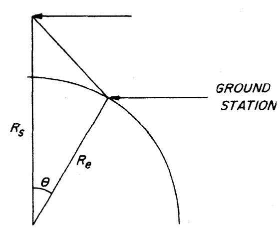


fig. 2. Calculation of the antenna elevation is based on this plane geometric figure. R_s is the orbital radius, referenced to the center of the earth.

which simplifies and rearranges to give the satellite latitude

$$\psi_s = \arcsin \{\sin[360°(T - T_{eqx})/P_s] \cos \alpha_{eqx}\}$$

The result is used again in the same equation to calculate angle C

$$\cos[360°(T - T_{eqx})/P_s] = \cos(90° - \psi_s) \cos 90° + \sin(90° - \psi_s) \sin 90° \cos C$$

which rearranges to give

$$C = \arccos\{\cos[360°(T - T_{eqx})/P_s] \cos\psi_s\}$$

The true longitude of the satellite is calculated by adding the longitude of the equator crossing reference point (λ_{eqx}), changing the sign of C when the satellite is in the Southern Hemisphere, adding the displacement due to the earth's rotation (0.25°/min), and compensating for a constant orbital drift factor (D) if necessary.

$$\lambda_s = (|\psi_s|/\psi_s) \arccos \left\{ \frac{\cos[360°(T - T_{eqx})/P_s]}{\cos\psi_s} \right\} + \lambda_{eqx} + (T - T_{eqx})(0.25 + D)$$

Again referring to **fig. 1,** angle A is now placed at the location of the ground station and is equal to 360° minus the correct antenna azimuth ($A = 360° - \alpha_{gs}$). Angle C is the difference between the longitudes of the satellite and the ground station ($C = \lambda_s - \lambda_g$). If C is greater than zero, then the true azimuth is 360° minus α_{gs}. Side b is the complement of the ground station's latitude ($b = 90° - \psi_g$), and side c is the distance from the ground station to the satellite in great circle degrees ($c = \theta$). Side c is calculated using the law of cosines.

$$\cos \ell = \cos(90° - \psi_s) \cos(90° - \psi_g) + \sin(90° - \psi_s) \sin(90° - \psi_g) \cos(\lambda_s - \lambda_g)$$

which simplifies to give the degrees distance

$$\theta = \arccos[\sin \psi_s \sin\psi_g + \cos\psi_s \cos\psi_g \cos(\lambda_s - \lambda_g)]$$

This value is used to calculate the azimuth

$$\cos(90° - \psi_s) = \cos(90° - \psi_g) \cos\theta + \sin(90° - \psi_g) \sin\theta \cos \alpha_{gs}$$

which rearranges to give

$$\alpha_{gs} = \arccos[(\sin\psi_s - \sin\psi_g \cos\theta)/\cos\psi_g \sin\theta]$$

The antenna elevation is calculated using a plane geometric model as shown in **fig.** 2. R_e is the earth's radius (6371.315km) and R_s is the orbital radius. According to Newton's law of gravitation, the radius of a circular orbit is a function of the orbital period. For any satellite in such an orbit, the centrifugal force ($F = mv^2/R_s$) is equal to the gravitational force ($F = GM_e m/R_s^2$). The velocity of the satellite is equal to the orbital circumference divided by the orbital

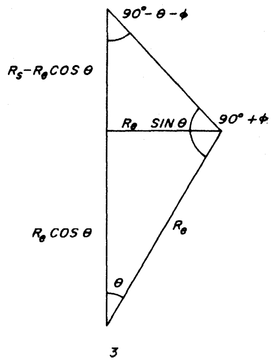


fig. 3. **By drawing a perpendicular line from the ground station to the orbital radius line, all angles and distances can be analyzed using basic trigonometric functions. As shown in the text, the actual antenna elevation angle is derived by this method.**

period in seconds ($v = 2\pi R_s/60P_s$), when P_s is in minutes. GM_e is the earth's gravitational constant ($398603 km^3/s^2$). Thus, we can calculate the orbital radius by equating the centrifugal and gravitational forces

$$m(2\pi R_s/60P_s)^2/R_s = GM_e m/R_s^2$$

$$R_s = \sqrt[3]{GM_e(30P_s/\pi)^2}$$

Fig. 3 shows the trigonometric relationships which result when a perpendicular is drawn from the ground station to the R_s line. Solving for the angle whose apex is at the satellite's position yields

$$90^\circ - \theta - \phi = arctan\ [R_e sin\ \theta/(R_s - R_e\ cost))]$$

which rearranges to give the antenna elevation

$$\phi = arctan\ [(R_s - R_e\ cos\theta)/R_e\ sin\theta] - \theta$$

If the satellite is below the horizon, the antenna elevation will be a negative number. This is useful as a conditional test to determine if the satellite is within range.

Ground distance is directly proportional to the arc distance and can be found by $D = k\theta$, where k is 111.14 for kilometers, or 69.06 for statute miles. Direct line-of-sight distance can also be determined by the law of cosines **(fig. 4)**.

$$D = \sqrt{R_s^2 + R_e^2 - 2R_s R_e\ cos\theta}$$

The constants for the Oscar 6 and 7 satellites are as follows:

	α_{eqx}	P_s	D
Oscar 8	**351.0100°**	**103.23162 min**	**0**
Oscar 7	**348.2990°**	**114.94483 min**	**0**

the SR-52 program

Due to the complexity of the calculations required, the program is divided into two modules.* The first module enters the appropriate reference and constant information into the data memory, while the second module computes the satellite's position and direction in terms of azimuth and elevation from a specified ground location. In the second module, the desired time is entered from which the calculator displays the elevation angle, indicating if the satellite is within range. The azimuth and arc distance can then be called from the calculator. To facilitate the tabulation of the results, one key has been programmed to advance or reverse the position of the satellite by a desired number of minutes. The following example demonstrates the use of the program.

The antenna aiming data for the first Oscar 7 pass of August 1, 1977 is desired. The first program card is read into the calculator, and the R/D switch is placed in the degrees mode. A reference orbit from January 1, 1977 is available. The satellite is found to cross the equator at 77.0° west longitude at 0148:49 on January 1. The station coordinates are 18°25′ north latitude and 65°58′ west longitude. Therefore, the following key sequence is executed:

enter	press	display	
7	A	7	(satellite #)
1	B'	2400	(enter date)
148.49	B	1548.82	
77.0	C	77.	
18.25	D	18.42	
65.58	E	65.97	

The second card is now read into the calculator. Since the exact time of acquisition is not known, a rough estimate can be made based on the fact that the satellite travels at about 3 degrees per minute and

*A copy of the program is available by sending a self-addressed, stamped-envelope to *ham radio.* Greenville, New Hampshire 03048.

it must be within 36 degrees of the station to be heard. First, determine if the satellite is within range at 0000 GMT on August 1 (213th day of 1977).

enter	press	display	
213	A'	511200	(enter date)
0	A	- 9.23	(enter time/ disp. elev.)
	C	45.8	(disp. distance)

The negative value for elevation showed that the satellite is below the horizon at 0000 GMT, while pressing C showed that it is about 10 degrees beyond the 36 degree range limit. If the satellite is approaching, it will take about four minutes to come within range. Advancing the satellite four minutes produces:

enter	press	display	
4	E	2.25	(elevation)
	B	147.8	(azimuth)
	C	33.4	(distance)

The satellite can be advanced at one minute intervals to produce a listing for this pass.

time	elevation	azimuth	distance
0004	2.2	147.8	33.4
0005	5.7	146.2	30.3
0006	9.4	144.3	27.2
0007	13.5	142.0	24.2
0008	18.1	139.0	21.2
0009	23.4	135.1	18.3
0010	29.3	129.8	15.5
0011	35.9	122.4	12.9
0012	42.7	111.5	10.6
0013	48.7	95.5	8.8
0014	51.6	74.3	8.0
0015	50.1	52.1	8.5
0016	44.8	34.3	9.9
0017	38.1	22.0	12.1
0018	31.3	13.6	14.6
0019	25.2	7.8	17.4
0020	19.7	3.7	20.3
0021	14.9	0.6	23.3
0022	10.6	358.1	26.3
0023	6.7	356.2	29.4
0024	3.2	354.7	32.4
0025	0.0	353.5	35.6

Results from this program have been compared with published equatorial crossing data, with the predicted coordinates accurate to within one tenth of one degree for both Oscar 6 and Oscar 7. This, for one year from a single reference orbit.

The user instructions for the program are straightforward, but the following hints are useful. A date need not be entered if the reference and unknown orbits occur within the same day. When entering a date, it must always be entered before the time. The dates of the reference and unknown orbits must be either within the same month, or else each day of the year must be assigned a consecutive number. The

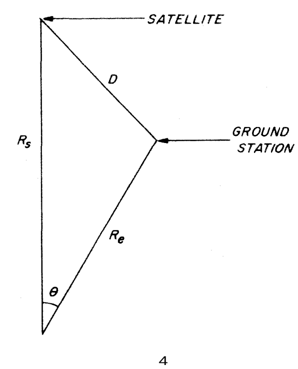


fig. 4. The direct line-of-sight distance to the satellite is calculated using the law of cosines and this figure.

"+ Time" key may be used to reverse the satellite's position by merely entering the negative value of the desired number of minutes. Finally, if you desire to enter a new time without using the "+ Time" key, it is necessary to re-enter the date (if used) before entering the new time. I can supply the program on magnetic cards with any station coordinates and reference orbit recorded for the cost of $5. This includes program documentation.

conclusion

By using the preceding equations, any person can build his own computerized satellite tracking station by interfacing a digital clock and station controls to a microcomputer. Even without a computer, this program offers valuable, accurate information for any satellite operator.

references

1. Joe Kasser. G2ZCZ, "Oscar Orbits on your Altair," 73, November, 1976, page 146.
2. Loring C. White, W1ODI, "Satellite Orbit Predicting," 73, August, 1976, page 113.
3. Kazimierz Deskur, K2ZRO, "Shoot Oscar with a Satellabe," 73, July, 1975, page 33.
4. Charles J. Harris, WB2CHO and Joel P. Kleinman, "The Oscarlocator," QST, November, 1976, page 51.
5. R. S. Burlington. *Handbook of Mathematical Tables and Formulas,* McGraw-Hill, New York, 1965. page 56.
6. R. C. Weast, editor, CRC *Handbook of Tables for Mathematics,* Chemical Rubber Company, Cleveland, 1967, page 581.

ham radio

protecting solid-state devices from voltage transients

There has been a great deal of discussion lately about solid-state equipment failing "for no apparent reason." If you have experienced such an occurrence and you can't explain it, chances are your solid-state gear has been zapped by a voltage transient. These are the little "gremlins" that can appear at any time, unannounced, and proceed to do their dirty work on your favorite electronic device (even when it is turned off).

While this problem may appear at first to be a recent phenomenon, it may surprise many of you to learn that voltage transients have been around for a long, long time. Why, then, have we begun to experience solid-state equipment failures (traceable to transients) only in the last few years?

The key to answering this question begins with an examination of the way equipment is built today as compared to 10 or 15 years ago. The one major and obvious difference is the use of solid-state devices in place of vacuum tubes. While transients have been around for a long time, they never bothered vacuum tubes, which can withstand momentary abuses without damage. Consequently, we have never paid much attention to the elusive voltage transient. To summarize our present situation, we know that:

1. High energy voltage transients can destroy solid-state devices.

2. We must design circuits more conservatively or add protective devices to suppress the transients.

3. We must try to eliminate some of the man-made causes of voltage transients.

In this article, I hope to provide you with some information on what voltage transients are, what causes them, why protection is necessary, and how to detect and suppress transients.

what are voltage transients?

To begin with, voltage transients are generally considered to be abnormally high voltages, occurring in a place where they don't belong, for an extremely short period of time (usually in the nanosecond range). These transients can be generated by a device within the electronic gear (relays or SCRs) or they may come from a number of external sources such as lightning or power line switching.

For example, if an electric current is made to increase or decrease rapidly in an inductive circuit, an extremely high voltage will be generated (directly proportional to the amount of inductance and the rate at which the current is attempting to change). While this all happens very fast, there is a possibility that enough energy will be generated to exceed the breakdown characteristics of the solid-state components in the conduit.

Another problem area exists with transients caused by outside sources that find their way into solid-state equipment through power connections, antenna lead-ins, and other signal inputs to the equipment. This type of transient is much harder to control than the internal transient, and yet can produce just as much damage.

sources of voltage transients

As discussed earlier, transients may be generated under all types of conditions, and may occur on an ac line as well as dc bus. An example of the generation of an internal transient is one that occurs in the windings of the transformer in a power supply. If the line switch on the primary side of the transformer is opened at the exact instant that the transformer core is beginning a downward swing of magnetization current (fig. 1A), a voltage spike is developed on the secondary winding of the transformer. The same thing can happen when the power supply is first turned on (fig. 1B). In this example, we are causing a rapid increase of current in an inductive circuit (the transformer winding) and the result is a momentary spike on the secondary circuit.

While the previous two examples involve ac circuits, the same phenomenon can occur with the interruption of dc circuits that supply inductive loads. A diode connected across the coil of a dc relay is a good example (fig. 1C). When power is interrupted, this circuit is designed to conduct and thereby clip any high-voltage transients that may be developed.

By W. J. Prudhomme, WB5DEP, 6419 Rosalie Court, Metairie, Louisiana 70003

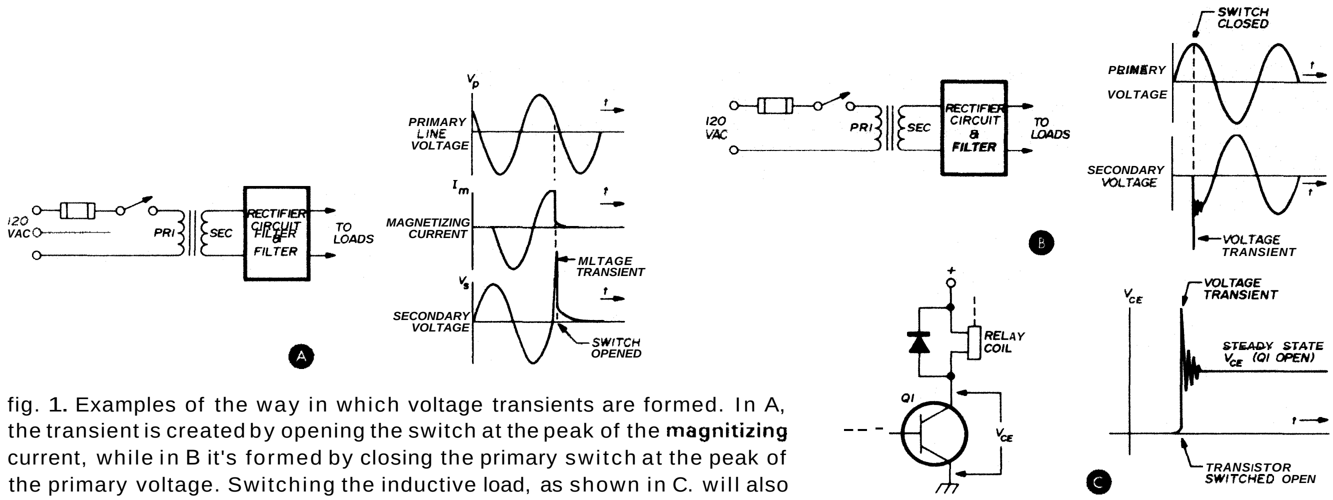
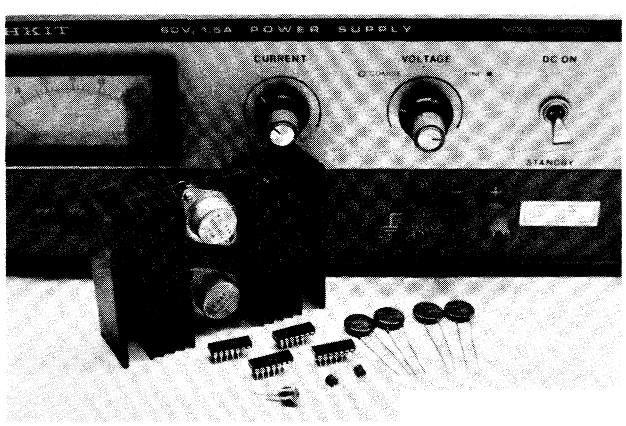


fig. 1. Examples of the way in which voltage transients are formed. In A, the transient is created by opening the switch at the peak of the magnitizing current, while in B it's formed by closing the primary switch at the peak of the primary voltage. Switching the inductive load, as shown in C. will also generate a transient on the V_{cc} line.

As far as outside sources of voltage transients, one of the most common is lightning striking a power line during a storm. Even when the strike is several kilometers away, it can readily travel through the utility's power lines and play havoc with solid-state equipment. Also, lightning in the area can generate voltage spikes on your antenna that easily find their way into the sensitive rf stages of receivers and transceivers. Here again, the actual lightning may be kilometers away and still generate enough transient energy to damage unprotected equipment.

When I was younger and didn't know better, I used to connect an NE-2 neon lamp across the lead-in of an unused antenna to detect these occurences. Even when an electrical storm was far awav (and althouah the NE-2 requires about 90 volts to ignite), it would flash with almost every lightning strike. Naturally, for safety reasons, I don't recommend that anyone try this experiment. However, it should give you some idea of the amount of energy an antenna system can absorb and what can happen if you don't protect sensitive rf circuits.



Transients are capable of destroying solid-state components such as IC's, rectifiers, and transistors (even power transistors usually assumed to be protected by heatsinks). To protect these sensitive components from voltage transients, it's necessary to use some type of energy absorbing device such as the metal oxide varistors shown in the photo. These are the four round shaped components on the right that look like disc capacitors. Commercial solid-state equipment such as the well designed power supply in the background may be protected by installing a varistor on the line side of the power transformer. See text.

Other forms of ac voltage transients may come from arcing contactors, incandescent lighting dimmers, electric drills, appliance motors with brushes, power-line switching, and many more. Also, air conditioning motor-starting contactors can produce damaging transients, as well as relays in the control circuits.

why is protection required?

To further illustrate the fact that sensitive solid-state equipment should be properly protected, refer to a study conducted by the General Electric Company in 1969.[1] In this study, which took two years to complete, surge voltages in both residential and industrial circuits were measured at 400 different locations in twenty cities.

Surprisingly, higher surge levels were recorded in the residential circuits than in the industrial areas. In addition, the results showed a peak as high as 2500 volts generated by a motor contactor within a residence, and nearby lightning generated a 5600-volt peak on a 120-volt residential service. There were also a significant number of surges in excess of 2000 volts occurring in homes on a repetitive basis.

The GE study proved conclusively that residential lighting circuits are subjected to severe transients, as well as any electronic equipment connected to those circuits. As a result, the solid-state equipment in your home is subject to damage from voltage spikes if not properly protected.

suppression of transients

There are several ways to suppress voltage transients, and depending on the circuit application, some devices are better than others. Some of the devices available for you to choose from include metal-oxide varistors, power zeners, and short-circuiting devices such as spark gaps and electronic *crowbar* circuits.

Metal-Oxide Varistors. These dynamic resistance devices feature both low cost and small size, and are capable of dissipating a considerable amount of energy for a short period of time. One source for metal-oxide varistors is the General Electric Company; their registered trade name for these devices is GE-MOV. These varistors are available from most electronics supply houses.

If you are going to use these devices, I would recommend that you obtain a copy of GE's "Transient Voltage Suppression Manual," ($2.50 from GE, Semiconductor Products Department, Electronics Park, Syracuse, New York 13201). This manual is a complete guide to the proper application of varistors.

Since varistors are bi-lateral devices, they may be used in both ac and dc applications. In low-voltage dc applications, however, they may not provide adequate protection due to their soft clamping characteristics. They are generally more suitable for ac line voltages, and consequently are excellent in suppressing transients on the line side of equipment power supplies. An example is shown in **fig. 2.**

Power Zener. While varistors may exhibit soft characteristics at low dc voltages, power zeners can clamp very hard and fast at those voltages. Also, high transient currents do little to raise the clipping

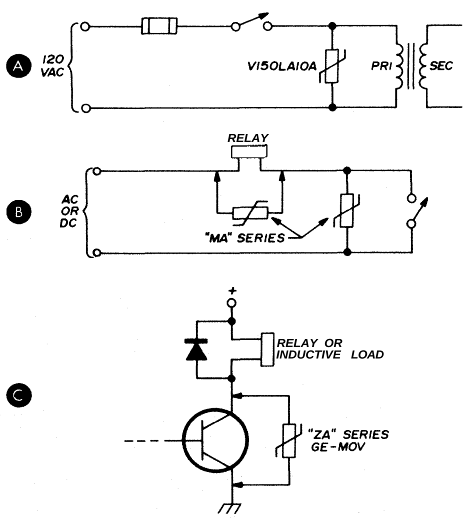
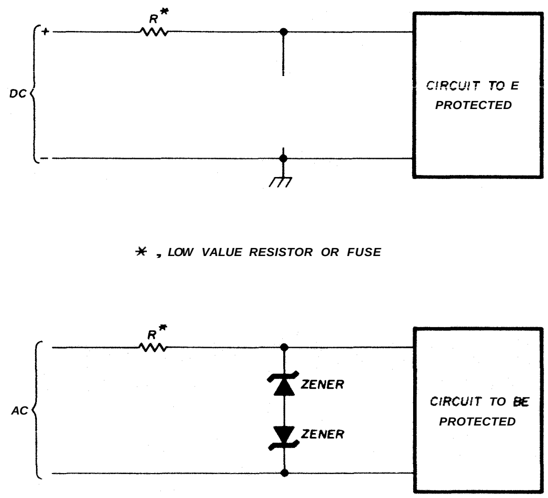


fig. 2. Application of commercial varistors will help eliminate line transients. The L series of GE-MOV varistors can handle up to 2750 amps (transient) and peak voltages from 95 to 1000 volts RMS (A). The MA series varistors are used primarily in low power applications, either ac or dc. In B, they're used to prevent contact arcing and line noise. To prevent transients while switching inductive loads, the ZA series of varistors are connected across the switching transistor as shown in C.



fig, 3. Instead of using commercial varistors, Zener diodes can also be used to provide protection against voltage transients. The two examples show ways in which zeners can protect both ac and dc circuits. For the dc circuit, the zener is rated at a voltage greater than the dc bus, but less than the maximum voltage of the circuit to be protected. In an ac application, they should be rated at slightly greater than the peak ac value.

level of most zeners, provided ratings are not exceeded. The most effective way to apply zeners is with some amount of series resistance to safely limit the current. Two schemes for protecting both ac and dc circuits are shown in **fig.** 3.

Short Circuit Methods. At first, this approach may seem contrary to normal applications. After all, a short circuit in most cases is a very undesirable situation. However, if it is properly controlled and occurs for very short periods of time (microseconds), the end result can be very beneficial in suppressing transients.

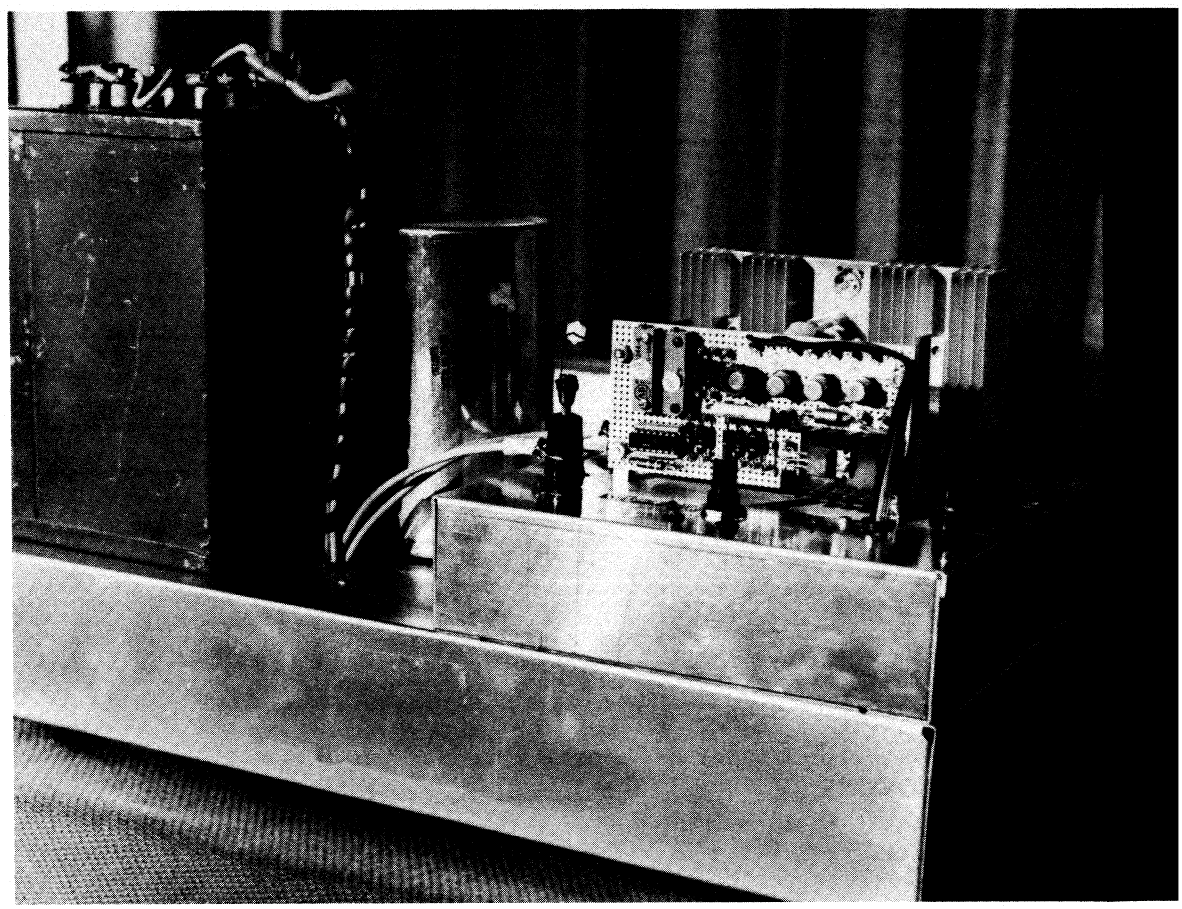
The two most commonly used devices in this category are spark gaps (both open and sealed units) and electronic crowbar circuits. The first is often used by amateurs for antenna applications, while the second is usually too complex for most hobby applications. The spark gap is adjusted to arc at a voltage above a certain level, effectively grounding the transient energy. Which method or combination of methods you choose, will depend on the application and what you are attempting to protect.

The best approach is to buy transient suppressors only in component form and apply them yourself using the guidelines in this article. Only then can you be assured of applying these useful devices for their intended purpose — protecting solid-state devices from voltage transients.

reference

1. IEEE Transactions on Power Apparatus and Systems, Vol. PAS-89, 1970.

ham radio



instantaneous-shutdown high-current regulated power supply

Protect both the new circuit and power supply by using this high-current instantaneously shutdown power supply

Anybody who builds a semiconductor circuit is faced with a potentially disastrous problem — what will happen when power is applied to the circuit? A short circuit could be catastrophic, for both the power supply and the newly-built circuit.

A high-current regulated power supply will continue to supply power, even with a short circuit, until one of the following events takes place.

1. The short circuit clears itself.

2. A semiconductor failure occurs (either the silicon or metal leads melts).

3. The power supply is damaged.

By **Alan Nusbaum, W6GB,** Dalmo Victor Operations, Bell Aerospace Textron, Division of Textron, Inc., 1515 Industrial Way, Belmont, California 94002

4. A fuse is blown, but usually too late to protect the circuit.

5. The power supply contains an extremely fast over-current sensing circuit that will cut off the current in less than 200 nanoseconds.

This article will describe just such an over-current shut-down circuit.

Over the past few years, there have been some new and novel voltage-regulator devices which will tolerate shorts or over-current loads by application of current fold-back circuits or thermal shutdown. Other versions which use a crow-bar device rely on blowing a fuse in the power supply primary. This method, however, is rather brutal treatment, and is often fatal to the pass transistor. I feel that a "graceful" but high-speed shutdown circuit is necessary to protect both the newly developed circuit and the power supply. Essentially, it must shutdown and not

As seen in **fig. 1,** a PNP pass transistor is driven by a complimentary NPN monolithic Darlington. The Darlington in turn is current driven by a medium-gain transistor which acts as a voltage error detector. This form of regulator has proven to be very sensitive and reliable in that it will hold to within 30 mV from no load to a 15 ampere load, with less than 10 mV ripple at full load.

The shutdown portion of the circuit uses an **SCR** to rob the base current from the Darlington pair. When the SCR is triggered from the voltage comparator, the point between the 470-ohm resistor and CR2 is effectively grounded, reducing the output voltage to zero. U1, a μA311 voltage comparator, is used to sense the voltage drop caused by over-current across the two 0.1-ohm resistors. The noninverting input of the comparator is fed from a separately regulated source (U2). The inverting input is the actual output voltage, after it has been appro-

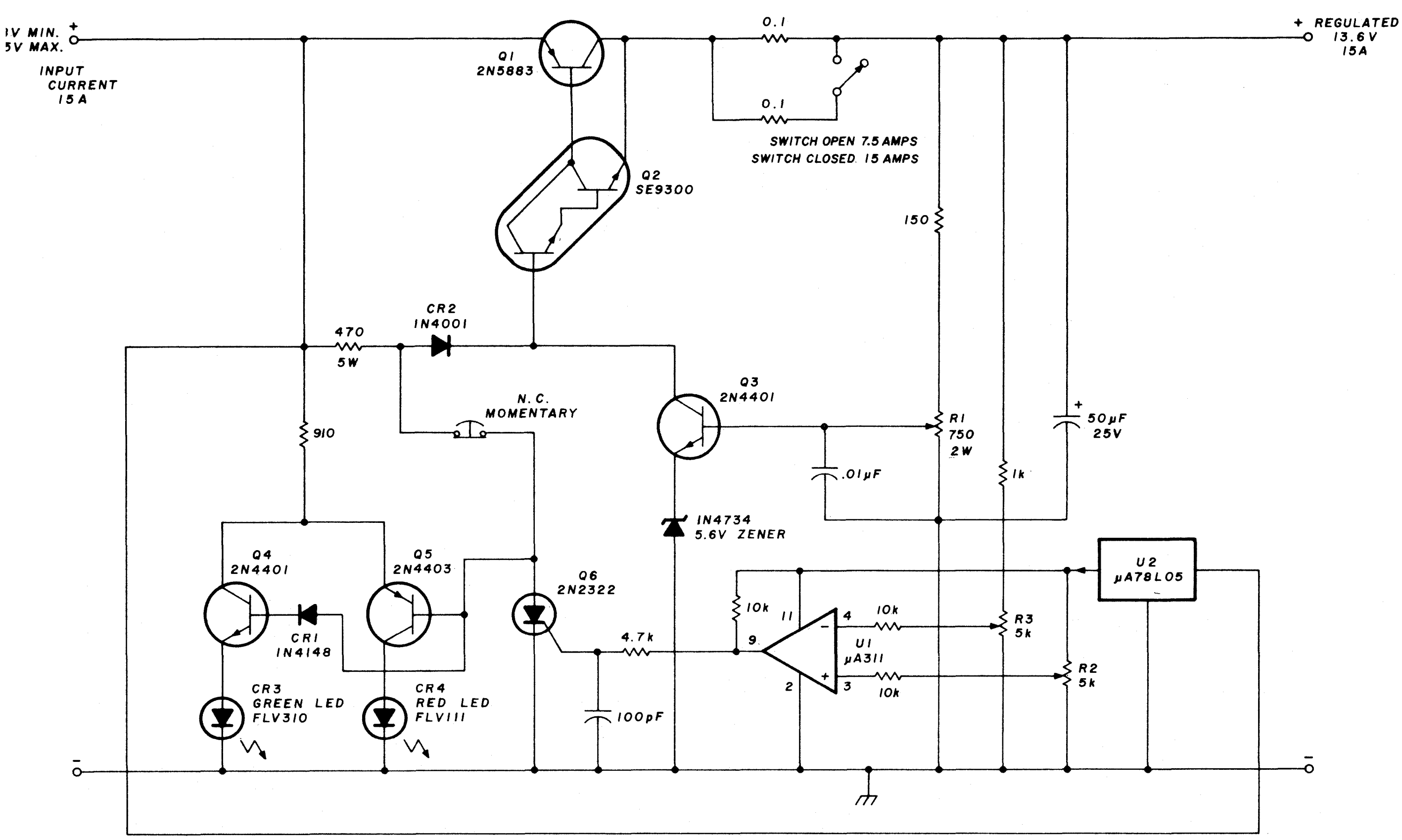


fig. 1. Schematic diagram of the high-current power supply. The separate 5-volt regulator is used to bias one input of the voltage comparator. The SCR is fired when the second leg of the voltage comparator detects a drop across the 0.1 ohm resistors. S1 is a normally-closed momentary type switch. Q1 must have an adequate heatsink (see text).

be allowed to restart after some pre-determined current level has been exceeded.

This power supply contains a high-gain error voltage control circuit and a voltage comparator that functions as a high-speed switch.

priately divided. When an under-voltage condition is created by over current, the output of the comparator will go high, turning on the SCR. To restart the regulator, it is necessary to momentarily disconnect the anode of the SCR, after clearing the over current condition. The green and red LEDs are used to indicate the status of the regulator, with red indicating that the SCR has been fired.

construction

The actual construction of the power supply is straightforward. However, the pass transistor must have an adequate heatsink. The power dissipated is

$$W_{diss} = (V_{in} - V_{out}) + V_{CEsat} \times I_{CE}.$$

To setup and adjust the supply, all components are connected with the exception of the SCR, Q6. A 100-ohm, 100-watt adjustable resistor can be used as a load. With the load resistor set for maximum resistance, the power supply is energized and R1 is adjusted for the desired output voltage (13.6 volts). An oscilloscope can also be connected to determine if there are any oscillations on the output waveform. If so, a 27-pF capacitor connected from collector to emitter of Q3 should eliminate the problem.

With a high-impedance voltmeter connected to the wiper arm, R2 should be adjusted to provide 2 volts to the comparator. This is the reference set point and should not be disturbed once set. After moving the voltmeter to the wiper of R3, this potentiometer should be set for a reading of 2.5 volts. With the meter still connected to R3, change the tap on the load resistor to the 50-ohm point. The 2.5 volt reading should not change.

The final step consists of checking the action of Q6. With the power supply switched off, connect the SCR. Also disconnect the load resistor and set the slider for a resistance of 1 ohm, but do not reconnect the resistor at this time. After re-energizing the supply, the red LED should be on; pressing the arming switch will turn on the green LED while extinguishing the red. If the LEDs do not function as indicated, remove the SCR and check the voltage on R3 to ensure that it's no lower than 2.5 volts. Also, check the voltage on the output of the comparator; it should be less than 100 millivolts. With the SCR installed and the circuit armed, the gate of Q6 should also measure less than 100 millivolts.

To test the circuit, momentarily connect the 1-ohm load resistor the to supply output. The shutdown circuit should operate instantaneously, and the red LED should come on. You should test this action several times to verify consistent shutdown.

At one ohm, the load current is 13.6 amps, but you can adjust R3 to actuate the shutdown at any current you desire. However, the voltage from R3 must not be lower than R2's voltage or the circuit will lock up in the fired mode. This system has been in use for over a year and has operated flawlessly during the entire time, even when powering tube-type mobile transmitters.

ham radio

command function debugging circuit

This device takes command information from standard PLL function decoders and applies a short time delay to provide debugged control commands

Our club had just finished installing an autopatch, exhausting all available funds, when the need for a better command decoder became evident. Each of the users had noted voice falsing of the decoder, and it was enough of an annoyance to be serious. More than once a telephone conversation had been terminated by a voice peak.

The investigation of causes and cures ran the gamut, until we finally decided that a fraction of a second delay on the command-function decoders would probably solve the problem. A number of designs were passed around and considered, and a few almost built, but the one described here was finally implemented — for three reasons: one, it works; two, it costs nothing to build; and three, it was built before any of the others. The design was based on the available components, and on the assumption that it might be necessary to drive a number of different types of loads. These considerations led to the use of transistor switches and reed relays to provide the outputs.

circuit description

The output circuitry from the tone decoder is shown in **fig. 1**. In our repeater, the debugging circuit was added without making any modifications to the autopatch. As shown in **fig. 2,** the debugging circuit consists of two portions, the transistor switches are wired to provide V_{CC} to their respective reed relays and the delay circuit which provides the ground return for the relay coils. Since only one relay

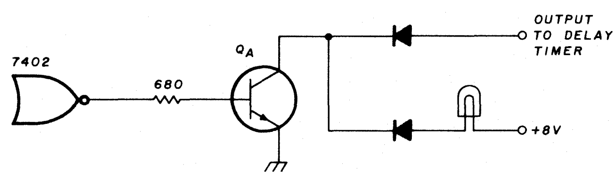


fig. 1. Output switching from the autopatch, prior to the installation of the debugger.

By J. Thomas Norman, WA7HFY, 1002 South 8th Street, Laramie, Wyoming 82070

is actuated at a time, only a single delay circuit is required. The diodes at the input of the delay timer form a multiple-input OR gate to actuate the delay after any digit has been decoded.

The timer itself is not very critical, especially since four different transistor types were used (even an unidentified NPN transistor). In its quiescent state, Q1 is biased on, charging C1 to approximately 8 volts. This level maintains Q2 and Q3 on, and Q4 off. When a function is decoded, Ql's base is effectively grounded back through the two diodes and Q_A. With the transistor now cut off, C1 will discharge through R3.

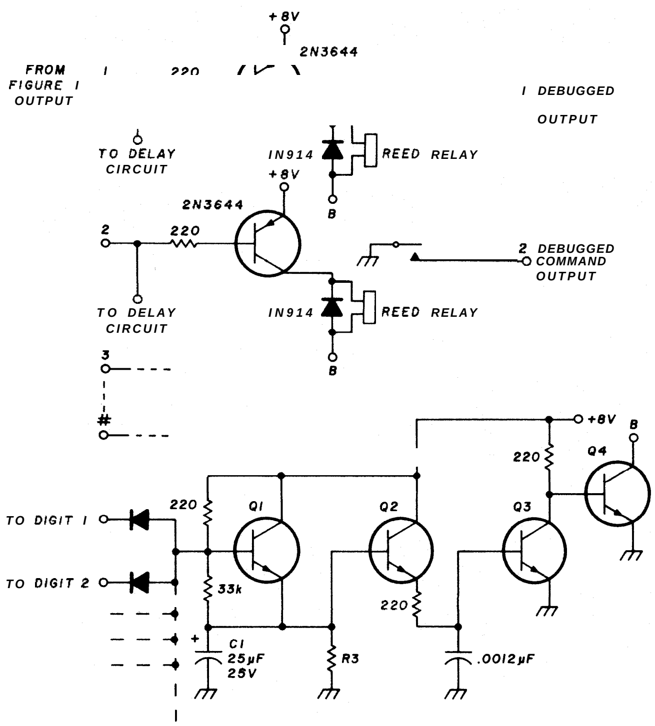


fig. 2. Schematic diagram of the add-on debugging circuit. Instead of the commands being controlled by transistor switches, they are controlled by reed relays. The delay portion of the circuit prevents the relays from picking up on noise spikes. The transistors can be almost any NPN type. R3 should be adjusted for the desired delay.

The value of this resistor can be adjusted for the desired amount of time delay, in our case approximately 0.1 second. When the voltage on the base of Q2 drops below about 1.2 volt, Q2 and Q3 are cut off and Q4 is turned on. With Q4 on, the reed relay is actuated, but when a noise spike is decoded as a pulse, the 2N3644 is turned on; since most noise pulses are very short in duration, they do not allow the timer to complete the circuit to actuate the relay.

ham radio

no-cost grid-dip meter

How to use no-cost parts salvaged from an old broadcast radio to build a grid dipper that tunes from 1 to 90 MHz

In these days of shortages and skyrocketing prices it seems improbable that anything could be free. However, except for a little spray paint, the instrument described here can, in truth, be free. Originally it was built for novice use to enable them to prune antenna systems, and to serve as an absorption wavemeter to insure output was on the correct band. It has proven so stable and reliable, however, that it can also serve a more experienced ham as a rugged instrument for general purpose work, and where considerably more output is needed than is available from soiid-state dippers.

The frequency range is limited to approximately 1-90 MHz due to the variable capacitor I used for tuning. If a fairly clean capacitor is used — one which is not corroded and has good wipers — tuning will be unusually smooth and free from spurious dips. From the photographs it is obvious that it is more bulky than most commercial dippers. The vacuum tube protruding from the end may not be pretty, but this arrangement contributes substantially to frequency stability.

The schematic, **fig. 1,** shows the basic circuit. The tuning capacitor is the type commonly used in older five-tube ac-dc broadcast sets. If you have no 150-volt supply easily available, you can add a series resistor in the B+ line. This resistor should be about 50k for each 100 volts higher than the desired 150-volt supply. For instance, if the dipper is to be run from a 350-volt supply, the supply voltage is 200 volts too high, so a 100k dropping resistor would be chosen. It can be any value from 82k to 120k, and should be rated at 2 watts. The heater voltage can probably be picked off the same supply.

If you don't have a milliammeter available, you can use a 1000 ohms-per-volt multimeter on the 1.0- or 1.5-volt scale, or you can install a 2.2k resistor in place of the meter and read across this resistor with a 20,000 ohms-per-volt multimeter or vtvm on a 1.0- to 1.5-volt scale. This hookup is shown in fig. **2.**

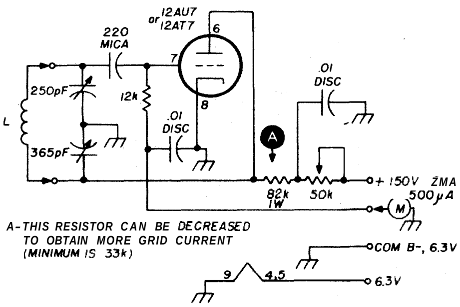
If you want to modulate the dipper, you can add the simple circuit shown in **fig.** 3. This transistor oscillator is powered by the dipper grid current and grid modulates the dipper. This makes it easier to locate the dipper signal during calibration, and allows the dipper to be used as a temporary signal generator for alignment work.

By Bill Wildenhein, W8YFB, 41230 Butternut Ridge, Elyria, Ohio 44035.

The dipper can also be used to check that your rig is tuned to the correct band. Switch off the B+ voltage to the dipper, but leave the heater voltage on. Poke the dipper into the rig near the output tank coil after you have tuned the rig, and tune the dipper for a *peak* instead of a dip. The peak will indicate which band you are actually on. Be careful of the high voltage!

plug-in coils

All the coils are wound on bases of defunct octal tubes. For those of you who have never smashed tubes, it can be done safely by holding the tube by the base with the glass envelope against the inside of a wastebasket. Give the glass a sharp rap with a hammer and the tube breaks without showering glass around as you would expect. Normally, the glass part cemented into the base remains relatively

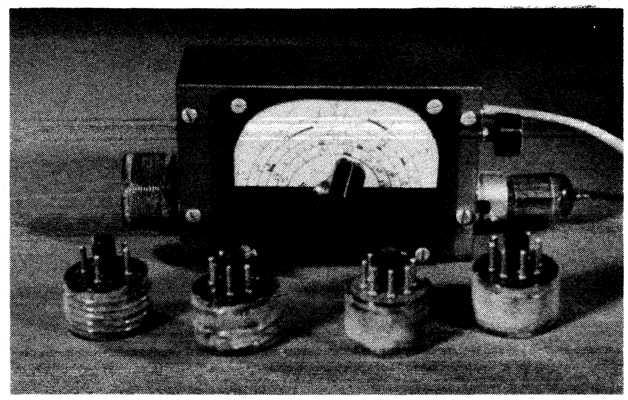


coil number	frequency range	wire size	number of turns
1	0.97 - 2.9 MHz	34 AWG (0.16mm)	83
2	2.8 - 8.7 MHz	24 AWG (0.51mm)	27
3	6.7 - 20 MHz	24 AWG (0.51mm)	12
4	15 - 45 MHz	20 AWG (0.81mm)	5
5	30 - 90 MHz	20 AWG (0.81mm)	2

fig. 1. All the parts for this "no-cost" grid dipper were salvaged from old radios and TV sets. All coils are wound on 32 mrn (1-1/4 inches) diameter octal tube bases; winding length is 14 mm (9/16 inch).

intact. This part of the glass can be broken up by punching it with a screwdriver. The remaining cement can easily be scratched out with a jackknife. Remove the leads from the pins by applying a soldering iron to the tips of the base pins while you pull on the leads with a pair of pliers.

Line up the tube bases after they are cleaned and select a pair of pins roughly opposite from each other (different tubes may have different base arrangements with some pins missing). Adjacent to one pin drill about a 1.5 mm (1/16 inch) hole as close to the bottom of the base as possible. Adjacent to the other pin, drill a 1.5 mm (1116 inch) hole 14 mm (9116 inch) higher up on the socket from the first



This view shows the GDO with the top of the enclosure and the dial face removed. The capacitor shaft fits through the notch in the top.

hole. Be sure the desired pins are cleaned of solder. It sometimes helps to run a piece of enameled wire down through the pin hole while heating the pin with a soldering iron. The wire sizes listed on the schematic need not be exactly as shown. Those are the wire sizes I used, but coils 2, 3 and 4 could use the wire commonly used in the yoke windings of TV sets. The first coil could use either number-34 or -36. This size range is often found in audio interstage transformers. If you are unwinding a transformer, it is easy to find out whether a particular wire size will fit. From the loose end of the transformer winding, measure 14 mm (9116 inch) and place a bit of masking tape there as a marker. As you unwind to this point, count the turns. If there are at least 80 turns to that point, the wire is small enough for the job.

As a last resort, if you can't find wire that small, you can find a size that allows perhaps, 60 turns in the total length. Wind the 60 turns, then continue winding over the first layer to get a smooth layer of about 15 turns. Then wind up over that second layer with a third layer totaling, perhaps, 10 turns. If you have never hand wound coils, here are some of the basics: First, carefully scrape the enamel insulation from one end of the wire. Poke the end through the hole in the base nearest the pin end. Now poke the wire down through the proper pin and solder it into the pin. Carefully pull the excess wire back through

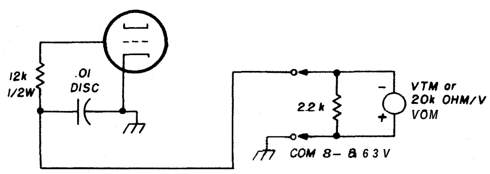
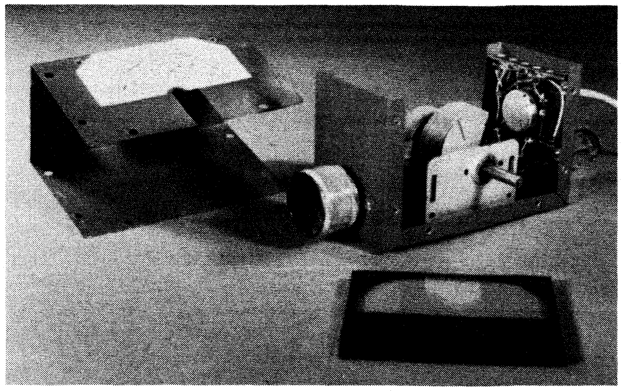


fig. 2. if you don't have a sensitive milliammeter, substitute a 2200-ohm resistor in its place and read the voltage drop across the resistor with a vtvrn as shown here.



The finished grid-dip oscillator with the coils for different tuning ranges.

the 1.5 mm *(1*/16 inch) hole. Avoid kinks as they seriously weaken the wire. If one of these kinks forms, straighten the wire as you pull it through the hole so the kink isn't pulled tight.

Now reel off a length of wire and fasten the spool in a vise. Pull the wire taut. Make sure there are no kinks, though. Holding the coil form with both hands, wind your way toward the vise, keeping tension on the wire. When the correct number of turns are on the form, hold the winding from slipping with one hand and cut off the wire to leave 15 to *20* cm (6 to *8* inches) of excess. With your free hand, scrape the enamel off enough of this end to get through the other pin. Poke this end of the wire through the top *1.5* mm *(1116* inch) hole, pull it up tight, and put your thumb over it to keep it from slipping. Poke the end down through the correct pin and pull it taut. Pull it over to the side to make a sharp bend at the end of the tube pin. This will prevent the wire from slipping back through while you solder it to the pin. Now you can trim up the pin ends with a file.

Coat the entire winding and winding surface of the coil form with plastic model cement for Styrene models. Put on a heavy coat. For perhaps 15 minutes you will have to alternately stand the coil on the plastic locator between pins, then tip it over and stand it on the open end of the form as the cement slowly tries to run off the form. When it sets sufficiently that it stops running, leave it overnight. The next day add a second heavy coat. This will form a protective coating that will securely hold the turns in place when inserting or removing the coil, assuring that your calibration remains accurate.

Several of the coils are space-wound. After the winding of these coils is completed, the turns can be adjusted slightly until they are uniformly spaced.

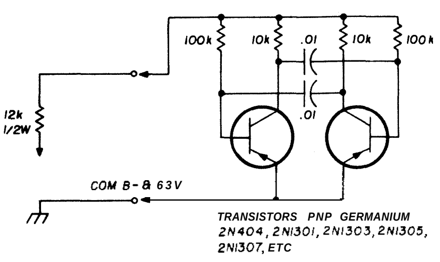


fig. 3. Simple modulator for the grid-dipper for fig. 1 uses a simple two-transistor oscillator which is powered by the grid current of the dipper.

enclosure construction

The photograph of the disassembled dipper shows the method of forming the sheet metal case. It was made of galvanized iron of the type used for furnace ducts. To form the sheetmetal box you need a couple lengths of angle iron about one foot long, and one piece about 5.7 cm *(2-114* inch) long. Cold-rolled steel barstock is even better. Stock about 13x19 mm (1/2 by *314* inch) is good. In case you have neither, pick up some scraps of oak flooring and have a friend cut off the tongue and groove to give you a nice sharp corner. You also need a vise and a fairly husky C-clamp. Be sure to lay out and bend up the main

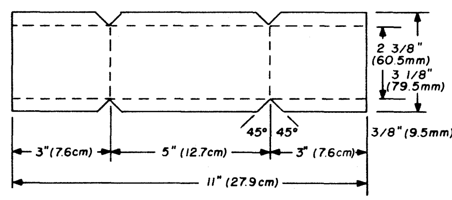


fig. 4. Sheet-metal layout used for the grid-dipper chassis.

chassis first. **Fig. 4** shows the layout I used. If your tuning capacitor is larger, make your box slightly larger. Take pains to make the layout very accurate. It saves headaches later when you bend up the lid.

Fig. 5A shows a cross-section if the bends are made correctly. Notice that the overall width is *6* cm (2-3/8 inch) plus two metal thicknesses. The reason for this is quite obvious if you look at **fig. 5B.** If the bending bar is placed exactly on the line, the outside edge of the metal extends one metal thickness beyond the line. Now look at **fig. 6A:** the metal is clamped differently than in **fig. 5B**. If you make one bend as in **fig. 5B** and the other bend as in **fig. 6A,** you will end up with the lopsided, inaccurate cross section shown in **fig. 6B.** Compare the dimensions with those in **fig. 5A;** to get these dimensions, you must clamp the bending bars as shown in **fig. 7.**

This leads to a simple sheet-metal rule: if you want the *inside* dimension of a U-shaped bend held to a size marked on a layout, the bending bars must also be on the inside, as shown in fig. *7.*

For this first piece, the outside dimension is relatively unimportant. The lid is the really important piece because it must fit around this first box, and the inside measurement of the lid must be accurate. After this first piece is bent up, carefully measure the width at all points, then lay out and bend the lid. Next, lay out the holes in the lid and drill them with about a 5 mm (3/16 inch) drill. After removing the burrs, place the two sections together and carefully mark the hole locations on the inner box. These can be drilled with a number 36 (2.5 mm) drill. Then the box can be assembled with number 6 (about 3.5-mm diameter) sheet-metal screws salvaged from scrap TV sets.

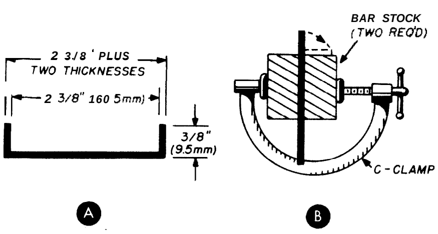


fig. 5. Proper method of bending the sheet metal used for the chassis (see text).

Locate and drill all socket holes, mounting holes and check the fit, then thoroughly clean both pieces with household cleanser. After they are dry they can be sprayed inside and out with gray primer paint before assembly and wiring is begun. The dial is made from a piece of card stock glued to the outer box. This is covered with a piece of clear plastic (small pieces are often available as scrap from hardware stores or glass shops). The "unbreakable" window panes are usually Plexiglass or Lexan, and they have very good rf insulating properties. For using a dial bezel, it is masked, then the layout drawn on the masking tape. With a sharp knife, cut away the part to be painted and spray with acrylic lacquer.

Your first homemade enclosure may take time and may be less than a professional job, but if you practice, using scraps of metal that can be picked up free, you will soon find it possible to do a fast, accurate job the first time. I usually spend less than a half-hour to make this sort of enclosure. The result is that you are no longer limited to standard box sizes or costs. Your dollar savings will pay for any investment in tools many times over. For example, you can get a combination square at a discount store for as little as a dollar, but in most cases they are anything but square! The aluminum or pot metal

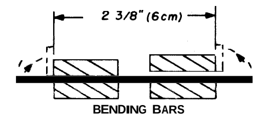


fig. 7. Proper setup for bending the flanges on a chassis to maintain an internal dimension (see text).

die-cast head may warp still more. For about $5.00 you can get one with a steel or cast-iron head that will be acceptably accurate for sheet metal work. With just a little care it will last you the rest of your life, and will guarantee a lifetime of accurate boxes.

If you have a very limited budget and can't afford anything but that discount store square, here is a way to check the square in the store, and pick the one that is square (see fig. **8).** Take along a piece of sheet metal that has one edge sheared true and straight. If the square isn't a true 90 degrees, it will show up immediately as shown in the drawing. Just place it on the metal as shown in position A and scribe a line on the metal. Swing it around to position B and see if it falls exactly on the previously scribed line. Watch out for a head with a crooked edge, as in fig. **9.** Strive for an accuracy of about 0.5 mm in 25 mm (1/64 inch in 10 inches), or better. It will make your work far easier.

calibration

Finally, let's cover calibration of the grid dipper. Up to about 30 MHz it can be checked against a general-coverage receiver. Above that frequency you may have to find a friend with a calibrated vhf dipper. Begin calibration at the lowest frequency and work toward the high end because general-coverage receivers commonly use a 455-kHz i-f which leads to poor image response at the higher frequencies. Starting at the low end, your calibration points will be accurate. If you are in doubt about one point, you can return to the last point to be sure you are correct, then carefully proceed higher in frequency. Another check is to watch the receiver's S-meter. The image signal is usually noticeably lower on the S-meter.

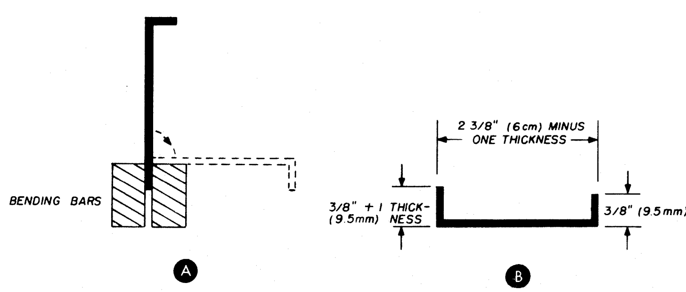


fig. 6. If bend bars are not used properly, one side of the chassis will have a different dimension than the other due to metal thickness (see text). Proper setup is shown in fig. 7.

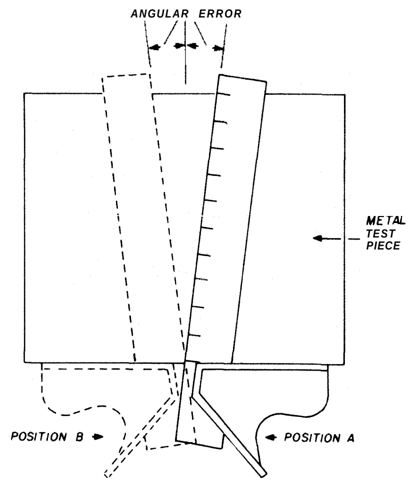


fig. 8. Simple method for checking the accuracy of a low-cost square.

Since the power output of this dipper will probably be as high as 200 milliwatts, it is inadvisable to attempt to use it to dip circuits connected to solid-state devices. You can easily blow a transistor. Remove the transistor before checking a circuit, but realize

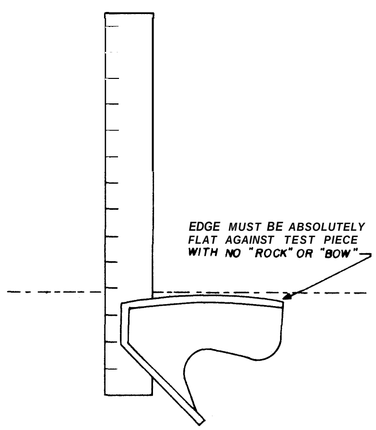
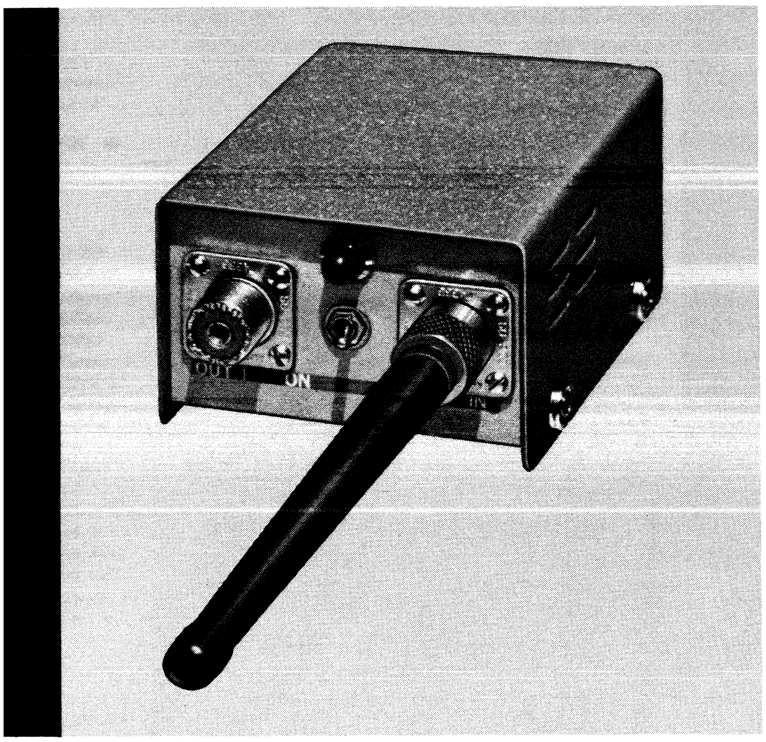


fig. 9. Beware of low-cost squares that are not perfectly flat — they should not rock against the test piece as shown here.

that the transistor itself adds capacitance to the circuit. I usually dip tank circuits before installing them, and include a small fixed capacitor to represent the capacitance of the transistor.

In conclusion, although this device is a homely looking gadget built from junk, it is still worth the effort to do the job right. I find it so stable and reliable that it has become one of the hardest working tools on my bench. I hope you enjoy the same benefits.

ham radio



neat packaging for vhf prescalers

Novel packaging for K4GOK's prescaler uses readily available parts

At least one thing is quite apparent from the popularity of technical literature in amateur radio circles. A large part of the hamming community is putting certain minimums of time into electronic gear construction. To a big segment of talented do-it-yourselfers in hamdom, this kind of involvement is necessary to get maximum enjoyment from the hobby. It's not easy. Even those who live for the smell of hot rosin and solder soon learn that workbench time is a scarce luxury. One result of the paucity of time is that we get only the highest priority stuff to the bench and quite simply, lean on the stories others tell (or write) to learn about out-of-reach or lower priority Items.

It is pretty well known that not being in a position to build a super hot new receiver doesn't spoil the fun of reading about how the author did it. The buzz word for this, of course, is *vicarious* and we'd be in rough shape without being able to enjoy things vicariously.

Even with the best intentions, many construction projects find their way into a file cabinet to await future action. Occasionally one may be pulled from the file and placed on the workbench for further thought, but every now and then lightning strikes and a construction article comes along to absolutely ruin your peace of mind until you can translate that writer's words into reality.

K4GOK's prescaler, as detailed in *ham radio* for February, *1976,"* was just such a bolt for me. It turned out so well and so easy to assemble that I thought it deserved a little extra mention in all the shacks where it is filed under *projects — future.*

There is probably a great number of frequency counters in use which conk out at 50 MHz — or less. The need, however, for accurate frequency measurement in today's hamshack is in the order of 300 MHz, and if not a requirement for 2-meter fm, it is certainly a great convenience.

"M. D. Kitchens, K4GOK, "VHF Prescalerfor Digital Frequency Counters." *ham radio,* February, 1976, page 32.

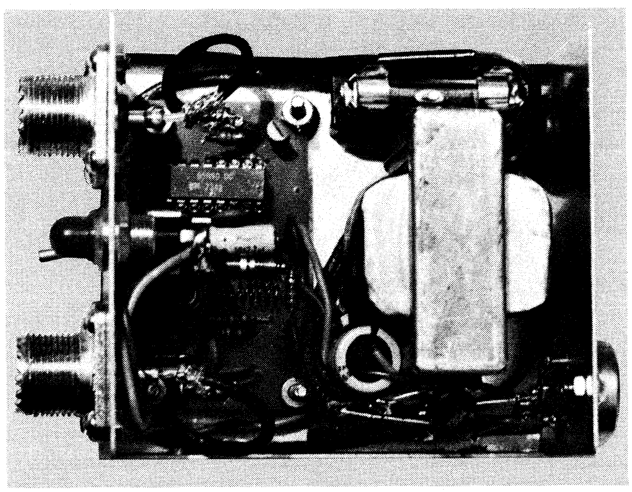
By Alan Smith, W8CHK, 3213 Barth Street, Flint, Michigan 48504

There is no problem if you are in a position to select your needs from a supplier's shelf, but if you are in the more common position and want to soup up a flea market special or your old reliable Heath 1101, this prescaler is for you.

The construction approach recommended here is to build the prescaler as a separate unit. It will be easy on time and gentle on your pocketbook. You can even get several parts at your local Radio Shack store.

The unit in the photograph was built by my son, WB8YOB, and is a self-contained copy of the circuit described by K4GOK. As shown in the photo, it is set up to prescale two-meter signals with a rubber duck antenna as the coupling element.

The case is an Archer 270-250 ($2.98 from Radio Shack). You can get the transistors at Radio Shack



Construction of the vhf prescaler showing the layout of the perfboard with the preamplifier and 95H90 decade scaler, left, and power supply components, right. Input and output coaxial connectors, left, give a good idea of the small size of the package.

by substituting 2N706 for 2N5179 and MPS-6533 for 2N5771. Otherwise the circuit can be assembled exactly as described in the original article.

The power supply is conventional. It requires a circuit-board transformer with an output voltage of 5.5 volts or more up to a maximum of 12 volts at 300 mA, a standard 500 mA bridge rectifier, one 1000 μF electrolytic capacitor, and an LM309K IC voltage regulator. The LM309K will keep the supply voltage at the required 5 volts.

These parts can be neatly fastened to a square of perf board and secured to the case, or the transformer can be fastened to the base and the remaining components simply hard wired. Add an on-off switch and a miniature red indicator light and you'll be ready to start prescaling.

ham radio

precision voltmeter calibrator

It's always nice to have a high-accuracy, digital voltmeter around, but unfortunately they're very expensive. To solve this problem, I decided to build a variable voltage source to accurately calibrate the dc voltmeters I had on hand. There have been many precision ten-volt sources described in current literature;[1,2,3] I have used the circuit shown in reference 1 as the basis for my design.

circuit details

The circuit shown in fig. **1** will produce from 0 to 10 volts with an error of less than 20 millivolts. In addition, it's also accurate with meter movements that have an internal resistance as low as a few thousand ohms. The 10 volts developed by the LM308 appears across R4. In my case, I used a ten-turn potentiometer and a suitably accurate ten-turn dial. The output from this potentiometer is fed into a 741 op amp connected for unity gain. In this way the op amp's high-input impedance will have negligible loading on the voltage source and its low output impedance will allow the calibration of meters with low internal resistance.

It will be necessary to null out the small dc offset error which appears at the output of the 741 op amp. This is accomplished by the 10k potentiometer, R5; ground the input of the 741 and adjust R5 for zero volt output. The nulling procedure must be completed before final calibration is attempted. **Table 1** shows the results of calibration against a known voltage source.

For operation, this calibrator requires plus and minus 15 volts, at less than 50 mA. The normally closed switch shown in **fig. 1** is used to start the calibrator by creating a transient through the 560 pF capacitor. If the power supply for the calibrator is already running, this switch can be eliminated. Finally, you should realize that this calibrator can not be used as a substitute for a variable voltage power supply.

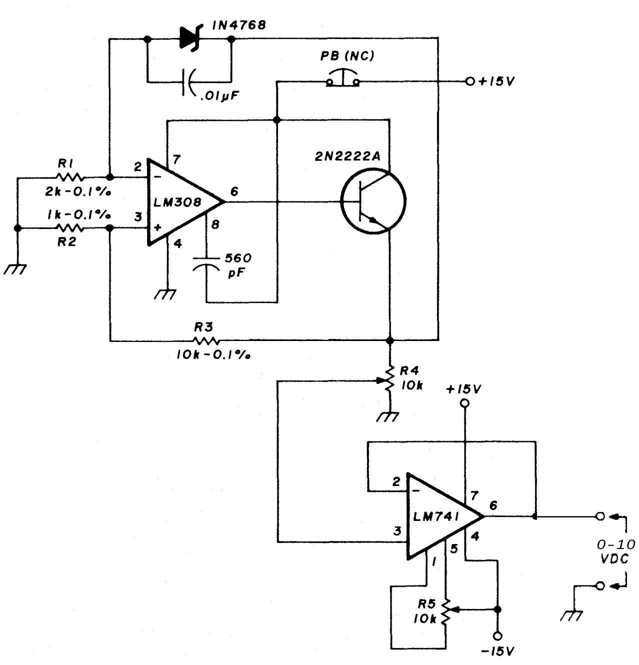


fig. 1. Schematic diagram of the 0 to 10 volt voltage standard. The pin numbers for the op amps are for the 8-pin mini-DIP package. R4 is a 10-turn potentiometer with a linearity of 0.1 per cent. The resistor tolerances should be closely observed to obtain a precise 10-volt output from the LM308.

table 1. Calibration results for the 0 to 10 volt standard.

nominal voltage as read on R4	laboratory potentiometer reading
10.00	10.00
9.00	9.00
8.00	**7.99**
7.00	7.00
6.00	6.00
5.00	5.01
4.00	4.00
3.00	3.00
2.00	2.00
1.00	1.00
0.00	0.00

By Hubert Woods, Calle Las Nubes, 1760, Guadalajara 5, Jalisco, Mexico

references

1. William Goldfarb, Electronics Circuit Designers Casebook, ***Electronics,*** June 7, 1975, page 107.
2. David W. Ishmael, WA6VVL, "Precision 10.000 V dc Reference Voltage Standard," 73, September, 1975, page 124.
3. ***Linear Data*** Book, National Semiconductor Corporation, Santa Clara, California 95051, page 2-1.

ham radio

the gyrator: a synthetic inductor

Using operational amplifiers you can simulate inductors in LC filter design

The Gyrator is an electronic device that inverts the impedance of a capacitor and therefore makes it take on the characteristics of an inductor. Through the use of such a device, some of the shortcomings of inductors can be eliminated. These shortcomings are large physical size, low Q, nonlinearity, and inter-winding capacitance.

A properly designed Gyrator will provide a synthetic inductor with high 4, wide bandwidth, inductance value independent of frequency, and good stability. Filters for frequencies up to 50 kHz can be designed using Gyrators.

LC filters

Since a Gyrator can be inserted directly into an LC filter in place of an inductor, let's first review how an LC filter works. An LC filter is a reactive, two-port, doubly-terminated device that reflects power back to the source in frequencies that fall outside its band-pass. As the number of individual elements making up the filter is increased, the ability of any one component to greatly change the resonant frequency decreases. It can be said, therefore, that a coupled

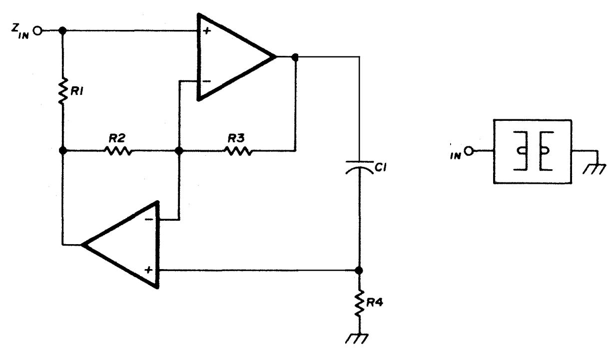


fig. 1. Basic Gyrator circuit and standard Gyrator symbol. Circuit is the basis for designing filters without inductors.

LC filter is basically rather insensitive to changes in values of either capacitors or inductors. What will happen (when changing a component) is that the amount of power reflected to the source will be decreased. As other component values are changed, the reflected power becomes less and less, and the stop bands become more and more poorly defined.

By John Loughmiller, WB9ATW, Route 1, Box 480C, Borden, Indiana 47106

designing a Gyrator filter

To implement an LC filter without an inductor, first calculate the required inductive and capacitive impedance for the filter you have in mind. If desired, the filter can be built using standard inductances to check your calculations before actually building the Gyrator.

Next, using the circuit in fig. **1,** construct a Gyrator replacement for the inductor. This circuit will simulate an inductor with the value of KC, where K is a constant derived from the resistors: $K = (R1 \bullet R3 \bullet R4)/R2$, and C is the capacitor whose impedance is being inverted (C1 in the circuit).

The resultant number so derived can be treated as if it were the value of impedance of an inductor, and the actual inductor can be removed and replaced with the synthesized inductor just created.

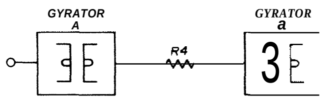


fig. 2. Method for simulating an ungrounded inductor using two Gyrators. Resistor R4 is the resistor that determines the simulated inductance.

A simpler approach is to make all four resistors the same value and use the formula R^2C. (The impedance presented to the input of the Gyrator is very nearly $j\omega R^2C$, hence the formula R^2C). Use a value of R large enough that the op amps won't be loaded, yet slightly lower than the differential input impedance.

The Gyrator circuit appears to be just another active filter circuit; however, there's a significant difference: in other active filter circuits amplifier phase shift degrades circuit Q. In Gyrators, the Q is enhanced. When the phase shift is greater than 90 degrees, the Q is actually *higher* than that of the capacitor itself.

table 1. Typical Q of various types of capacitors at 1 kHz.

capacitor type	Q at l kHz
Mica	600
NPO ceramic	1500
Glass	1500
Polystyrene	2000
Polypropylene	3000

Speaking of capacitors, **table 1** indicates the Q to be expected (at 1 kHz reference) for various types of capacitors. You can see that a high-Q inductor can be synthesized by using one of the high-Q capacitors, since the Gyrator inverts its capacitive impedance into an inductive-type impedance.

With regard to temperature, the NPO ceramic is least affected by changes of this variable. Higher stability could therefore be expected if NPO-type capacitors were used; however, there would be a tradeoff in Q. For high Qthe choice is polystyrene or polypropylene, with the latter perhaps a better choice as it has a better temperature coefficient.

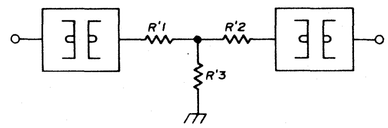


fig. 3. Synthesized T filter using Gyrators, which performs as if it were constructed as a standard T filter using inductors.

floating inductors

Perhaps the principal drawback to Gyrators is the fact that a floating (ungrounded) inductor can't be simulated by a single Gyrator. It's possible, however, to use two Gyrators as in fig. 2 and successfully simulate such an inductor.

The formula is $L = [(R1 \bullet R3 \bullet C)/R2]R4$, and $R4$ is a shared resistor between Gyrator A and B, as in fig. *2*. $R4$ is the resistor that determines the simulated inductance. So if $R4$ becomes a loaded port, the simulated inductance will be dependent on the value of resistance connected to the port. This being the case, resistor $R4$ could (if desired) be divided into sections and T or pi filters simulated, such as shown in fig. 3. Here, a T filter is created synthetically, which would perform as if it were constructed in the manner shown in fig. **4.**

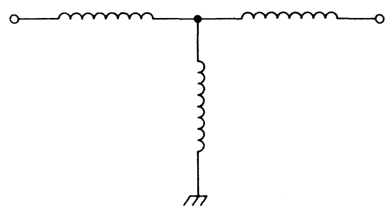


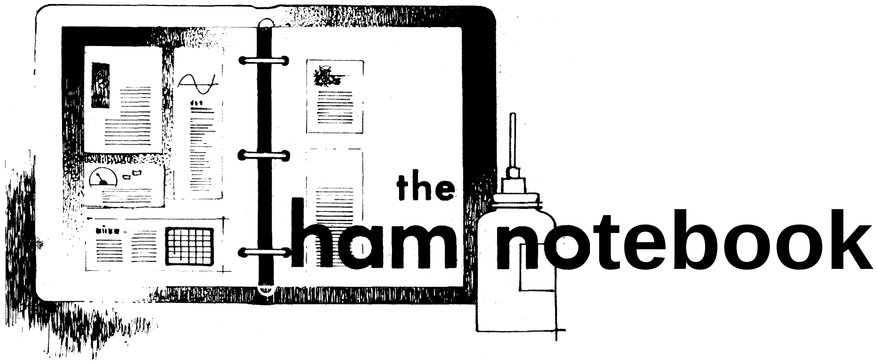
fig. 4. T filter using inductors in the usual manner, which can be simulated by Gyrator techniques.

When building multisection filters, a quad op-amp is recommended. A 4136 is the choice of Mr. Thomas Lynch, from whom much of the information in this article was obtained.

bibliography

1. Thomas H. Lynch, "The Right Gyrator Trims the Fat Off Active Filters," *Electronics,* July 21, 1977, page 115.
2. William C. Sutherland, "Op-Amp Gyrator Simulates High-Q inductor," National Aeronautics and Space Administration Tech Briefs, Vol. 2, No. 3, Fail, 1977, page 318.

ham radio



pi-network rf choke

The great majority of pi-network rf amplifiers use shunt feed for the high voltage dc to the plate. Therefore, great dependence is placed on the isolation properties of the rf choke. Most chokes are not suitable for this purpose due to resonances, especially when the amplifier is a multiband unit.

The most successful shunt feed rf chokes are single layer solenoid wound, preferably using resistance wire to lower the Q and damp out potential resonant frequencies. One of the best chokes was the one made by Collins for the ART-13 transmitter. They are difficult to obtain, however.

To avoid the problem, series feed at a low rf potential point as shown in fig. **1** is suggested. In this circuit any of garden variety rf chokes are suitable. The plate blocking capacitor is now at the low rf potential part of the circuit; it must, however, be able to withstand the high voltage dc and be able to carry the tank circuit rf current without overheating. Transmitting ceramic capacitors are recommended.

Gary Legel, W6KNE

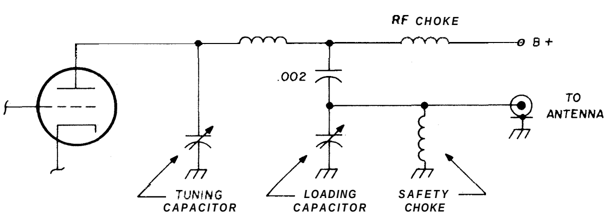


fig. 1. Suggested placement for rf choke in multiband rf power amplifiers. Since this is a low-impedance point in the output circuit, it reduces the performance requirements of the rf choke.

calculating feedline loss with a single measurement at the transmitter

Much has been written about the amount of insertion loss of various types of feedline and connectors, and about how the loss in some types of coax increases sharply with age and exposure to the elements. What is not widely known in amateur circles is that the loss in any run of cable can be determined with only an swr bridge or wattmeter, a transmitter or other drive source, and a shorted connector.

If a piece of cable is terminated with a short or an open circuit, the vswr measured at the transmitter will be infinite if the cable has zero insertion loss. In the real world, any cable has some loss, and this loss will cause the vswr measured at the transmitter to be less than infinity. The greater the cable loss, the lower the vswr.

To determine the cable loss, disconnect the antenna from the feedline and replace it with a shorted connector. This produces the infinite vswr at the transmitter end of the cable more reliably than simply leaving the cable unterminated. Connect the swr bridge or wattmeter at the transmitter and measure the vswr of both forward and reflected power, using the lowest possible transmitter power to avoid possible damage to the final amplifier. The reflection coefficient, ρ, is found by using one of the following formulas:

$$\rho = \sqrt{\frac{P_{reflected}}{P_{forward}}} \text{ or } \rho = \frac{SWR-1}{SWR+1}$$

This reflection coefficient is a numerical ratio which must then be converted to decibels:

$$\rho(\text{dB}) = 20 \log_{10}\left|\frac{1}{\rho}\right|$$

Note that a vswr of ∞ corresponds to $\rho = 1$ or $\rho(dB) = 0\ dB$, and a vswr of 1.0 corresponds to $\rho = 0$ or $\rho(dB) = \infty\ dB$. The cable loss is determined by using the general equation:

$$\rho(dB)_G = \rho(dB)_L + 2A_o$$

where $\rho(dB)_G$ is the reflection coefficient at the transmitter, $\rho(dB)_L$ is the reflection coefficient at the load, and A_o is the cable loss in decibels. Since $\rho(dB)_L = 0$ in this case,

$$A_o = \frac{\rho(dB)_G}{2}$$

This same general formula can also be applied if vswr or wattmeter readings are taken at both ends of the cable with the antenna connected instead of the shorted termination. The advantage of using the shorted termination is that it eliminates the need to take readings at the antenna, which is usually a two-man job.

Once the feedline loss is known, the antenna is reconnected, and it is now possible to accurately determine the vswr at the antenna based on the reading taken at the transmitter:

$$\rho(dB)_L = \rho(dB)_G - 2A_o$$

$$\rho_L = \frac{1}{\text{antilog}_{10} \frac{\rho(dB)_L}{20}}$$

$$SWR = \frac{1+\rho}{1-\rho}$$

For most accurate results with multiband antennas, a feedline loss calculation should be made for each band.

John E. Becker, K9MM

R M terminal modification

After using the Mini Micro Mart RM Terminal Unit for a number of months, some deficiencies were noted which made it a little unwieldy to use, and sometimes downright difficult! I decided to pull all the guts out of it, because I didn't know how it worked, and if it broke, about all I could do is use it for weight in the back of my car in the winter. This modification is used to convert the parallel output of the original keyboard into special data for insertion into the loop of a RTTY system. In addition to this, it gives you end of line (EOL), letters, and figures indications through the use of an LED. The EOL feature is especially nice if you are using a video display which has less than the 66 characters necessary for a complete line of hard copy. In addition to this, it provides parallel data out, still in Baudot code, just in case you need it to feed an ASCII converter.

Looking at the schematic of the converter itself, the data from the RM terminal is fed into the UART and strobed into it through a 74121. The strobe from the keyboard is only a high, and it is necessary to have a pulse, so I used the 74121 to make it. The UART is clocked with a NE555, which can be set at any speed. The required frequency for various speeds is:

speed	baud	clock frequency
60	45.45	728 Hz
66	50.00	800 Hz
75	57.00	912 Hz
100	73.70	1179 Hz

It can be seen that the clock speed necessary for the UART is the Baud rate times 16.

Coming out of the UART is the serial data, which is fed into a transistor driver, which keys an optical isolator. These devices are good for more than 20 mA, so if used in an application where the loop is less than this, no problems should be noted. Please note the polarity.

Also coming out of the UART from pin 24 is a terminal which says in effect, "one character is finished;" this is normally high, and when the character is complete, it goes low for a moment. This pulse is fed into a CD4020 binary counter, which can be set up for any count up to 2^{14}. In this instance we are counting up to 64, which is close to the 66 count maximum. Fortunately, the CD4020 is also equipped with a reset, and when the carriage return key on the keyboard is struck, a positive pulse resets the chip, and the process starts all over again. The output from the 4020 is used to drive a flip-flop made from a 7400, but in the configuration shown, another section is used to drive a LED or audible alarm to show that the end of line has arrived. The other flip-flop shown is used for indication of which shift you are in.

This modification can be made quite easily, and the additional circuitry can be put on perf board or a separate printed-circuit board. I built the unit on a wire-wrap card because it was incorporated into another system.

Tim Ahrens, WA5VQK

switching inductive loads with solid-state devices

Recently, while attempting to interface my 8080A microprocessor to a model 26 *Teletype* machine, I destroyed the 8080A and a 2N2222 transistor. The following suggestions may help others from encountering problems when trying to switch inductive loads with solid-state devices. My first attempt at a switching circuit is shown in **fig. 1.** What happens is this:

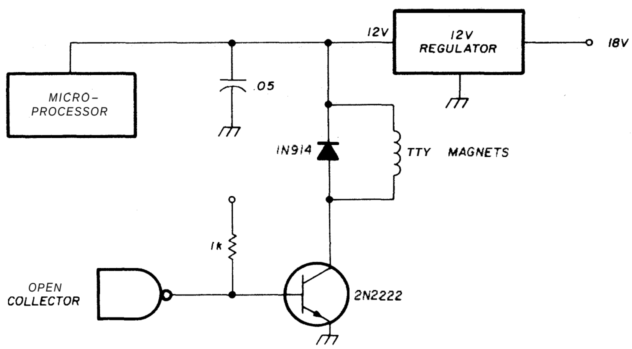


fig. 1. Initial circuit used by WA6ROC to interface an 8080A microprocessor with a teleprinter. This arrangement destroyed both the expensive 8080A and the 2N2222.

1. Assume that the 2N2222 is conducting, the collector of the transistor is at 0.2 volt (it is saturated).

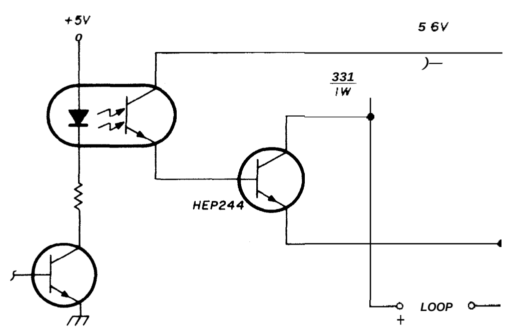


fig. 2. Basic opto-isolator circuit used by WA6ROC to provide protection to the delicate microprocessor.

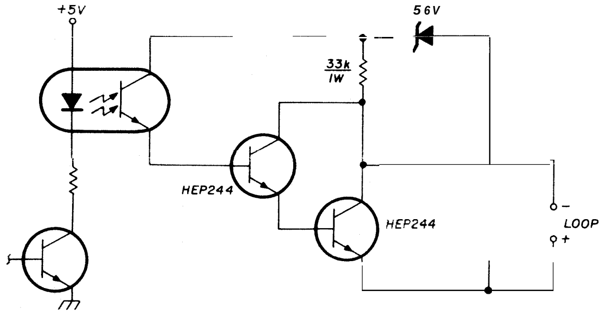


fig. 3. Since circuit of fig. 2 would not key low voltages, a Darlington pair was used, as shown here. Design of the circuit is discussed in the text.

2. The gate tries to turn off the 2N2222.

3. The inductor tries to maintain its current of 60 mA, thus the collector voltage of the 2N2222 rises, as it turns off.

4. When the collector reaches 12.7 volts the 1N914 starts to conduct. The +12 volt regulator, however, will *not* clamp its output at +12 volts (indeed most series-regulated power supplies will allow the output voltage to climb if you try to force current into the output).

5. Thus, the collector voltage of the 2N2222 climbs very high, and it experiences secondary voltage breakdown (fatal). Also, the +12 volt line follows the collector voltage (but it is one diode drop less). Thus the large spike on the +12 volt line destroys the microprocessor (and maybe other devices).

It was at this point that I decided to use an opto insulator — to provide a means of protection of the delicate microprocessor (or other solid-state parts). I tried the circuit which appeared in the November, 1976, issue of ***ham** radio** (see fig. **2).**

This circuit is optically isolated, but it has one problem: it will not key low voltages. Since all I had at the time was a low-voltage power supply, a simple modification was made to allow this circuit to key a considerable amount of current, over a wide range of voltages.

The reason that this circuit will not key low voltage loops is that an appreciable voltage will exist from the collector to the emitter of the HEP244 when it is turned on. Suppose the HEP244 is supposed to key a 60 mA loop, let us further suppose it has a current gain of 60. Therefore, to key the loop, the collector current will be 60 mA and the base current must be

$$\frac{I_c}{B} = \frac{60}{60} = 1\,mA$$

(actually $I_c = 59\ mA$, $I_B = 1\ mA$, $\beta = 59$)

To sustain conduction, the collector to emitter voltage must be

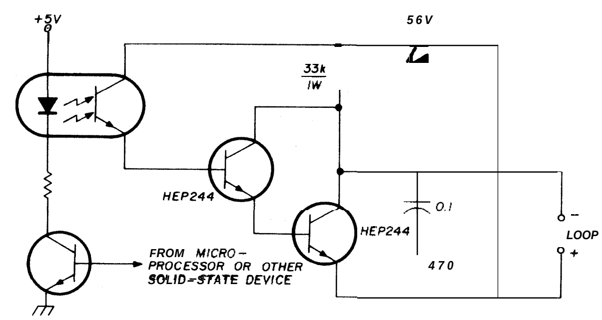


fig. 4. Final circuit configuration used by WA6ROC to interface a teleprinter with an 8080A microprocessor. The RC filter across the Darlington pair speeds up the release time of the print magnets.

V_{be}(HEP244) + V_{CESAT} (opto-isolator transistor) + 1 mA x 33 kilohms = 0.7 + 0.2 + 33 = 33.9 volts.

The obvious way to reduce the collector-to-emitter voltage is to reduce the voltage drop across the 33k resistor. If the current gain of the HEP244 were higher, then it would draw less base current, and the voltage across the 33k resistor would be less. The best way to do this is to add another HEP244 and make the two transistors into a Darlington configuration as shown in fig. **2.**

The effective current gain of the Darlington pair is β^2 so the current through the 33k resistor is about

$$\frac{60\ mA}{60 \times 60} = 167\ \mu A.$$

The collector-to-emitter voltage of the keying transistor(s) is

$$V_{BE2} + V_{CESAT} + 167\mu A \times 33k = 0.7 + 0.7 + 0.3 + 0.55 = 2.25\ volts$$

(versus 33.0 volts before). This circuit will satisfactorily key both low and high voltage loops, and will handle quite a bit more current than 60 milliamps.

Now, as to the matter of protecting the keying transistor from secondary voltage breakdown. The common trick is to clamp the collector of the keying transistor to the magnet supply. If you do this, be sure to follow these two precautions:

1. Use a diode capable of carrying several amps, as the current through the suppressing diode can be several orders of magnitude greater than the magnet current, and small signal diodes (such as the 1N914) will often be destroyed by the large surge.

2. Make sure the magnet supply has sufficient capacitance to adequately absorb the energy of the spike without allowing the voltage to climb.

Sometimes, however, the clamping diode will slow down the release time of the magnets (because the diode allows the current flow to continue for some time). In a case where this is important, an RC filter across the keying transistor will usually solve the problem. Thus the final configuration (fig. **4**) is the circuit I now use to key my teleprinter magnets.

Thomas C. McDermott, WA6ROC

*K Ebneter, K9GSC, and J. Romelfanger, K9PKQ, "RTTY Test Message Generator," *ham radio*, November, 1976, page 30.

comments

satellite tracking

Dear HR:

The article in your September issue, "Tracking Oscar Satellites," was especially interesting, and I might make some comments regarding the nautical mile. Most Maritime nations, including the United States, have adopted the International Nautical Mile, which is exactly 1852 meters, or 6076.10333 . . . U.S. feet.

For most navigational purposes, the nautical mile is one minute of latitude or any other great circle. On the Clarke spheroid of 1866, used for mapping North America, the nautical mile varies from 6046 feet at the equator to 6108 feet at the poles. The length of one minute of a great circle of a sphere having an area equal to that of the earth is 6080.2 U.S. feet. This was the U.S. standard nautical mile prior to the adoption of the International Mile of 1852 meters.

One of the first attempts to establish a standard of length was made by the Greeks, who used the length of their Olympic stadium as a unit and called it, naturally, the stadium. It was 600 Greek feet (607.9 U.S. feet), or almost exactly one-tenth of the International Nautical mile. The Romans got into the act with a 625-foot stadium (606.3 U.S. feet). This is quite close to the British Cable of 608 feet. The Roman Mediterranean mile of 4859.59 feet was gradually replaced by the Greek unit and was probably the mile referred to in the Bible (Matthew 5:41). The word mile comes from the Latin *mille* (thousand) – the one-thousand paces of the Roman mile.

For convenience in short-range plotting in the Navy and Maritime Service, the Radar Plotting Sheet and Maneuvering Board use a 6000-foot (2000-yard) mile, which differs from the Nautical Mile by slightly over 1 per cent. It greatly simplifies range instruments, gear ratios, and computations, and is within the range accuracy of most electronic navigation systems.

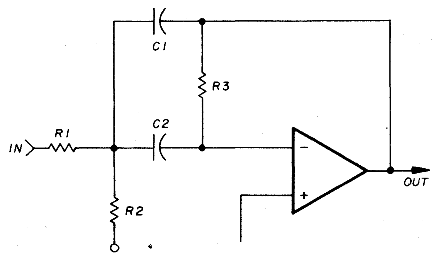
I have written several navigational programs for my HP-97 calculator, and, in great-circle distances and bearing calculations, I use the 1852-meter International Nautical Mile.

I. L. McNally, K6WX
Sun City, California

active filters

Dear HR:

I would like to compliment W4IYB on the fine article concerning active RC filters in the October 1976, *ham radio.* He has presented three basic filter configurations, each with dif-



ferent adaptations. Personally I prefer the second configuration; I have used it in thousands of modems sold to users of the telephone network. I ran off a computer tracing of the bandpass characteristics of the filters, both single and 4-unit combinations, based on a Q of 6. My curves come quite close to what is shown in W4IYB's fig. 2. I should mention however, that the equations you presented are difficult for the average ham radio reader. I use the following simpler equations:

$$R_1 = \frac{Q}{C \cdot f \cdot g \cdot 2\pi}$$

$$R_2 = \frac{1}{2 \cdot (Q - 9/Q) \cdot C \cdot f \cdot 2\pi}$$

$$R_3 = \frac{2 \cdot Q}{C \cdot 2\pi \cdot f}$$

where

$$C = C_1 = C_2$$

g = gain from input to output (usually set to 1 or 0.5)

Q = not more than 10 or 20 for 741s *in* the audio *range*

f = frequency in the audio range

Robert H. Weitbrecht, **W6NRM**
Redwood City, California

noise interference

Dear HR:

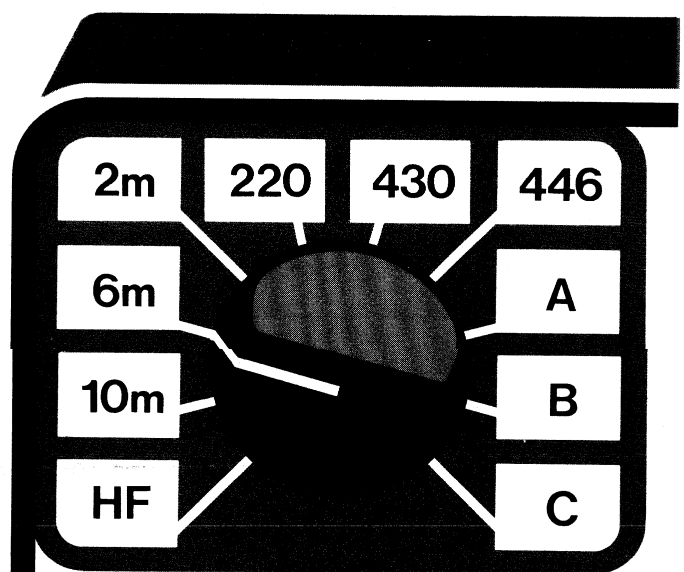
I ran across an unexpected source of interference not long ago. A noise sounding like a machine gun was creating tremendous interference. It covered up to 900 kHz, and with multiples of 900 kHz, up to 30 MHz. The source of the noise was finally found to be originating from the telephone lines, with faulty battery chargers at the substations being the cause. The intensity of the interference was great enough to cover local broadcast stations, even though I live more than 6 miles (9km) from the substation. Even after locating the source, it is very difficult to solve the problem because the telephone company is very reluctant to admit blame.

Keith Olson, **W7FS**
Belfair, Washington

$1.50

ham radio magazine

JULY 1978



vhf communications receiver

ham radio magazine

JULY 1978

volume 11, number 7

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

United States: one year, $12.00
two years, $22.00; three years, $30.00

Canada: one year, $14.00
two years, $26.00; three years, $36.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

rth America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

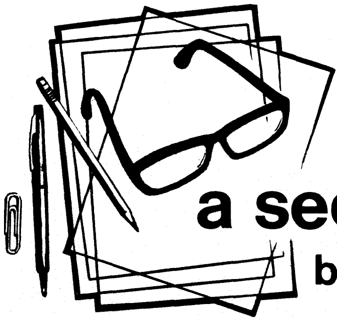
Foreign subscription agents are
listed on page 117

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

In our modern day world of solid-state electronic gadgets and centralized urban living, it's the rare amateur who hasn't been troubled at one time or another by interference complaints. As often as not the interference is caused by some other source, but if you have a tower in your backyard, you're a likely suspect and the first one to whom they turn when the local taxicabs (or whatever) tear up your neighbor's favorite television show or come booming through their quadraphonic stereo system.

As I have mentioned in this column several times in the past, the problem can be effectively cured only by proper design and construction of home-entertainment equipment at the manufacturing level. The consumer electronics business is highly competitive, however, so the manufacturers are reluctant to add filtering and lead bypassing that would increase the sales price of their equipment. For many years the manufacturers contended that less than 5% of home entertainment equipment operated in an rf environment which required special attention – but with the proliferation of two-way radio systems as well as higher power a-m and fm broadcasting stations and high-speed digital systems which can cause interference, I doubt that many consumers would agree.

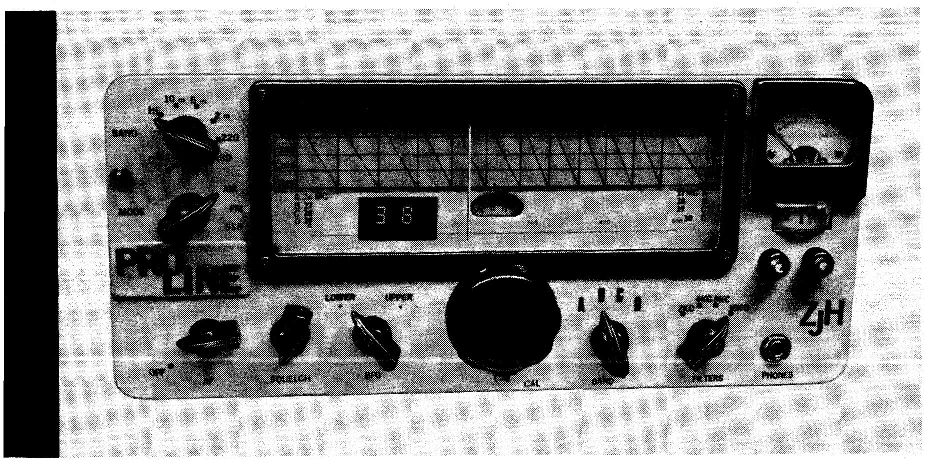
Several bills have been introduced into Congress which would give the FCC authority to regulate the manufacture of home-entertainment devices to reduce their susceptibility to interference from nearby radio transmitters, but none have passed. Now Senator Goldwater is sponsoring a Bill which would require better RFI rejection; the Bill, S-864, has been referred to the Senate Subcommittee on Communications and hearings began in Washington on June 14th. Among those invited to testify were the ARRL, FCC, Institute of High Fidelity, and Heath. Although there's no chance that the Goldwater Bill will make it to the Senate floor during this session, the hearings will help pave the way for speedier action on future RFI legislation.

Consumers are becoming increasingly aware of the RFI problem, so the time is right for legislation such as that proposed by Senator Goldwater. Radio amateurs have known for a long time that the majority of RFI problems are not due to interference per se, but are caused by the interception of signals by devices which were not designed to operate in today's strong rf environment. The only way to eliminate 90% of the RFI problems is through legislation such as S-864 which would eventually require the manufacturers to correct those design deficiencies which lead to unnecessary interference.

Individual amateurs can help toward the eventual passage of a bill requiring better RFI rejection by letting their Senators know of their support for S-864, particularly if one of their Senators is a subcommittee member. In addition to Chairman Hollings (South Carolina), the members are Griffin (Michigan), Magnuson (Washington), Cannon (Nevada), Inouye (Hawaii), Ford (Kentucky), Durkin (New Hampshire), Zorinsky (Nebraska), Riegle (Michigan), Stevens (Alaska), Packwood (Oregon), Schmitt (New Mexico), and Danforth (Missouri). Letters to the Senators addressed to the United States Senate, Washington, D.C. 20510, will reach them promptly and may help considerably.

The letters do not have to be long, although background information on your (or your neighbors') RFI problems could be important. Even a note to the effect that you support S-864 would be a valuable contribution. Remember that previously introduced RFI legislation never made it through Congress – now that Senator Goldwater has started the ball rolling again, let's make sure it has enough momentum to become law. Now is the time to lend your support to this vital effort; write today and make your voice heard.

Jim Fisk, W1HR
editor-in-chief



general-purpose vhf receiver

Design details of a receiver that covers the popular vhf ranges, in one convenient package

Any uhf enthusiast can appreciate a receiver that monitors all vhf frequencies and modes in one small, convenient package. Alas, this sort of receiver doesn't exist in the amateur marketplace. For years at my station, a Collins 75A2 – supported by a bewildering array of converters – did the job. Soon after my station was remodeled, I developed a strong desire to replace the large, unwieldy (and ugly) receiver rack with smaller and modern equipment. I'm an avid homebrewer always looking for new projects to occupy limited time and pocket money, so plans for a new receiver were soon germinating.

design features

Hf operators and shortwave listeners alike have always enjoyed the convenience of general-coverage receivers, so why not something similar in nature, only intended for the vhf regions and tailored to today's needs for diversified vhf operation? Doug DeMaw[1] was on this track some years back when he described a tunable i-f receiver for use with converters. While its abilities fell short of my receiving requirements, several weeks of daydreaming produced on paper a receiver better able to meet my goals, which would incorporate the following features:

By Peter J. Bertini, K1ZJH, 20 Patsun Road, Somers, Connecticut 06071

1. Four-MHz coverage, through a 26-30 MHz i-f range, to allow full reception of the 6-, 2- and 1-1/4-meter bands without changing converter crystals. Dial readout was desired to at least I-kHz resolution with mechanical and electrical stability for smooth CW and ssb reception.

2. Multimode detection for a-m, ssb, and fm with squelch to allow monitoring all the popular vhf modes.

3. Several selectivity positions for mode compatibility and operating convenience.

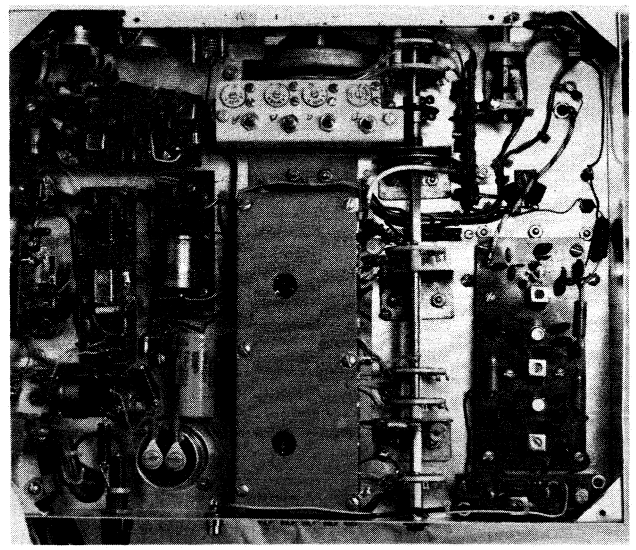
4. All components self-contained in one neat package including all converters, power supplies, and speaker.

A pretty tall order to fill, and obviously some corn-promises must be reached. Adequate coverage of 4 MHz was best done in four 1-MHz segments, starting at 26 MHz and ending at 30 MHz. This allows for quick scanning across a band while maintaining a tuning rate comfortable enough for ssb reception.

The range of 26-30 MHz was chosen for the i-f because many converters come equipped for this range and the i-f is also high enough for good converter image-rejection.

Performance data for the basic receiver, covering the 26-30 MHz range, is presented in **table 1**. I'd like to point out that no pretense is made of using this receiver as the nucleus for a moon-bounce, scatter, or other demanding station-receiver role. Those so inclined will do better with a special-purpose receiver. Templates or board layouts are not available, and this receiver is not intended as a beginner's project.

Schematics of the vhf receiver are shown in **figs. 1** through **10.** The basic receiver, not considering the vhf converters, is a dual-conversion design using the standard frequencies of 10.7 MHz for the first i-f and 455 kHz for the second i-f. Motorola MFE 121 dual-gate mosfets were used in the 26-30 MHz rf preselector and in the first- and second-mixer stages. No peaking of the preselector is required across the range on any of the 1-MHz receiver bands. The vfo bandswitch also selects a set of preselector trimmers for each of the four bands; stagger tuning provides broadbanding and uniform gain over each 1-MHz segment.



Bottom view of the author's receiver. The two boards in the bottom right are the i-f amplifier and the agc detector/amplifier. In the upper left is the bfo/product detector board. Mounted on the left side of the main chassis are the a-m detector and squelch boards. Below and to the right of the product detector is the audio amplifier board. To the left of the audio board is the fm limiter and detector.

The basic receiver input allows for direct monitoring of frequencies between 26-30 MHz should the 10- or 11-meter bands be of interest. Special attention to

table 1. Performance data for the basic receiver.

frequency coverage	26-30 MHz in four 1-MHz bands
circuit	superhet, dual conversion; 10.7 MHz first i-f; 455 kHz, second i-f
sensitivity	0.12 μV detectable in a-m/fm mode; 0.1 μV detectable in ssb mode
noise figure	not measured — estimated at $\approx$2 dB
stability	after I-hour warmup in stable atmosphere, less than 500 Hz per hour
spurious responses	all greater than 50 dB down
i-f rejection	80 dB down
dial accuracy	1 kHz digital resolution. Dial mechanical backlash less than 200 Hz
IMD performance	two 1-mV signals separated 20 kHz required to produce a third order product equivalent to 1.5 μV
agc range	agc action begins at 0.3 μV; i-f distortion at 15 mV
selectivity	2 kHz, 4 kHz, 8 kHz, 16 kHz (13 kHz actual) 455 kHz filters. 10.7-MHz IMD filter is 13 kHz
modes	fm, a-m, ssb and CW detectors
squelch	noise operated, all modes f-m generator and Hewlett-Packard 608D a-m generator used for performance analysis
vhf ranges	2-meter, 6-meter, 1-1/4-meter, 430-434 MHz, 446-450 MHz coverage inboard; three external converter provisions

adequate shielding and power-line bypassing is encouraged. The recent proliferation of 27-MHz CB units increases the likelihood of annoying i-f breakthrough from strong signals in the 27-MHz range. Converters feeding the receiver should be of low or near unity gain to preserve receiver dynamic range. Modern designs without rf amplifiers, especially those employing hot-carrier diodes in double-balanced mixers or mosfet mixer circuits, are ideal. Of course, you can use your own converters; but converters with excessive gain should be followed by an appropriate T-pad attenuator to prevent receiver overload.

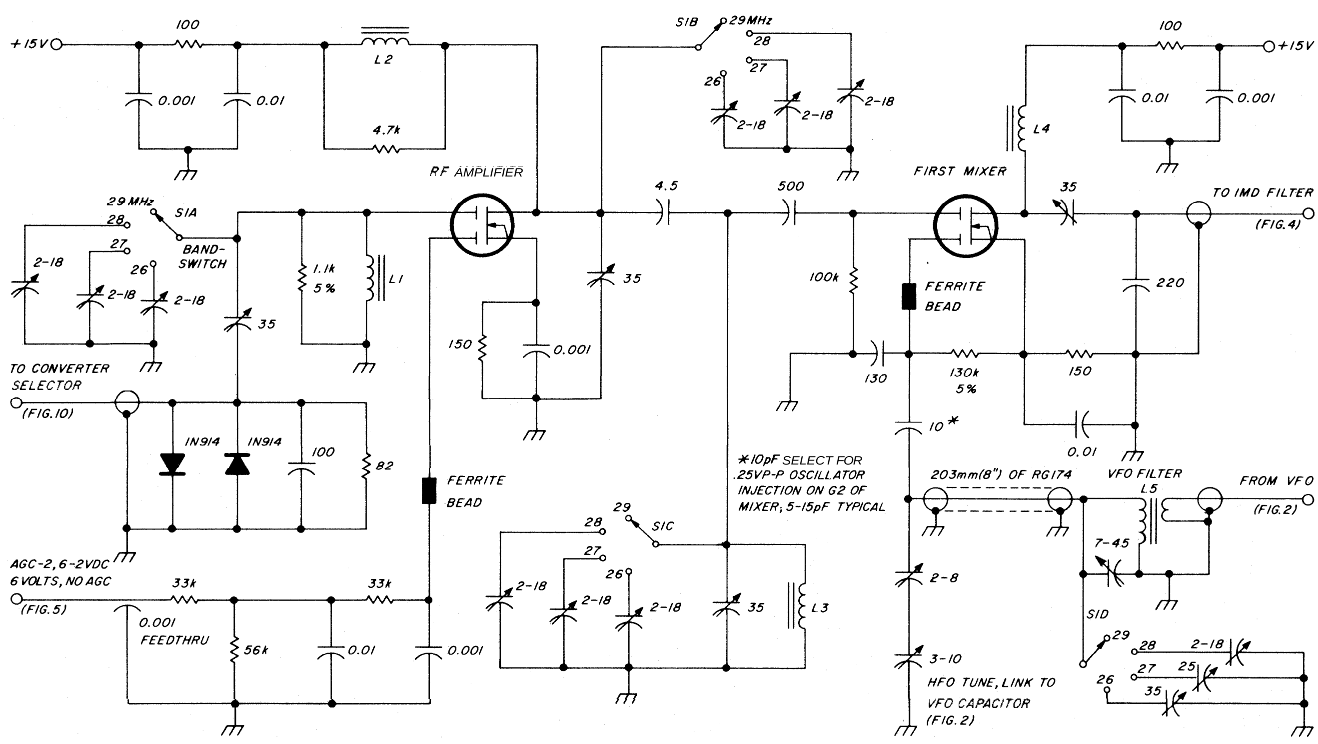
An alternative to the i-f attenuator pads to bring the converters to or near unity gain was suggested by *Hamtronics*. The *Hamtronics* C25-series converters produce between 10-20 dB gain, depending upon the band and device alignment.

The cascade front-end stage in the C25 converters is broadbanded; slight stagger tuning of these stages yields the desired 4-MHz bandwidth. However, the i-f output transformer at 28 MHz has a comparatively narrow passband. Resistive loading of the i-f transformer primary broadens the i-f passband while also decreasing converter gain, which eliminates the need for external attenuators. The approach used on the C25 converters should be adaptable to other makes of converter that exhibit a restricted i-f passband and excessive gain.

The *Hamtronics*-series converters designed by Jerry Vogt, WA2GCF, were used for the vhf converter front ends of this receiver. Kits are available at modest cost. Three *Hamtronics* converters cover the three vhf bands: a P25-50 for 6 meters, a P25-150 for 2 meters, and a P25-220 for 1-1/4 meters.

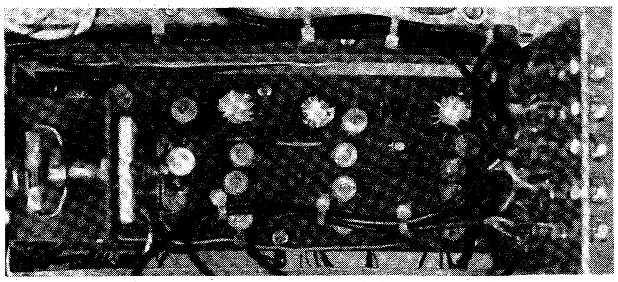
Since the 314-meter band is 30 MHz wide, two uhf converters were needed to monitor this band adequately. One is for the 430-434-MHz DX segment; the other is for the 446-450-MHz range, covering the 400-MHz fm repeater output channels for the northeast corridor of the U.S. These are unity-gain converters, and, when used without an external rf amplifier, don't require the T-pad i-f attenuators.

Receiver use is not limited to amateur frequencies. Suitable converters provide many enjoyable hours monitoring commercial and military air traffic, police, radio-telephone, weather bulletins, municipal and federal government, and much other interesting vhf



L1 - L5	wound on 13 mm (0.5 in.) OD type E toroid cores
L1	12 turns 0.5 rnrn (no. 24) wire
L2	12 turns 0.5 rnrn (no. 24) wire
L3	12 turns 0.5 rnrn (no. 24) wire
L4	30 turns 0.5 mm (no. 24) wire
L5	16 turns 0.8 rnrn (no. 22) wire; link to vfo is 1 turn 0.8 mm (no. 22) wire
S1	6-pole, 4-position bandswitch (see text)

fig. 1. Rf amplifier, first mixer. and vfo filter. The fets are MPF121, SK3050, or equivalent.



Top view of the receiver showing the rf amplifier and first mixer.

activity. Even the hf frequencies can be up-converted, as DeMaw did in his "Receiving Package," reference 1, to produce a truly all-band receiver.

mixers and filter arrangements

Vfo injection from 15.3-19.3 MHz is supplied to gate 2 of the first mixer, (an MFE/MPF121). A Piezo Technology Model 1433 crystal filter, which has 13-kHz bandwidth with a 10.7-MHz center frequency, follows the mixer and acts as an IMD filter, which protects the second mixer from strong out-of-pass-band signals. The 13-kHz bandwidth of this filter sets the maximum receiver selectivity. (It's electrically similar to the KVG XF9A filter.)

A 10.245-MHz crystal-oscillator signal, mixing with the 10.7-MHz i-f signals in the second mixer stage (fig. 4), produces the 455-kHz i-f. Four 455-kHz Collins mechanical filters follow, which select the desired 455-kHz i-f bandwidth. Selectivity positions of 16, 8, 4 and 2 kHz are provided by the four filters.

The use of so many expensive mechanical filters may appear extravagant, but they permit versatility. The 4-, 8-, and 16-kHz filters were salvaged from a demolished R390A i-f strip purchased at a hamfest for $5.00. The 2.1-kHz filter was purchased at another for only $18.00. A 2- or 3-kHz filter will serve the majority of ssb and CW vhf requirements, and a simple LC bandpass filter, made up from i-f transformers, will be adequate for fm or a-m reception if inexpensive mechanical filters aren't readily available. Note that in this receiver, the 16-kHz filter passband is limited to 13 kHz by the selectivity of the Piezo Technology filter. A 20-kHz, 10.7-MHz filter would have improved this situation, but I used materials on hand. The additional cost of a new filter was not justified.

The vfo (fig. 2) operates in the 15.3-19.3-MHz region in four bandswitched 1-MHz segments. Mechanical rigidity and electrical stability are paramount watchwords for a vfo working this high in frequency. Careful mounting of all vfo components and elimination of chassis flexing and dial backlash are important for good vfo performance. The chassis was rigidly reinforced. The Eddystone dial assembly serves admirably.

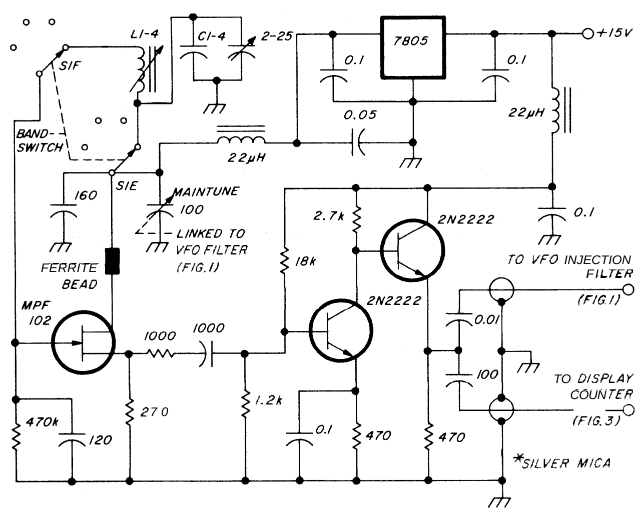
The use of polystyrene caps and good-quality ceramic coil forms and trimmers help contribute to vfo stability. The vfo, built on 3-mm-thick (1/8 in.) glass-epoxy board, was mounted beneath the receiver away from heat-producing components and drafts. The jfet oscillator is powered by a dedicated 5-volt regulator; the low voltage was helpful in reducing drift from rf component heating. The end result is a vfo exhibiting freedom from microphonics and drift, which permits extending monitoring periods without frequent and annoying retuning.

Extensive filtering of the vfo second harmonic was necessary after a problem surfaced during monitoring of 29.6 MHz. Instead of amateur signals, several local CB operators were heard. When monitoring 29.6 MHz, the vfo second harmonic is at 37.8 MHz. When mixed with the 27.1-MHz CB signals spurious responses were produced at 10.7 MHz, the first i-f!

frequency counter

Despite the excellent performance of the Eddystone 898 dial, the 1-MHz spread didn't permit the desired 1-kHz dial resolution, partially because of the physical limitations involved and also because of small inconsistencies in linearity between band segments caused by the bandswitched vfo circuit.

An ideal solution would have been a counter com-



C1 43 pF NPO (15.3116.3 MHz vfo; 26127 MHz receiver)
C2 27 pF NPO (16.3/17.3 MHz vfo; 27128 MHz receiver)
C3 27 pF NPO (17.3/18.3 MHz vfo; 28129 MHz receiver)
C4 not used (18.3/19.3 MHz vfo; 29/30 MHz receiver)
L1 - L4 9.5 mm (318 in.) OD core ceramic. 7 turns 0.6 mm (no. 22) enamel wire

fig. 2. Vfo and buffer amplifiers.

puting the vfo, bfo, hfo, and vhf converter oscillator frequencies to give an exact frequency readout. The cost and complexity of such a counter, and the likelihood of generating spurious signals from the counter circuits, quickly ruled it out. The decision was made to use a counter, but to count and display only the vfo frequency **(fig. 3)**. Up to the tens of kHz position, there is a direct correlation between the vfo and operating frequency, so a two-digit display supplies a direct readout of the tens of kHz and the receiver operating frequency (in kHz). Above 10 kHz, the Eddystone dial-calibration points and bandswitch position supply the hundreds of kHz and MHz readings. It's easy to include a third display for hundreds of hertz, but remember that, unless the other conversion oscillators are extremely accurate and set on frequency, the cumulative error makes this resolution meaningless. Of course, even a 1-kHz readout requires careful frequency setting and regulated power supplies.

The basis for the counter was a circuit in the January 1976 issue of ***ham radio***.[2] Its simplicity, small size, and low cost made it appealing for this application. It's built on a small 102 x 102 mm (4 x 4 in.) square of glass epoxy vectorboard and is sandwiched between two aluminum plates that provide shielding and a ground plane for the counter. No birdies from the counter were heard in the finished receiver.

I used an MD-640 incandescent 7-segment display in place of LED displays. The MD-640s are brighter, cheaper, and don't require current-limiting resistors for each segment as in the case of their solid-state counterparts. The display was also more uniform than that produced by most bargain-basement LED

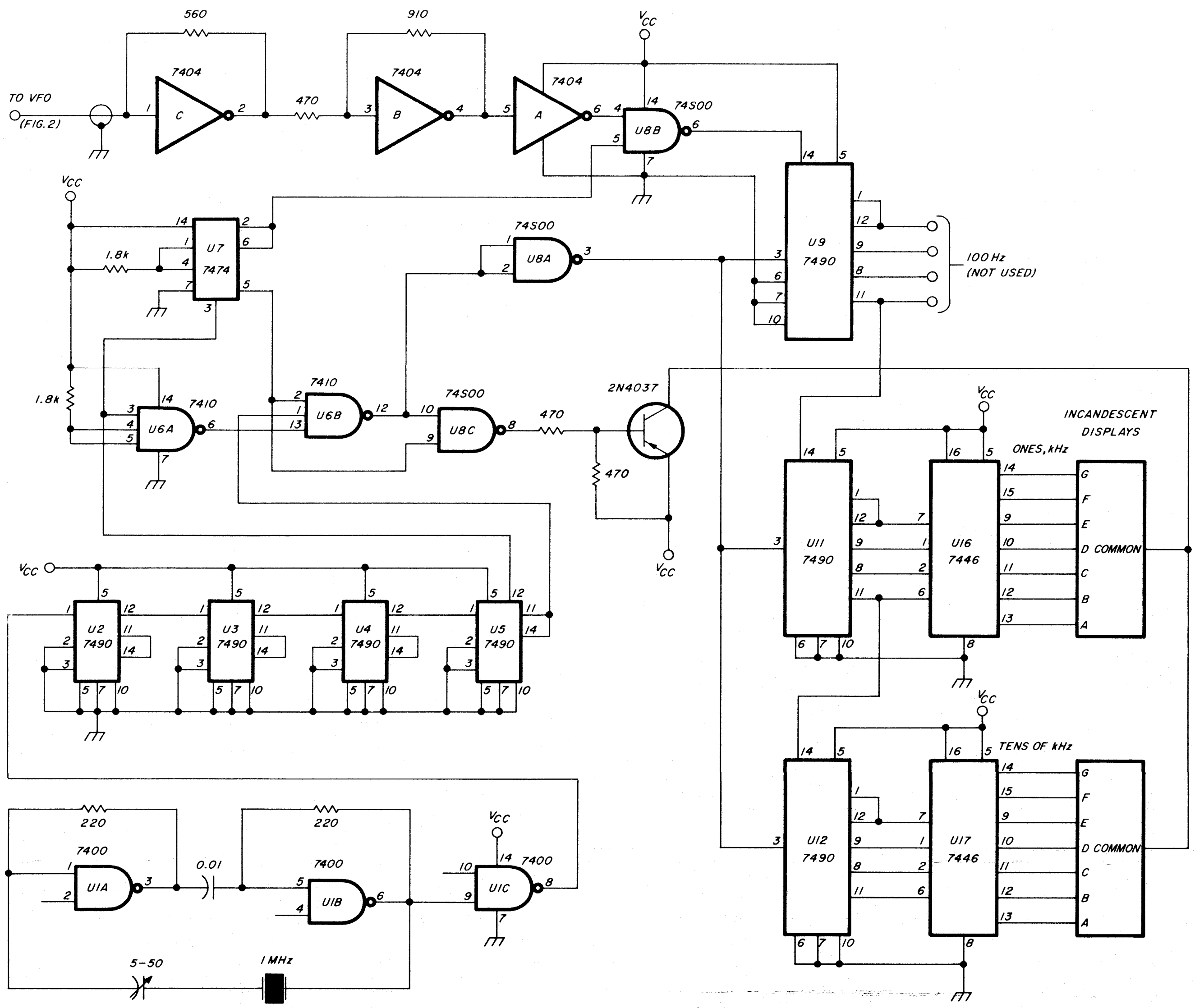
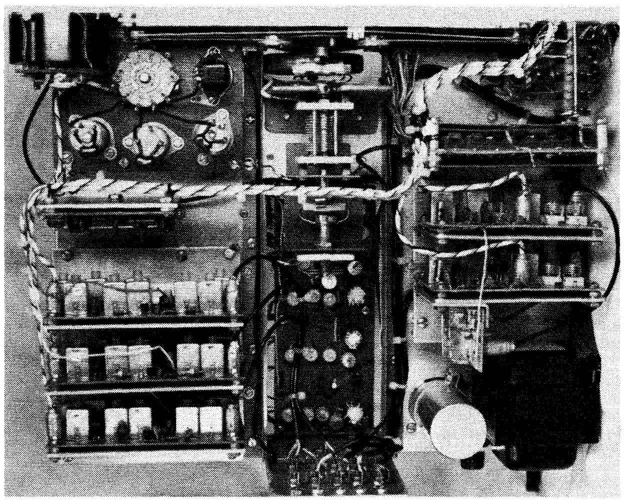


fig. 3. Frequency display. IC numbers refer to the article in the January, 1976, issue of *ham radio* (reference 2).



Overall top view of the receiver. The i-f filters are mounted in the upper left. The ICs for the counter are mounted in the upper right. Note the shielding on both sides of the counter board.

displays. A window for the readouts was carefully milled through the steel Eddystone dial plate. A piece of gray plexiglass behind the window provides contrast for the display readouts.

i-f strip, agc detector, and amplifier

The output of the selected 455-kHz filter goes to our first 455-kHz stage, an MPF102 fet **(fig. 5).** This stage provides some gain and a good termination for the relatively high filter impedance, but primarily it's an agc attenuator.

The next two stages of i-f amplification employ two high-gain MC1550 ICs. The i-f interstage transformers, from the Radio Shack general replacement line, are resistive loaded rather heavily for stability and to ensure the i-f amplifier is broadband enough not to restrict the broader selectivity positions. (While aligning the i-f amplifiers, it may be necessary to remove the loading resistors to see the peak at resonance.) The last i-f amplifier is a single 2N3904 transistor stage. It is not under agc control and is designed to have sufficient output to drive the agc detector and demodulators.

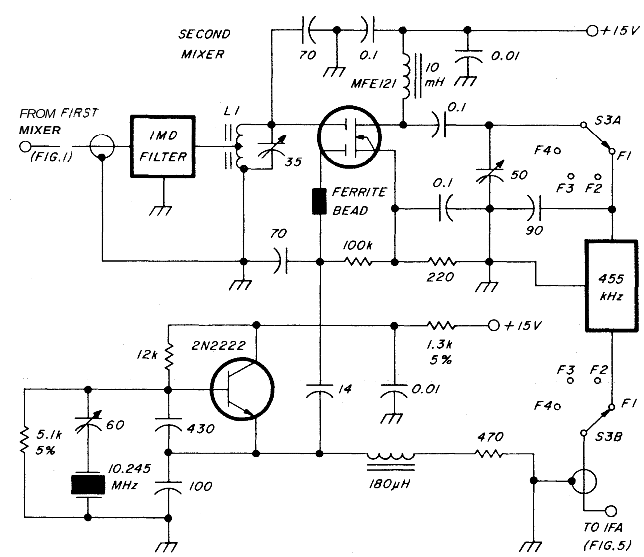
A pair of germanium diodes in a voltage doubler rectifies the i-f signal and presents a proportional dc level to a 2N3053 dc amplifier. The 2N3053 provides an increasing voltage potential during periods of agc action for reducing the MC1550 i-f stage gain and conversely, a decreasing dc potential for gain reduction in the MPF102 i-f amplifier and MFE121 preselector.

A selectable RC time constant controls the agc response; for simplicity the mode switch selects the agc time constant appropriate for the mode selected.

Agc voltage is also used to provide the signal strength meter reference voltage. If the gains of all the converters are equalized, the meter may be calibrated in microvolts instead of just providing a relative signal-strength indication. My converter selector switch also provides agc voltage to the converters, but external agc was not advised for use with the *Hamtronics* converters.

Ssb or CW reception is accomplished with a hot-carrier-diode product detector circuit inspired by another article.[3] The bfo is on the same board. Because of the high cost of 455-kHz crystals and the advantages of a variable bfo, the tunable bfo route was taken. Use of Radio Shack transformers was again made in the product detector and bfo circuits **(fig. 6)**. Note that several volts p-p of bfo energy are required for diode saturation and proper operation of the detector. The bfo signal is amplified to prevent pulling and to develop ample bfo injection voltage. Recovered audio is clean sounding and not fatiguing, indicating low harmonic distortion from this circuit.

The a-m detector **(fig. 6)** is simple and requires little explanation. A half-wave rectifier, consisting of a slightly forward-biased hot-carrier diode for improved low-level signal detection, rectifies and detects the a-m signal. A low-noise audio preampli-



F1 - F4	Collins 455-kHz mechanical filters (2, 4, 8, and 16 kHz used)
IMD filter	Piezo Technology model 1433; 6-dB bandwidth: 13 kHz
L1	wound on 13-mm (0.5 in.) OD type E toroid (red) core. 26 turns 0.8 mm (no. 22) enameled wire; tap 10 turns from cold end
S1	4-position, 2 poles; shield between wafers
S3	filter switch; 2-pole, 4-position wafer

fig. 4. Second mixer.

fier increases audio level. As with the product detector, a-m audio is clean and pleasant sounding.

fm detector

The 455-kHz i-f signal directly feeds the fm detector board, bypassing the mode-selector switch that feeds the a-m and ssb detectors **(fig. 7)** as selected. A single Motorola MC1355 i-f amplifier and limiter 14-pin IC performs all fm signal-processing functions.

hybrid module designed for use in their Tac-Tec series vhf-uhf fm portable communication radios **(fig. 8)**. Unfortunately, the exotic device is available only directly from RCA or one of their authorized two-way service centers. Distributor cost is around $28.00; user suggested price is close to $38.00.

The 432141 is noise operated. A 390-pF coupling capacitor from the recovered fm ratio detector audio provides the high-frequency audio noise components

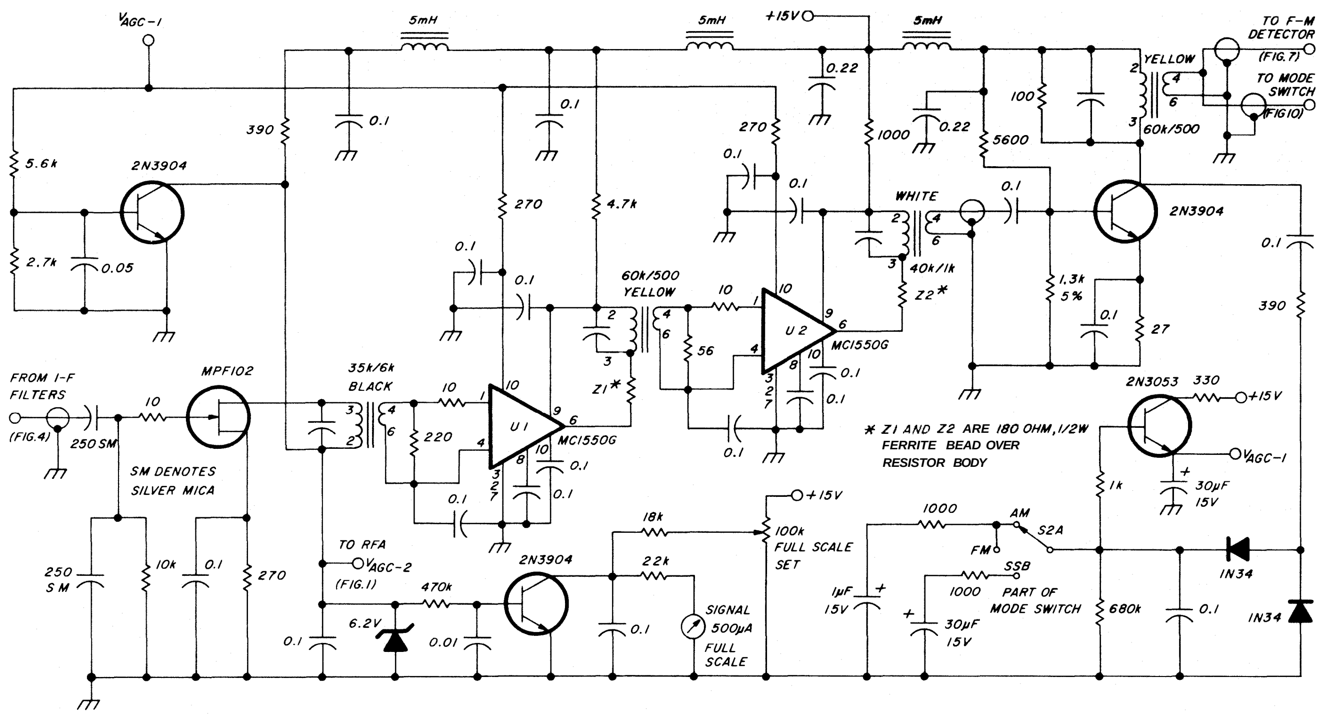


fig. 5. I-f amplifiers and agc detector and amplifier.

Originally I had planned to use the Miller type 8806 discriminator transformer with the MC1355, but after a two-month wait on a back order my distributor shipped me the 8805 ratio detector as a substitute. Minor circuit changes will allow use of either transformers with comparable results.

Both the ratio detector and discriminator circuits provide a plus or minus dc voltage to indicate proper tuning of the received frequency. The detector will drive a zero-center microammeter directly. The meter I used had a zero-center, ± 6 V movement. A dc operational amplifier was necessary to drive it.

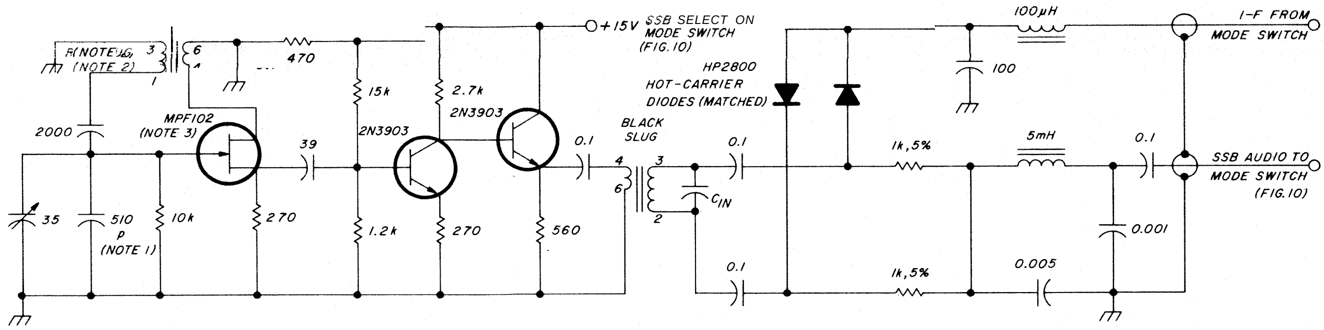
Center-tune meters are commonly associated with tunable fm receivers, although they're useful in tuning a-m signals as well. For this reason, and for squelch operation, the fm detector is operational in all modes and isn't affected by mode-switch position except for the selection of fm audio.

The squelch circuit centers around a single RCA

for squelch operation. A 50k front-panel pot allows setting the squelch threshold point.

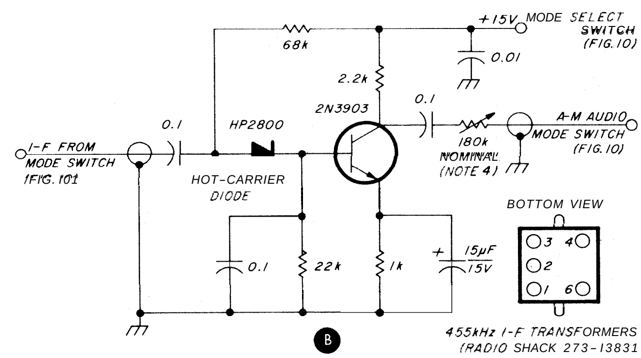
The design of the LM-380 audio amplifier **(fig. 9)** provides a convenient method of squelch control. One pin of the LM-380 is for optional bypassing of an internal voltage divider supplying operating bias to early amplifier stages of the LM-380. The squelch-gate output (pin 12) of the 432141 module, **fig. 8,** holds this bias point at ground to mute the receiver. Because of the dc-coupling design of the LM-380, a simple RC time constant between the squelch module and audio PA prevents an annoying speaker "pop" during squelch action. External receiver rnuting is also provided by supplying an external ground to the same point on the LM-380.

The versatile 432141 squelch module also has provisions for a time constant, provided by an RC network, to prevent receiver squelch action while receiving rapidly fading signals from mobile stations. In this



NOTES
1 p DENOTES POLYSTYRENE CAPACITORS
2 RED CORE DENOTES BROADCAST-BAND LOCAL OSCILLATOR COIL
3 MPF102 SHOULD BE HIGH GRADE A 2N5486 CAN BE SUBSTITUTED
4 REDUCE FOR MORE AUDIO

A



B

fig. 6. Ssb beat-frequency oscillator and product detector, A, and a-m detector, B.

receiver the time constants are mode-switch selected for best performance. A 120 microsecond squelch dropoff is used for a-m and fm signals; an appreciably longer delay is provided for ssb signals. Pin 4 of the 432141 module is an inverted output of the pin-12 squelch gate, which mutes the LM-380 audio PA. I used pin 4 to light a front panel call lamp through a dc amplifier to indicate band activity.

Recovered fm audio is fed through an active audio filter in the 432141 module (3 dB gain) for conventional 6-dB-per-octave de-emphasis audio processing of the received signal. Note that pins not shown on the schematic for the 432141 module are active and are used for special applications of the RCA radios: quiet channel and fast mute. All unused pins should be unterminated.

As the 432141 requires only 10 volts for proper operation, a 5-volt zener drops the 15 V dc supply bus to a suitable level. The RCA-module pins are not keyed; refer to **fig. 8** for pin alignment. Caution: The chip can be installed 180 degrees around, and will be damaged if done so.

Since the receiver could be used for casual monitoring of various citizen and amateur services over its basic 26-30 MHz i-f tuning range, sensitivity and noise figure were contributing factors in its design criteria. More often than not, converters for frequencies above 400 MHz employ passive mixing devices, often without the aid of an integral rf preamplifier. Since these converters exhibit negative gain, not only does the mixer noise figure play a large role in system performance, but also the noise figure and

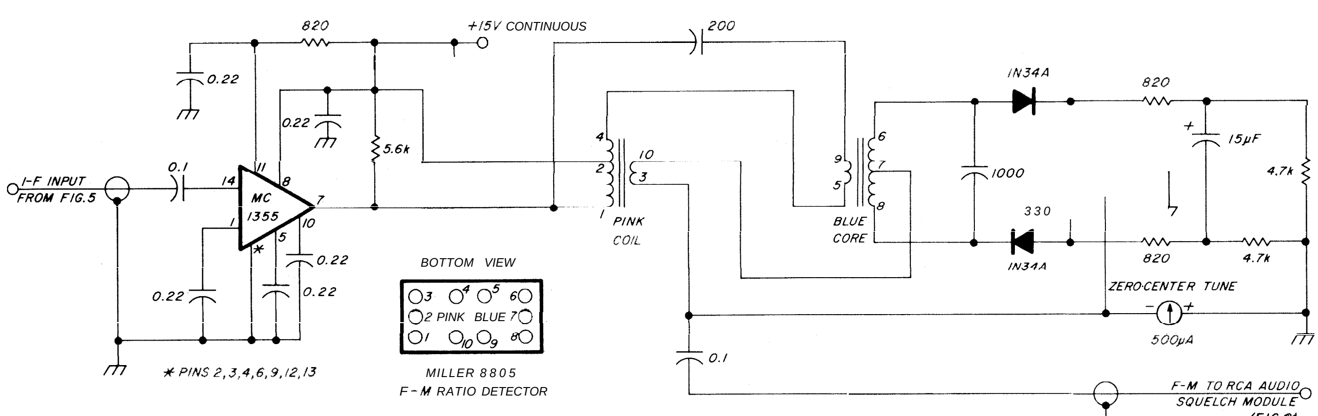


fig. 7. Fm limiter and ratio detector.

sensitivity of the i-f strip are important if optimum results are to be realized.

A single stage rf amplifier is used, using a Motorola MFE121 dual-gate mosfet **(fig. 1)**. Agc control over the RFA is through the dc biasing level on gate 2. A small ferrite bead directly on the gate-2 lead inhibits parasitic uhf oscillations. The 20-30 MHz input and output coils of the RFA are resistive loaded to improve bandwidth, stability, and to reduce front end gain. While the resistive loading provides sufficient bandwidth to allow operation over each 1-MHz range without cumbersome preselector tracking capacitors, additional trimmers are bandswitched on the

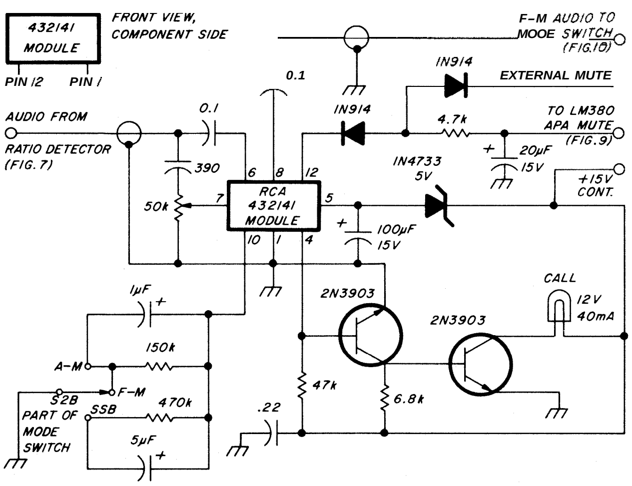


fig. 8. Noise squelch circuit for fm, a-m, and ssb reception.

lower three receiver ranges for proper RFA operation.

A *Minilabs* MLA-1 double-balanced mixer was tried in the first version of the receiver. Exotic power fets for impedance matching, high local-oscillator-injection requirements, and cost soon eliminated this scheme. The old axiom "simplest is often best" was proven in the final circuit used for the first mixer. Another MFE121, using conventional gate-2 local-0scillator injection, is employed. A ferrite bead again is required on gate 2, as in the RFA stage. Impedance transformation between the mixer output and the 10.7-MHz filter is through a capacitive divider across the mixer output tank circuit.

Vfo injection to the MFE121 mixer is filtered through a simple single LC toroidal stage, which reduces vfo harmonics and subsequent spurious receiver responses, as mentioned later. Because the vfo range covers 15.3-19.3 MHz in four 1-MHz steps, bandswitching of trimmers, as in the RFA stages, was required to resonate the filter on three lower ranges. The relatively low vfo injection frequencies

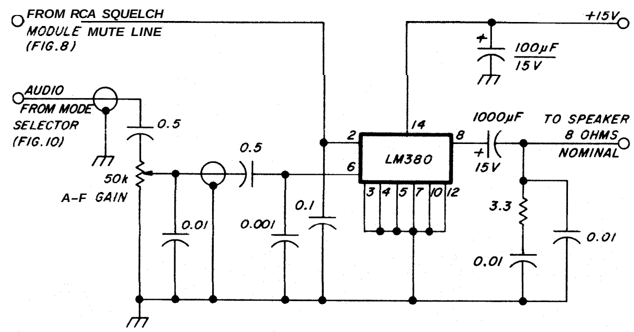
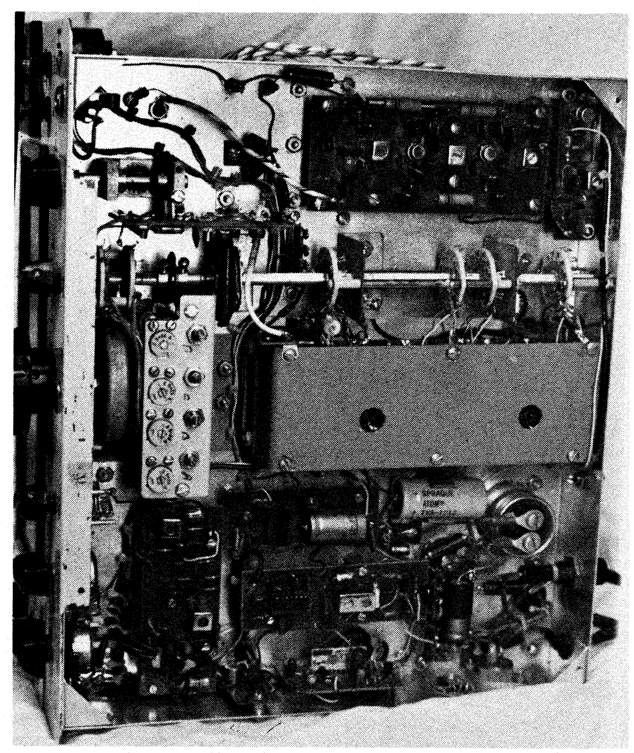


fig. 9. Receiver audio circuit.

and the desired high circuit *Q* prevented broadbanding of this stage. As a solution, a 3-10 pF variable capacitor mechanically linked to the vfo main tuning capacitor provides filter tracking with the vfo frequency. The purpose of the 2-18 trimmer in series with the vfo tracking capacitor **(fig. 1)** is to set a 1-MHz tuning range for the filter.

During alignment considerable back-and-forth tuning and peaking are required to adjust the vfo ranges, tracking, and LC-filter range.

The 10 pF capacitor coupling the LO injection to the mixer was empirically chosen. At 29.6 MHz, the



The right-angle gear drive is used to select an i-f filter mounted on the top of the chassis. The four trimmers, behind the dial assembly, are used in the main receiver vfo to select the different frequency segments.

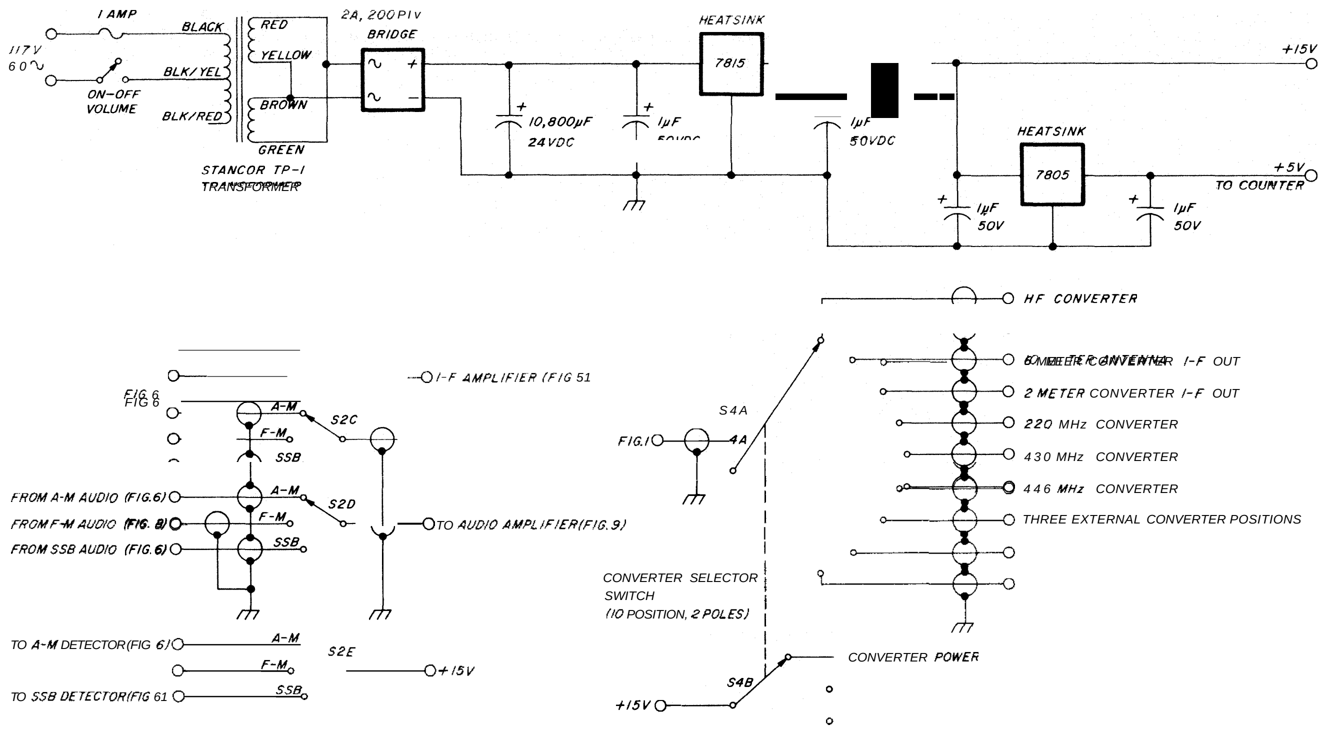


fig. 10. Power supply and switching arrangements.

vfo operates at 18.9 MHz. The second harmonic of the vfo is 37.8 MHz. If sufficient 37.8-MHz harmonic energy reaches the mixer, signals at 27.1 MHz are readily converted to the 10.7 i-f output. If the coupling capacitor is too large the harmonic injection becomes excessive; if it's too low mixer gain suffers. My receiver (worst case) has close to 50 dB of spurious rejection, or a 300-microvolt signal on the spurious frequency will produce a signal equivalent to 1 microvolt on the operating frequency.

construction

The photographs show placement and mounting of the major receiver components. The cabinet and chassis is a LMB type CO-1 enclosure. The five *Hamtronics* converters were mounted vertically on aluminum plates for space conservation and rf shielding. The two uhf converters were mounted on the left top side of the chassis, while the three vhf units flank the right side. The vfo counter is mounted vertically between the front panel and the uhf converters; the counter board is sandwiched between two aluminum plates for shielding. The chassis center was used for the 28-MHz rf amplifier and first mixer. The board is recessed below the chassis for access to the bandswitch assembly.

Behind the S-meter and the 455-kHz Collins filters, another vertical shield supports the PC-board assembly for the 10.7-MHz IMD filter, second mixer, and the second conversion oscillator.

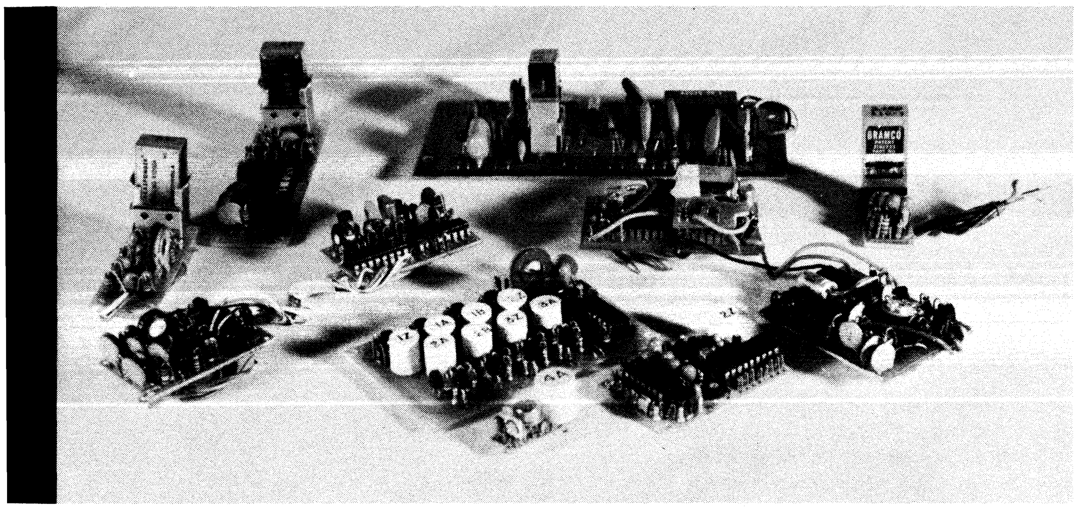
The bottom of the chassis is dedicated to the power supply components, left rear; the 455-kHz i-f stages and agc detector, right side; the vfo components, front center; multimode detectors, the squelch, and bfo, left front.

The chassis was reinforced along the cutout for the 28-MHz front end to minimize chassis flexing. The vfo tuning-capacitor supports are of heavy-gauge metal for mechanical rigidity. The bandswitch assembly transverses the entire width of the chassis, front to rear. It was constructed from several disassembled switches salvaged from flea markets. Low-loss ceramic wafers are recommended. The first two wafers are for vfo bandswitching; the third is for the vfo injection-filter trimmers. Wafers 4, 5, and 6 are for rf amplifier bandswitching. L-shaped aluminum brackets were placed between wafer sections 2 and 3, 3 and 4, and 5 and 6 for mechanical support and rf shielding.

references

1. Doug DeMaw, W1CER, "Receiving Package for 30 to 144 MHz," The *Radio Amateur's Handbook,* American Radio Relay League, 1974 edition.
2. Jim Pollock, WB2DFA, "Six-Digit 50-MHz Frequency Counter," *ham radio,* January, 1976, page 18.
3. Mike Goldstein, VE3GFN, "A Practical Discussion of Product-Detector Operation," *ham radio,* October, 1969, page 12.

ham radio



subaudible tone encoders and decoders

A review of the latest two-meter directory confirms the impression I obtained from vacation trips and from amateurs visiting the greater Cleveland, Ohio area: Most amateur repeaters are still carrier accessed. However, in localities where unoccupied two-meter pairs have become scarce and intermod problems on all bands more prevalent, some form of tone access is becoming more common. This situation is particularly noticeable in the larger east- and west-coast metropolitan areas and along the Great Lakes. Some repeaters have optional guard systems that are turned on and off automatically or by the control operator, as conditions require.

Both tone-burst and ***Touch-Tone*** access control are used, but the most popular method seems to be continuous subaudible tone, commonly known as ***PL,*** from the Motorola trade name for the system "Private Line." A selectable guard system[1] that uses ***PL*** has been in use on the Cleveland 16/76 repeater for some time and has led to considerable interest in various types of encoders and decoders. This article covers experiences that other club members and I have obtained about encoders and decoders we've bought or built, tried and discarded, or adapted to our use.

First, let's look at some of the reasons for using ***PL.*** The advantages on a control or link frequency to which access is strictly limited are obvious. Anyone who is a repeater control operator in an area where more than one machine can be heard on the same frequency can appreciate the advantage of having an encoder on the repeater transmitter and a decoder on his monitor receiver.

PL on the repeater input also helps to minimize interference caused by intermod and sources other than amateur transmitters. In crowded areas, individuals or small groups looking for a frequency on which they can experiment or operate a special-purpose repeater, can share the same channel with much less distance between their stations than would be required without ***PL.*** My point is that many reasons exist for using continuous subaudible tone on input or output other than a wish to operate a closed repeater.

reed-type encoders

The encoders most used on control and link frequencies, and by operators with converted commercial gear, are those in which a resonant vibrating reed establishes the tone frequency. The advantages of these encoders over other types of low-frequency oscillators include 1) reliability and stability under temperature extremes and supply voltage changes, and 2) ability to change frequency by merely plugging in a new reed. In addition, the reed encoder generates a pure sine wave, which does not need filtering.

By Pat Shreve, W8GRG, 2842 Winthrop Road, Shaker Heights, Ohio 44120

Disadvantages include the cost of reeds (particularly if the user wishes to work several repeaters with different *PL* frequencies), and size. It is difficult if not impossible to fit a reed encoder into many of the popular hand-held and mobile transceivers.

Early popularity of reed encoders and decoders led to the adoption of the standard commercial subaudible tone frequencies for amateur use, shown in **table 1.** These have carried over into the design and production of other types of equipment. Two reed encoders, a decoder, and a combination encoder/decoder are shown in the photographs. The circuit of the Communications Specialists* miniature encoder is shown in **fig. 1**. With a Motorola Vibrasponder reed, it produces a clean sine wave to 3.4 volts rms; it will go higher, but the wave peaks will be clipped. These test results and those that follow were obtained with a 12.5-volt regulated supply voltage; Output was measured with a precision ac voltmeter, and the wave form was evaluated by

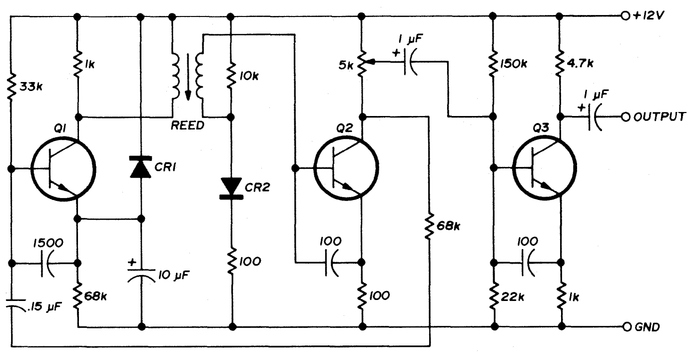


fig. 1. Communications Specialists miniature reed encoder. CR1 and CR2 are silicon diodes. Q1, Q2, and Q3 are general-purpose npn silicon transistors.

comparison with the output of a Heath IG-1B wave generator using a dual-trace oscilloscope. Tone frequency was 110.9 Hz.

Fig. 2 is the circuit of a subminiature encoder built by our club according to a design used in some Motorola equipment. It is smaller than the original or the Com Spec unit. Where space is a problem, the reed and socket can be separated from the PC board. It produces an equally good waveform, but has much lower output: from 0.22 to 0.35 volt rms, depending on the reed. It works well if the transmitter has sufficient audio amplification between the PL injection point, which should follow any speech filters, and the modulator.

tunable oscillators

A number of tunable oscillator circuits have been

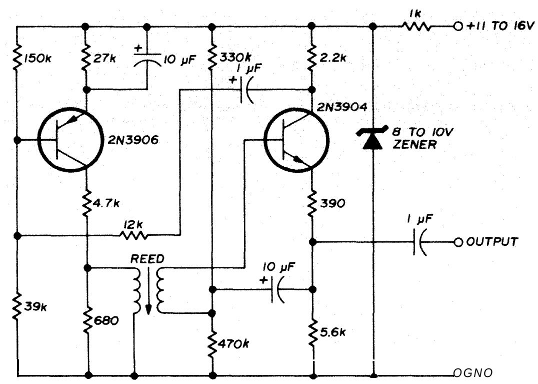


fig. 2. Lake Erie ARA subminiature reed encoder.

tried as PL encoders by repeater groups in or near Cleveland. The most popular was the twin-T circuit shown in **fig.** 3, which came to our club from the Great Lakes repeater group in Detroit. It is compact, inexpensive, and can be assembled from readily available parts by anyone with a minimum of experience or equipment. For satisfactory performance the frequency-determining capacitors must be molded Mylar or polycarbonate components, and precision 1% resistors should be used where shown. Even then, the circuit needs to be retuned occasionally and will give trouble in a mobile installation parked in a Lake Erie winter or in a desert sun. The wave shape is satisfactory, but the load and bias resistors may have to be changed for different output frequencies to prevent distortion.

I also experimented with tunable encoders designed around a function generator such as the Intersil 8038. Several pilot units showed promise, but none fully overcame problems of rf sensitivity and need for a more stable supply voltage than was easily obtainable in a mobile installation.

digital encoders

Use of a crystal-controlled oscillator to generate a stable frequency is nothing new; but until multistage dividers on a single IC chip became readily available,

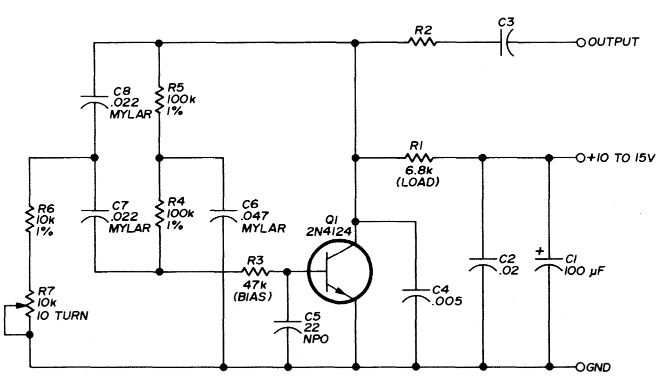
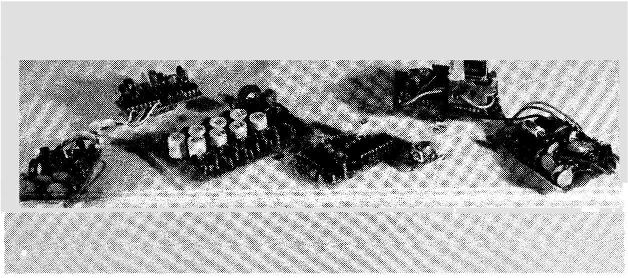
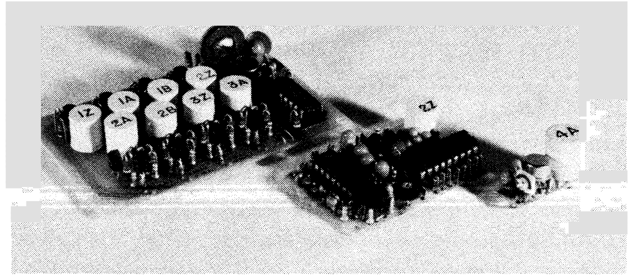


fig. 3. Twin-T oscillator. C1 is not required in battery-powered portables. Select R2 to give desired modulation level.

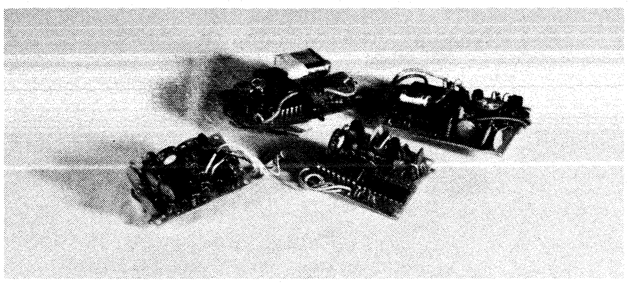
"Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92667.



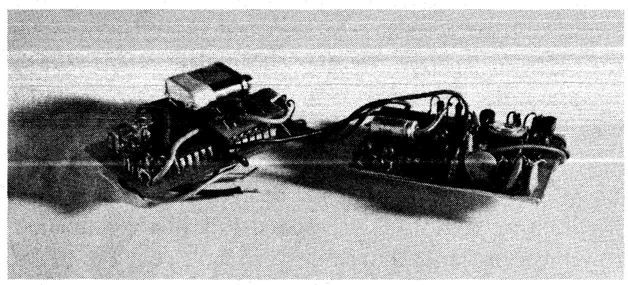
Ceramic resonator and crystal-controlled encoders and encoder-decoders.



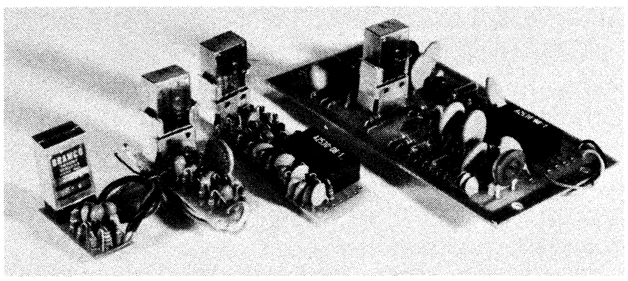
Communications Specialists eight-frequency encoder, encoder-decoder, and sub-miniature encoder.



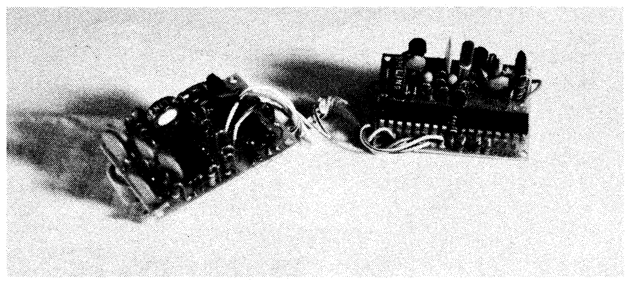
Crystal-controlled encoder-decoders.



Encoder-decoder designed by the author.



Two reed encoders, reed decoder, and encoder-decoder.



Adcom crystal-controlled encoder-decoder.

it was not practical to use high-frequency crystals to generate the low frequencies used in a subaudible tone encoder. Development of CMOS ripple counters, capable of division by factors in the thousands or millions, eliminated the need for bulky divider chains in low-frequency generators and timers and at the same time eliminated the need for a regulated 5-volt power source and sometimes difficult shielding against rf and external noise.

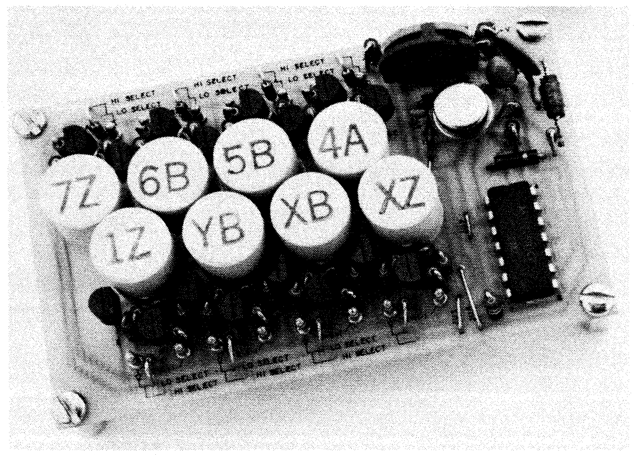
Descriptions of the operation and application of a number of these dividers are found in manufacturers' publications.[2,3] Those of most interest for *PL* use are the 4020 and 4060, both capable of division by 16,384 (2^{14}). The 4020 will accept input frequencies to 7 MHz; the 4060 to 4 MHz. The 4060 includes an oscillator circuit that can be crystal controlled.

Two approaches to the use of a multistage divider to reduce a crystal frequency to the *PL* range are possible: Use the full range of the divider and select a crystal that will give the desired output, or program the divider to give any desired output from an available crystal.

Let's look at the unprogrammed divider, in which the crystal is selected for a specific output frequency. Only two ICs are needed, a 4020 divider and a 4030 exclusive-OR gate, which serves as crystal oscillator and digital-to-analog (D-A) converter. The crystal frequency is the desired output multiplied by 16,384. The circuit in **fig. 4** is such an encoder, designed and used by members of the Lake Erie Amateur Radio Association (LEARA), which operates the Cleveland 16/76 and 28/88 repeaters. The choice of whether to use a 1.0- or 2.2-μF output-filter capacitor depends on whether you want a stronger signal (use 1.0 μF) or a cleaner waveform (use 2.2 μF). The unit leaves something to be desired in both respects. A better design could be worked out with a 4060 using the internal oscillator and substituting an operational amplifier wired as a lowpass filter for the 4030 D-A converter.

Some may ask why a D-A converter is needed at all. Certainly a square-wave digital output will modulate the transmitted signal; many solid-state CW identifiers use such an output. The trouble is, that on most amateur transmitters, square-wave modulation is far from subaudible. Many of the harmonics in the square wave fall in the audible range, and the resulting signal can be very unpleasant. Reed encoders have a clean sine-wave output.

Another device equivalent to crystal control is used in several Communications Specialists encoders. It is a ceramic resonator much like an i-f filter operating between 250 and 500 kHz. The small size of the resonator and a special IC make possible the Com Spec microminiature ME-3 tone encoder, which is hardly larger than a postage stamp. The ME-3 circuit is shown in **fig. 5.** The special IC contains the oscillator, divider, and gates, which form a lowpass square wave to sine-wave converter. The output is a clean sine wave adjustable to any level to 3.2 volts rms. The output frequency can be changed by plugging in a different resonator. Similar circuitry is used in the ME-8 encoder, which provides for selection of one of eight frequencies by electronically switching the resonators, and in the combination encoder-decoder discussed later.



Eight-frequency encoder with ceramic resonators.

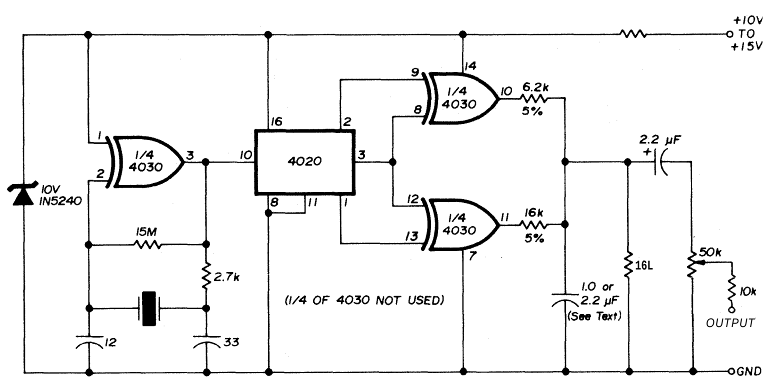


fig. 4. Lake Erie ARA crystal-controlled encoder.

Encoders in which a single crystal is used to generate multiple output frequencies can also be built with a 4020 or 4060 CMOS divider. Both these ICs can be reset to zero at any point in their counting cycle by a high-level input to the reset inverter. Since outputs are available from all divider stages from **4** through 14, diodes can be used to combine outputs to give a reset pulse after any combination of 16 oscillations of the crystal.

In my experiments, the range of crystal frequencies has been limited on the high side by divider capability and on the low side by crystal cost. For reasons explained later, the last divide-by-four step is performed by a separate device, so the divider output should be $4f_{PL}$, where f_{PL} is the desired encoder output.

The lowest of the standard tones in **table 1** is 67.0 Hz. The maximum capability of the divider is 2^{14} or 16,384, so the top limit on the crystal frequency is $67.0 \times 4 \times 16{,}384 = 4.391$ MHz.

My lower limit is 3 MHz, based on the price of an International Crystal general-purpose crystal, which is lowest in the range between 3.0 and 10.99 MHz. To illustrate how the divider is programmed, assume a crystal frequency of 3.066 MHz and a desired output of 110.9 Hz. The division factor is

$$\frac{3{,}066{,}000}{110.9 \times 4} = 6912$$

subtract 2^{13}	4096
	2816
subtract 2^{12}	2048
	768
subtract 2^{10}	512
	256
subtract 2^{9}	256
	zero

This example shows that when Q9, Q10, Q12, and Q13 outputs are all high at once, the counter will have divided by exactly 6912. If four diodes are connected with anodes to these outputs of the IC, and a common cathode lead is connected to the reset input, the counter will reset to zero after dividing by 6912. A similar calculation will show that diodes con-

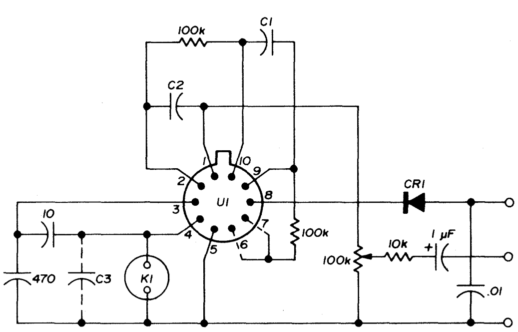
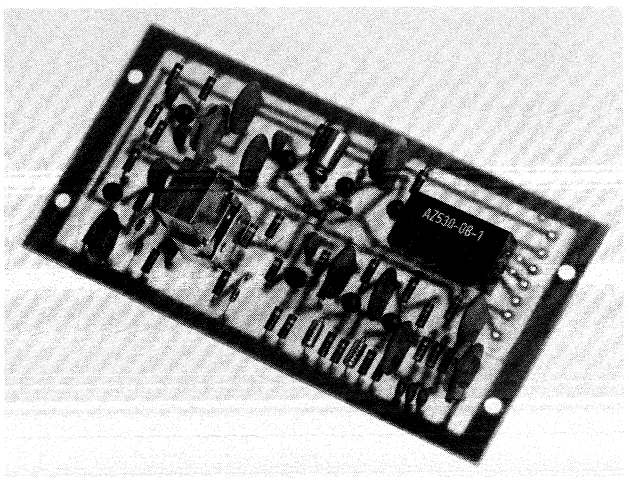


fig. 5. Communications Specialists micro-miniature encoder. Values of C1, C2, and C3 and connection to pin 6 or 7 of U1 depend on frequency. U1 is a custom-made IC; K1 is a ceramic resonator.



Reed encoder-decoder circuit board.

nected to 010, 011, 012, and Q13 will result in a division factor of 6780 and an output of $3{,}066{,}000 \div 7680 = 399.21$ *Hz*, which is 4×99.8 *Hz* — well within tolerance of a 100 Hz *PL*.

The reason for the external division by four is that a divider output programmed in this way is not the 50% duty cycle square wave desired for easy conversion to a sine wave. If the counter output is used to drive a dual flip-flop, such as a 4013, it will divide by four and give the desired wave; or if the encoder is to be coupled with the decoder described later, the divide-by-four operation can be performed by the decoder shift register.

One encoder of this type is made by Avcom*. No circuit diagram is available and the ICs are unmarked, but the encoder apparently uses a 4020 divider and a 4035 shift register to divide the output of a 3.334-MHz fet crystal oscillator. A 50% duty cycle square wave from the shift register is converted to an approximate sine wave by an RC lowpass filter and amplified by an npn output transistor. Maximum output of the unit I tested is 2.85 volts rms. The output waveform is reasonably good between 1 and 2.4 volts, but peak clipping occurs at higher levels and distortion appears below 0.8 volt.

decoders

Only two types of subaudible tone decoder I've tested have given consistently satisfactory results: the reed and the digital. I've heard of designs that use a linear IC such as the NE567V, frequently used to decode *Touch Tone,* but I've never seen one that will perform satisfactorily at PL frequencies.

To work as a PL decoder, the circuit should be sufficiently sensitive to respond to any signal that will quiet the receiver, have stability equal to a reed encoder, and have a bandwidth sufficiently narrow not to be triggered by a *PL* on an adjacent standard frequency **(table 1)**. The circuit should have a "hang-up" connection that will release the receiver squelch, so that the operator can receive signals that don't have *PL* and can also monitor the frequency before transmitting. Outputs that will permit use with either pull-to-ground or pull-to-V+ squelch circuits are desirable.

*Avcom, Inc., P.O. Box 29153, Columbus, Ohio 43299.

reed decoders

The receiver on LEARA'S 16/76 repeater has two Motorola reed decoders. One is on the 110.9-Hz access tone and the other discriminates against the 100-Hz *PL* used across the lake in Detroit, which minimizes interference from there when the Cleveland repeater is operating in the fully open carrier access mode. The Motorola circuits are not reproduced here, but part and circuit diagram numbers are given for those interested.[4,5]

Sensitivity is quite adequate for the excellent receiver with which they are used. Capture bandwidth is less than ±1 Hz on a signal with a low PL level; but once captured, the decoder will follow a shifting tone approximately 2 Hz either side of the nominal frequency. Tone filters are provided to eliminate the subaudible tone from the receiver output. The enable/disable function can be remotely controlled without difficulty.

A reed decoder similar to the Motorola units is obtainable from Communications Specialists either as a separate miniature model or as part of a combination encoder-decoder using the same reed for both functions. The decoder circuit diagram is shown in **fig. 6.** Sensitivity is 2.5 millivolts at the reed frequency at

table 1. Standard EIA subaudible tone frequencies. Higher frequencies not listed are not commonly used by amateurs.

frequency (Hz)	code	frequency (Hz)	code
67.0	XZ	118.8	2B
71.9	XA	123.0	3Z
74.4	WA	127.3	3A
77.0	XB	131.8	3B
79.7	SP	136.5	4Z
82.5	YZ	141.3	4A
85.4	YA	146.2	4B
88.5	YB	151.4	5Z
91.5	ZZ	156.7	5A
94.8	ZA	162.2	5B
97.4	ZB	167.9	6Z
100.0	1Z	173.8	6A
103.5	1A	179.9	6B
107.2	1B	186.2	7Z
110.9	2Z	192.8	7A
114.8	2A	203.5	M I

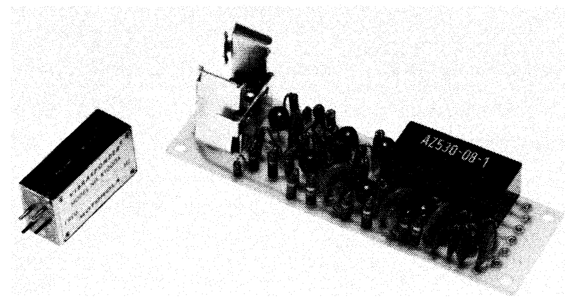
which I tested it (nominally 110.9 Hz). Capture range is ±0.15 Hz at this signal level and ±1.5 Hz at 20 millivolts. Once captured the decoder will stay locked to a 20 millivolt signal to within ±2.0 Hz.

digital decoders

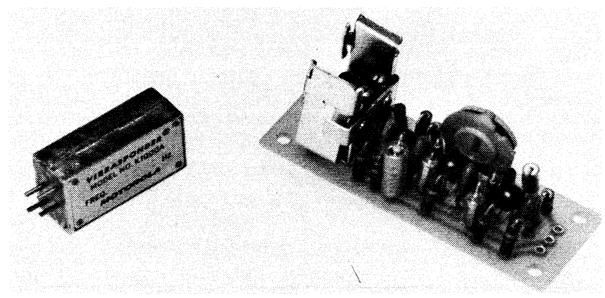
Reed decoders generate a usable output when an incoming signal drives the reed at its mechanical resonant frequency. Digital decoders are not resonant circuits. They produce an output when the incoming signal frequency matches that of a signal generated locally. The usual source is a digital encoder such as those described earlier. I've not found any published material on how or why the circuits operate but have built one that works. The diagram is shown in **fig. 7.** Here is what I think it does:

The output of the fet crystal oscillator, Q1, is fed to U1, a 4020 divider, which is diode-programmed to output at four times the desired *PL* frequency. The diodes are on a plug-in matrix board, permitting quick and easy frequency change. A 4060 used for the divider would eliminate the need for a separate oscillator. The divider output drives U2, a divide-by-four flip-flop, which in turn controls the frequency of an 8038 function generator. The divider also drives U5, the decoder shift register. The shift register is wired to supply V+ to each of the four control inputs of U6 in succession. Since its input is at four times the *PL* frequency, the shift register drives each input of U6 high for 1/4 of a *PL* cycle.

U6 is a quad bilateral switch. When one of its inputs is high, the corresponding 1-μF capacitor is connected to U7, which is a quad operational amplifier. The incoming audio signal from the receiver discriminator is filtered and amplified. It is a square wave at the point of connection to U6. When its frequency matches the rate at which U6 is being cycled by the shift register, U5, the third and fourth stages of the op amp act as a switch to turn off Q2, ungrounding the squelch connection. When the "hang-up" switch is closed, Q2 grounds the squelch connection unless a signal with *PL* is received. An inverter transistor can be added if V+ is required to control squelch.



Miniature reed decoder.



Miniature reed encoder.

The Com Spec encoder-decoder works substantially in the same way as that described above. It has several advantages, however — smaller size (because of the ceramic resonator and special ICs,

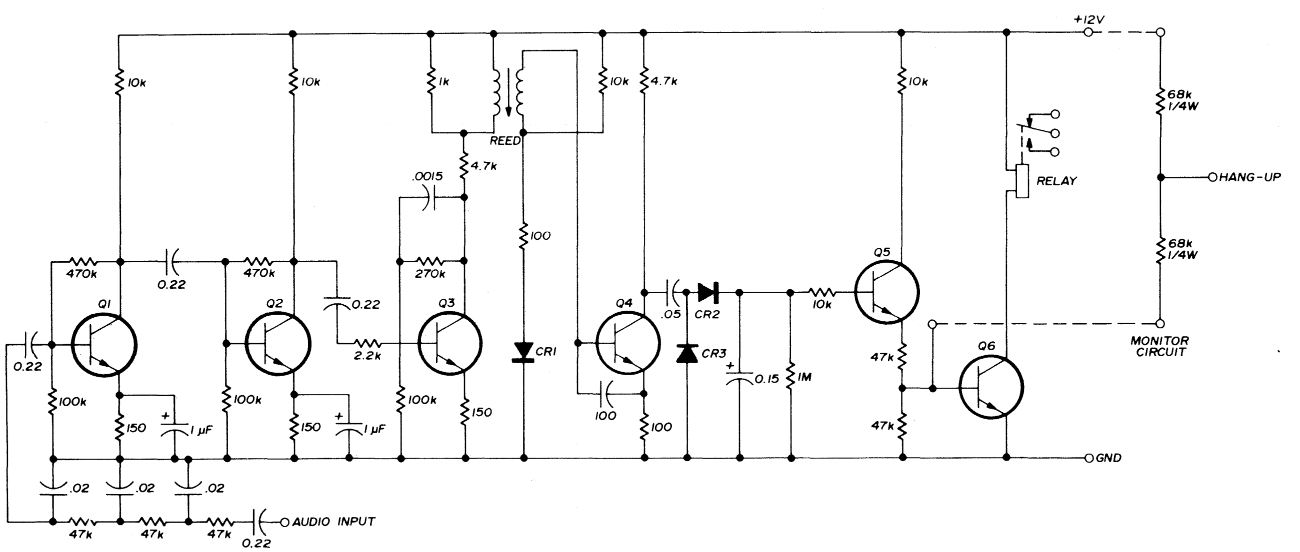


fig. 6. Reed decoder. Transistors are general-purpose silicon npn. CR1 and CR2 are silicon signal diodes. With the monitor circuit connected, the receiver will respond to a signal without PL when the "hang up" terminal is ungrounded.

which combine several functions on one chip); lower cost; and a built-in audio filter to remove the *PL* tone from the receiver audio output. Instructions on how to connect it to most amateur equipment are furnished on request. Bandwidth of all digital decoders I've tested is comparable to the reed types. Sensitivity is a little less but adequate for all the receivers on which I have tried them.

decoder-detector and tunable encoder

The encoder shown in **fig. 7** uses a phase comparator and function generator to provide a sine wave output. This is because I designed the circuit as a tunable *PL* detector, which permits the operator to match and retransmit an unknown *PL* frequency. The block diagram is shown in **fig. 8. A 4PDT** switch is added to the circuit of **fig. 7,** and the VCO portion of the 4046 is used.

In the detect mode, the crystal oscillator is disconnected. The VCO runs at four times the frequency of the function generator; this relationship is maintained by the phase comparator, and the LED connected to pin 1 of the 4046 illuminates when the loop is in lock. The frequency-adjusting potentiometer on the 8038 function generator is accessible to the operator.

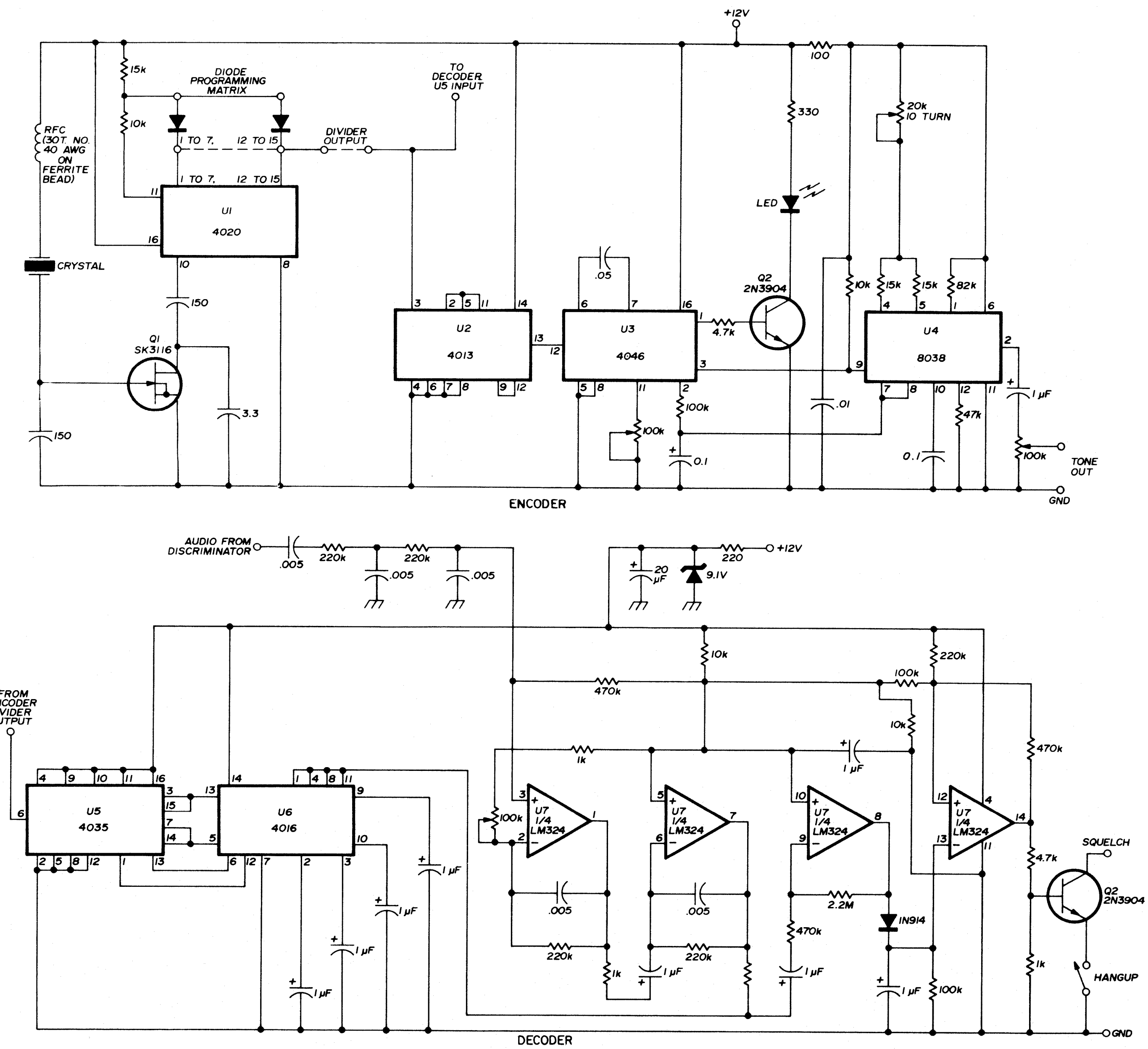
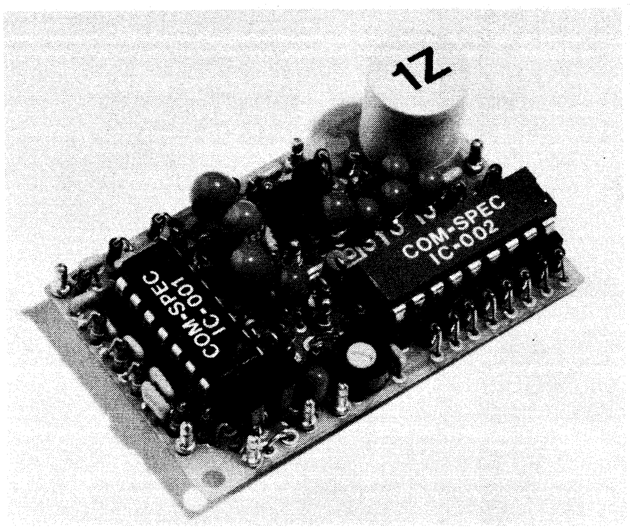


fig. 7. Encoder-decoder with programmable divider. The encoder schematic is at top; LED indicates when phase comparator loop is in lock. The decoder circuit is below; the 100k pot is used to adjust sensitivity.



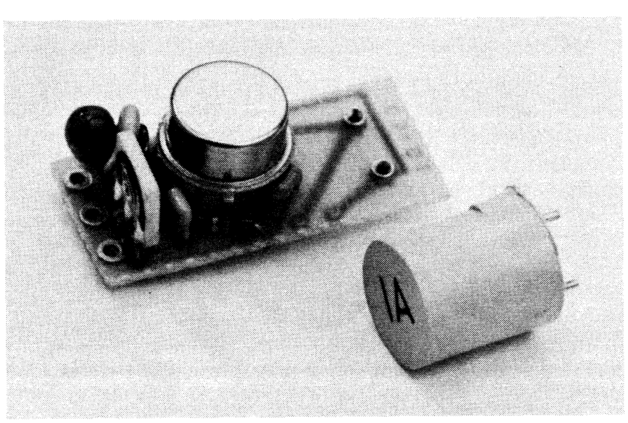
Communications Specialists encoder-decoder.

To match a received *PL* frequency, the switch is thrown to *detect* and the frequency of the free-running function generator adjusted until the decoder output LED shows a frequency match. The encoder will then transmit the same frequency as that received.

It's not ordinarily necessary to adjust the vco, which will hold its lock over a wide range of frequencies. Although the function generator is free-running it will remain within *PL* tolerances for several transmissions. The LED will show the need for readjustment whenever the incoming tone is received.

The greatest limitation of this system is that to acquire a repeater with an unknown access tone, you must be able to hear another station on the input frequency unless the tone is being retransmitted. With the switch in the *crystal* position, the unit operates as a normal digital encoder-decoder.

Micro-miniature encoder with ceramic resonator.



Each of the subaudible tone encoders and decoders described has its advantages and disadvantages. Where space restrictions are not a factor and frequencies are not changed often enough to make the cost of reeds prohibitive, a reed-type unit is hard to beat for stability and clean output. The units with ceramic resonators are much more compact, however, and are comparable in performance. They cost less overall if many frequencies are wanted.

My tunable model is for the experimenter or those who enjoy something different. It's not really as valuable to the traveling ham as one might think — even if you succeed in matching the unknown *PL* on that closed repeater that has been tantalizing you, you probably won't get anyone to talk to you when you do get in!

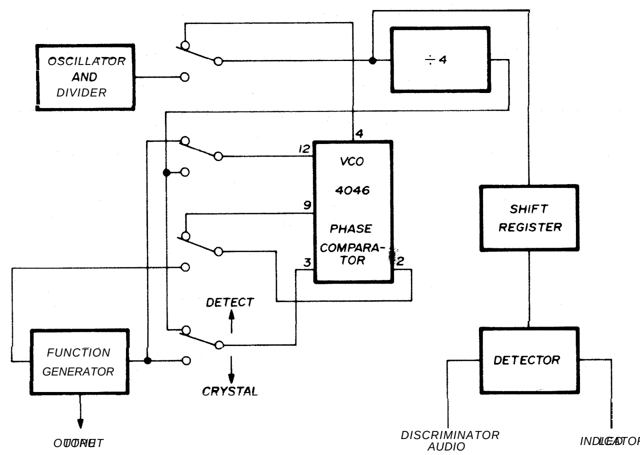


fig. 8. Block diagram of detector and tunable encoder based on encoder-decoder shown in fig. 7.

acknowledgement

I'd like to express my thanks to Spence Porter, WA6TPR, of Communications Specialists, Inc., for the opportunity to test and evaluate their products; for his description of the nature and functions of special components; and for permission to reproduce the circuit diagrams in this article.

references

1. R. B. Shreve, W8GRG, "Automatically Controlled Access to Open Repeaters," *ham radio,* March, 1974, page 22.

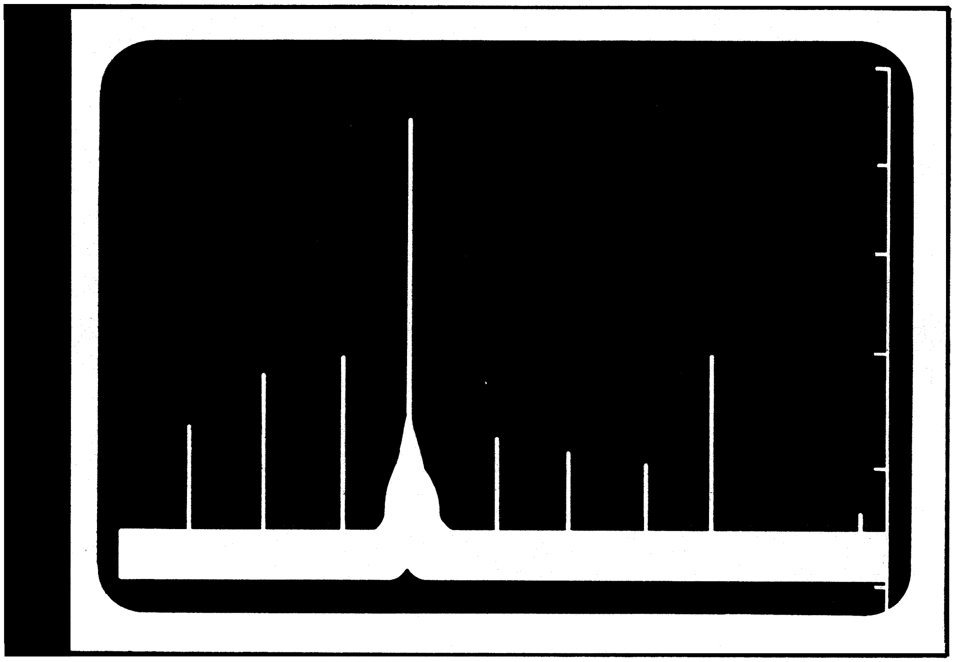
2. *McMOS Handbook,* 1st edition, Motorola Semiconductor Products, Inc., Phoenix, Arizona, pages 4-76.

3. *COSIMOS Integrated Circuits Databook SSD-203,* RCA Communications. Inc., pages 109 and 291.

4. *Private Line" Decoder Model TLN 8401A Used in 136-174 MC MOTRAN Radio Sets,* Motorola Publication EPD-14779C (T-2422A).

5. *"Private Line" Encoder and Decoder Model TLN 4181A (Schematic and Circuit Board Detail),* Motorola Publication PEPS-3531-0.

ham radio



pseudo-logarithmic display

for the microwave spectrum analyzer

This pseudo-logarithmic circuit for your home-built microwave spectrum analyzer provides good resolution and 40 dB dynamic range

In a recent article I described a microwave spectrum analyzer which covered dc to 2.5 GHz with up to 2 GHz of dispersion, 2 MHz resolution, and 50 dB of dynamic range.[1] This analyzer was built almost completely from surplus materials and has been well received by the amateur microwave community. However, the instrument has one drawback: the display graduations are linear rather than logarithmic. This limitation was discussed in the original article, and reader suggestions were solicited.

Before my spectrum analyzer article appeared (but after the manuscript was finalized) ***ham radio*** published a very fine article by Jeff Walker, W3JW, on the design and construction of a high resolution high-frequency spectrum analyzer.* In that article Walker described a simple and effective circuit for providing his analyzer with a pseudo-logarithmic display which allowed him to view 40 dB dynamic range at one vertical deflection setting. It seemed to me that this circuit would, with suitable modification, greatly enhance the performance of my analyzer. I am pleased to report that it did just that.

circuit description

Walker's circuit, shown in **fig. 1,** consists of an audio-frequency detector, lowpass filter, and a unique nonlinear diode limiter arrangement. My analyzer already included an i-f detector diode, the out-

By H. Paul Shuch, N6TX, Microcomm, **14908** Sandy Lane, San Jose, California **95124**

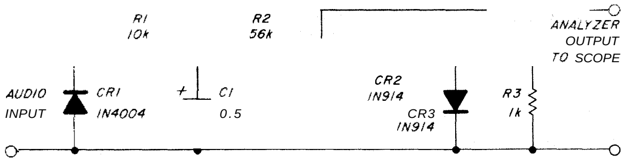


fig. 1. Pseudo-logarithmic signal-processing circuit developed by W3JW for use in a high-frequency spectrum analyzer.2

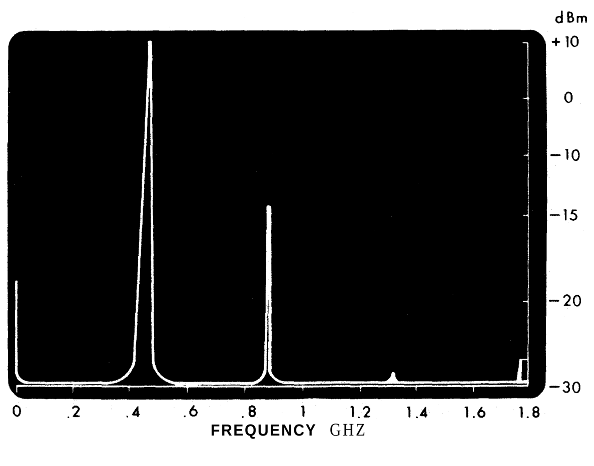
put of which I applied to Walker's filter/limiter circuit. However, I found it necessary to change the value of C1 to achieve the desired video frequency response at high sweep speeds (a value of 1000 pF is acceptable for sweep speeds of up to 60 Hz), For the logarithmic shaper circuit I replaced the 1N914 switch diodes with general-purpose Hewlett-Packard hot-carrier diodes. The final circuit values are shown in **fig. 2.**

Note that the detector circuit I used in my original analyzer provides a positive-going video output. If one of the more common negative-output detectors were used, it would be necessary to reverse the polarity of the Schottky diodes in the logarithmic shaper circuit.

performance

This shaper circuit enabled me to easily view 40 dB dynamic range (+10 to – 40 dBm), with an unusual response which is very nearly logarithmic at 10 dB/cm at very low (– 20 to – 30 dBm) and very high (– 10 to +10 dBm) signal levels. Intermediate amplitudes are "stretched" somewhat, as seen in the scope photograph. However, it is possible to measure signal amplitudes to within one or two dB over the entire 40 dB range, once you get the hang of it. It is possible to view spectral components as far down as – 40 dBm, but scale compression at the low end is so great that you can only guess at the actual amplitude.

Spectrum display of a 450-MHz signal source, as viewed on the microwave spectrum analyzer with logarithmic video processing. The desired signal is at +10 dBm; second harmonic is down 23 dB at – 13 dBm. Fourth harmonic is clearly visible at 40 dB down (– 30dBm). Also visible is a third harmonic component at approximately –35 dBm. Total display dynamic range easily exceeds 40 dB. Note the non-uniform vertical deflection graduations, discussed in the text.



calibration

The display response indicated in the photograph was achieved on my analyzer with i-f attenuation set at a minimum and video sensitivity at 50 mV/cm. The display was calibrated with the aid of a stable 10 mW signal source and a calibrated step-attenuator, by observing changes in the display amplitude as various amounts of attenuation were switched in. Since every analyzer is likely to exhibit its own transfer characteristics, it's a good idea to perform a similar calibration yourself if you duplicate this project.

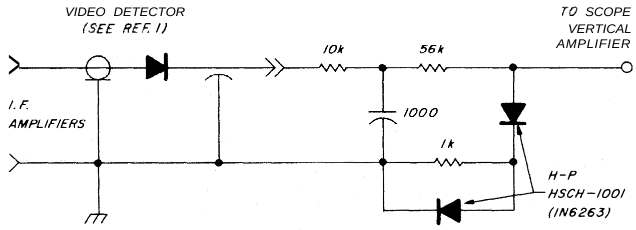


fig. 2. Signal-processor circuit as modified by N6TX for use with his microwave spectrum analyzer.'

One further point: When I change from low-band (dc to 2 GHz) to high-band (500 MHz to 2.5 GHz) coverage, the vertical scale calibration changes con- *siderably.* This is due to the difference in i-f gain with the i-f amplifiers operating at 2 and 1.5 GHz, respectively. Once the analyzer is recalibrated, however, I find it possible to easily resolve signal amplitudes over at least a 40 dB range, with the analyzer operating in either band.

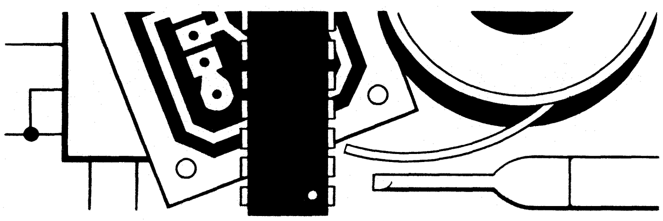
Any feedback from readers who attempt to apply this or other signal-processor circuits would be greatly appreciated. All correspondence which includes a stamped, self-addressed envelope will be answered.

references

1. H. Paul Shuch, WA6UAM, "Low-Cost Microwave Spectrum Analyzer," ham radio, August, 1977, page 54.
2. Jeff Walker, W3JW, "High-Resolution Spectrum Analyzer for Single Sideband," hamradio, July, 1977, page 24.

ham radio

the weekender



1.2 ampere variable-voltage power supply

Whether you are a neophyte just getting started with electronics, an old-timer who hasn't built anything since the days of the 807, or an amateur in need of a handy bench supply, here is project that you can complete in a weekend, yet does not contain any exotic or hard-to-find parts. To make the project even easier, an etched and drilled printed-circuit board is available. The components are available from standard parts houses such as Allied, James Electronics, Lafayette, and Radio Shack. This should take the hassle out of getting the parts together to start the project. The finished product is a neat package that you can be proud to put your call letters on, and will find extensive use in your shack or on your work bench.

circuit description

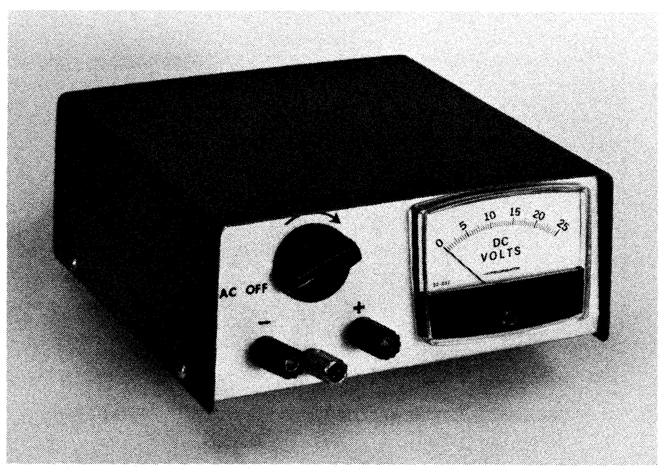
The power supply furnishes a regulated dc output that is variable from 1.5 volts to 24 volts at 1.2 amperes. The regulation is excellent and the ripple is so low that you can power just about any type device with it, from a high gain op amp to a little QRP rig. Although the unit is configured as a bench supply, don't overlook its use for new equipment designs, as well as for powering portable or small mobile rigs in the shack. The little supply will even run the kids' HO trains as I found out last Christmas when their power pack went sour on Christmas Eve. They used the meter on the power supply for a speedometer to see how fast the trains would go before jumping the track.

The circuit, depicted in **fig. 1,** consists of three basic sections: a standard dc supply, a modern three-terminal regulator, and a metering circuit. The ac input (117 Vac 60 Hz) enters through a three-wire cord for safety, placing the case of the supply at ground potential. A fuse is placed in the hot side of the ac line in case of a catastrophic failure, such as a shorted power transformer. The power supply is turned off and on by S1, which is coupled to the output voltage level control R3. With this arrangement you will not be so apt to connect a five-volt device to the power supply and flip on the power switch with the level control set at twelve volts. This feature can save a part or two from an unexpected smoke test.

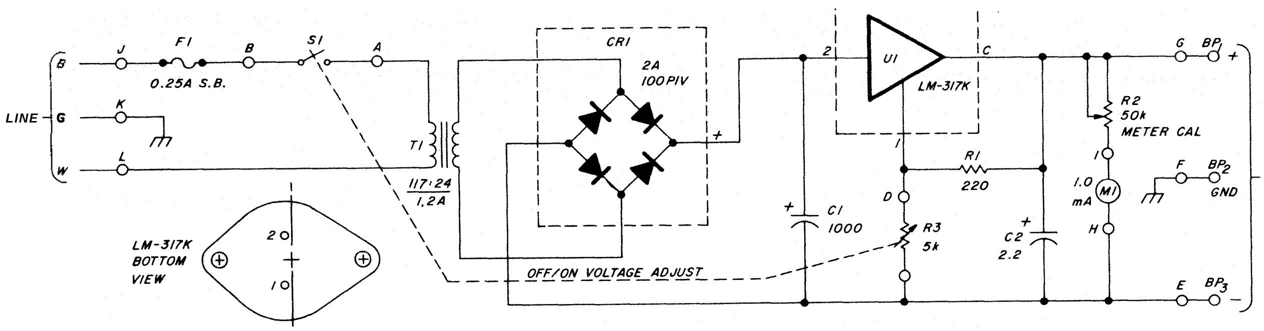
The power transformer steps down the 117 Vac to 24 Vac and isolates the circuitry from the ac line. The transformer output is applied to a full-wave bridge rectifier circuit, CR1, which rectifies the 60 **Hz** ac and furnishes 120 **Hz** pulsating dc. The dc is then filtered by the input filter capacitor C1. The basic power supply furnishes about 35 volts dc when lightly loaded.

The output from the basic power supply is applied to the input of the voltage regulator, U1. The output of the voltage regulator is controlled by a voltage divider network formed by resistors R1 and **R3.** As the

The author's completed power supply.



By Ken Powell, WB6AFT, 6949 Lenwood Way, San Jose, California 95120



R1	220-ohm ½-watt resistor (RS 271-000)	FH1	Fuse Holder (RS 270-739)
R2	50k-ohm trimmer resistor (RS 271-219)	F1	¼-A slow blow fuse (RS 270-1288)
R3	5k-ohm w/switch (RS 271-1714 & 271-1740)	HS1	Heat Sink (Allied Radio 957-2650)
C1	1000 μF/35Vdc capacitor (RS 272-1032)	U1	LM-317K (James Electronics LM-317K)
C2	2.2 μF/35Vdc capacitor (RS 272-1407)	Case	(RS 270-254)
T1	24 volt/1.2A transformer (RS 273-1480)	Spacers	(RS 270-1393)
CR1	2A/100 PIV bridge rectifier (RS 276-1152)	Heat Sink	
M1	0-1 mA meter (RS 22-052)	Compound	(RS 276-1372)
BP1, 2, 3	Binding Posts (RS 274-662)		

fig. 1. Schematic of the variable-voltage power supply. The lettered terminals are used to indicate where the leads enter and leave the printed-circuit board. All resistors are 1/2 watt tolerance; capacitors are rated at 35 volts dc. RS part numbers are available from Radio Shack.

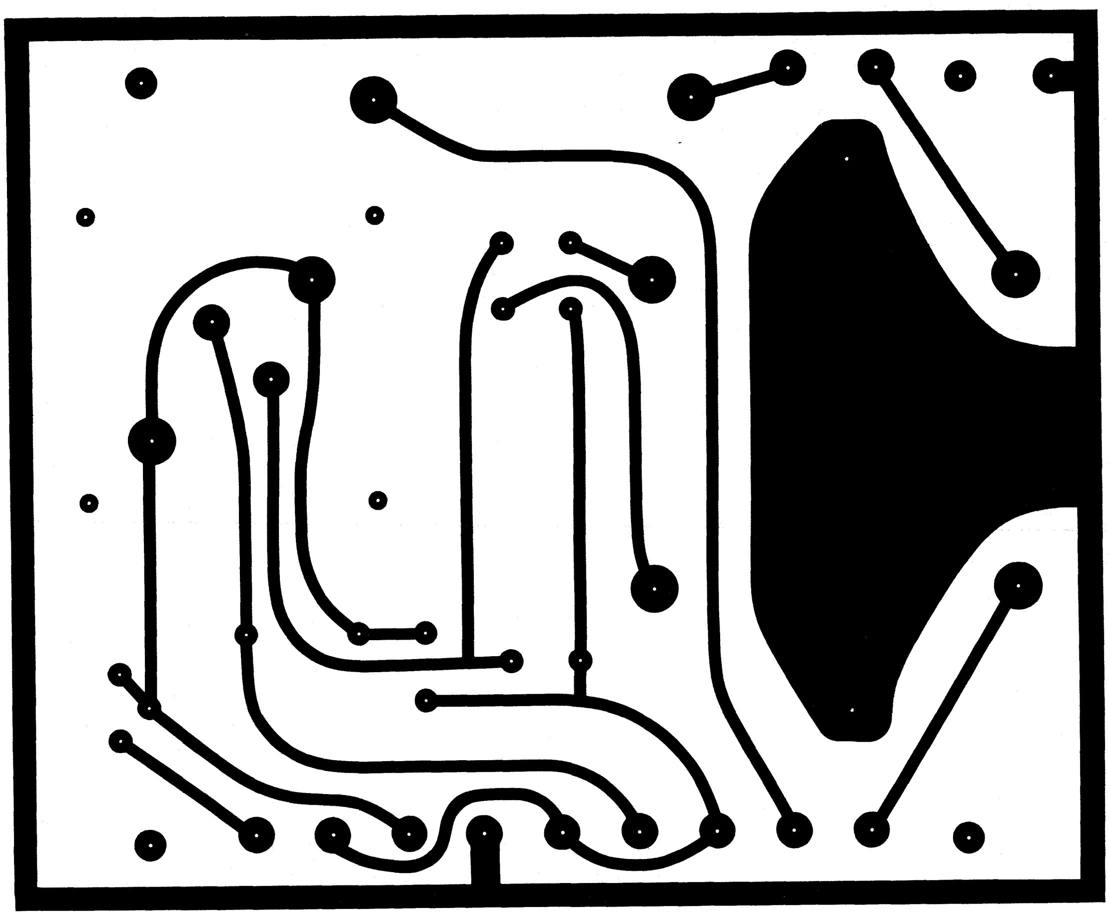


fig. 2. A full-size foil layout for the printed-circuit board. An etched and drilled board is available for $4.00, postpaid, from J. Oswald, 1436 Gerhardt Avenue, San Jose, California 95125.

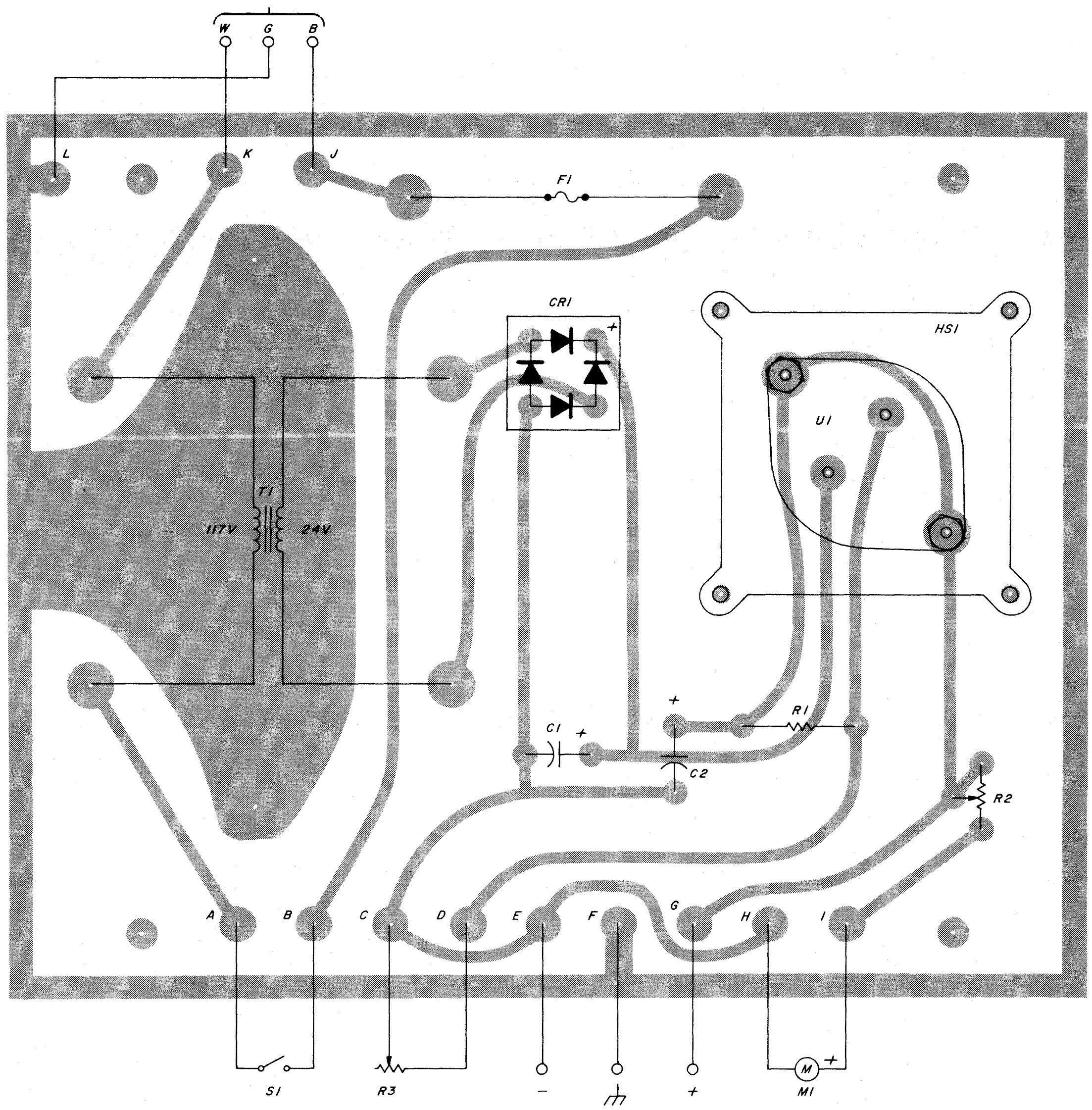


fig. 3. Parts placement diagram for the circuit board.

value of R3 is varied, the output voltage from the regulator varies accordingly; C2 is added to improve the performance of the regulator. A metering circuit is included to indicate the output voltage. A 0-1 mA meter was chosen since this seems to be the most common value available, with surplus units being advertised as low as $1.50.

A small variable resistor, R2, in series with the meter provides an accurate means of calibration. The power supply outputs, both plus and minus, are isolated from ground so the unit may be used as a positive or negative power supply. A ground terminal is also brought out to the front panel should its use be required under certain conditions.

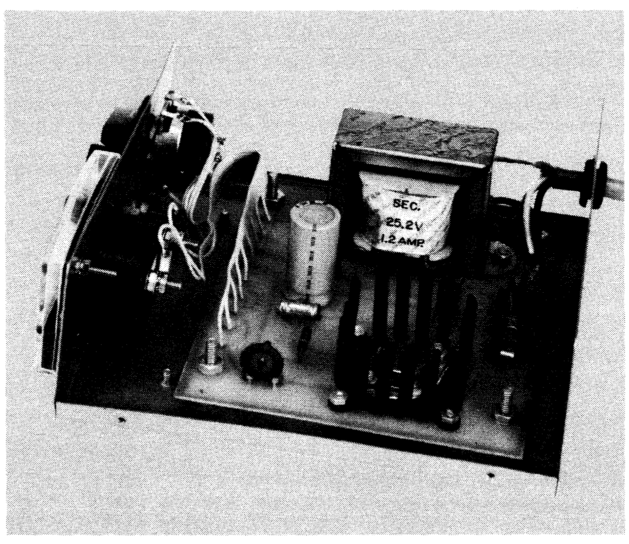
A full-sized printed-circuit board layout is shown in **fig.** 2. This layout assumes the components are the same size as the ones specified in the parts list. If you etch your own board, I would advise using glass-epoxy board, rather than the lighter phenolic type board, since it must support the weight of the power transformer. The heavier board will provide a sturdy and stable package.

When starting construction I temporarily mounted the four corner screws and standoff spacers to the board to protect the foil side of the board while it was handled during construction. Next, mount the power transformer as this will make a sturdy base to hold the board while the smaller parts are mounted and soldered. Coat the bottom side of U1 with heat-sinking compound to form a good thermal junction between it and the heat sink HS1, and mount them to the board. The remainder of the components can now be mounted and soldered. **Fig.** 3 illustrates the component layout and care should be taken to observe the polarity of C1, C2, and CR1. This completes assembly of the basic board.

If you are going to install the printed-circuit board in a chassis box as shown in **fig. 4,** it is best to install the interconnect wiring and the ac line cord prior to mounting the board. Slip a grommet over the line cord and solder the cord to the board. Next, solder the wires to the interconnect terminals at the front edge of the board and run them off to the left edge of the board and then double them back to the right edge of the board. Now, install the printed-circuit board in the chassis box and solder the wires from the front edge of the board to their respective components on the front panel, breaking them out at right angles to the board, parallel to their respective components. The loop left in the wiring between the terminals and the front panel components will allow the board to be removed and turned over for service, should it ever be requried. The ac line cord and grommet are now placed in the cutout at the left edge of the rear panel. Again this is done to facilitate service to the board without unsoldering any wires.

If you use the meter shown in the parts list, and wish to convert the scale to read volts rather than the original milliamperes, remove the plastic cover from the meter and the two small screws retaining the meter face. Then, you can erase the numbers with a typewriter eraser. With rub-on or decal numbers, replace the original markings as follows: change 0.2 to5, 0.4 to 10, 0.6 to 15, 0.8 to 20, and 1 to 25, leaving the zero digit alone. With a little care, you can do a very nice job on the meter and the neatly graduated

Inside view of the power supply showing the printed-circuit board and most of the internal components. The bridge rectifier is mounted behind the heatsink for U1.



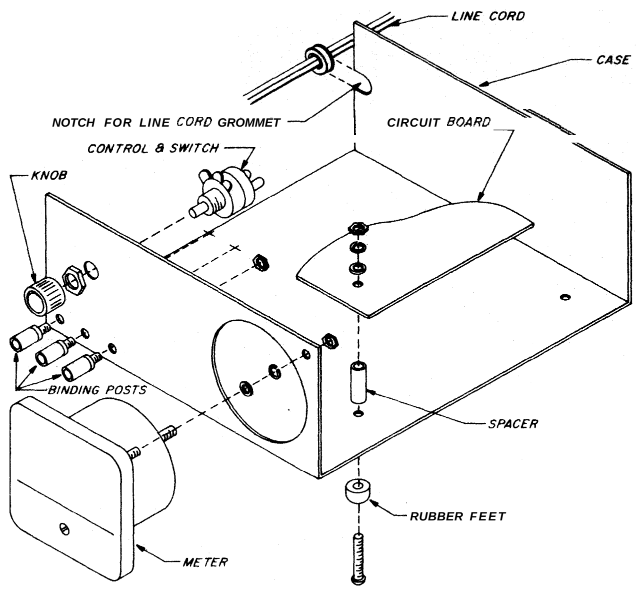


fig. 4. Mechanical details of the enclosure.

scale will be 0.5 volts per division. If you want to skip the meter work, install a 0-25 volt dc meter such as the Lafayette 99P51039V, but in this case be sure to set the calibration trimmer R2 to its minimum resistance position.

test and calibration

The first step, providing you have used a 0-1 milliampere meter, is to set the calibration resistor to its maximum resistance position. Now set the meter to zero with the meter adjusting screw on the front plastic meter cover. Connect a VOM or VTVM, set to 25 volts dc or higher, to the front panel output jacks. Plug the power supply into 117 Vac, advance the output level control to turn on the power supply, and adjust the control until the VOM or VTVM reads 25 volts. Now, adjust the calibration trimmer, **R2,** for a full scale reading of 25 volts on the panel meter M1. Next, check the readings at 20, 15, 10, and 5 volts. The panel meter should track your VOM or VTVM readings quite closely, with the greatest accuracy being achieved at the upper end of the scales.

To check the load regulation, set the power-supply output at 6 volts and apply a load, such as three no. **47** pilot lamps in parallel, to the output jacks. No change in the meter readings should take place as the load is applied and removed. If you have a scope, you can look at the power supply output under load, but under moderate load it is virtually ripple free. In the absence of a scope you can listen to the output with a pair of high-impedance headphones coupled through a 0.1 μF capacitor, with silence being the rule. The ripple on both of the supplies I've con-

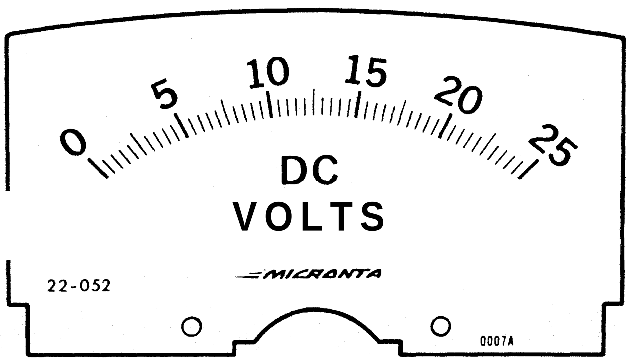


fig. 5. Full-scale meter face after modification.

structed was so low that I could not measure it with my old scope. If the supply meets the above parameters it is time to put it to work; don't worry about hurting it because it can take just about all the abuse you can dish out.

I have built two of these units and use them on the work bench, as I always seem to need both plus and minus voltages at the same time. Both supplies have been excellent performers. The esthetics of the finished product is proportional to the effort you put into it, but I found that you can actually build one of these supplies in a single weekend, have it look as good as a commercial product, and still have time for a late night QSO or two.

There are many variations that could be made, such as placing two boards in a single enclosure and making a dual output supply, or adding a switch and a meter shunt to allow the reading of output current. The fact that all the components are easily obtained, a ready made board is available, and there are no critical adjustments make this bench supply an enjoyable project; I hope you get as much satisfaction out of building and using it as I have.

ham radio

keyboard cleaning on the HP-35 calculator

Owners of HP-35 and equivalent pocket calculators may be experiencing some problem with keyboard operation. Problems such as a double entry or intermittent function is usually due to dirt under the keyboard contacts and is easily corrected.

The Hewlett-Packard series uses thin spring strips for key switches separated by a single, thin plastic sheet from the buttons. The sheet provides a barrier to prevent dirt and moisture from entering the contact area and will wear through after a year or two of operation. The sheet is the major source of trouble, not the contacts.

Plastic sandwich bags of polyethylene are a good source of replacement material for the sheet and may be used in one or two-layer thicknesses.* A common problem is how to open the case.

Models 35, 45, and 55 all use six screws for the main case. Two are easily accessible in the battery compartment, two are under the bottom feet, and the remaining two are hidden by the instruction label. The label is made of aluminum foil stock and its adhesive backing allows easy removal; if you have had one this long, you don't need the instructions. Keep the keyboard side down when removing the screws. When open, the small circuit board screws are easily visible but be careful of the double-wire contacts joining it to the main board.

*Mylar and Teflon sheet has been tried by some but does not work as well as the polyethylene sandwich bag plastic. *Glad* bags and *Baggies* material has been very successful locally.

The main board is screwed to the case top and removal will expose the barrier sheet and key buttons. Use the old sheet for a pattern, tracing the outline and access holes with a felt marker (Sanford Sharpie or equivalent). Be sure to keep the old sheet for future repair.

A clean, fine-bristle artist's brush is good for cleaning the area between switch spring strips and contact surface. It is better to work "dry" than to use commercial cleaners since these usually leave a residue. Isopropyl alcohol is good and ordinary rubbing alcohol is suitable even though it contains some water. Inspect the contact area with a magnifier for any stray hairs; a good artist's brush will have bristles firmly attached but some may break off.

This is also a good time to clean the buttons and case front. A lattice-like frame of plastic holds the buttons from the back. Use extra caution in removing this. Once removed, the buttons will simply fall out. Ordinary hand soap and water is an excellent cleaner and will not harm the plastic or markings. Use a bowl to contain the buttons and soapy water — all buttons are individual and it is too easy to lose one or two down a basin drain. The slide switch has a separate contact with special lubricant and the contact must be removed and set aside.

On reassembly, check button locations with the owner's manual. Do not force the screws into the plastic case or use too much torque; the original threads are quite adequate.

Leonard H. Anderson

radio sounding system

An unusual application of amateur radio for atmospheric studies

How **many** of **you** vhf enthusiasts have experienced the thrill of "an inversion DX contact" and later wondered just what caused it? Such a phenomenon is caused by weather. Here's a sounding system that you can use to find out what's happening in your area. The heart of the system is called a ***radiosonde.***

A radiosonde (or sonde) is a remote weather sensor that uses radio signals to furnish, by telemetry, data to a ground-based receiver and recorder. A radiosonde is usually carried aloft by a helium-filled balloon. From high above it sends back atmospheric information. Weather services use such devices daily all over the world to develop weather forecasts. While the radiosonde described here is a scaled-down version of its bigger cousins, it will allow interested vhf experimenters to study the atmosphere up to several thousand feet (or kilometers). Maybe you can forecast the next big tropo opening!

system description

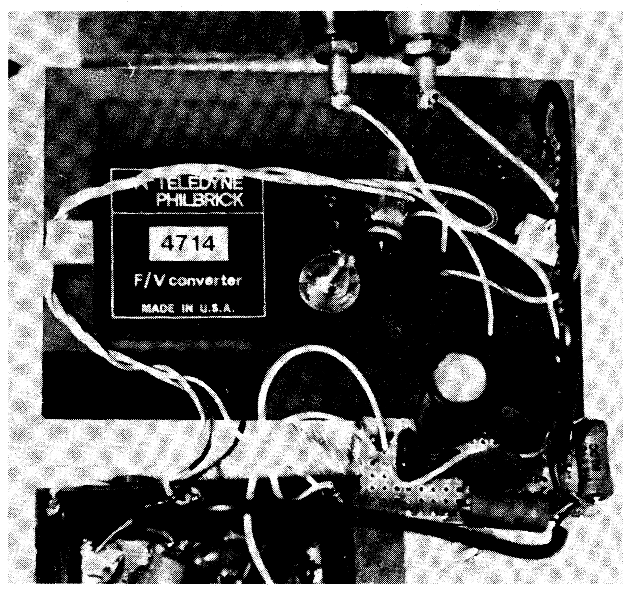
Fig. **1** illustrates the amateur weather telemetry system. **It** consists of the airborne radiosonde and a ground-based station that includes a uhf converter, i-f stage, oscilloscope, frequency-to-voltage (F/V) converter, and a chart recorder.

Radiosonde. A schematic of this little unit appears in fig. **2. It** consists of a sensor, modulator, and a uhf transmitter that operates in the 420-425 MHz portion of the amateur 70-cm band.

Anything set aloft on a balloon doesn't stand much chance of being seen again, so I've kept the circuits simple and the costs down. This is especially important if you're planning to use these circuits in any quantity. (A parachute design is included to help increase the odds of retrieval.)

The sonde shown in fig. 2 was originally modeled after one built by the Argonne National Laboratory for use in the 403-406 MHz band. It operates around 422 MHz and has few circuit modifications. The transmitter doesn't drift more than ±1 MHz, so operation near the band edge is quite safe.

The sonde measures temperature changes and transmits the data to the ground station. A small thermistor, RT, changes value with temperature. Thermistor RT and capacitor CX form an RC circuit that provides an audio signal, which varies as a func-



View showing the inside of the audio-processing stage and the frequency/voltage converter.

tion of temperature. This tone modulates Q2, the transmitter, which provides a uhf fm signal. Although power output is only milliwatts, when the sonde is several thousand feet (or several km) up, its signal can be heard for hundreds of square miles.

The sondes are constructed on small epoxy PC boards (fig. 3). Half the board holds the compo-

By Larry L. Jack, KL7GLK/3, 1 East Lake Drive, Bay Ridge, Annapolis, Maryland 21403

nents, while the other half is a convenient place to tape the 9-volt battery that powers the sonde.

Receiver. Now what's needed is something to receive these interesting weather signals. I use an inexpensive fm broadcast tuner (Lafayette ST-22) for a variable i-f amplifier. It has afc and the wideband capability to receive the sonde's broad, drifting signal. It costs less than the parts needed to build a comparable i-f; but if you have the urge, feel free to experiment.

A Vanguard Labs model 408 uhf converter feeds 422 MHz, which is down converted to 90 MHz, to the i-f amplifier. You now have a receiver over which you can hear the varying temperature-dependent tone from the radiosonde.

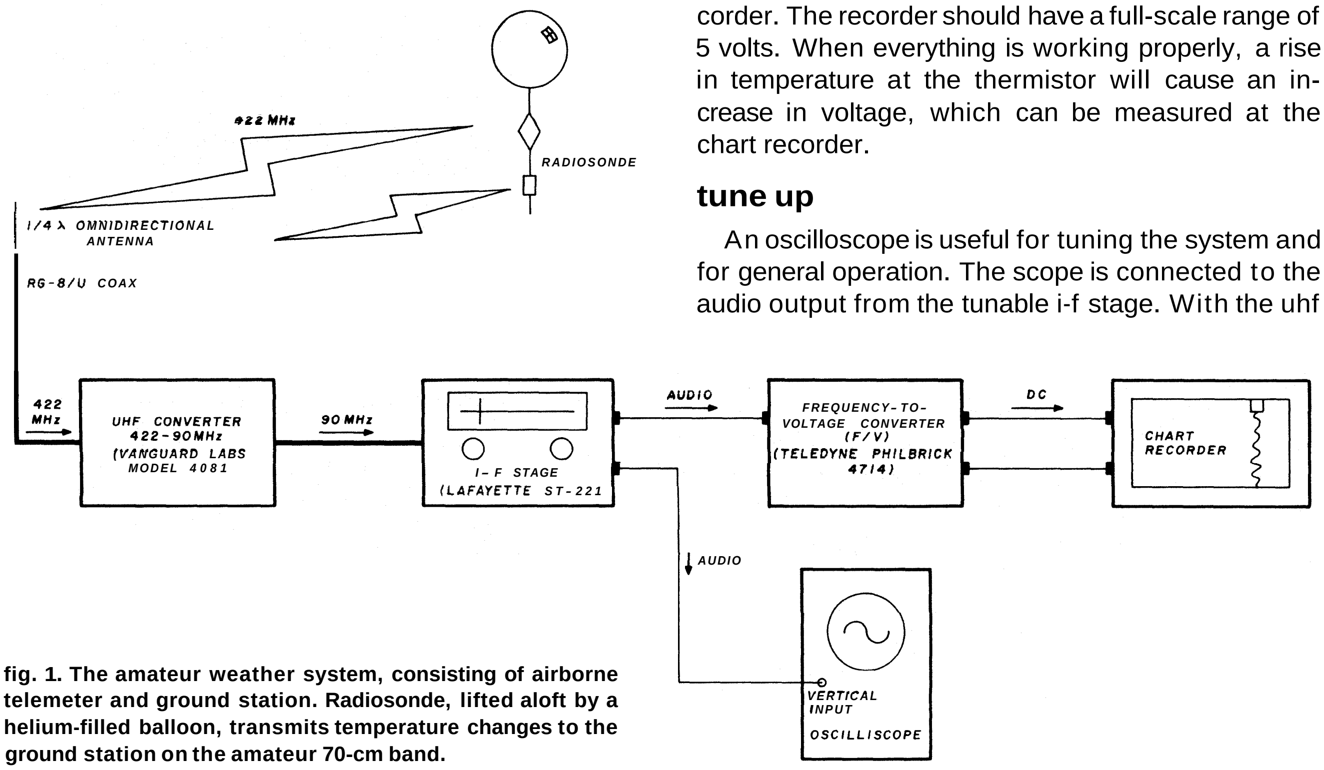


fig. 1. The amateur weather system, consisting of airborne telemeter and ground station. Radiosonde, lifted aloft by a helium-filled balloon, transmits temperature changes to the ground station on the amateur 70-cm band.

FV converter

The next stage accepts the audio signals from the i-f amplifier and converts them to a dc voltage. It's a model **4714** frequency-to-voltage converter made by Teledyne Philbrick and is driven by two audio stages that provide limiting and amplification. (See **fig. 4.**) A dc voltage from the FV converter drives a chart recorder. The recorder should have a full-scale range of 5 volts. When everything is working properly, a rise in temperature at the thermistor will cause an increase in voltage, which can be measured at the chart recorder.

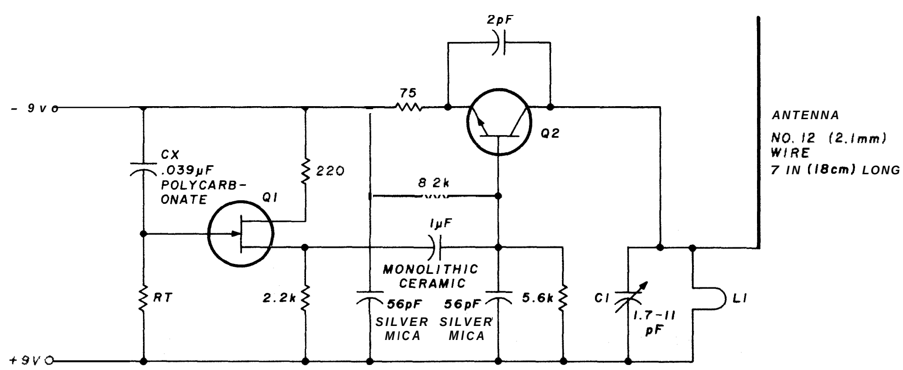
tune up

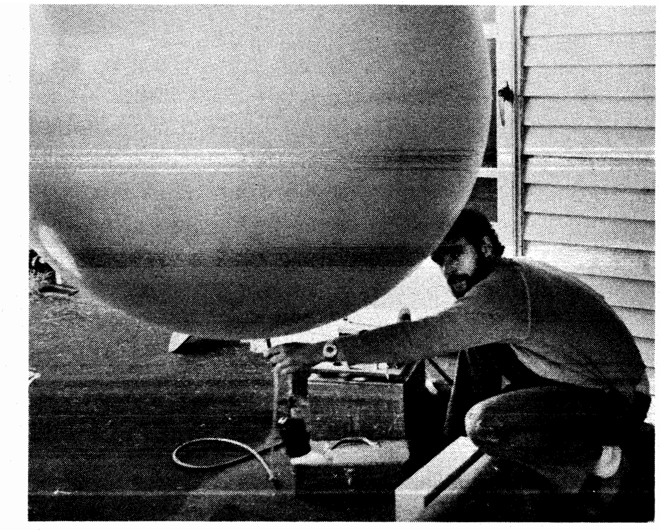
An oscilloscope is useful for tuning the system and for general operation. The scope is connected to the audio output from the tunable i-f stage. With the uhf converter and the i-f stage on, a characteristic noise signal will appear on the scope. With the FV converter stage on, a noise trace will appear on the chart recorder. R1 (50k) in **fig.** 4 is adjusted to give a 2.5-volt trace for 200 Hz into the FV converter. This gives a noise trace at approximately 4.5 volts on the chart recorder. R2 allows fine adjustments at the full-scale end of the chart recorder.

Place the sonde on an elevated nonmetallic stand,

C1 Johanson **9301** trimmer
L1 on PC board
Q1 2N4852
Q2 2N3563
RT Fenwall thermistor GA45J1
Resistors are 5% tolerance 1/8 W

fig. 2. Radiosonde schematic. Circuit was modeled after one built by the Argonne National Laboratory. Parts count and cost are kept low, because retrieval chances are small.





Author KL7GLK filling balloon with helium before launch.

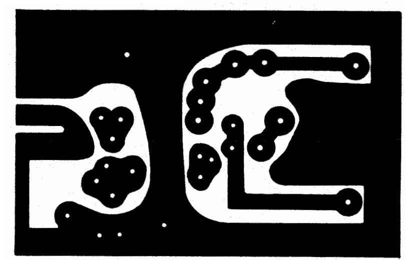
such as a small cardboard box, and connect a 9-volt battery. Tape the battery in place on the sonde. Adjust the transmitter output tuning capacitor (C1, **fig. 2**) using an insulated tool. Watch the scope and chart recorder. At a point on C1 a sawtooth wave will appear on the scope. The recorder trace will smooth to a straight line between 60-70 per cent of full scale.

Where each sonde operates in this range will be a function of air temperature and the tolerance of RT, CX, and other components. A warm breath of air on the thermistor will cause the trace to increase in amplitude then decrease as the thermistor cools.

If you wish to measure other weather data, other resistance-variable sensors can be used; for example, a hygristor can be substituted for the thermistor to measure humidity.

calibration

When the sonde and receiver are working properly, the sonde is ready for calibration. Begin by using a



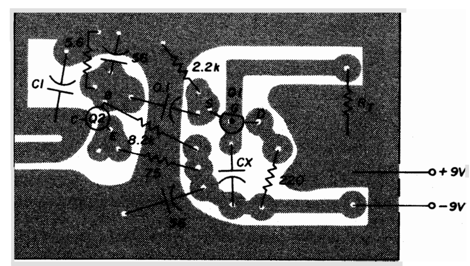
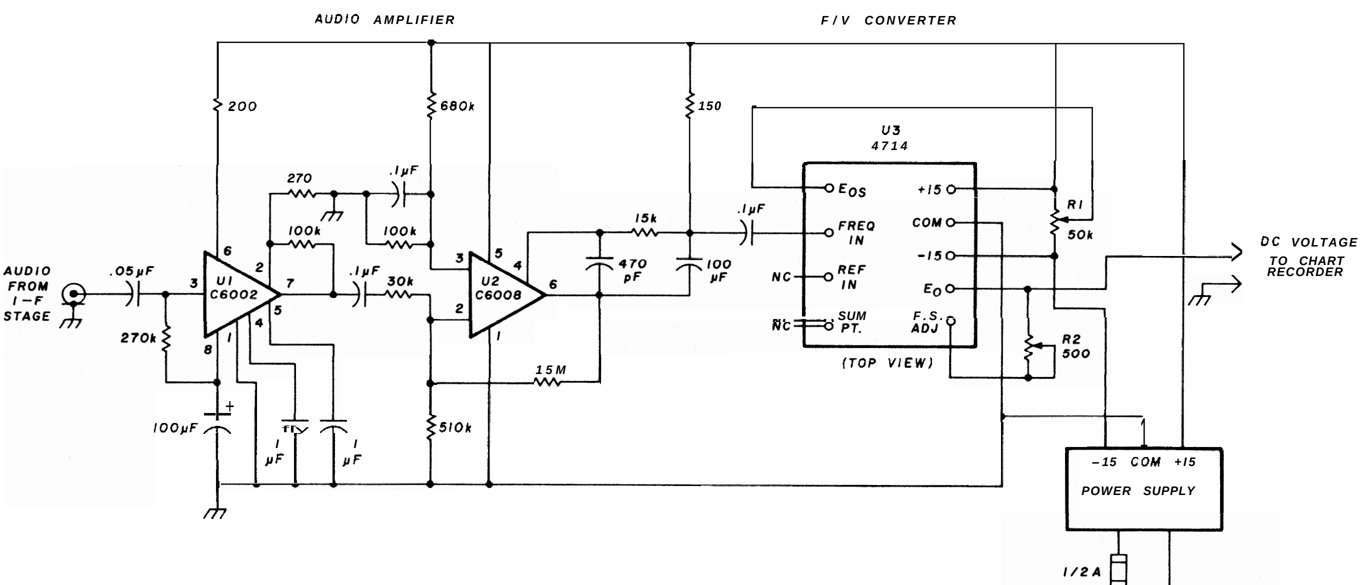


fig. 3. PC-board layout, A, and component placement, B, for the radiosonde.



U1 HEP C6002
U2 HEP C6008
U3 Teledyne-Philbrick 4714 frequency-to-voltage converter
power supply Teledyne 2212 +15 - 0 – 15 Vdc 100 mA

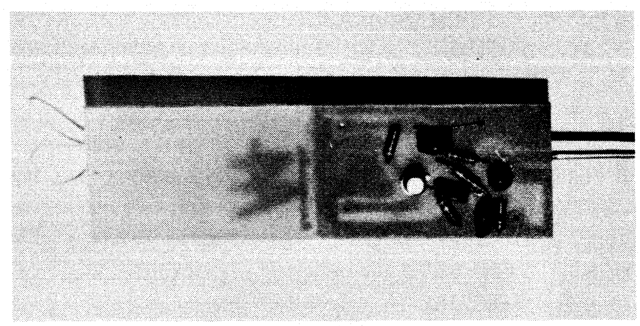
fig. 4. Schematic of the audio amplifier and frequency-to-voltage (FV) converter. The FV converter drives the chart recorder.

thermometer to measure the room temperature that the sonde is monitoring. Mark this value on the chart recorder alongside the trace. Place the sonde in a chamber that can be cooled (by ice, for example) to 68F (20C) below room temperature. When the sonde stabilizes at this new temperature, mark this value on the chart next to its corresponding trace.

The system has a linear temperature response from 23 to 86F (−5 to 30C). Using the two calibration points to form a temperature-to-voltage slope, any temperature point along this slope can be interpolated. A calibration factor can be calculated by dividing the change in temperature by the change in voltage. For example, if the two calibration points were 36 and 75F (2 and 25C), for a voltage change of 1.0 volt there would be a 36F (2.3C) change per 100mV; *i.e.*, $25 - 2 = 23$, and $23/1.0 = 2.3$.

Most chart-recorder paper is divided into 100 lines. On a 5-volt range, each line is 50 mV, so a change of ±1 of these divisions is a change of ±3.6F (±2.3C). This number, (2.3C or 3.6F), is assigned to the sonde as its ***calibration factor.***

Before flying the sonde, measure the outdoor air temperature and mark it on the chart. Once the sonde is in flight, the calculated temperature can be subtracted from this point, giving the temperature



The 422-MHz radiosonde without the 9-volt battery attached.

the sonde is measuring. Because of differences in tolerances of each sonde's components, each sonde will have a slightly different calibration factor. It's therefore a good idea to calibrate each sonde individually.

preflight

The question of determining altitude for the corresponding sonde data now arises. It's necessary to know the rate of the balloon's ascent, so that the altitude can be calculated as a function of time, as measured by the chart recorder; *i.e.*, chart speed =

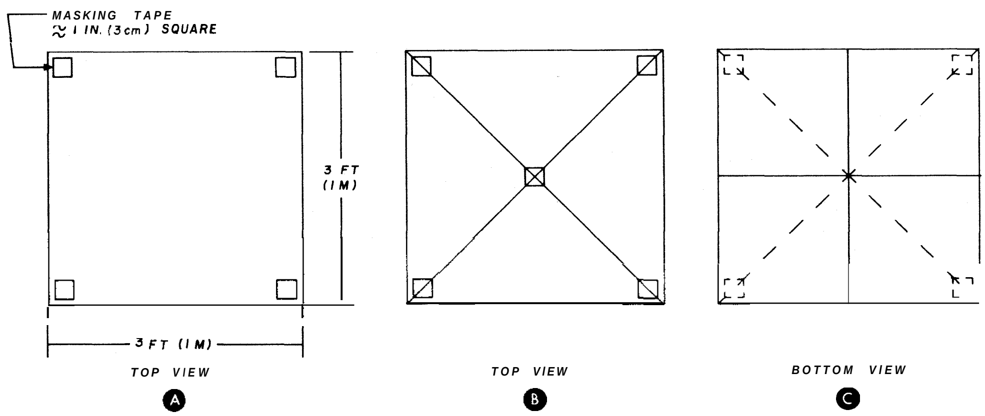
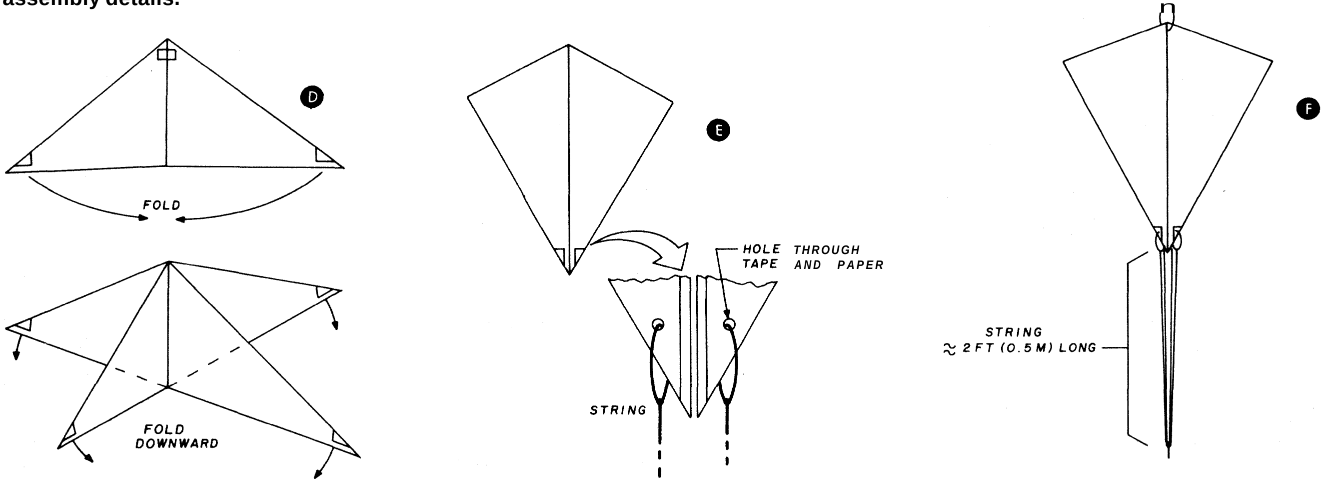


fig. 5. Construction details for the parachute. Sketches A through C show top and bottom views; D through F show assembly details.



2 inches (51mm) per minute. A rough rule of thumb is: A l-ounce (30-gram) helium-filled balloon, filled to just lift a 4.9-ounce (139-gram) weight, will ascend alone at 613 feet (187 meters) per minute.

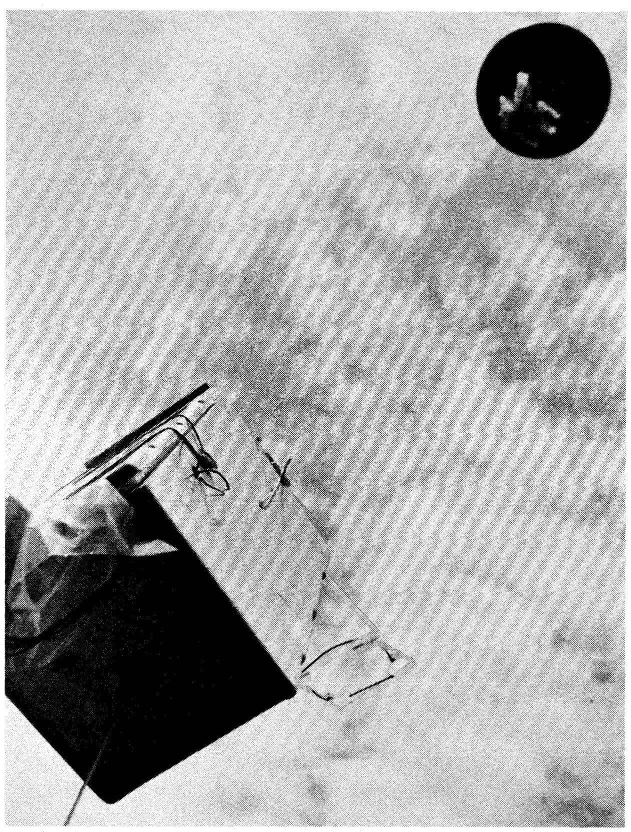
Attaching a sonde and a parachute will upset these figures, so if this method of determining the rise rate is used, some experimentation will be necessary. The best method of determining altitude is to double track the balloon with theodolites and calculate the altitude by triangulation.

The balloons used to lift the radiosondes and their parachutes (there is a good reason for the parachutes, by the way) are meteorological balloons."

A 1-ounce (30-gram) balloon is overfilled to lift the sonde but it works. These overfilled balloons will burst sooner than others (that's why there's a parachute), but usually long after they have drifted out of radio range. As mentioned before, the balloons are filled with helium. Small cylinders of helium are available from firms selling compressed gases.

the parachute

When the balloon bursts, your little sonde could hurtle down through someone's property if the parachute and sonde didn't descend nice and slowly. If



Radiosonde balloon and parachute just before launch. This is the type of equipment used by the United States Weather Service (Beuker's Laboratory model 1207).

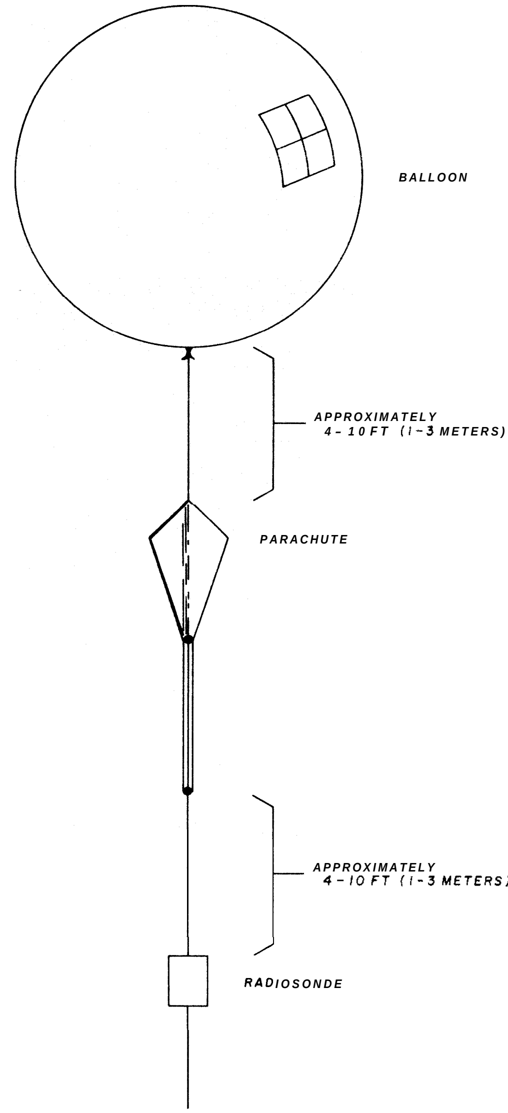


fig. 6. Telemeter assembly showing the balloon, parachute, and radiosonde. When the balloon bursts, the parachute opens, allowing the sonde to drift gently to earth. Your name and address attached to the sonde should increase the chances of its retrieval.

you attach a return address label, maybe someone will mail it back. We get a good number returned this way. You can make your own parachutes with some paper folding and string (see **figs. 5** and **6**).

flight

The system is now ready for use. I'll describe the procedure my compatriots and I use to fly the sonde. First of all it's necessary to file a *Notice to Airmen* by calling the Federal Aviation Administration (FAA).

"Available from Weather Measure Corporation, P.O. Box 41257, Sacramento, California 95841.

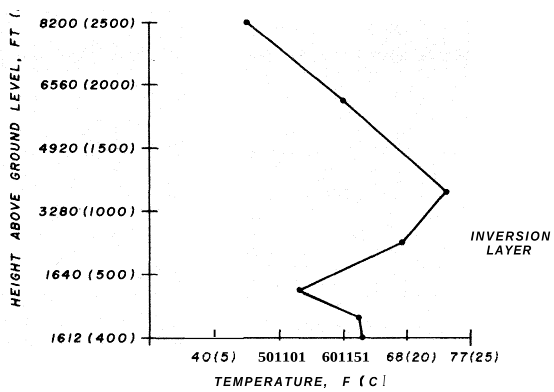


fig. 7. Inversion sounding trace showing temperature changes as a function of altitude between 400 and 2500 meters (1312 and 8200 feet).

Look for a flight service station in your telephone book. They request you tell them the number and times of the flights and the location of the launch. There's nothing to get upset about, but be sure you file your notice.

All right, the balloon is filled, and the train of string, parachute, and sonde is then attached.

Turn on the receiving and recording equipment, attach the radiosonde battery, and tune the receiver to the sonde frequency. Make a note on the chart of the sonde's calibration factor. Then make a side-by-side comparison of the sonde temperature with that of a thermometer; mark the thermometer reading beside the chart-recorder trace. If theodolites are being used to track the balloon, alert the operators. Release the balloon and mark this moment on the chart recorder. As the sonde ascends tune the receiver to follow any transmitter frequency drift.

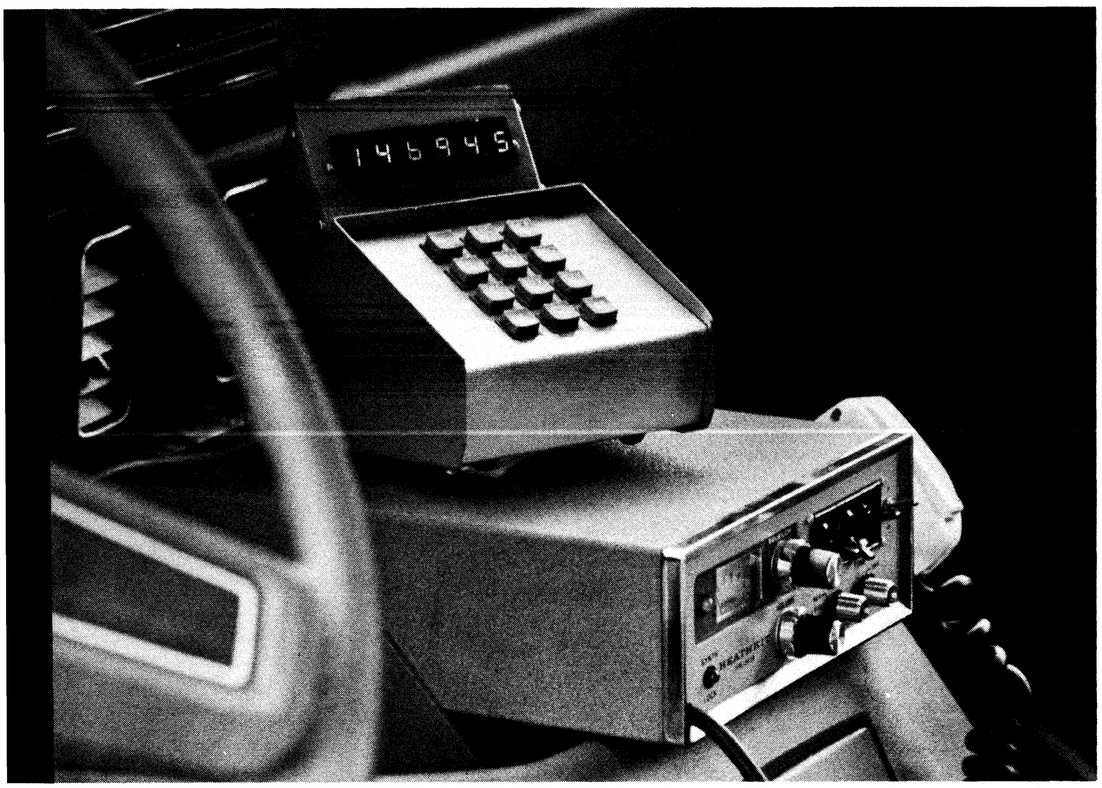
The chart recorder will begin to show the changes in the air temperature as it takes a cross section, or profile of the atmosphere. Depending on conditions, the temperature changes will range from mild to dramatic.

For instance, the inversion phenomenon mentioned in this article's introduction would look similar to the profile in **fig. 7.**

onward and upward

From here on, I refer you to the vast number of meteorology books. As more experience is gained using radiosondes, your questions on weather phenomena will quickly outdistance the scope of this article. The system described won't put you in competition with the National Weather Service, but it will provide an interesting medium for a personal study of the atmosphere and an unusual use of arnateur radio. Have fun with it.

ham radio



outboard LED frequency display

for the Heath HW2036

Operating the HW2036 no longer has to be done in the dark — this outboard display shows the frequency set in the switches

Picture yourself driving along the freeway one evening and your Heath HW2036 comes alive with an interesting QSO on the 01/61 repeater. Just before the punch line, the operator decides to move to 147.18 simplex. If you want to hear the rest of the joke, you'll have to either turn on the dome light in your car so you can see where to set the thumbwheel frequency-selection switches on your rig or you can set the switches to a common starting point and count the clicks as you advance them to the correct frequency. By that time, though, the joke is past history. Having owned a HW2036 for a year now, I found that the only real problem was quickly changing frequency at night. There are no lights provided on the front of the panel to illuminate the thumbwheel switches. After many hours of fumbling in the dark with the frequency switches, it occurred to me that the switches must provide binary-coded information to program the synthesizer variable dividers. Why not use that same information to drive some LEDs, thus displaying the selected frequency?

By Bill Stephens, WB8TJL, 217 West Reed Street, Bowling Green, Ohio 43402

circuit details

Switches S3, S4, and S5 provide BCD information to the variable divider ICs (U401, U402, U403 respectively). If you are setting your frequency to 146.94, for example, you dial up 6 9 4, since the 1 and 4 remain constant across the band.

I found it was possible to tap off of the four leads at the rear of the three thumbwheel switches and use that information as inputs to the three 7447 ICs (see fig. 1). The 7447s convert the BCD data from the switches into a seven-segment format necessary to drive the LEDs. This in no way alters the operation of the synthesizer, nor does it degrade the performance of the rig. All of the outboard components can be easily housed in a small box and set on top of or beside the HW2036. A flat cable containing 15 leads is used to connect the HW2036 and the outboard unit.

display LEDs

Since LEDs 1 and 2 always displayed the digits 1 and 4, it was decided to conserve costs and space and not use a 7447 decoder/driver. Instead, they are permanently wired to display 1 and 4. LEDs 3 thru 6 are driven by U1 through 4, respectively. Each of the LEDs has a 100-ohm 1/2-watt resistor in series with pin 3 and the 5-volt line. This resistor limits the current to each segment to no more than 25 mA. It would be better to connect one resistor in series with each of the seven leads of each of the six LEDs, but I didn't find the prospect of wiring 42 resistors too appealing. Therefore, I decided to limit the current going to the common-anode connection of each LED; that required only six resistors.

The only drawback to this technique is that as different digits are displayed, using more or less segments, the overall intensity of that particular digit will be slightly brighter or dimmer than other digits. For instance, if the digit 1 is displayed, only two segments of the LED are lit (segments b and *c*). On the other hand, if **8** is displayed, all seven segments (*a, b, c, d, e, f, g*) are used and they must share the same amount of current as the two segments in the first example. With a 100-ohm resistor and a 5-volt line this works out to 25 mA per segment when two segments are on and 7 mA per segment when all seven are lit. In practice this poses no real problem because the change in intensity is barely noticeable.

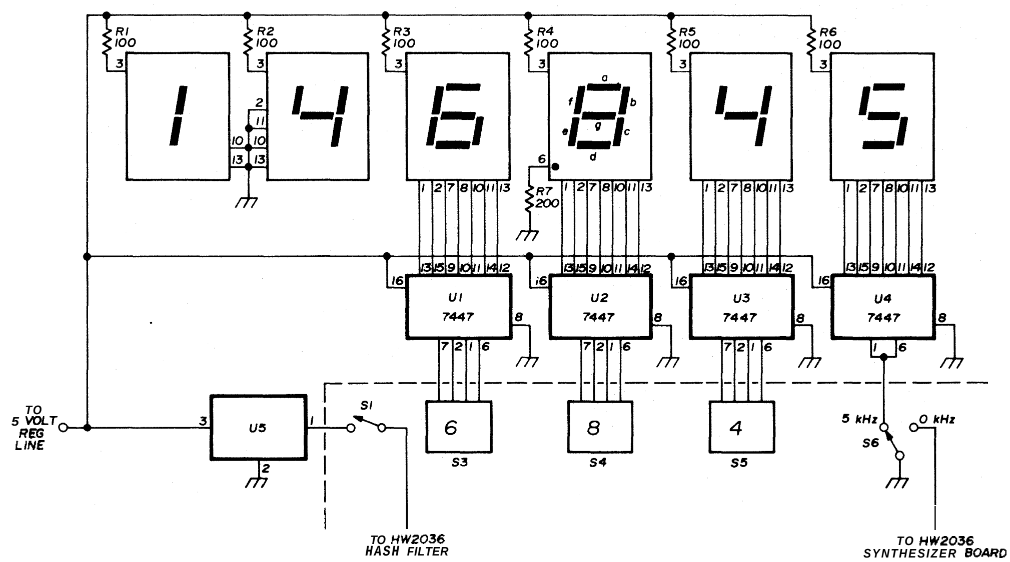


fig. 1. Schematic diagram of the addition to the HW2036 to add an outboard frequency display. Instead of using 7 resistors for each LED, a common current-limiting resistor is inserted in the anode line. The LEDs must be of the common-anode type. The 5-volt regulator is a 7805.

obtaining power

Referring to the schematic of the HW2036, 12 volts is obtained from lug 2 of the on-off switch S1. This line and a chassis ground are fed thru a cable to the outboard display unit's 5-volt regulator. An LM309K could be used, although the smaller package TO-220 7805 was used in my unit. The 7805 is rated at one ampere of current and is capable of powering the four 7447s and the six display LEDs, although it does run a bit warm at times. Mounting the regulator on a small heatsink or bolting it to the display unit case should overcome this problem.

decoder/drivers

Each switch has a 1, 2, 4, and 8 lead coming from it; these are connected to pins 7, 1, 2, and 6, respectively, of the 7447s. When a thumbwheel switch is set to program the digit 6, it grounds the two unused leads of the 1 and 8 lines. With pins 6 and 7 grounded, the outputs ground the appropriate leads of the associated LED, causing it to form the digit 6. Each of the 7447s is connected to its associated LED by seven leads, one for each of the seven segments in the LED.

If the operator decides to program in a split channel, for instance 146.945, S6 is used to control U4. Within the rig, S6 grounds one lead of U405 when the frequency ends in a zero; it lifts that same lead above ground when a split channel is selected. To display the digit 5, when the switch is set in the 5 kHz position, bring a lead off of the unused side of S6 and tie it to pins 1 and 6 of U4. Thus, when the switch is in the 0 position there is no input to U4 and no digit appears. When the switch is thrown to the 5 kHz position, pins 1 and 6 are grounded, displaying a 5.

construction details

I built my unit on perf board with 2.5 mm (0.1 inch) hole spacings. One board, measuring about 9 cm^2 (3-1/2 inch square) contains the four 7447s and the voltage regulator. A second board, measuring about 9 cm x 3.8 cm (3-1/2 inches x 1-1/2 inches), contains the six LEDs and is connected to the first board by 30 jumpers. While perf board is entirely acceptable for this project, it is suggested that printed-circuit techniques be used due to the relatively large number of board-to-board jumpers and the small space in which to work. The many connecting jumpers tend to make the interconnecting harness very stiff and difficult to work with.

My unit is housed in a homebuilt box measuring 5 x 9 x 10.2 cm (2 x 3-1/2 x 4 inches) with a sloping front panel for easier viewing of the LEDs. The size and shape of your box will be determined by where and how you are going to mount it. Four-conductor flat cable was used throughout the construction. Four lengths of cable were used to bring the 15 leads out of the back of the HW2036 case (they fit nicely between the top of the PA board assembly and the case). Individual strands of the same wire were stripped and used for the 5-volt line and ground bus on the two perf boards.

If the unit is to be mounted in an outboard case, it is recommended that a plug be installed to disconnect the outboard unit from the HW2036 for ease of servicing.

I wish to thank WB8NQW and K8TT for their help and thoughts in completing this project.

ham radio

phase-locked loops

basic building blocks for frequency manipulation

A basic discussion of phase-locked loops and how they are used in communications systems

In electronic systems, information can be expressed in four ways: voltage, current, frequency, and phase angle. In modern electronics, operational amplifiers have become the basic building blocks for circuits which manipulate voltages and currents. But what about frequency and phase angle? Enter the phase-locked loop or PLL.

The first widespread use of the phase-lock system was in TV receivers to synchronize the horizontal and vertical sweep oscillators to the transmitted sync pulses. Lately, narrowband phase-locked receivers have proved to be of considerable benefit in tracking weak satellite signals. This is due to the superior noise immunity of PLL systems. Although it's not well known, the synchronous reception of radio signals using the PLL technique was first described in the early 1930s; it was known as a "homodyne" receiver.

In the early days, applications using PLLs had to be implemented using discrete components. Even after the advent of transistors, the PLL circuit was considerably complex. Thus, the use of PLL methods in most electronic systems was both expensive and impractical.

In the late 1960s Signetics Corporation developed monolithic circuit versions of the PLL system. This development of single chip PLLs changed things considerably. A single package device, used with a few external components, offers all the benefits of PLL operation while making their use practical, uncomplicated, and economical.

PLL theory

Just what is a PLL and how does it do all this frequency manipulating? The op amp is a voltage/current feedback system; that is, a portion of the output is fed back to the input. In an op amp circuit, this feedback component is a current. The PLL is a feedback system but in this case the component fed back is a frequency. **Fig. 1** shows a block diagram of a feedback system.

A phase-locked loop is basically an electronic servo loop. The function of a PLL is to detect and track small differences in phase and frequency existing between the input and a reference signal. A block diagram of a basic PLL system is shown in **fig. 2.** In this circuit the voltage-controlled oscillator is driven in the direction that will minimize the error signal. Note the similarities between **figs. 1** and **2.**

Like most other complex circuits, phase-locked loops have special terms associated with them; understanding their operation is easier when you become familiar with the language. The following is a brief glossary of terms encountered with PLLs:

Capture range. The range of frequencies over which the loop can detect a signal on the input and respond to it. This is sometimes called the lock-in range (lock-in range refers to how close the signal must be to the center frequency before acquisition can occur; thus it is one-half the capture range).

Current controlled oscillator. An oscillator in which the frequency is determined by an applied current.

Damping factor. In a PLL this refers to the ability of a loop to respond quickly to an input frequency step without excessive overshoot.

Free-run frequency (f_o). Also called the center frequency; it is the frequency of the vco with no input signal.

By Bob Marshall, **WB6FOC,** Analog Applications, Signetics Corporation, Post Office Box 9052, Sunnyvale, California 94086

Lock range. The range of frequencies over which the loop will remain in lock; also called tracking range.

Loop gain (K_v). Product of the dc gains of all the loop elements; in units of sec^{-1}.

Loop noise bandwidth. A loop relating to damping and natural frequency which describes the effective bandwidth of the input signal.

Lowpass filter. A filter which permits only dc and low frequencies to travel around the loop; it determines the capture range of the loop.

Natural frequency. The characteristic frequency of the loop (not to be confused with free-running frequency).

Phase detector gain factor (K_d). The conversion factor between the phase detector output voltage and the phase differences of the input and vco signals; expressed in volts-radians.

Phase detector. A circuit which compares the relative phase between two inputs and produces an error voltage dependent on the difference. This error voltage corrects the vco frequency during tracking. Sometimes called a phase comparator or mixer.

Quadrature phase detector. A phase detector operated in quadrature (90° out of phase) with the loop detector.

VCO conversion gain (K_o). Conversion factor between vco frequency and control voltage in radians/sec/volt.

Voltage Controlled Oscillator or VCO. An oscillator whose frequency is determined by an applied control voltage.

loop operation

As was mentioned earlier, the PLL is a feedback system; therefore, it can be characterized mathematically by the same equations that apply to other, more conventional feedback systems. However, the parameters in the PLL equations deal with phase rather than a current or voltage.

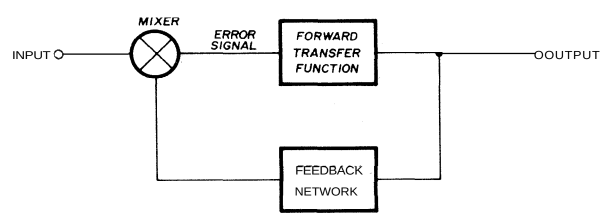


fig. 1. Block diagram of feedback system shows the basic arrangement of both operational amplifiers and phase-locked loops. Current feedback is used in an op amp whereas a frequency component is fed back in a phase-locked loop.

A mathematical analysis of a PLL can get pretty hairy but a qualitative analysis will explain the basic principle of PLL operation. During the following discussion, it will be helpful to refer to **fig.** 3.

With no input signal applied, the error voltage V_d is zero. The vco will operate at a set frequency f_o or the *free-run* frequency. When an input signal is applied to the system, the phase detector compares the phase and frequency of the input with the vco frequency. This generates an error voltage $V_{e(t)}$ that is related to the phase and frequency difference between the two signals; this error voltage is then filtered, amplified, and applied to the control terminal of the vco. In this manner, the control voltage $V_{d(t)}$ forces the vco frequency to vary in a direction that reduces the frequency difference between f_o and the input signal.

If the input frequencyf, is sufficiently close to f_o, the feedback nature of the PLL causes the vco to synchronize or lock with the incoming signal. Once in lock, the vco frequency is identical to the input signal except for a finite phase difference. This net phase difference θ_o is necessary to generate the corrective error voltage V_d to shift the vco frequency from its free-running value to the input signal and thus, keep the PLL in lock. This self correcting ability of the system allows tne PLL to track frequency changes of the input signal once it is locked.

Another way of describing the operation of the PLL is to observe that the phase detector is, in actuality, a multiplier circuit that mixes the input signal with the vco signal. The mixer produces the sum and difference frequencies $(f_1 \pm f_o)$. When the loop is in lock, the vco duplicates the input frequency so that the difference frequency component $(f_i - f_o)$ is zero; hence the output of the phase comparator contains a dc component. The lowpass filter removes the sum frequency component $(f_i + f_o)$ but passes the dc component which is amplified and fed back to the vco. Notice that with the loop in lock, the difference frequency component is dc and independent of the band edge of the lowpass filter.

lock and capture

What happens before the loop is locked? Let's assume for a moment that there is a frequency on the input to a PLL. The phase comparator mixes this incoming frequency with the free running vco frequency. If the difference frequency $(f_L - f_o)$ is greater than the band edge of the lowpass filter, the input to the vco is still zero so the vco remains at its free-run fre-

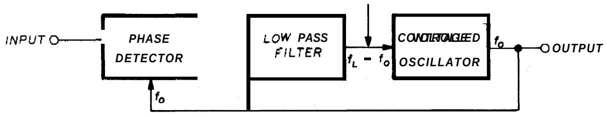


fig. 2. Basic phase-locked loop consists of a phase detector, lowpass filter, and voltage-controlled oscillator (vco). The vco is driven in a direction which minimizes the error signal.

quency. As the input frequency approaches that of the vco, $f_1 - f_0$ decreases and approaches the band edge of the lowpass filter. Now some of the difference component is passed to the vco control. This in turn decreases the frequency difference component which allows more information through the filter. This positive feedback mechanism causes the vco to snap into lock with the input signal. Thus, the capture range is again defined as the "frequency range centered about the vco free-run frequency over which the loop can acquire lock."

Once the loop is locked, $f_1 - f_0$ is essentially dc and thus unaffected by the lowpass filter. The lock range is limited by the range of the error voltage that can be generated and the corresponding deviation in vco frequency which is produced.

It is important to distinguish the difference between *capture* range and ***lock*** range. Lock range is defined as the frequency range, usually centered about the initial vco frequency, over which the loop can track the input signal once lock has been achieved. The total time required for the vco to obtain a locked condition is called the pull-in time. Pull-in time depends on the initial frequency and phase difference between the two signals and the overall loop gain as well as the lowpass filter.

effects of the lowpass filter

In the operation of the loop, the lowpass filter serves a dual function. First, by attenuating the high-frequency error component $(f_L + f_o)$ at the output of the phase comparator, it improves the interference-rejection characteristics; second, it provides a short term memory for the PLL and ensures a rapid recapture of the signal if the system is thrown out of lock because of a noise transient. Reducing the lowpass filter bandwidth has the following effects on system performance:

1. The capture process becomes slower and increases the pull-in time.

2. Capture range decreases.

3. Interference-rejection properties improve since the error voltage caused by an interfering signal is attenuated further by the lowpass filter.

4. The transient response of the loop (the response of the PLL to sudden changes of the input frequency within the capture range) becomes undamped.

This last effect also produces a practical limitation on the lowpass loop filter's bandwidth and roll-off characteristics from a stability standpoint. A detailed analysis of a PLL under lock condition using Laplace transforms will prove that if either the loop gain or the filter time constant is too large, the loop itself will break into sustained oscillations.

The lock range of the PLL, f_L, can be shown to be numerically equal to the dc loop gain K_v

$$4\pi f_L = 2\omega_L = 2\,K_v$$

Since the capture range, f_c, denotes a transient condition, it is more difficult to derive, but with a simple lag filter the capture range can be approximated as

$$4\pi f_c = 2\omega_c \simeq \sqrt{\frac{2\pi f_L}{\tau_1}} = \sqrt{\frac{K_v}{\tau_1}}$$

where τ_1 is the time constant of the loop and f_L is the lock frequency. Thus, the capture range increases as the time constant of the filter decreases, while the lock range is a function of the dc loop gain.

Fig. 4 shows the typical frequency-to-voltage transfer characteristics of the PLL. The input is assumed to be a sine wave whose frequency is swept slowly over a broad range of frequencies. The vertical scale is the corresponding error voltage of the loop.

Fig. 4A shows the loop error voltage with an increasing frequency. The loop does not respond until the frequency reaches f_1, which corresponds to the lower edge of the capture range. At that time the loop suddenly locks on the input and causes a negative jump of the loop error voltage. As the frequency continues to increase, the loop error voltage increases. Notice that V_d is zero when the incoming signal f_i equals f_o. The loop continues to track the incoming signal until the frequency equals f_2; this corresponds to the upper edge of the lock range. The PLL then loses lock and V_d returns to zero.

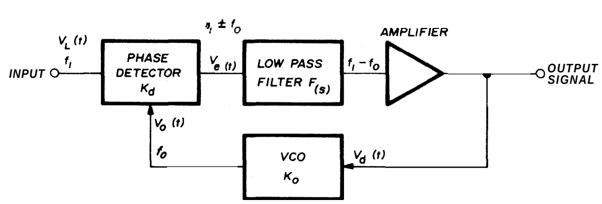


fig. 3. Diagram of a phase-locked loop showing the frequency and voltage relationships in the system. See text for description of operation.

The slope of V_d is equal to the reciprocal of the vco gain $(1/K_o)$ measured in volts/radians/sec. If the input frequency is swept slowly back (illustrated by fig. **4B**), nothing happens until the incoming frequency reaches f_3. The loop continues to track until f_4 where it breaks lock and the error returns to zero. The total lock and capture ranges of the system are:

$$2f_L = f_2 - f_4 \ (lock)$$

$$2f_c = f_3 - f_1 \ (capture)$$

As indicated by the transfer characteristics of fig. 4, the PLL has an inherent selectivity about the center frequency set by the free running vco frequency f_o. It can also be seen that the loop will respond to input signal frequencies that are separated from f_0 or f_L depending on whether the loop starts with or without an initial lock condition. The linearity of the frequency-to-voltage conversion characteristics for the PLL is determined solely by the vco conversion gain. For most PLL applications, the vco is required to have a highly linear voltage-to-frequency transfer characteristic.

functional applications

Now that you are familiar with what a PLL is, you probably wonder what it all means and what it can do for you. As a functional building block the PLL is suitable for a wide variety of frequency related applications. These applications generally fall into one or more of the following catagories: fm demodulation, frequency synthesis, frequency synchronization, signal conditioning, a-m demodulation, and frequency modulation.

fm demodulation

If the PLL is locked to an fm signal, the vco will track the instantaneous frequency of the input signal. The filtered error voltage V_d, which forces the vco to maintain lock with the input signal, then becomes the demodulated fm output. The linearity of the vco's voltage-to-frequency transfer characteristic determines the linearity of the demodulated signal. Some typical fm demodulation applications are discussed below.

Broadcast fm detection. In this application, the PLL can be used as a complete fm i-f strip, limiter, and fm detector. It can be used to detect wide- or narrow-band fm signals with greater linearity than can be obtained by other means for frequencies within the range of the vco (presently up to about 50 MHz). One increasingly popular use of the PLL is in scanning receivers where a number of broadcast channels may be sequentially monitored by simply

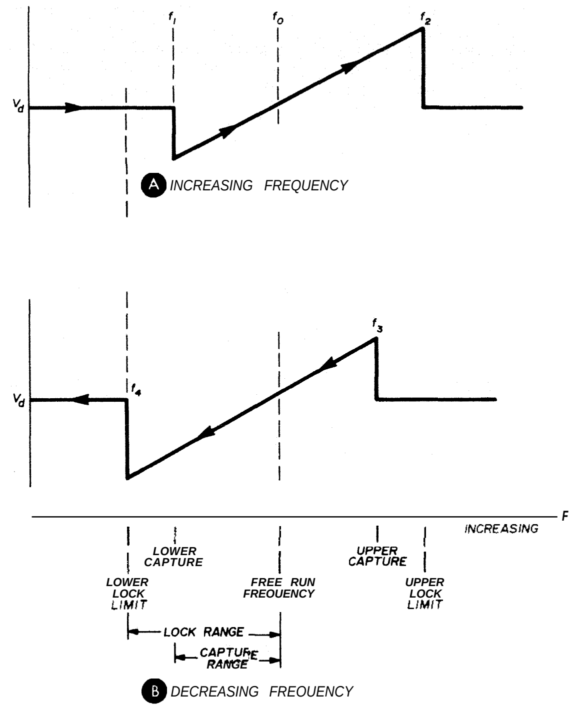


fig. 4. Frequency-to-voltage transfer characteristics of a phase-locked loop for increasing input frequency, A, and decreasing input frequency, B.

varying the vco free-run frequency. (Scanning receivers are also using a digital PLL technique which is, in principle, similar to linear PLLs).

Fm telemetry. This involves demodulation of a frequency-modulated subcarrier. One example is the use of the PLL to recover the SCA (storecast music) signal from the combined signal of commercial fm broadcast stations. The SCA signal is a 67-kHz fm subcarrier which puts it above the frequency spectrum of the normal stereo or monaural program material.

Frequency-shift keying **(FSK)**. This is essentially digital fm. Frequency-shift keying is a means for transmitting digital information by a carrier which is shifted between two discrete frequencies (as in RTTY, for example). In this case, the two discrete frequencies correspond to a digital 1 (mark) and a digital 0 (space), respectively. When the FSK signal is connected to a PLL, the demodulated output (error voltage) shifts between two discrete voltage levels which correspond to the demodulated binary output.

frequency synthesizer

Frequency multiplication can be achieved with the PLL in two ways: locking to a harmonic of the input signal, or insertion of a counter (digital frequency divider) in the loop. Harmonic locking is the simpler

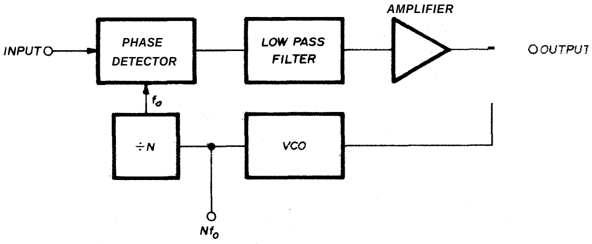


fig. 5. Phase-locked loop frequency synthesizer using a digital frequency divider.

and is achieved by setting the vco free-run frequency to a multiple of the input frequency and allowing the PLL to lock. A limitation of this scheme, however, is that the lock range decreases as successively higher and weaker harmonics are used for locking. This limits the practical harmonic locking range. For large multiples, the use of a digital frequency divider provides better results; the basic arrangement is shown in **fig. 5.** The loop is broken between the vco and the phase detector and a counter is inserted. In this case, the fundamental of the divided vco frequency is locked to the input frequency so the vco is actually running at a multiple of the input frequency. The amount of multiplication is determined by the counter.

In frequency multiplication applications, it is important to remember that the phase comparator is actually a mixer and its output contains the sum and difference frequency components. The difference frequency component is dc and is the error voltage which drives the vco to keep the loop in lock. The sum frequency components (basically twice the frequency of the input), if not well filtered, will induce incidental fm on the vco output. This happens because the vco is running at many times the input frequency. The sum frequency component appearing at the control voltage input of the vco causes a periodic variation of its frequency about the desired multiple. For frequency multiplication, it is generally necessary to filter quite heavily to remove this sum component. The tradeoff is reduced capture range and a more underdamped loop transient. The size of the loop filter limits the minimum input frequency.

frequency synchronization

Using a PLL system, the frequency of a less precise vco can be phase locked with a low level but highly stable reference signal. The vco output reproduces the reference signal at the same per unit accuracy but at a much higher power level. In some applications, the synchronizing signal can be a low duty cycle burst at a specific frequency. The PLL can be used to regenerate a coherent CW reference frequency by locking onto the short synchronizing pulse. An example of this is phase-locked chroma-reference generators in color television receivers.

In digital systems the PLL can be used for a variety of synchronization functions. For example, two clocks can be phase locked to each other so that one can function as the backup for the other. Other popular applications include locking to National Bureau of Standards' station WWVB to generate an inexpensive laboratory frequency standard.

signal conditioning

By selecting the proper vco free-run frequency, the PLL can be made to lock to any number of signals present at the input. The vco output tracks that frequency while attenuating the undesired frequencies of sidebands present at the input. If the loop bandwidth is sufficiently narrow, the signal-to-noise ratio at the vco is better than that at the input.

a-m demodulation

A-m demodulation can be achieved with a PLL by using the scheme shown in **fig. 6.** In this mode of operation the PLL functions as a synchronous a-m detector. The PLL locks onto the carrier of the a-m signal so the vco output has the same frequency as the carrier, but with no amplitude modulation. The vco will track the input but with a 90° phase shift; if the input is now sent through a 90° phase-shift network and fed into a multiplier, the output of the second multiplier will be directly proportional to the amplitude of the input signal. The PLL still exhibits the capture phenomenon; thus the loop maintains a high degree of selectivity centered about the free-run vco frequency. Since this method is essentially a coherent detection technique which involves the averaging of two companded signals, it offers a higher degree of noise immunity than a conventional peak-detector type a-m demodulator.

fm modulation

Since the PLL has a voltage-controlled oscillator, it's possible to inject a signal into the loop and cause the vco to change frequency. This signal can be injected at the lowpass filter or across the timing component; the per cent of modulation is controlled by

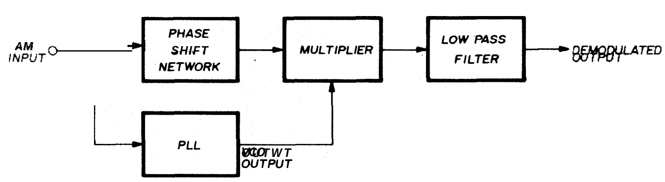


fig. 6. Coherent a-m detector circuit using a phase-locked loop.

the amplitude cf the injected signal — linearity is a function of the voltage-to-frequency transfer characteristics of the vco.

PLL design considerations

Many integrated circuits use phase-locked loops along with specialized circuitry to form specialized circuit functions. The versatility of these special systems has been sacrificed for convenience and circuit optimization. Circuits such as the μA758 Stereo Multiplex Decoder, TDA2541 Video I-F System, and others have built-in PLLs which have been optimized expected input frequency range. Since the loop's capture ability is a function of the difference between the incoming and free-running frequencies, the band edges of the capture range are always centered about the free-run frequency. Typically the lock range is also centered about the free-run frequency. By choosing a center frequency offset from the incoming signal, the detection or tracking range of the loop is limited to one side. This permits rejection of an adjacent higher (or lower) frequency signal and still permits wideband operation (narrowband operation reduces tracking speed).

table 1. User's guide to phase-locked loop ICs.

	upper freq (MHz)	max lock range (%F_o	fm distortion	output swing ±5% deviation (volts p-p)	center frequency stability (ppm/°C)	frequency drift/w supply voltage (%/V)	input resistance	a-m output avail	typical supply current (mA)	supply voltage range (volts)
NE560	30	40%	0.3%	1	±600	0.3	2kt	no	9	+16 to +26
NE561	30	40%	0.3%	1	±600	0.3	2kt	yes	10	+16 to +26
NE562	30	40%	0.5%	1	±600	0.3	2kt	no	12	+16 to +30
NE564	50	30%			±200					+5 to +10
NE565	0.5	120%	0.2%	0.15	±200	0.16	5kt	no	8	±6 to ±12
SE565	0.5	120%	0.2%	0.15	±100	0.08	5kt	no	8	±6 to ±12
NE567	0.5	14%	5%"	0.20	35±60	0.7	20kt	yes"	7	+4.75 to +9
SE567	0.5	14%	5%"	0.20	35±60	0.5	20kt	yes*	6	+4.75 to +9
NE566	0.5	N/A	0.2%	30%/V§	±200	0.16			7	+12 to +26
SE566	0.5	N/A	0.2%	30%/V§	±100	0.08			7	+12 to +16

• A-m and fm outputs are available, but are not optimized for linear demodulation

t Input biased internally.

§ Figure shown is vco gain in per cent deviation per volt.

to perform best in those applications. But there are also PLLs which can be used as basic building blocks; these offer the most flexibility in circuit design. Signetics offers three basic classes of single-chip PLL circuits: the general purpose PLL, a PLL with an added multiplier, and the PLL tone decoder. National Semiconductor has a general-purpose PLL and a tone decoder. A more complete list of the PLLs available from various manufacturers is detailed in **table 1.**

To obtain the optimum performance from a PLL circuit it is important that the user become familiar with the tradeoffs that can be made. To be more specific, the following discussion will be directed at the 560, 561, 562, 564, 565, 566, and 567 phase-locked loops. The tradeoffs and loop conditions, however, will hold true for all basic PLLs. Generally speaking, the user is free to select the frequency, tracking or lock range, capture range, and input amplitude.

center frequency selection

Setting the center frequency is accomplished by selecting one or two external components. This free-running frequency is usually set in the center of the

As was mentioned earlier, the loop uses a multiplier in which the input signal is multiplied by a unity square wave at the vco frequency. The odd harmonics present in the square wave permit the loop to lock to input signals at these odd harmonics. Thus the center frequency may be set to 3 times or 5 times the input signal. The tracking range will, however, be considerably reduced as higher harmonics are used.

It should also be noted that the PLL can lock onto the harmonic of the desired signal. If this is not acceptable, steps must be taken (such as prefiltering) to prevent undesired lockup of the loop to harmonically related signals.

In evaluating the loop for a specific application, compute the magnitude of the expected signal component nearest f_o. This magnitude can be used to estimate the lock and capture range.

The PLLs are stabilized against center frequency drift due to power supply variations and the 565 and 567 are temperature compensated over the full military temperature range (−55 to 125°C). All of the loops are affected by external components which must have equal (or better) stability over the desired operating temperature range.

Two things limit the lock or tracking range of a PLL. First, any vco can only swing so far; if the input signal frequency goes beyond this limit, lock will be lost. Second, the voltage developed by the phase detector is proportional to the product of both the phase and the amplitude of the in-band component to which the loop is locked. If the signal amplitude decreases, the phase difference between the signal and the vco must increase to maintain the same output voltage and, hence, the same frequency deviation.

It often happens with low input amplitudes that even the full $\pm 90°$ phase range of the phase detector cannot generate enough voltage to permit tracking over wide deviations. When this occurs, the effective lock range is reduced. Therefore, if the input signal is weak, you must give up some tracking capability and accept greater phase errors. Conversely, a strong input signal will allow you to use the entire vco swing capability and keep the vco phase (referred to the input signal) very close to $90°$ throughout the range.

Note that tracking range does not depend on the lowpass filter. If a lowpass filter is in the loop, however, it will have the effect of limiting the maximum rate at which tracking can occur. Obviously, the voltage across the lowpass filter capacitor cannot change instantly, so lock may be lost when sufficiently large step changes occur. Between the constant frequency input and the step-change frequency input is some limiting frequency slew rate at which lock is just barely maintained. When tracking at this rate, the phase difference is at its limit of $0°$ to $180°$. It can be seen that if the lowpass filter's cutoff frequency is low, the loop will be unable to track as fast as when the lowpass filter's cutoff frequency is higher. Thus, when maximum tracking rate is needed, the lowpass filter should have a high cutoff frequency. However, a lowpass filter with a high cutoff frequency will attenuate the sum frequencies to a lesser extent so the output contains a significant and often bothersome signal at twice the input frequency, (Remember that the multiplier forms both the sum and difference frequencies; during lock the difference frequency is zero, but the sum frequency at twice the locked frequency is still present.) If necessary, the unwanted sum frequency component can be filtered out with an external lowpass filter.

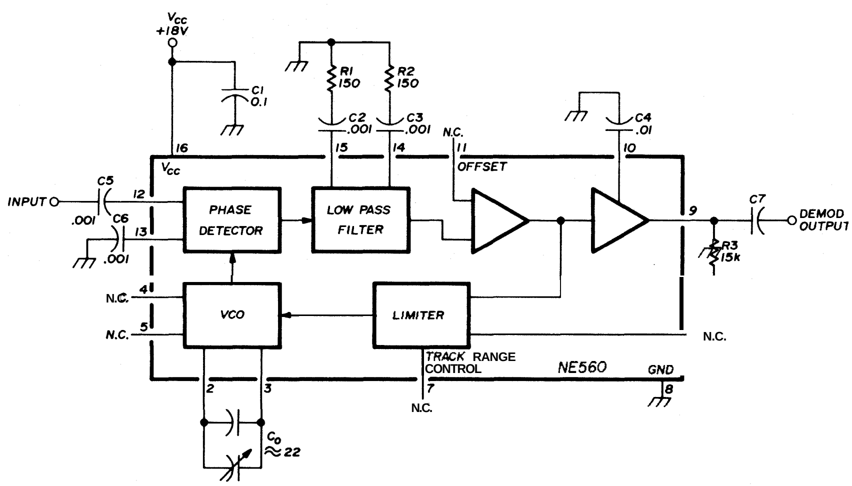


fig. 7. 10.7-MHz fm demodulator using an NE560 PLL IC. Design details are given in the text.

capture range control

There are two main reasons for making the lowpass filter time constant large. First, a large time constant provides an increased memory effect in the loop so that it remains at or near the operating frequency during momentary fading or loss of signal. Second, the large time constant integrates the phase detector output so that increased immunity to noise and out-band signals is obtained.

In addition to the lower tracking rates of large loop filters, other penalties must be paid for the benefits gained; the capture range is reduced and the capture transient becomes longer. Reduction of capture range occurs because the loop must use the magnitude of the difference frequency component at the phase detector to drive the vco toward the input frequency. If the cutoff frequency of the lowpass filter is low, the difference component amplitude is reduced and the loop cannot swing as far. Thus, capture range is reduced.

choice of input level

Whenever amplitude limiting of the in-band signal occurs, whether in the loop input stages or prior to the input, the tracking (lock) and capture ranges become independent of signal amplitude.

Better noise and out-band signal immunity is achieved when the input levels are below the limiting threshold since the input stage is in its linear region and the creation of cross-modulation components is reduced. Higher input levels will allow somewhat faster operation due to greater phase detector gain and will result in a lock range which becomes constant with amplitude as the phase detector gain becomes constant. Also, high input levels will result in a linear phase-versus-frequency characteristic.

lock-up time and tracking speed control

In tracking applications, lock-up time normally has little consequence, but occasions do arise when it is desirable to keep lock-up time short to minimize data loss when noise or extraneous signals drive the loop out of lock. Lock-up time is of great importance in tone decoder type applications. Tracking speed is important if the loop is used to demodulate an fm signal. Although tthe following discussion dwells largely on lock-up time, the same comments apply to tracking speed.

No simple expression is available which adequately describes the acquisition or lock-up time. This will be better appreciated after you have reviewed the following factors which influence lock-up time:

1. Input phase
2. Lowpass filter characteristic
3. Loop damping
4. Deviation of input frequency from center frequency
5. In-band input amplitude
6. Out-of-band signals and noise
7. Center frequency

Fortunately, it is usually sufficient to know how to improve the lock-up time and what must be traded off to obtain faster lock-up. Suppose you have set up a loop or tone decoder and find that occasionally the lock-up transient is too long. Remember all the factors that influence lock, what can be done to improve the situation?

1. Initial phase relationship between the incoming signal and the vco; this is the greatest single factor influencing the lock time. If the initial phase is wrong, it first drives the vco frequency away from the input frequency so the vco frequency must walk back on the beat notes. The only way to overcome this variation is to send phase information all the time

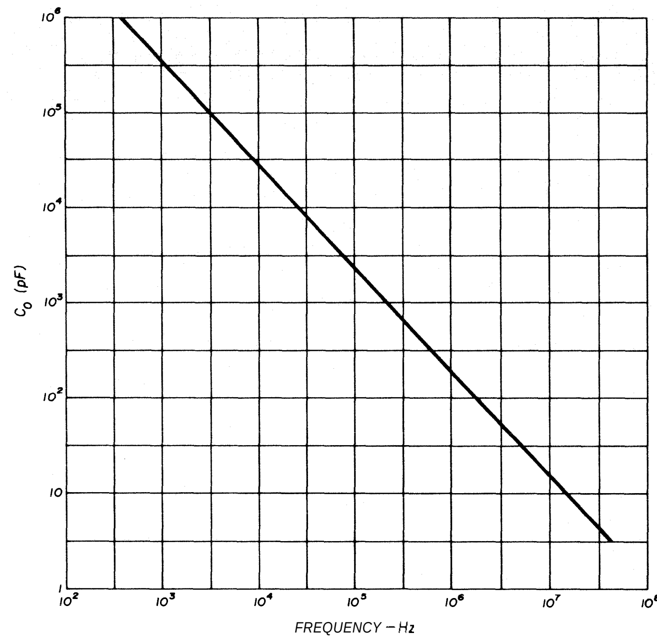


fig. 8. Free-running oscillator frequency as a function of vco timing capacitance.

so that a favorable phase relationship is guaranteed at $t = 0$. Usually, however, the incoming phase cannot be controlled. Using two TTLs with the voltage-controlled oscillators synchronized 90° apart reduces worst-case lock-up time by one-half because the input can never be more than 45° out of phase with one of the loops.

2. Lowpass filter. The larger the time constant of the lowpass filter, the longer the lock-up time. You can reduce the lock-up time by decreasing the filter time constant, but in doing so you sacrifice some of the noise immunity and out-of-band signal rejection which caused you to use a larger filter in the first place. You must also accept a sum frequency (twice the vco frequency) component at the lowpass filter and greater phase jitter resulting from out-of-band signals and noise. In the case of the tone decoder (where control of the capture range is required since it specifies the device bandwidth) a lower value lowpass capacitor automatically increases bandwidth. You gain speed only at the expense of added bandwidth.

3. Loop damping for a simple time constant lowpass filter is:

$$\zeta = \frac{1}{2}\sqrt{\frac{1}{\tau K_v}}$$

Damping can be increased not only by reducing τ, as discussed above, but also by reducing the loop gain K_v. By using the loop gain reduction to control band-

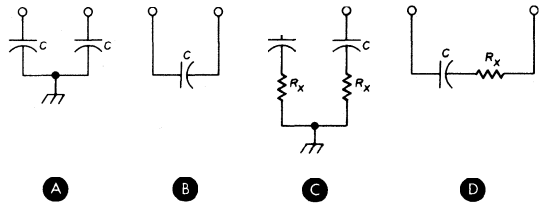


fig. 9. Typical RC lowpass filters. Approximate formulas describing their frequency response are presented in the text.

width or capture and lock range, you achieve better damping for narrow bandwidth operation. The penalty for this damping is that more phase detector output is required for a given deviation so that phase errors are greater and noise immunity is reduced. Also, more input drive may be required for a given deviation.

4. Input frequency deviation from free-running frequency. Naturally, the further an applied input signal is from the free-running frequency of the loop, the longer it will take the loop to reach that frequency because of the charging time of the lowpass filter capacitor. Usually, however, the effect of this frequency deviation is small compared to the variation resulting from the inital phase uncertainty. Where loop damping is very low, however, it may be predominant.

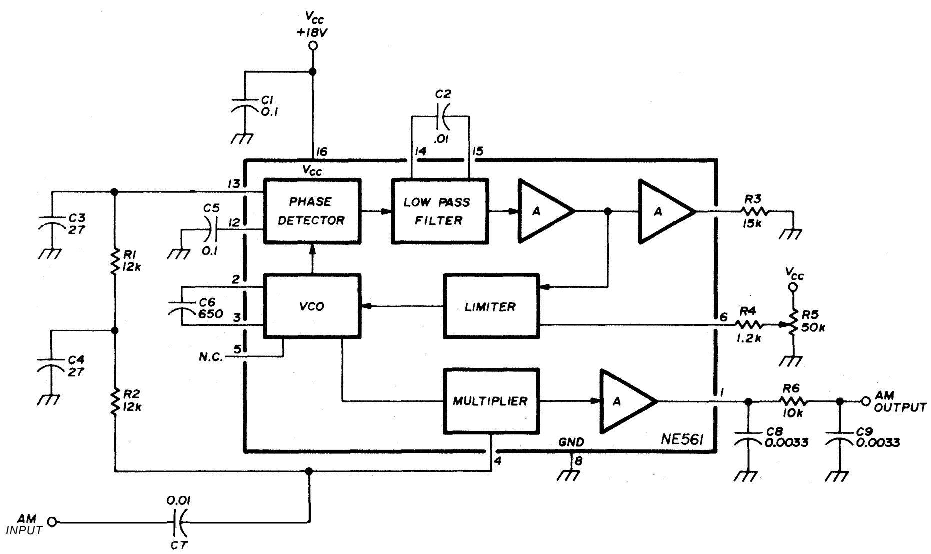


fig. 10. A-m detector for 455-kHz i-f using the NE561 PLL.

5. In-band input amplitude. Since input amplitude is one factor in the phase detector gain K_d, and since K_d is a factor in the loop gain K_v, damping is also a function of input amplitude. When the input amplitude is low, the lock-up time may be limited by the rate at which the lowpass capacitor can charge with the reduced phase detector output (see **4** above'.

6. Out-band signals and noise. Low levels of extraneous signals and noise have very little effect on the lock-up time, neither improving or degrading it. However, large levels may overdrive the loop input stage so that limiting occurs, at which point the in-band signal starts to be suppressed. The lower effective input level can cause the lock-up time to increase, as discussed in **5** above.

7. Center frequency. Since lock-up time can be described in terms of the number of cycles to lock, fastest lock-up is achieved at higher frequencies. Thus, whenever a system can be operated at a higher frequency, lock will typically be faster. Also, in systems where different frequencies are being detected, on average the higher frequencies will be detected before the lower frequencies. Because of the wide variation due to initial phase, however, the reverse may be true for any single trial.

In all of the above design considerations, it is important to remember that the PLL is a dc loop. Any dc level change injected into the loop will affect its operation. The loop is also sensitive to temperature changes because most voltage-controlled oscillators have a temperature coefficient of around 600 ppm/C°. The resistors and capacitors used in the frequency determining network also have temperature coefficients which must be considered when designing circuits using the PLL.

design examples

Let's take a look at a practical design example using the NE560 PLL as a 10.7-MHz fm demodulator (see **fig. 7**).

Supply voltage. Generally, the operating voltage is determined by the available power supply or the device data sheet. The manufacturers specify the V_{CC} at which the device parameters were measured. This simplifies the V_{CC} selection since the device is tested at an optimum voltage; for the NE560 this voltage is 18 volts. Capacitor C1 is a decoupling capacitor for the supply and should be located as close as possible to the V_{CC} pin of the IC.

VCO free-run frequency. Since this is a 10.7-MHz detector, the approximate timing capacitance can be found on the data sheet graph. This graph **(fig. 8)** shows that the timing capacitor should be about 22 pF. An approximate value for the timing capacitor can be found from:

$$C_o \cong \frac{300}{f_o}$$

where C_o is in pF and f_o is in MHz. Using this formula, the capacitor is calculated to be about 28 pF. The design example uses a 22-pF capacitor in parallel with an 8-pF trim capacitor for fine tuning.

Lowpass filter. The output of the phase detector is the sum and the difference of the input fm signal and the vco frequency. The loop filter must remove the sum component. Because the modulation on the incoming signal contains the information desired, it is necessary to retain the difference frequency. In addition, the attenuation of the high-frequency component increases the interference rejection characteristics. To maintain loop stability at all signal levels, the loop cannot cause more than 12 dB per octave rolloff.

Fig. 9 shows several lowpass filter configurations. The circuits of **figs. 9A** and **9B** can provide **6** dB per octave rolloff at the desired bandwidth frequency; resistor R_x in **9C** and **9D** will break the response up at higher frequencies. R_x is typically between 50 and 200 ohms. Because of the complexities of the transfer functions, which many designers use to characterize these filters, it is usually easier to use approximation formulas for the lowpass filter. For **fig. 9A** and **9B** the formula is

$$C = \frac{26.6}{f}\,\mu F$$

where f is the desired cutoff frequency. The lowpass filters of **fig. 9C** and **9D** can be approximated by

$$C = \frac{13.3}{f}\,\mu F$$

At frequencies greater than 5 MHz, the filters of **fig. 9C** and **9D** will provide greater loop stability.

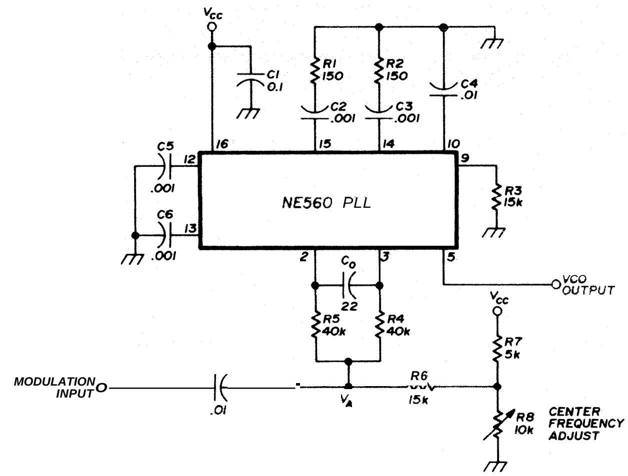


fig. 11. 10.7-MHz fm modulator using the NE560 PLL. Resistor R8 provides ±10% variation of the center frequency. For best waveform linearity resistors R4 and R5 should be equal; minimum value for R4 and R5 is 20k.

Assuming that the desired frequency bandwidth of the demodulated signal is 15 kHz, the filter capacitor values (C2 and C3 in **fig. 7**) are

$$C = \frac{13.3 \times 10^{-6}}{15 \times 10^{3}} = 886\ pF$$

or approximately 0.001 μF

Resistors R1 and R2 were selected to be between 50 and 200 ohms; I used 150-ohm resistors.

De-emphasis filter. Standard fm broadcast stations use a de-emphasis time constant of approximately 75 microseconds. According to the NE560 data sheet, the internal resistance into pin 10 is 8k. The 3 dB rolloff frequency is given by

$$f_{3dB} = \frac{1}{2\pi R_D C_D} \text{ where } R_D \text{ is } 8000$$

For a time constant of 75 microseconds

$$C_D = \frac{1}{2\pi f R_D} = \frac{75 \times 10^{-6}}{R_D} = 0.0094\ \mu F$$

In practice, we can use a 0.01 μF capacitor since the internal resistance has a ±20 per cent tolerance.

Output. The output, pin 9, has a 15k dc load and ac coupled output. The detected output amplitude for 75 kHz deviation, according to the data sheet, is 30 mV minimum. The demodulated output is emitter coupled and requires a dc path to ground.

Input level. The input level required for a constant tracking range is 2 mV. Also, the a-m rejection char-

table 2. Dedicated ICs which use phase-locked loops.

function	part number	manufacturer
FSK modulator/ demodulator	XR210	Exar
CB synthesizer	HCTR0340	Hughes
	MC15104	Motorola
	MM55104	National
	NE575	Signetics
	SP8920	Plessey
TV synthesizer	NC6410	Nitron
Tone decoder	NE567	Signetics
	LM567	National
	XR567	Exar
Stereo decoder	μA758	Signetics
	LM1310	National
	LM1800	National
VCO function generator	NE566	Signetics
	LM566	National
	11C58C	Fairchild
	MC4024	Motorola

acteristic for high input signal levels is reduced for signals greater than 30 mV, so the input signal should be between 2 and 20 mV.

synchronous a-m detector

Because the NE561 has a multiplier incorporated on the chip, it is possible to accomplish synchronous a-m detection; **fig. 10** shows the circuit. The vco tracks the input to the phase detector with a 90° phase shift. By shifting the input signal 90° the vco and the input become in phase. These signals are mixed in the multiplier — the output is the a-m signal (due to the amplitude difference of the input, the constant amplitude of the vco, and the 90° phase difference).

To accomplish this phase shift of the input, a simple RC lag network is used. The values of R1, C3, R2, and C4 are calculated using the relationship

$$R1C3 = R2C4 = \frac{1}{2\pi f_o} \text{ (assume } R1 = 12k\text{)}$$

$$\text{Thus } C3 = C4 = \frac{1}{2\pi 455x10^3 x12x10^3} = 29\,pF$$

(Use a standard value, 27 pF.)

The timing capacitor is calculated as

$$C_6 \approx \frac{300}{f_o}\,pF \text{ (}f_o\text{ in MHz)}$$

$C_6 \approx 660\,pF$ (use a standard value, 680 pF)

By controlling the current into pin 6 the vco can be fine tuned so the absolute value of C6 is not critical. By injecting or loading current into pin 6, the vco frequency can be adjusted ±20 per cent (with 4 mA control current). Since no information is taken from the lowpass filter, the filter capacitor, C2, is not critical; the value is selected to maintain loop stability. I used a 0.01 μF for C2 which worked quite well.

Capacitors C8, C9, and resistor R_c form a post a-m filter to eliminate carrier feedthrough; R3 is optional because that output is not used.

frequency modulation

Since the PLL IC contains a vco, frequency modulation is accomplished by summing a modulating signal with the error signal or by changing the dc level to control the vco at the modulating rate. Referring to **fig. 10,** if the current into pin 6 becomes a variable current, the vco will change frequency at the modulated rate. The vco output is taken at pin 5.

Another method of changing the frequency is to insert offset currents across the timing capacitor. To build a 10.7-MHz fm modulator, the circuit of **fig. 7** can be modified as shown in **fig. 11.** Note that the input is ac grounded and the vco output is taken from pin 5. Frequency deviation can be approximated by the formula given below.

$$f = f_o \left[1 - \frac{V_A - 6.4}{1300\,R} \right]$$

By incorporating a switching arrangement the circuit can be used as both an fm demodulator and modulator.

summary

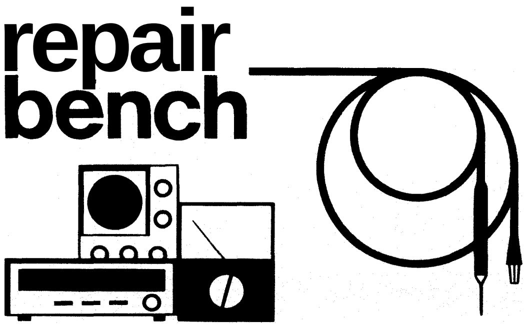
In addition to the basic PLL, many manufacturers have ICs for dedicated applications which incorporate PLLs into the design. For example, the CA3089 is an fm i-f strip which contains all the functions of an i-f amplifier and audio detector. Because these parts are dedicated to a specific application, however, they lose flexibility. **Table 2** is a partial list of parts which use PLLs in dedicated applications. This list is by no means complete, but is an indication of the various types of circuits which are presently available.

In summary, the PLL is a versatile building block for use in frequency manipulating circuits. By careful design, an awareness of the PLL's limitations, and knowledge of the design tradeoffs, you will find them as easy to use as operational amplifiers.

bibliography

1. ***Signetics Analog Data Manual,*** Signetics Corporation, Sunnyvale, California, 1977.
2. F. M. Gardner, ***Phase Lock Techniques,*** John Wiley & Sons, New York, **1966.**
3. J. A. Mattis, "The PLL — A Communications System Building Block," ***Broadcast Engineering,*** February, 1972.

ham radio



Bob Stein, W6NBI

semi-precision voltage calibrator for digital voltmeters

If you are relying on your digital voltmeter for accurate voltage measurements, you may have a problem which you have never considered — how well is your voltmeter calibrated? The specifications for the instrument tell you its guaranteed accuracy, and should also state the period of time that this accuracy will be maintained after calibration. Typically, the period is between three months and one year, depending on the manufacturer's tests or specmanship. After that, you must accept the DVM's accuracy on faith, or have it recalibrated. If you are fortunate enough to have access to a calibration-standards laboratory, recalibration is no problem. Otherwise, you must return the voltmeter to the manufacturer's service center at a cost of at least $20 plus shipping, plus the inconvenience of being without it for several weeks.

An alternative is to build the voltage calibrator described in this article. Although it will not permit the same degree of calibration accuracy which can be achieved with precision voltage standards, it will indicate whether or not your DVM requires recalibration and will allow you to perform the calibration if you are willing to accept about 0.3 per cent accuracy on dc and 1 per cent on ac.

Because of its simplicity, the calibrator has several limitations, all of which are based on the assumption that most digital voltmeters owned by hams are inexpensive, 3-112-digit types. First of all, both the ac and dc calibration adjustments on the DVM must be made by a single control for each, on the lowest ac and dc voltage ranges (accuracy of the higher voltage ranges is dependent upon the accuracy of the input voltage dividers in instruments of this type). Second, the DVM calibration voltages specified in the instruction book must be within 25 per cent of either 100 millivolts or 1 volt; the voltage calibrator is designed to allow you to choose one of these two calibration voltages. And third, the dc input resistance of the DVM must be at least 10 megohms, although lower input resistances can be accommodated at the expense of accuracy.

In addition to the voltage calibrator, an audio-frequency [illegible]ato[illegible] ha[illegible]ing a l[illegible] output [illegible] is required. This generator must supply a voltage equal to the specified ac calibration voltage, at the frequency specified by the manufacturer of the DVM. Other than that, the voltage calibrator is self-contained and self-calibrating.

circuit description

The voltage calibrator circuit is shown in **fig. 1.** Two LM308 op amps are used in a precision full-wave rectifier circuit to obtain an ac calibrating voltage by substitution. A reference voltage source, to be described later, provides the dc calibrating voltage and a reference for the rectifier circuit. Either a 100.0-millivolt or a 1.000-volt reference voltage source is used, depending on the requirements of the DVM undergoing calibration. It can be seen that when switch S1 is in the DC CAL position, the reference voltage source is connected directly to the DVM, permitting calibration of the dc voltage range.

Since the precision full-wave rectifier is dc coupled, it will respond to dc as well as ac; its output can therefore be calibrated from the reference voltage source. The ac signal from the audio generator is adjusted to provide an equivalent dc output from the rectifier, which is measured using the DVM (already calibrated on its dc range). Since the ac input is now known in terms of the dc output from the rectifier, the DVM is connected directly to the audio generator and calibrated for ac voltage.

Let's now examine the precision rectifier, described by Dobkin in **reference 1,** in more detail. LM308 op amps were selected because of their low input-bias currents, ensuring that the bias current supplied from the reference voltage source would not exceed 14 nanoamperes. When an ac voltage is applied to the inverting inputs of both op amps, U1 functions as a half-wave rectifier and produces the output shown in **fig. 2B.** The positive half-cycle of the input is inverted in the op amp and applied to the inverting input of U2 through CR1 and R6. The inverted negative half-cycle of the input is clamped at approximately 0.7 volt by CR2, but is isolated from U2 by CR1. Therefore the output of U1 which reaches U2 is a true half-wave rectified version of the input, having the same peak amplitude, e_{pk}.

To simplify the explanation, it is best to consider the two inputs to U2 separately. The output of U1 is applied to U2 through R6. Therefore the gain of U2 for this input is established by the ratio (R7 + R8)/R6. If this gain is set at 2.222 by means of GAIN control R8, the output resulting from this input would be as shown in **fig. 2C.** Simultaneously, the original ac input is applied to U2 through R4, with the gain of U2 for this input being determined by the ratio (R7 + R8)/R4. Since R4 is twice the value of R6, within the tolerances of 1-per cent resistors, the gain becomes 1.111 for the same setting of R8; the resultant output is shown in **fig. 2D.** Waveforms **C** and **D** are summed (more rigorously, the input currents are summed), producing the actual output shown in **fig. 2E.** Note that this is a full-wave rectified version of the ac input, but amplified by a factor of 1.111. The dc level at the output of the rectifier circuit is equal to the average value of the input signal times the gain of the circuit. Since the average value of a sine wave or a full-wave rectified sine wave is 0.6366 times the peak value, the dc output becomes $0.6366 \times 1.111e_{pk}$, or $0.707e_{pk}$, which is the rms value of the ac input.

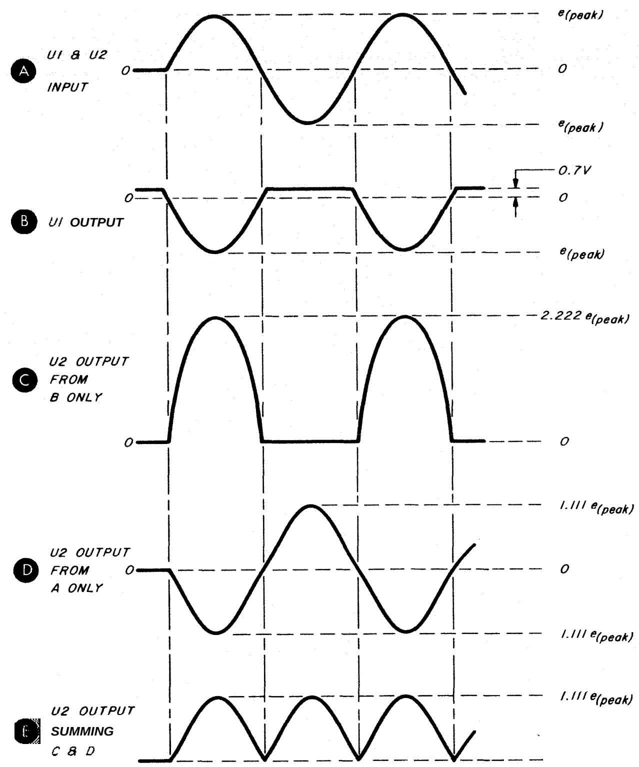


fig. 2. Waveforms of the precision full-wave rectifier circuit.

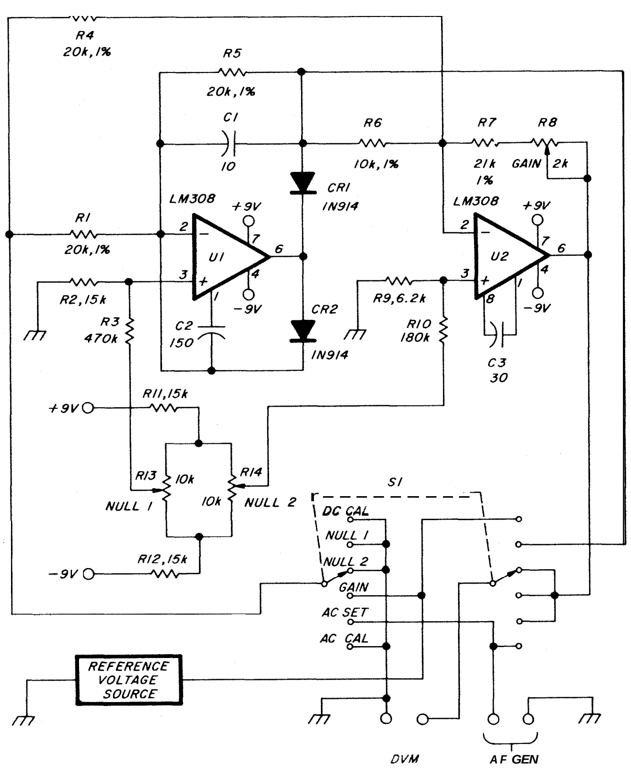


fig. 1. Schematic diagram of the semi-precision voltage calibrator. Switch S1 must be a nonshorting type rotary switch. The reference voltage source is shown in fig. 3.

Note that there are no coupling capacitors within the rectifier circuit, allowing it to operate as a dc amplifier having a gain of 1.111. Therefore a known dc voltage, E, applied to the input will produce an output equal to $1.111E$. It is this characteristic which permits ac calibration using a dc source.

The input offset voltages of U1 and U2 are nulled by means of potentiometers R13 and R14, respectively, in order to establish zero output voltage for zero input. The entire circuit may be powered by two inexpensive 9-volt transistor batteries, or a dual-voltage power supply can be used; total current drain is less than 2 milliamperes.

The sequence of operation is as follows: The DVM is calibrated on its dc range, as described previously, when switch S1 is set to DC CAL. When S1 is set to NULL 1, the output of U1 is nulled by means of R13, and when S1 is set to NULL 2, the output of U2 is nulled with R14. In the GAIN position of S1, the dc reference voltage source is applied to the rectifier input and GAIN control R8 is adjusted for a dc output, indicated on the DVM, of 1.111 times the reference voltage. Next, S1 is switched to its AC SET position, which connects the external af generator to the input of the rectifier. The amplitude of the audio-generator signal is then adjusted to provide a dc reading on the DVM which is equal to the dc reference voltage;

therefore the rms value of this ac signal is now equal to the dc reference. Finally, S1 is set to **AC CAL,** which connects the audio generator directly to the DVM. Changing the DVM function switch to **AC** VOLTS then permits calibration of its ac range.

referencevoltage source

As you may have deduced from the discussion so far, the accuracy of the voltage calibrator is dependent upon the accuracy of the reference voltage source. Fortunately, the source utilizes an inexpensive and commonly available 1.35-volt mercury cell. Suitable types, the most expensive of which costs about $2, are listed below. All are 1.35-volt reference cells. Mercury cells are also made with a 1.4-volt emf; they are not usable in this application.

Mallory	Eveready	Burgess
RM12R	E12N	HG12R
RM401R	E401N	HG401R
RM502R	E502	HG502R
RM601R	E601	HG601R

Although the battery voltage is specified to only two decimal places, it's been determined by a large equipment manufacturer that the terminal voltage of a new battery under about 0.1-milliampere load is within a millivolt or two of 1.354 millivolts. This value has been used as a basis for the reference voltage.

Fig. 3A shows the circuit of the 100.0-millivolt source, and **fig. 3B** shows the 1.000-volt source. To establish a reference voltage with an accuracy of ±0.2 per cent, it would be necessary to use selected or special values of 0.1 per cent resistors. To eliminate this obviously impractical requirement (for the average user), standard values of 1 per cent resistors have been specified. However, these must be bridged or measured on a calibrated digital ohmmeter which is accurate to within ±0.1 per cent. Since it is the ratio of the two sections of the voltage divider across the battery, (rather than the absolute resistances) which determines the reference voltage, it is possible to calculate a value for resistor R when the values of R101 and R102 are accurately known. Selecting a standard 1 per cent value which is closest to the calculated value of R will always result in a divider within the required accuracy."

For example, assume that you have borrowed a Wheatstone bridge and have determined that the values of RIOI and R102 for a 100.0-millivolt source are 9984.7 and 803.5 ohms, respectively. From **fig. 3A,**

$$R_p = \frac{RIOI}{12.540} = \frac{9984.7}{12.540} = 796.2\ ohms$$

***Refer to the end of this article for a source of these resistors.**

Also from **fig. 3A,**

$$\frac{1}{R} = \frac{1}{R_p} - \frac{1}{R102} = \frac{1}{796.2} - \frac{1}{803.5} = 0.00001141$$

$$R = 87\ 637\ ohms$$

The nearest standard 1 per cent resistance value is 86.6k ohms, which is then connected in parallel with R102. If you recalculate R_p using the limiting values of an 86.6k, 1 per cent resistor (*i.e.*, 85 734 and 87 466 ohms), and then calculate the ratios of $RIOI/R_p$, you will find that they are well within 0.1 per cent of 12.540. This holds true for any value of ***R*** required for the ranges of R101 and R102.

To simplify the selection of a value for R, R_p has been plotted against the resistance of R101 in **fig. 4** for both the 100.0-millivolt and 1.000-volt reference voltage sources. If the measured value of R102 is within 0.1 per cent of R_p, R will not be required. If the value of R102 is higher than R_p, R can be selected as described above. If R102 has a measured resistance less than R_p, it cannot be used, necessitating a change in either RIOI or R102.

When the voltage calibrator is in use, either the DVM or the precision rectifier input is connected in parallel with the bottom section of the voltage divider. The effect of this on the 100.0-millivolt source is negligible because of the low value of R_p. However, the effect on the 1.000-volt source cannot be disregarded because of the relatively high value of R101 in that circuit. Thus the equation for the selection of R_p in **fig. 3B,** as well as the graphical representation in **fig. 4,** has taken into account the shunting effect of a typical 10-megohm DVM input resistance.

The equivalent input resistance of the precision rectifier circuit (including the effect of the op amps'

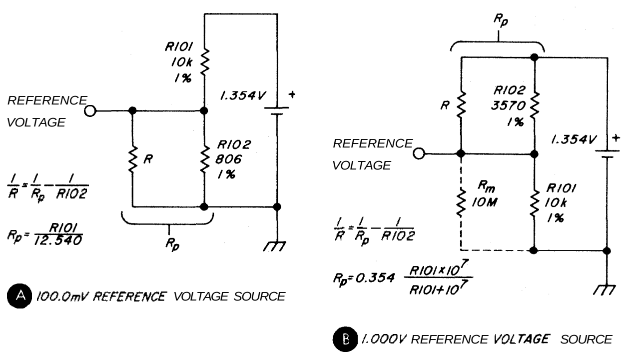


fig. 3. Reference voltage sources for use with the voltage calibrator. In each circuit, R is selected so that it, in parallel with R102, will satisfy the equations shown next to each circuit, within k0.2 per cent. R_m in circuit B is the input resistance of the DVM. Refer to the text for suitable battery types.

input-bias currents) will be between 5 and 65 megohms, depending on the individual devices which are used. When connected to the reference voltage source, the input voltage to the rectifier circuit will differ from that applied to the DVM by less than 0.1 per cent, which is insignificant compared to the 1 per cent accuracy of the precision rectifier.

Little need be said about building the voltage calibrator. The parts layout and arrangement are not at all critical, allowing any type of construction to be used. Point-to-point wiring on perf or copper-clad board will suffice, as will a printed-circuit board.

It is recommended that potentiometers R13 and R14 be multiturn types, because the nulling adjustments are critical. Since the calibrator will be used infrequently, these controls as well as GAIN potentiometer R8 can be multiturn, screwdriver-adjust trimmer pots.

One precaution – do not solder the mercury cell into the circuit. Rather, buy or build a suitable holder. A fresh battery should always be used for calibration; the battery must be easily replaceable.

DVM calibration

Connect the DVM to be calibrated to the DVM terminals of the voltage calibrator, and set it to the appropriate dc range. Connect an audio-frequency generator to the AF GEN terminals of the calibrator, with the generator output turned down to zero. Apply +9 and –9 volts dc to the calibrator, and allow the calibrator and DVM to warm up for at least 15 minutes.

The complete calibration procedure is as follows. The numerical values given in each step are based on using the 100.0-millivolt reference; equivalent values for the 1.000-volt reference are in parentheses.

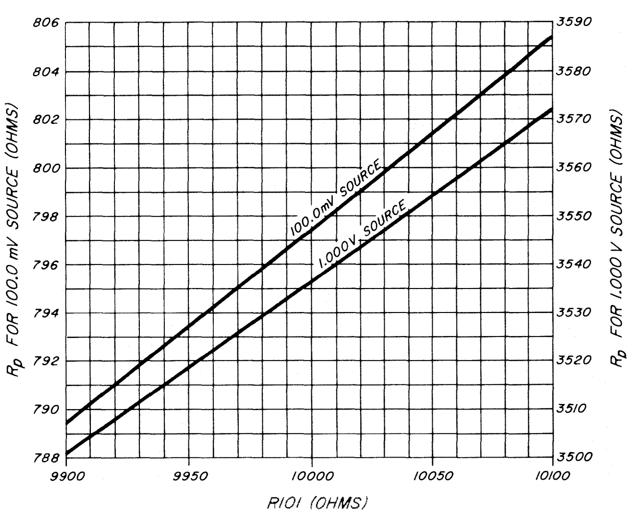


fig. 4. Values of R_p plotted against the measured resistance of R101, from the formulas given in fig. 3. R102 must be equal to or greater than R_p, as explained in the text.

1. Set S1 to DC CAL and adjust the DVM dc calibration control for a reading of 100.0 millivolts (1.000 volt).

2. Set S1 to NULL 1 and adjust NULL 1 potentiometer R13 for a reading on the DVM as close to zero as possible.

3. Set S1 to NULL 2 and adjust NULL 2 potentiometer R14 for a reading on the DVM as close as possible to zero.

4. Repeat **steps 2** and 3; the nulls should be within 0.1 millivolt (1 millivolt) of zero.

5. Set S1 to GAIN and adjust GAIN control R8 for a reading of 111.1 millivolts (1.111 volts) on the DVM.

6. Set S1 to AC SET and adjust the af generator output for a reading of exactly 100.0 millivolts (1.000 volt) on the DVM. Note that the DVM is still set to read dc voltage.

7. Set the DVM to read ac volts and set S1 to AC CAL. Adjust the DVM ac calibration control for a reading of 100.0 millivolts (1.000 volt).

The DVM is now calibrated within normally acceptable standards for most amateur work. As the title of this article indicates, the calibrator is only a semi-precision substitute for a voltage calibration standard. But for most ham and experimental applications, its accuracy should suffice.

source of precision resistors

Precision resistors for the voltage calibrator can be supplied by Melvin Sales Company, Post Office Box 5283, San Mateo, California 94402. The following sets will be made available:

description	postpaid cost, USA & Canada	postpaid cost via air, other countries
Matched set (±0.2%) of voltage-divider resistors for 100.0-mV source	$2.50	$3.00
Matched set (±0.2%) of voltage-divider resistors for 1.000-V source	2.50	3.00
Set of five 1% resistors for precision rectifier	1.50	2.00
*Set of five resistors (nominally 100, Ik, 10k, 100k, and 1M ohms) individually calibrated to ±0.1% for use as resistance standards	4.00	4.50

*Not required for voltage calibration, but are suitable for checking or adjusting resistance ranges of a digital volt-ohmmeter.

reference

1. Robert C. Dobkin, "Precision AC/DC Converters," Linear Brief LB-8, National Semiconductor Corporation, Santa Clara, California 95051, August, 1969.

ham radio

multiband J Antenna

Construction of a novel multiband J antenna for home or mobile use called the "J on a J"

Need a tri-band, vertical vhf antenna? Does your car look like an excited porcupine? Do you have a 30-meter tower and want omnidirectional coverage from a single antenna structure for all the repeater bands without the nulling effects of side mounting? The antenna system described here can be made for 10, 6, and 2 meters, 220 and 450 MHz, or all five bands. Antennas for the higher frequencies, 144, 220, and 450 MHz, can also be mounted as a single antenna structure on a plastic car (the antennas don't require a ground plane). Interested? Read on.

J antennas

The J antenna has been around a long time. It is primarily a ham antenna, never having gained acceptance in the commercial world, although it has been sporadically manufactured by several companies. The J antenna is a half-wavelength radiator mounted vertically and end-fed with a quarter-wavelength resonant transmission line (basically an end-fed Zep). Its angle of radiation and gain are the same as a half-wavelength antenna; electrically it is similar to the coaxial or sleeve antenna. The horizontal radiation pattern of the J antenna is almost a perfect circle, even when mounted on a metal roof of a car, because it doesn't require a ground plane and can be mounted well above the vehicle's roof.

The J antenna is fed at the appropriate point on the quarter-wavelength resonant line portion of the antenna (see **fig. 1**). Since the impedance on a resonant quarter-wavelength line goes from zero at the shorted end to infinite at the open end, it's possible to provide a good match to almost any transmission line. For best isolation from the supporting mast, however, it's necessary to feed the antenna with a balanced feedline; if you use coax, this requires a balun. If a balun is too much bother, the antenna can be fed directly with coax at the proper impedance matching point. In this case there will be some interaction between the antenna and the mast.

multibanding

After noting that the J antenna doesn't care what's below the shorting block at the bottom of the quarter-wave section, and it doesn't care about the diameter of the radiator (within reason), I concluded that a 450-MHz J antenna could be located on top of a 144 MHz J as shown in **fig. 2.** Chris Bushman, WB6EEQ, took me seriously and built just such a system. That was two years ago and it's still working very well.

Expanding this concept to include other bands would give you a single antenna structure, to install in that critical spot on the top of your tower, that in-

By Bob Thornburg, WB6JPI, 13135 Ventura Boulevard, Studio City, California 91604

cludes 10 meters, 6 meters, 2 meters, 220 MHz, and 450 MHz, Dimensions are given in **table 1** for all these although only a 144-220-450 mobile version has actually been built and tested by WA6VSK.

design

The design of the J antenna is very simple. The length of the quarter-wavelength matching section is given by

$$L = \frac{1134}{f_{MHz}}\ (cm) = \frac{2880}{f_{MHz}}\ (inches)$$

The half-wavelength radiator is approximately twice this length or

$$L = \frac{2181}{f_{MHz}}\ (cm) = \frac{5540}{f_{MHz}}\ (inches)$$

The spacing between the mast portion of the quarter-wave section and the quarter-wavelength rod is not critical as long as the two are parallel.

construction

A mobile 144-220-450 MHz J antenna can be easily built from a CB whip. A base station vhf/uhf J that includes 10 and 6 meters should be built with aluminum tubing. Since the antenna should not be insulated, a push-up mast would work well for the 10 meter and up antenna.

The mobile J antenna is made from an ordinary CB whip. For strength cut 60 cm (2 feet) from the top of the antenna and use this piece for the 144-MHz stub. The shorting blocks are aluminum stock drilled as shown in **fig.** 3. The spacers to maintain the stub spacing for 144 and 220 MHz are made from plexiglass, drilled the same as the corresponding shorting block. No plexiglass spacer is needed for the 450-MHz section. Cut the quarter-wavelength stubs

the quarter-wave 200-ohm point is an arbitrary length of TV twinlead. Other forms of feedline can also be used.

The 50-ohm coax may be directly fed to the stub at the appropriate 50-ohm point with the shield tied to

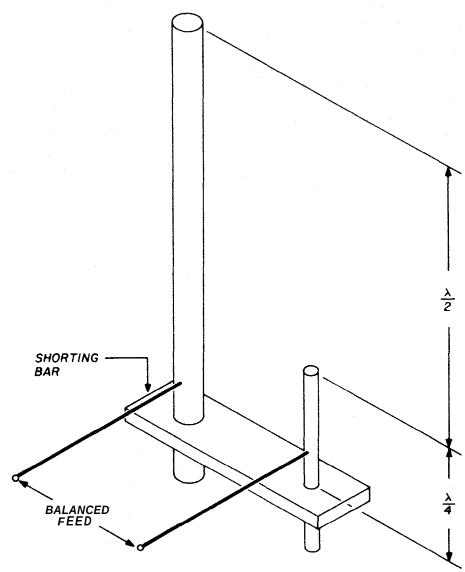


fig. 1. Basic J antenna consists of a half-wavelength vertical radiator and quarter-wavelength matching section. The balanced feedpoint (use balun with coax feedline) is moved along the quarter-wave section until a match is obtained. Unbalanced feed may also be used, as described in the text. The base of the antenna (shorting bar) may be grounded.

the opposite side. The stub could also be insulated from the shorting block and fed like a quarter-wave antenna (which it isn't). In general, the J antenna is relatively tolerant of the actual feedpoint or technique.

table 1. Construction dimensions for the half-wavelength J antenna.

frequency (MHz)	A	B	C	D	E	F
29.0	7.52 rn	2.51 m	15 cm	31.8 cm	6.4 cm	3.42 rn
52.5	4.16 m	1.38 rn	10 cm	19.1 cm	3.8 cm	1.89 m
147.0	1.49 m	49.5 cm	5 cm	6.4 cm	1.9 cm	67.3 cm
223.5	97.2 cm	32.1 cm	3.8 cm	4.1 cm	1.3 cm	44.1 cm
440.0	48.9 cm	16.5 cm	2.5 cm	2.1 cm	1.3 cm	22.4 cm
29.0	296.5″	98.75″	6″	12.5″	2.5″	134.5″
52.5	163.75″	54.50″	4″	7.5″	1.5″	74.25″
147.0	58.50″	19.50″	2″	2.5″	0.75″	26.50″
223.5	38.25″	12.63″	1.5″	1.63″	0.5″	17.38″
440.0	19.25″	6.50″	1.0″	0.83″	0.5″	8.83″

about 8-10 cm (3-4 inches) longer than necessary to allow tuning and as a convenient place to store the balun (under the shorting block).

The balun is simply a half-wavelength section of 50-ohm coaxial cable as shown in **fig. 4.** The feed to

The feedlines for the higher-frequency antennas actually become part of the lower-frequency antenna system. For this reason they must be tied closely to the main antenna pole. This provides capacitative coupling through the coax jacket and ensures that the outer shield electrically follows the fields on the main antenna; for all practical purposes it becomes part of that structure.

tuning

As noted previously, the specific location of the feedpoints is not too critical. However, the length of the stub is very critical and by adjusting the stub length, nearly any antenna length or feedpoint can be matched. The recommended procedure is to fully assemble the antenna, connect an rf source to the highest frequency coax, and adjust the appropriate stub for minimum vswr. If the vswr at the edge of the band is less than 2:1 (and more-or-less symmetrical) move on to the next highest frequency/coax and tune the stub. After adjustment is completed, re-check the lower frequency antenna because there is some interaction.

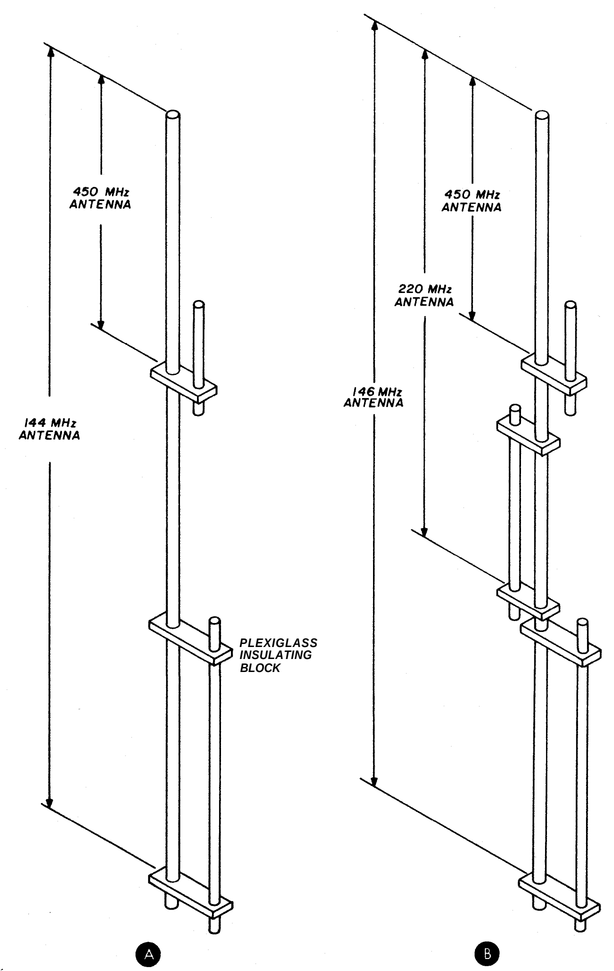


fig. 2. Multiband J antennas for 146 and 450 MHz, A, and for 146,223, and 450 MHz, B. Antennas built by the author were fed with separate coaxial feedlines and baluns.

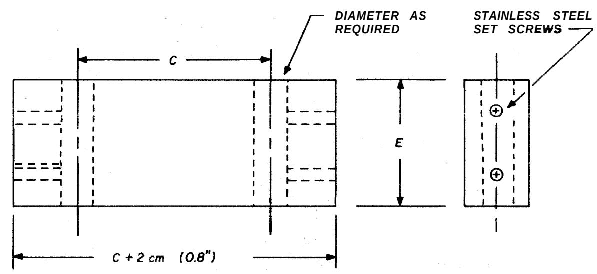


fig. 3. Layout of the aluminum shorting bars and plexiglass spacers. Dimensions C and E are given in table 1.

If the bandwidth is too narrow for your application, load the tuned section more by moving the feedpoint toward the antenna portion. The 2:1 swr bandwidth should be at least 10 MHz at 450 MHz, 4 MHz at 220 MHz, and 3 MHz at 144 MHz.

operation

The original design of the multiband J antenna was intended to allow operation on one band at a time, with the unused coax connectors left open. There is a reasonable amount of coupling between the sections so that if you are transmitting on 144 MHz, for example, some of this energy will be received by the 450-MHz antenna. Measurements have shown this coupled power to be 15-20 dB down, but that's still sufficient to damage a sensitive receiver frontend.

I don't recommend you try two-band duplex operation with this antenna system, although with the excellent preselectors in the Motorola ***Motrac,*** it has been done on 450 MHz and 146 MHz without damage.

ham radio

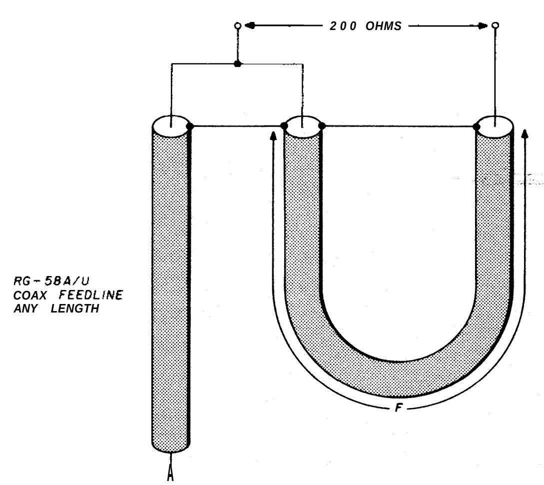


fig. 4. Construction of the half-wavelength coaxial balun. Length F is given in table 1.

Colpitts oscillator design technique

Concise, accurate design technique for Colpitts oscillators eliminates the need for empirical design

Almost everyone interested in radio communications has at one time been confronted with the need for a high-frequency oscillator. Unfortunately, little information is available to help those individuals design their own oscillator circuits. This article describes a technique for the design of a Colpitts oscillator which is applicable to both crystal and LC oscillators. The technique to be described is simple, and yet has been proven to be more effective than empirical solutions. Basic oscillator design at the lower frequencies is covered in many textbooks and, therefore, will not be described in detail.

Although this article will deal exclusively with the Siliconix U310, the design equations are also applicable to bipolar transistors. For reasons to be mentioned later, bipolar transistors are not as desirable in some applications. Subjects such as oscillator noise and stability are covered elsewhere and will not be described in detail here. More specifically, what will be dealt with is what it takes to get the oscillator to ***start*** oscillating. The technique offered is simple enough so that you need not know how oscillators oscillate. Since scientific calculators are now selling for less than fifty dollars, it is justifiable to do away with some of the old "rule-of-thumb" solutions.

basic assumptions

The Colpitts oscillator is perhaps most commonly seen in the configuration shown in **fig. 1.** To simplify the analysis, those components necessary for dc operation have been omitted **(fig. 2).** To further simplify the ac analysis, the oscillator is redrawn in the configuration shown in **fig.** 3. This is the general form of a common-gate amplifier, with a feedback capacitor, C1, between the input and the output. The amplifier is considered with no signal applied from an external source; the input is shunted only by C2 and the source-bias resistor. Note that if the capacitors were replaced with inductors and vice versa, the circuit would be the common Hartley oscillator. An analysis is somewhat easier with the circuit in **fig.** 3 since it now appears as a resonant tank circuit with an amplifier connected to it. However, one more circuit element needs to be added: the load resistance, R_L. This is the element which will accept power from

By Larry Leighton, WB6BPI, Siliconix Incorporated, 2201 Laurelwood Road, Santa Clara, California 95054

the oscillator. For this circuit to oscillate, all that's necessary is enough energy be tapped from the tank circuit, amplified, and routed back to the tank circuit to compensate for the energy absorbed by R_L.

The first step is to determine what output power is required. This, of course, depends upon the application, but in most cases it will be relatively low, particularly when frequency stability is of prime importance. Crystal oscillators generally have relatively low output levels, mainly to prevent the crystal from fracturing. In one case, a crystal ordered from a prominent manufacturer had a rated power dissipation of one milliwatt. To obtain more power from an oscillator with a fixed-supply voltage, a lower load impedance is required. If a lower load impedance is applied, however, either the gain of the transistor has to be increased, or feedback has to be increased to maintain oscillation. Since the crystal is in series with the feedback signal, care must be taken when considering how much output power you can expect from the oscillator.

Another consideration regarding output power is that inductors and capacitors do consume some power. They always have some associated series resistance which can be minimized by using higher-quality components. When rf current passes through these components, heat is generated from the power dissipated across the resistance. This heat causes a change in the values of the inductor or capacitor and, hence, a change in frequency. The effects of these changes and the changes associated with various components are covered in reference 1.

The transistor also has power-limiting characteristics. From turn-on of the oscillator, until a steady-state condition has been reached, the transistor parameters will change. The amount of change must generally be determined empirically, but can be minimized by operating the transistor at relatively low power, limiting output from the oscillator.

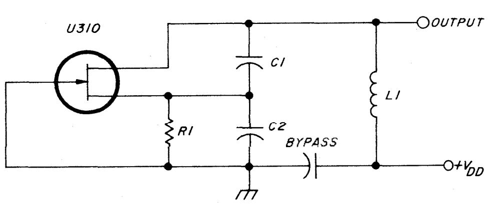


fig. 1. Schematic diagram of a basic common-gate Colpitts oscillator.

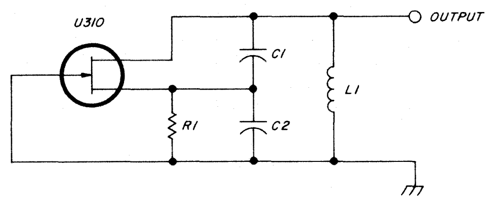


fig. 2. Ac model of the common-gate Colpitts oscillator shown in fig. 1. For analysis, the dc components have been eliminated.

In this article, the Siliconix U310 jfet is characterized at 9 volts drain-to-gate voltage and 2 mA of drain current in the common-gate configuration. This is thought to be a fair compromise between output power and parameter changes. Consider a class-A oscillator in which the drain current does not appreciably change from the oscillating to not-oscillating condition. If optimized, it will be less than 50 per cent efficient, with a maximum output power of 9 milliwatts (9 V•2 mA/2). Oscillator design almost always requires some compromises, so there is nothing binding regarding the 9 volt, 2 mA operating point; it can be changed to meet the needs of your application.

oscillator design

The first step to consider in oscillator design is the required output power; the second is to determine the load resistance necessary to obtain the required output. The load the oscillator requires is almost always different than the amplifier, buffer, or mixer it must drive. Initially, only the oscillator load will be considered.

Only class-A oscillators will be discussed in this article because the transistor parameter changes are more easily defined. If the supply voltage and drain current are known, the solution for the load resistance, R_L, is

$$R_L = \frac{(V_{DS} - V_{DS(sat)})^2}{2\,P_{out}}$$

$V_{DS(sat)}$ can be obtained from the data sheet; in this case it is 2 volts. Earlier it was determined that 9 milliwatts would be a desirable output. Given the operating point of 9 volts:

$$R_L = \frac{(9-2)^2}{2(.009)} = 2722 \text{ ohms}$$

Since $V_{DS(sat)}$ is only an approximation, rounding R_L to 2700 ohms for use in the following calculations is justified.

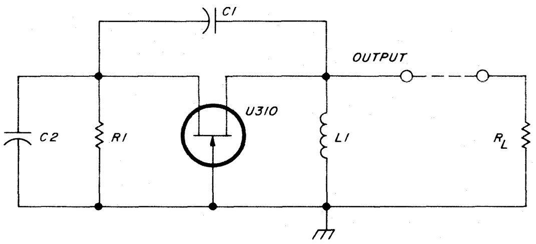


fig. 3. To further simplify analysis of the oscillator, this diagram shows the circuit redrawn into the general form of a comrnon-gate amplifier.

The third item to consider is transistor selection. The Siliconix U310 has been demonstrated to operate as an oscillator up to 900 MHz.[2] It has an advantage over more commonly available vhf and uhf jfets because of its high zero-bias drain current and g_m (forward transconductance). This means that the U310 has potentially higher output power and more stable characteristics when operated at a lower-drain current. Another distinct advantage of the U310 is that the gate lead, and in this case, the intrinsic bulk of the transistor, is connected to the metal case of the transistor package. Therefore, the U310 can be used with a heatsink, minimizing the change in transistor parameters. Although not recommended by the factory, an alternative heatsink method is to solder the U310's case to the chassis. This is convenient only when used in the common-gate configuration (be careful when soldering, as the U310 can be destroyed by overheating).

Perhaps the most critical parameter in ***amplifier*** design is the gain stability factor of the transistor itself.[3] The Siliconix U310, when operated in the common-gate configuration at 20 volts and 10 mA, is unconditionally stable at almost all frequencies. In fact, when loaded with practical, somewhat lossy external components, the U310 could be considered unconditionally stable at all frequencies. Unconditional stability, by definition, means the transistor will not oscillate when presented with any positive real source or load impedance; stability, therefore, is a measure of the transistor's ability to oscillate.

It is desirable that the stability factor of the transistor is such that it will not oscillate. This may seem contradictory to the design goal, but is used to emphasize that the source of feedback necessary to sustain oscillation should be the option of the ***designer*** and not the transistor. In the common-gate configuration, the U310 has been optimized to the point where the intrinsic feedback elements of the transistor are so small that the values of feedback necessary to make the U310 oscillate ***are*** the choice of the designer.

Most rf bipolar transistors, in the equivalent common-base configuration, are only conditionally stable and tend to be so at many frequencies. Special precautions must be taken to insure oscillation at only the desired frequencies. This requires more components, which increases cost and decreases reliability. Also, very few bipolars, ***designed*** for the common-base configuration, are available.

The next item to consider is the operating or loaded Q (&) of the resonant tank circuit. The value chosen for Q_L is very much dependent upon the application. In addition, there are upper and lower restrictions on &. In an attempt to make (& low, the transistor's reactive components would assume a higher percentage of the total circuit values. This is undesirable in many applications since stability would be sacrificed as the transistor parameters change with temperature. Additionally, the harmonic content of the output increases as Q_L decreases. There are many applications where Q_L should be very high. One example would be when a low-noise oscillator is required to drive a sensitive mixer or product detector.

The upper limit for Q_L is established by the unloaded Q (Q_u) of the inductor and how critical the tuning adjustment can be; this is particularly true for crystal oscillators.

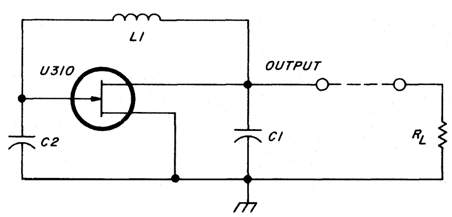


fig. 4. A cornmon-source oscillator shown as an ac model.

At very high frequencies, the Q_u of the capacitor must also be considered, especially when using capacitors generally designed for use at lower frequencies. While Q_L is used as a starting point for calculations, the final value of Q_L will be different than the initial assigned value. In general, the small change will not affect circuit performance.

The Q_u of a reactive element is determined by the intrinsic resistive element. In circuits where higher-quality capacitors are used, this resistance generally can be disregarded. Inductors, however, with only a few exceptions, must always be considered as having some associated resistance. The unloaded Q is defined as:

$$Q_u = \frac{X_L}{R_S}$$

where R_S is the series resistance of the inductor, and X_L is the inductive reactance at the operating frequency f_o. Q_u can be measured, but in those instances where it can't, assuming a Q_u between 150 to 250 is safe, if standard inductor winding techniques are used.

In those cases where a Q_L of 10 or greater is chosen, it is beneficial to include 4, in the calculations necessary to determine total R_L, since the circuit performance will be altered. When time permits, it is advisable to include Q_u for all designs. The effects of Q_u can best be emphasized by the following:

$$Insertion\ loss = 20\ log_{10} \left[1 - \frac{Q_L}{Q_u}\ (dB) \right]$$

For a Q_L of 10 and a Q_u of 100, the insertion loss would equal 0.92 dB. This means that 0.92 dB of the total power is dissipated across the inductor rather than delivered to R_L. A side effect is that as 4, gets smaller, the change in frequency from initial turn-on increases because the power dissipated across the inductor causes it to change reactance. This is especially true in those cases where a ferrite core is used for tuning.[1] Many of the commercially-available inductors, particularly the smaller molded variety, are not designed for use in frequency-determining applications where stability is of concern. The Q_u of many of the smaller molded inductors is as low as 50, which means with a $Q_L = 10$, the insertion loss of the inductor is 1.94 dB. The point to keep in mind is that wherever frequency stability is the prime consideration, Q_u is *very* important, even though its effects may have to be determined empirically.

y parameters

This design technique will use what is commonly referred to as admittance, or y parameters. The oscillator designs presented here require the use of algebra and trigonometry with the y parameters. Review material can be found in any of the standard references.[4,5]

One item which should be understood concerning y parameters is their relationship to ohms and reactance. Resistance is a measure of the opposition to current flow. Its counterpart in terms of admittance is conductance, which is a measure of the acceptability to current flow. They are simply related; one thousand ohms is equal to 111000 mho (the mho is designated as the *siemens).*

Reactance is a measure of the opposition to changing current flow and its counterpart in terms of admittance is susceptance; these are also similarly related: 1000 ohms reactance is equal to 1/1000 siemens of susceptance. Just as inductors and capacitors have different signs attached to them in terms of reactance, they also have different signs in terms of susceptance.

Inductors in terms of susceptance have negative signs placed in front of them and capacitors have plus signs in front of them. Therefore, a capacitive reactance of 1000 ohms is equal to +1/1000 siemens, and an inductive reactance of 1000 ohms is equal to −1/1000 siemens. Another simple relationship is that 1000 ohms equals 1 millisiemens or mS, ($1\ 000\ ohms = 1/1000\ siemens = 0.001\ siemens = 1\ millisiemens$). This article will use the term millisiemens or mS often so it is important to be familiar with the relationships. The relationships previously described regarding reactance and susceptance are correct and usable as relates to this article.

Referring to **figs.** 3 and **4,** the oscillator designs will determine the component values necessary to ***start*** the initial oscillations. For oscillations to start, the ac resistance measured at the junction of C1, L1, R_L, and the drain of Q1 must be infinite.[5] But, in terms of conductance, it would be zero. Therefore, the conditions necessary to start oscillations are:

$$y_t = y_{22} - \frac{y_{21} y_{12}}{y_{11}} = zero$$

where

y_t = terminal conductance (junction of C1, L1, R_L, and the drain of Q1)
y_{22} = output admittance of the circuit
y_{21} = forward transconductance of the circuit
y_{12} = reverse transconductance of the circuit
y_{11} = input admittance of the circuit

These parameters also include the y parameter data of the transistors, which in most cases are available from their data sheets. The parameters might be listed as shown in **table 1.**

Listed below are the equations which relate common source, gate, and drain parameters to each other. These equations also relate to bipolar transistors, except the source becomes the emitter, the gate becomes the base, and the drain becomes the collector. To convert common-gate parameters to common-source parameters:

$$y_{11s} = (g_{ig} + g_{fg} + g_{rg} + g_{og}) \pm j(b_{ig} + b_{fg} + b_{rg} + b_{og})$$

$$y_{21s} = -[(g_{fg} + g_{og}) \pm j(b_{fg} + b_{og})]$$

$$y_{12s} = -[(g_{rg} + g_{og}) \pm j(b_{rg} + b_{og})]$$

$$y_{22s} = g_{og} \pm b_{og}$$

where the non-subscript g represents conductance and the non-subscript b represents susceptance. To convert from common source to common gate, exchange s and g subscript values. That is

$$y_{21g} = -[(g_{fs} + g_{os}) \pm j(b_{fs} + b_{os})]$$

From this point on, unless stated otherwise, all numerical terms will be in millisiemens (mS); instead of using the term 1000 ohms or 111000 siemens, 1 mS will be used. Further, in many cases the mS will be ***implied*** and, therefore, just the number with its sign will be used. If no sign is given, it is assumed to be positive. As an example: $y_{11} = +7.66 \times 10^{-3}$ *siemens* $+j\ 1.59 \times 10^{-3}$ *siemens* will be shown as $y_{11} = 7.66 + j\ 1.59$, where the first term is the conductance and the second term the susceptance.

common source design

This design example will use the common-source oscillator shown in **fig. 4.** Earlier calculations for the common-gate oscillator determined the R_L to be 2700 ohms, which will also be used in the common-source example in addition to an f_o of 100 MHz. A & of 10 has been selected since it yields practical component values. Referring to **fig. 5,**

$$Q_L = \frac{R_L}{X_{L1}} \text{ or } X_{L1} = \frac{2700}{10}$$

therefore, X_{L1} equals 270 ohms or -3.7 mS. At 100 MHz, this is an inductance of 430 nH, or approximately 10 turns of no. 22 AWG (0.6 mm) enameled wire wound on a 6.5 mm (114-inch) drill bit. An actual inductor was wound and found to be 422 nH for an X_{L1} of 265 ohms. The number used for the calculations is therefore -3.77 mS. The Q_u of the inductor was measured to be 188, so the resistance in shunt with the inductor is equal **to 0.02 mS,** $[(X_L)(Q_u) = R_p$, $(265)(188) = 50k$ *ohms or* 0.02 *mS*$]$. The inductive susceptance (b_L) and conductance (g_L) can be added to the transistor's y parameters in the following manner:

$$y_{11} = [(g_{11s} + g_L)] \pm j[(b_{11s} + b_L)]$$

$$y_{21} = [(g_{21s} - g_L)] \pm j[(b_{21s} - b_L)]$$

$$y_{12} = [(g_{12s} - g_L)] \pm j[(b_{12s} - b_L)]$$

$$y_{22} = [(g_{22s} + g_L)] \pm j[(b_{22s} + b_L)]$$

The 100 MHz y parameters, from **table 1,** for the Siliconix U310 are:

$$y_{11s} = -0.919 + j\,3.34$$

$$y_{21s} = 8.5 - j1.8$$

$$y_{12s} = -0.0455 - j1.13$$

$$y_{22s} = 0.129 + j\,1.18$$

Adding the inductive susceptance ($b_L = -3.77$) and shunt conductance ($g_L = 0.02$) yields:

$$y_{11} = [(-0.919 + 0.02)] \pm j[(3.34) + (-3.77)]$$

$$y_{21} = [(8.5) - (0.02)] \pm j[(-1.8) - (-3.77)]$$

$$y_{12} = [(-0.0455) - (0.02)] \pm j[(-1.13) - (-3.77)]$$

$$y_{22} = [(0.129 + 0.02)] \pm j[(1.18) + (-3.77)]$$

or

$$y_{11} = -0.0899 - j\,0.435$$

$$y_{21} = 8.48 + j\,1.97$$

$$y_{12} = -0.0655 + j\,2.64$$

$$y_{22} = 0.149 - j2.59$$

Any slight discrepancies noted can be attributed to round-off errors. All of these calculations were done with an HP67 programmable calculator with automatic rounding.

table 1. y-parameter listing for the Siliconix U310 operated at 9 volts and 2 mA. The unit, in all cases, is mS.

common gate	common source
$y_{11} = y_{ig}$ input admittance	$y_{11} = y_{is}$ input admittance
$y_{21} = y_{fg}$ forward transconductance	$y_{21} = y_{fs}$ forward transconductance
$y_{12} = y_{rg}$ reverse transconductance	$y_{12} = y_{rs}$ reverse transconductance
$y_{22} = y_{og}$ output admittance	$y_{22} = y_{os}$ output admittance

common gate

freq. (MHz)	y_{11}	y_{21}	y_{12}	y_{22}
100	7.66 + j 1.59	−8.62 + j 0.615	−0.0831 − j 0.0512	0.129 + j 1.18
125	8.75 + j 1.92	−8.56 + j 0.824	−0.0914 − j 0.0714	0.111 + j 1.49
150	8.98 + j 2.34	−8.47 + j 0.816	−0.0797 − j 0.0861	0.129 + j 1.81
175	9.06 + j 2.80	−8.55 + j 0.646	−0.0669 − j 0.0972	0.126 + j 2.13
200	9.07 + j 3.24	−8.84 − j0.112	−0.0602 − j0.119	0.124 + j 2.42

common source

freq. (MHz)	y_{11}	y_{21}	y_{12}	y_{22}
100	−0.919 + j 3.34	8.50 − j 1.8	−0.0455 − j1.13	0.129 + j 1.18
125	0.209 + j 4.16	8.45 − j 2.31	−0.0197 − j 1.41	0.111 + j 1.49
150	0.563 + j 4.88	8.34 − j2.62	−0.0490 − j 1.72	0.129 + j 1.81
175	0.576 + j 5.48	8.42 − j2.77	−0.0589 − j 2.03	0.126 + j 2.13
200	0.294 + j 5.43	8.71 − j 2.31	−0.0633 − j 2.3	0.124 + j 2.42

The next step is to add the load conductance $(0.370\,\text{mS})$ to g_{22}, the ***real*** part of y_{22}.

$$\begin{aligned} y_{22} &= (0.149 + 0.37) - j\,2.59 \\ &= 0.519 - j\,2.59 \end{aligned}$$

Change the sign of the susceptance and record it as $b_{22} = 2.59$ mS. This is the starting value of susceptance which will be used to tune the circuit to resonance, whereupon b_{22} equals zero. The new set of y parameters from the previous operations are:

$$y_{11} = -0.0899 - j\,0.435$$

$$y_{21} = 8.48 + j\,1.97$$

$$y_{12} = -0.0655 + j\,2.64$$

$$y_{22} = 0.519 \pm j\,0$$

The next step is to find the value which when added to the input of the transistor will make the circuits output conductance zero. This can be found by solving:

$$\begin{aligned} y_{11t} &= \frac{y_{21}y_{12}}{g_{22}} - y_{11} \\ &= \frac{(g_{21}g_{12} - b_{21}b_{12}) + j(g_{21}b_{12} + g_{12}b_{21})}{g_{22}} - (g_{11} + j\,b_{11}) \\ &= \frac{-0.00576 + j\,0.0222}{0.519} - (-0.899 - j\,0.433) \\ &= \frac{0.023\ \angle\ 105^\circ}{0.519} - (-0.899 - j\,0.433)^* \\ &= 0.0442\ \angle\ 105^\circ - (-0.899 - j\,0.433) \\ &= (-11.1 + j\,42.8) - (-0.899 - j\,0.433) \\ &= -10.2 + j\,43.3\ mS \end{aligned}$$

Since the solution produces a real part which is negative, the addition to the input of the transistor cannot be performed with passive components. The real part is not always negative; in many cases it can be a positive resistance, but if added to the circuit it increases the cost and adds complexity. I'll demonstrate that the real component of y_{11t} can be neglected, with little error added to the calculations.

*The numerator in the expression $y_{21}y_{12}/g_{22}$ must be converted to polar form before it can be divided by g_{22}. When a polar-to-rectangular conversion key is not provided on the calculator, the following rules must be applied: If the real part of the numerator is greater than zero, the angle is equal to $\tan^{-1}(Im/Re)$. If the real part is less than zero or negative, the angle is $180^\circ + \tan^{-1}(Im/Re)$, where Im is the imaginary part of $y_{21}y_{12}$ and Re is the real part. To divide, retain the angle of $y_{12}y_{21}$ and divide by 0.519. The polar form of this expression must now be converted back to rectangular form to subtract y_{11}. The imaginary part is $(\sin 105^\circ)(0.0442) = 42.8\,mS$ and the real part $(\text{cosine } 105^\circ)(0.0442) = -11.1\,mS$.

In this design, the real part will be disregarded and only the imaginary part $(43.3\,\text{mS})$ will be used.

The b_{11t} value of 43.3 mS is the susceptance of the input shunt capacitance, $C2$. At 100 MHz, 43.3 mS

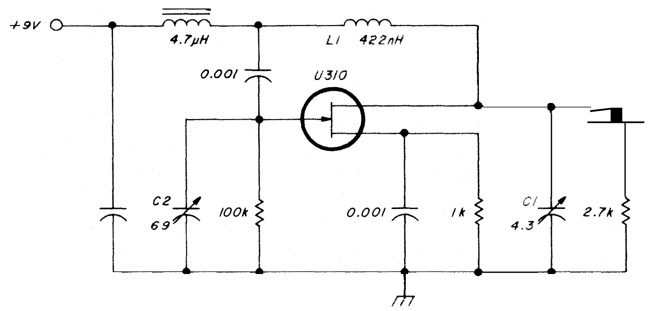


fig. 5. A common-source Colpitts oscillator designed using the procedure described in the text.

susceptance is equal to an X_{C2} of 23.11 ohms, which is 69 pF. Add 43.3 mS (b_{11t}) to b_{11} for a new total b_{11} of 42.8 mS. Subtract the load conductance, 0.370 mS, from g_{22} for a new g_{22} of 0.149 mS. And finally, solve the equation:

$$\begin{aligned} Y_{out} &= y_{22} - \frac{y_{21}y_{12}}{y_{11}} \\ &= 0.149 - \left[\frac{(g_{21}g_{12} - b_{21}b_{12}) + j(g_{21}b_{12} + g_{12}b_{21})}{-0.899 + 42.8}\right] \\ &= 0.149 - \left[\frac{-0.00576 + j\,0.0222}{-0.899 + 42.8}\right] \\ &= 0.149 - \left[\frac{0.023\ \angle\ 104.5}{42.81\ \angle\ 91.2}\right] \\ &= 0.149 - (5.357 \times 10^{-4} \quad 13.3) \\ &= 0.149 - (0.521 + j\,0.123) \\ y_{out} &= -0.373 - j\,0.123\ \text{millisiemens} \end{aligned}$$

To tune to resonance, change the sign of the imaginary part and add this value to the first recorded value of b_{22} $(2.59 + 0.123 = 2.71\ mS)$. This is the total susceptance of C1 at 100 MHz. The reactance of C1 is 369 ohms, or 4.31 pF at 100 MHz. Note that the real part of Y_{out} is equal to -0.373 mS or -2681 ohms. The desired value was -2700 ohms, an error of less than 1 per cent. This is the error introduced by not using the real part of y_{11t}.

After all the calculations were performed, the circuit in **fig. 5** was constructed and tested. The value of $C2$ required to start oscillations was found to be 45 pF instead of the calculated value of 69 pF. The feedback network consisting of L1 and the dc blocking capacitor were removed, and the circuit was measured again. The lead lengths, added value of the dc blocking capacitor, and the fact that the inductor was slightly distorted from its original shape, changed the total feedback circuit to be equivalent to

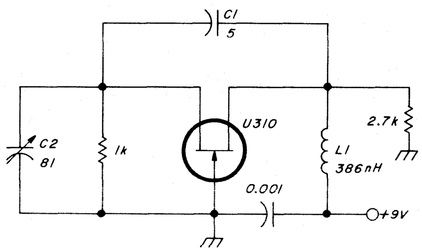


fig. 6. A common-gate oscillator designed with the generalized design technique.

497 nH or −3.2 mS. This changes the calculated value of C2 from 69 pF to 53 pF. Further, if the value of g_m (forward transconductance) were reduced by a factor of 10 per cent, a conceivable situation, the value of C2 would change to 48 pF, not far from the measured 45 pF.

The previous calculations assume that all measurements are absolutely accurate. Since this is not possible in practice, C1 and C2 should be variable to compensate for inaccuracies in measurements, as well as changes in transistor parameters.

The values calculated for the passive components in the circuit are those values required to start oscillation. As the circuit oscillates, the net parameters of the transistor change and consequently, the values of C1 and C2 will change. The equations presented in this article only provide a starting point, but are a preferred alternative to the empirical approach.

common gate design

The common-gate oscillator in **fig.** 3 is easily designed by starting with the values of C1 and L1 obtained from the common-source oscillator problem and adding them to the common-gate parameters. The common-gate y parameters for the U310 at 9 volts and 2 mA are:

$$y_{11g} = 7.66 + j\,1.59$$

$$y_{21g} = -8.62 + j\,.615$$

$$y_{12g} = -0.0831 - j\,0.0512$$

$$y_{22g} = 0.129 + j\,1.18$$

Add the source resistor (in this case 1000 ohms or 1 mS) to g_{11}; y_{11} then equals 8.66 + j 1.59. Using the value of capacitance for C1 obtained in the first design, select the closest standard value, fixed capacitor available. Since a 5-pF capacitor is generally more available, this value is used for C1 in **fig. 6.** Since both terminals of C1 are at some rf potential, a tuning tool will generally change the total value of the feedback capacitance by some unknown amount, therefore necessitating the use of a fixed value, instead of a variable capacitor. The 5-pF value is equal to 3.14 mS at 100 MHz, and can be added to the new y parameters in the following manner:

$$y_{11} = g_{11} \pm j\,(b_{11} + b_f)$$

$$y_{21} = g_{21} \pm j\,(b_{21} - b_f)$$

$$y_{12} = g_{12} \pm j\,(b_{12} - b_f)$$

$$y_{22} = g22 \pm j\,(b_{22} + b_f)$$

which yields

$$y_{11} = 8.66 + j\,4.73$$

$$y_{21} = -8.62 - j\,2.53$$

$$y_{12} = -0.0831 - j\,3.19$$

$$y_{22} = 0.129 + j\,4.32$$

To the new value of y_{22}, you should add the value of R_L (0.370 mS) to g_L and also X_{L1} (−3.77 mS) to b_{L1}, yielding y_{22} = 0.499 + j 0.551. The value of b_{22} should be recorded for future use. As in the common-source example, set b_{22} equal to zero and calculate y_{11t}. The calculated value of y_{11t} (−23.4 + j 50.9) serves as a starting value for C2. The next step consists of adding 50.9 mS to b_{11}, subtracting the g_L of 0.370 mS from g_{22}, and solving for y_{out}, which equals −0.338 − j 0.205 millisiemens. The final value for L1 can be determined by solving the following equation:

$$\begin{aligned} X_{L1} &= (-1)(b_{out}) + (-1)(b_{22}) + (b_{L1}) \\ &= (-1)(-0.205) + (-1)(0.551) + (-3.77) \\ &= -4.12\ mS \end{aligned}$$

which equals 243 ohms or 386 nH.

Note that g_{out} = −0.338 mS (−2.95k ohms), which is not the desired −2.7k ohms. The reason for the er-

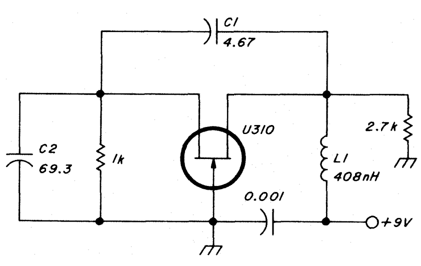


fig. 7. The same oscillator shown in fig. 6, except designed with the more accurate method.

ror is the same as before – the real part of y_{11t} was disregarded. This can easily be compensated for by decreasing the susceptance of C2. As an example, by reducing the susceptance of C2 by 10 per cent, g_{out} becomes – 2626 ohms which is sufficient to start oscillation. The circuit for the common-gate configuration, before adjustment of C2 for the necessary g_{out}, is shown in **fig. 6.** In this solution, the inductor should be variable, or the inductance decreased and a variable shunt capacitor added for adjustment.

For those cases when a more accurate result is necessary, the following procedure may be used. *Use those component values determined in the design for **fig. 5** as a starting point* and proceed with the following steps.

1. Start with the U310 common-gate parameters.

2. Add the source resistor to g_{11} (in this case, 1 mS)

3. Add the susceptance of C2 to b_{11}
$y_{11} = 8.66 + j\,44.9$

4. Add the susceptance of C1 to b_{11} and b_{22}.

5. Subtract the susceptance of C1 from b_{21} and b_{12}, which produces new y parameters of:

$$y_{11} = 8.66 + j\,47.6$$
$$y_{21} = -8.62 - j\,2.1$$
$$y_{12} = -0.0831 - j\,2.76$$
$$y_{22} = 0.129 + j\,3.89$$

6. Add the susceptance of L1 to b_{22}

7. Add the load conductance to g_{22}
$y_{22} = 0.499 + j\,0.124$

8. Record the value of $(-1)(b_{22})$ or -0.124 mS

9. Set b_{22} equal to zero

10. Solve the following equation for g_f and b_f

$$g_f \pm j\,b_f = \frac{A + jB}{C + jD}$$
$$= \frac{0.0094 - j\,0.255}{-0.451 - j\,42.7}$$
$$= 0.00365 + j\,0.220$$

where

$$A = (g_{22}g_{11} - g_{12}g_{21} + b_{21}b_{12} - b_{11}b_{22})$$
$$B = (g_{22}b_{11} + b_{22}g_{11} - g_{12}b_{21} - g_{21}b_{12})$$
$$C = (-1)(g_{12} + g_{21} + g_{11} + g_{22})$$
$$D = (-1)(b_{12} + b_{21} + b_{22} + b_{11})$$

11. Record the value of b_f

12. Add b_f (0.220 mS) to b_{11} and b_{22}

13. Subtract b_f from b_{21} and b_{12} to obtain

$$y_{11} = 8.66 + j\,47.8$$
$$y_{21} = -8.62 - j\,2.32$$
$$y_{12} = -0.0831 - j\,2.98$$
$$y_{22} = 0.499 + j\,0.220$$

14. Solve for y_{11t}, which yields $0.118 + j\,0.260$

15. Record b_{11t} (0.260 mS)

16. Add b_{11t} to b_{11}, $y_{11} = 8.66 + j\,48.1$

17. Subtract the load conductance 0.370 mS from g_{22}, $y_{22} = 0.129 + j\,0.220$

18. Solve for y_{out}, which equals $-0.371 + j\,0.00109$ mS

19. Record the value $(-1)(b_{out}) = -j\,0.00109$ mS

20. Final values for C1, C2, and L1 are determined as follows:

C1 Starting value of C1 from **fig. 5**	2.71	mS
value of b_f	0.22	mS
	2.93	mS

C1 final value = 2.93 mS or 4.67 pF

C2 Starting value of C2 from **fig. 5**	43.3	mS
value of b_{11t}	0.26	mS
	43.6	mS

C2 final value = 43.6 mS or 69.3 pF

L1 Starting value of L1 from **fig. 5**	– 3.77	mS
plus $(-1)(b_{22})$ from **step 8**	– 0.124	mS
plus $(-1)(b_{out})$ from **step 19**	– 0.00109	mS
	– 3.9	mS

L1 final value = – 3.9 mS or 409 nH

The final circuit diagram, with component values, is shown in **fig. 7.** Note that g_{out} from **step 18** is within 0.32 per cent of the desired g_{out} of – 0.37 mS or – 2.7k ohms. Again, the reason for the slight

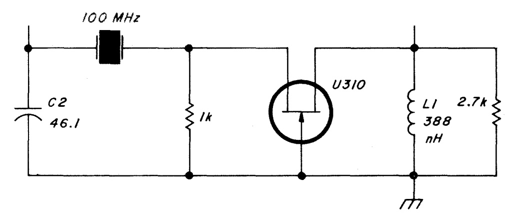


fig. 8. A 100-MHz crystal oscillator which was designed using the same methods as the Colpitts oscillators.

discrepancy is caused by disregarding the real parts of y_{11t} and g_f, and is of no consequence.

crystal oscillators

This oscillator design technique can be modified for use with crystal oscillators, particularly overtone crystals **(fig. 8)**. When operated in the series-resonant mode, the crystal has some series resistance, which must be added to the transistor's common-gate parameters, along with the 1000-ohm source resistor. Given the common-gate y parameters:

1. Add the source resistor to g_{11}.

2. Convert the y parameters to z parameters and add the series resistance of the crystal directly to the real part of z_{11}. A typical value of crystal series resistance is 80 ohms for a seventh overtone crystal; this data is available from the crystal manufacturer.

3. Convert the z parameters back to y parameters.

4. Convert the new set of common-gate parameters to common-source parameters.

5. Design the circuit using the technique described for **fig. 5.**

6. Using the common-gate parameters from **step** 3, and the component values determined from **step 5,** complete the design by using those steps outlined for **fig. 7.**

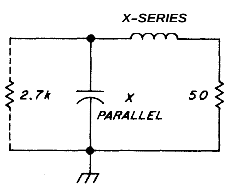


fig. 9. A lowpass L network for matching between the high source resistance of the oscillator and a 50-ohm load.

The necessary equations to convert y parameters to z parameters are:

$$z_{11} = \frac{y_{22}}{\Delta y} \qquad z_{21} = \frac{-y_{21}}{\Delta y}$$

$$z_{12} = \frac{-y_{12}}{\Delta y} \qquad z_{22} = \frac{y_{11}}{\Delta y}$$

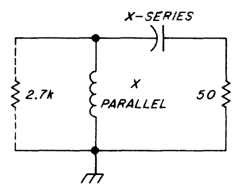
where

$$\Delta y = y_{11}y_{22} - y_{21}y_{12}$$

To convert z parameters back to y parameters, interchange the y and z values.

It is helpful to know what effects the three circuit components have on the oscillator circuit. In general, when the input shunt capacitor is increased in value, the frequency of the oscillation is decreased, while the negative output resistance increases. Increasing the feedback capacitance lowers the frequency and also the negative output resistance. Changing the output reactance, theoretically, only changes the frequency and does not effect the output conductance.

fig. 10. A highpass L network, in addition to impedance matching, also provides dc isolation for the load.



The tuning procedure for the oscillator is quite easy (refer to **figs.** 3 and **4**). Apply dc power to the oscillator. If it immediately oscillates, tune the output shunt element to the desired frequency. If the oscillator does not start, or ceases oscillation when tuning, decrease the capacitance of C2, which adjusts the output conductance. When it has been determined that the oscillator is tuned to frequency and oscillating, the input shunt element (C2) can be adjusted for the desired output level and the output shunt element tuned for the correct frequency.

The crystal oscillator is tuned in the same manner. The exception is when the parallel capacitance of the crystal is relatively high. In that case the procedure is to increase the capacitance of C2 until oscillation ceases and then decrease the capacitance in small increments until the circuit oscillates again. This procedure should be followed since in some cases, the parallel capacitance of the crystal provides enough feedback to allow the circuit to oscillate at frequencies very close to f_o, but not as a function of the series-resonant mode of the crystal. For breadboard designs, it is easiest to insert a resistor, equal to the series resistance of the crystal, in place of the crystal; the same tuneup procedure is used except that the resistor is replaced with the crystal during the last stages of tuning.

As an added advantage, when using the U310 in the crystal oscillator circuit, the inductor sometimes shunted across the crystal to prevent spurious oscillations is not necessary. This inductor is almost always necessary when using a bipolar transistor as a Colpitts overtone oscillator.

impedance matching

Throughout this article, a load impedance of 2700 ohms was used. The actual load will generally be some other value. Quite often this load will be 50 ohms resistive. In this case, the 2700-ohm load resistance necessary for oscillation will have to be matched to the 50-ohm load. The simplest solution is the L network shown in **figs. 9** and **10.** The value for the series reactive element can be found with the following equation:

$$X_{series} = R_s \sqrt{\frac{R_p}{R_s} - 1}$$

where

R_s = series resistance of load

R_p = parallel resistance to be transformed

In this example

$$X_{series} = 50 \sqrt{\frac{2700}{50} - 1} = 364 \text{ ohms}$$

The value for the shunt element can be determined from:

$$X_{parallel} = \frac{R_s R_p}{X_{series}} = \frac{(50)(2700)}{364} = 371 \text{ ohms}$$

Fig. 11 shows the highpass matching network **(fig. 10)** applied to the common-gate oscillator shown in **fig. 6.** The 364-ohm reactance at 100 MHz is equivalent to 4.37 pF. Any error introduced by using the nearest standard capacitance value (5 pF) can be compensated for by L1 and C2. The shunt inductance of the matching network must be added to L1 for a new total inductance for L1. This is easily accomplished by adding the susceptance of $X_{parallel}$ and L1 [-4.1 mS + (-2.7 mS) = -6.8 mS or 147 ohms inductive reactance which at 100 MHz equals 234 nH]. The circuit in **fig. 11** was constructed and the test results for 100 MHz are: Value of C2 to start oscillation = 43 pF; final values after tuning are C1 = 5 pF; C2 = 25 pF; and L1 = 215 nH. Maximum power output = 9.2 milliwatts. With C2 set to 12 pF, the circuit oscillated with crystals ranging from 95 MHz to 116 MHz. The value of L1 was changed to accommodate the different frequencies.

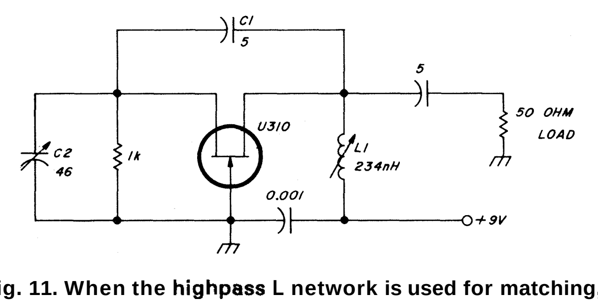


fig. 11. When the highpass L network is used for matching, the shunt inductor is combined with inductance in the oscillator. In this case, L1 is a slug-tuned 6.5 mm (1/4 inch) coil, wound with 6 turns of no. 18 AWG (1 mm) wire. The turns are spaced to occupy 12.5 mm (1/2 inch).

concluding comments

This article is not intended as a construction project. However, those circuits shown with inductance and capacitance values have been built and tested, and performed very close to predictions. This design technique has been used for many oscillator designs and has been found to be superior to the empirical approach, particularly if a programmable scientific calculator is available.

The techniques presented here appear to be equally valid at the lower frequencies. The data necessary are g_m, c_{rss}, c_{oss}, and c_{iss}. This information is almost always obtainable from the transistor data sheet. These parameters can be substituted in the common-source y parameters by assuming the input and output resistance of the fet is very high and can, therefore, be disregarded.

$$y_{11s} = 0 + j\left(\frac{1}{X_{c_{iss}}} + \frac{1}{X_{c_{rss}}}\right)$$

$$y_{21s} = g_m - j\frac{1}{X_{c_{rss}}}$$

$$y_{12s} = 0 - j\frac{1}{X_{c_{rss}}}$$

$$y_{22s} = 0 + j\frac{1}{X_{c_{oss}}} + \frac{1}{X_{c_{rss}}}$$

Low-frequency oscillator design is available from many sources. This technique might not be as usable as others, but it does allow a close approximation for low-frequency design.

Expressing an idea is often difficult for me. I am gratified by the many personal, and professional friends who have helped expand this idea and also provided the additional enthusiasm and necessary technical expertise, especially Ed Oxner, manager of Special Projects Engineering at Siliconix, Will Alexander, WA6RDZ, Earl McCune, WA6SUH, and Bonnie (The Boss).

references

1. Doug DeMaw, W1FB, "Vfo Design Techniques for Improved Stability," ham radio, June, 1976, page 10.
2. Ed Oxner, "High Performance Fets in Low-Noise VHF Oscillators," *Siliconix Design Idea, DI 73-1*, Siliconix Incorporated, Santa Clara, California.
3. Roy Hejhall, "RF Small Signal Design Using Admittance Parameters." Motorola *Application* Note AN-215, Motorola Semiconductor Products, Phoenix, Arizona.
4. H. Paul Shuch, WA6UAM, "Solid-State Microwave Amplifier Design," ham radio, October, 1976, page 40.
5. Noden Westlake, Applied Mathematics for Electronics, Prentice-Hall, Englewood, New Jersey.
6. Mohammed Shuaib Ghasi, Principles and Design of Linear Active Circuits, McGraw-Hill, New York.

ham radio

visual aids

for working on microcircuits

Devices are getting smaller and smaller — consider these visual aids before you give up on a construction project with today's ICs

With the introduction of transistors and miniaturized circuits most everyone dealing with them has probably experienced difficulty seeing components and circuits because of their small size. Today, with integrated-circuit devices requiring even smaller printed-circuit-board design, the visual or seeing requirement is even greater.

the problem

When you consider the visual anomalies found in the general population such as nearsightedness, farsightedness, astigmatism, or combinations of these, and when you include problems of binocularity (the ability or inability of the eyes teaming together, such as one eye receiving a larger image than the other), poor eye-muscle coordination, cataracts, and any ongoing or past history of ocular diseases, it's no wonder many of our IC projects end up with solder bridges, improper or unsoldered connections, and components mounted improperly. The fact is that many simply can't see well enough to avoid these pitfalls.

I recently completed a project that used 29 ICs, three of which were LSI devices, all mounted on four PC boards, three of which had circuit paths on both sides. Even though I have 20-20 vision at distance and near, and have never had any pathological problem causing visual impairment, I experienced great difficulty seeing my work. I therefore came to the conslusion that if I were having these kinds of problems with what's generally considered normal vision, there must be many electronic hobbyists who are having even greater problems. With this in mind, I offer the following observations to help in understanding these visual tasks and to allow most to see with better efficiency.

lighting and magnification

First, realize that vision requires light. But it has to be *useful* light; that is, not too dim (below the visual threshold) or too bright (above the threshold and therefore saturating the visual system, especially the retina). Second, visual acuity (how well you see) depends on the image size impinged onto the retina. Visual acuity is directly related to object size. The conclusion is that, if you have the proper lighting

Reprinted with permission from the May-June, 1977, issue of RTTY *Journal.*

By Dr. Robert Sullivan, Optometrist, K9SRL,
410 NE 7th Street, Linton, Indiana 47441

conditions and magnification, visual acuity generally improves, However, there are limiting factors. For example, as the image size on the retina gets larger, the field of view gets smaller.

optical aids

If you're nearsighted (can't see well at a distance without glasses or contact lenses), you might do better visually, at the near distances used in electronic construction, without your glasses or contacts. Since a nearsighted eye without correction in place is in effect too strong, removing the spectacles has the same effect as looking through a magnifying lens. You'll notice that near objects (within about 20 cm [8 inches] of your eyes depending upon your prescription) look larger. However, this might not work if you have astigmatism, as your vision could be distorted.

If you're farsighted (can't see nearby objects well without glasses or contacts), you should wear your correction at all times for near electronic work. Since a farsighted eye is a weak eye without correction, wearing glasses or contact lenses, in effect, makes your eyes stronger. Also, far sightedness involves a problem with the eye's focusing mechanism and without correction, eye fatigue and headaches are more common.

If you normally wear glasses full time, and if they're the bifocal or trifocal type, you should wear them for near electronic work.

work glasses

You might consider having a special pair of glasses made especially for electronic work. I made a pair with one lens having a +16 diopter power and the other lens opaque, which forces me to use one eye only, since such a large prescription for both eyes creates a condition that makes binocularity impossible. This is a common problem with some of the available optical aids recommended for near visual tasks. These devices are usually binocular in nature (both eyes are used). To maintain this binocularity, weaker lenses are used and the resulting magnification is less.

I put the +16 lens on my right eye since I'm right-eye dominant. With these glasses, approximately 4X magnification is achieved and the field of view is about 7 cm (3 inches). The focal point is 6.25 cm (2.5 inches), so I must hold things close. I found that with these glasses and a handheld penlight (I use the popular disposable type), I could examine PC boards for errors in component mounting, find solder bridges, and perform general inspection with great ease. Just one thing more: these suggestions should be implemented *after* you've checked with your eye doctor.

ham radio

solving RFI problems
in home-entertainment devices

How to dispose of RFI problems quickly and easily — ideas from an overseas amateur that will work in your station

Radio-frequency interference (RFI) from amateur transmitters to television, broadcast, and hi-fi sets is still a problem. Most such cases of RFI can be attributed to *fundamental overload* of the home-entertainment device from the transmitter, especially if the transmitter is running high power (500 watts or more).

The remedies are well known for this type of RFI, but I'd like to summarize some of the cures:

1. Install a filter in the ac line to the device.

2. Install a 0.01-μF capacitor across the speaker terminals.

3. Install an rf choke in the speaker leads. Such a choke can be made by winding the speaker leads onto a ferrite rod.

4. Make sure the home-entertainment device has a good rf ground.

5. Use shielded cable for the speaker leads, and ground the shield.

6. Install a good-quality 0.01-μF capacitor between the device chassis and ground.

7. Ground the device's antenna coaxial-cable shield.

8. Wind 10-15 turns of the device's antenna coax cable onto a ferrite rod. This will form an effective rf choke.

9. Install a good highpass filter directly at the device's antenna terminals. Make sure the filter is shielded, and ground the shield.

By John DeVoldere, ON4UN, 215 Poelstraat, B-9220 Merelbeke, OV Belgium

In all cases of reported RFI, I've eliminated the problem by one or more of the methods mentioned above without having to make any changes inside the home-entertainment device.

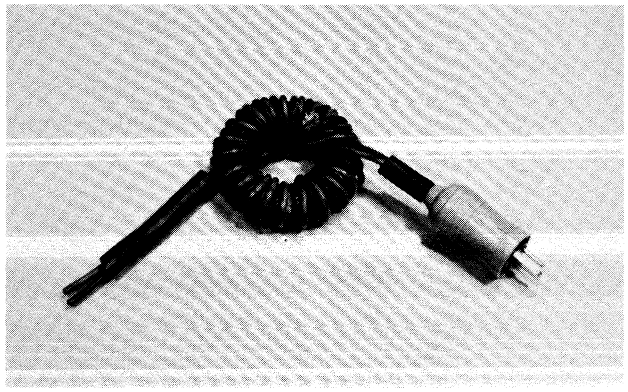
There's nothing original here so far, but I affer some other hints that you may find useful in your own RFI problem:

I always keep a selection of filters available for any RFI complaint. The set of filters includes a good-quality ac filter for the power line, a highpass filter for twin lead or coax-cable, ferrite-rod filters for speakers, an rf choke for coax cable, and some wires and clip leads for grounding purposes in tests

responding to RFI complaints

Almost all interference complaints I receive are by telephone. Here's what I do. I ask the complainant to leave the phone off the hook and tell him I'll be right over to check the problem. Then I switch on my phone patch, turn down the receiver gain control so the receiver won't trip the VOX in my transmitter, and adjust my transmitter for VOX operation through the telephone. Then I grab my assortment of filters, hop into my car, and ring the complainant's door bell within minutes after his telephone call.

Now here's where you need some diplomacy. Don't be aggressive, but explain that you're genuinely interested in resolving the RFI problem as a mutual endeavor. Tell the complainant you're there to investigate the problem and want to use his telephone to put your transmitter on the air. He may raise an eyebrow at this remark, but pick up his phone and put your transmitter on the air. Talking into the complainant's phone will trip your VOX, which makes it possible for ***you*** to make the diagnosis of the RFI problem, if any. No help from the outside, and you can do it right away.



Ferrite or powdered-iron toroid can also be used to form in-line rf chokes with loudspeaker leads.

TVI suggestions

In cases of TVI complaints, proceed as follows:

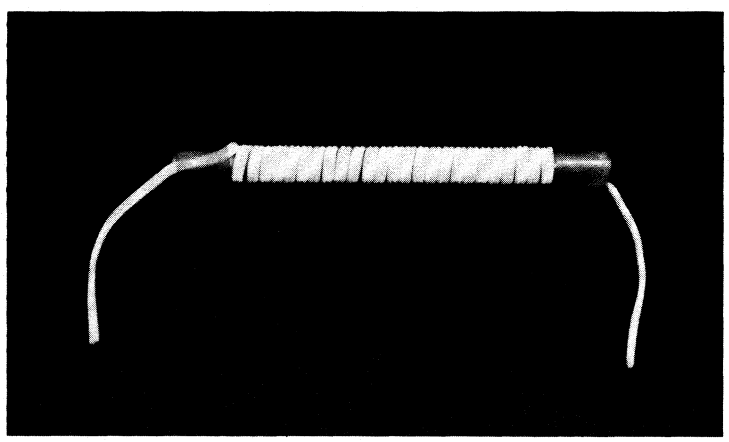
1. Disconnect the device's antenna feedline. If TVI continues, install a filter in the device's ac line. If this doesn't help, you're probably in big trouble, because you have a case of direct pickup by the device.

2. If disconnecting the antenna feedline stops the TVI, try the following:

A. Install a highpass filter

B. Install a coaxial-cable rf choke (10-15 turns on a ferrite rod).

C. Ground the antenna coax-cable shield at the TV-set chassis (through a 0.01-μF capacitor if necessary) to a good rf ground.



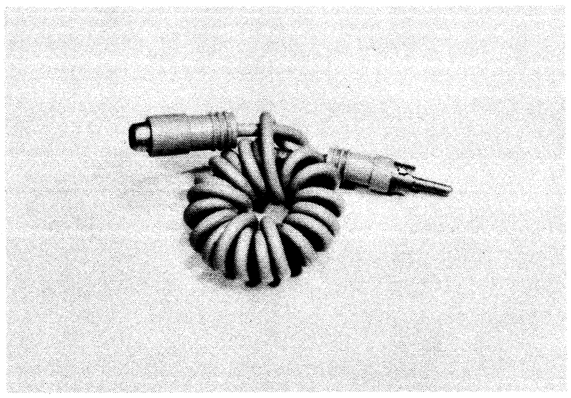
Loudspeaker lead can be wound into an rf choke by winding the lead around a ferrite antenna rod.

suggestions for hi-fi RFI cures

If you have to deal with a complete hi-fi system including tuner, amplifier, turntable, and recorder-deck, there's only one logical approach that can be used.

Disconnect all units from each other and from the ac line – except for the amplifier (just leave the ac cord and speaker leads connected). If interference still persists, disconnect the speakers and try the unit on headphones. If the interference disappears, the problem probably lies in pickup through the speaker cables (the RFI is being fed to the preamplifier stages through the audio feedback circuit).

Install the ferrite-rod filters in the speaker leads (10-15 turns). If necessary, connect 0.01-μF capacitors across the speaker terminals to ground. In very stubborn cases you may have to go all the way and use shielded speaker cable and ground the cable shield to a good rf ground.



In some cases interference can be reduced by winding external audio input leads on toroidal cores.

If disconnecting the speaker cables doesn't kill the RFI, the pickup must be coming through the ac line, so you must install a good filter between the set ac input and the house wiring. Again, a good rf ground may have to be connected to the ac filter.

If the hi-fi amplifier plus the speakers (by themselves) don't show any interference, connect all other pieces of equipment, one-by-one, to determine where the RFI is appearing. If connecting the tuner brings up the RFI, try highpass (or coax rf-choke) filters on the antenna lead; then try an ac line filter. A similar approach can be used when connecting other pieces of equipment, such as tape decks.

summary

Using your phone patch to solve the RFI complaint ***quickly*** and ***independently;*** using a logical approach when checking the complainant's set; having a ready assortment of anti-RFI filters available at all times – all these will go a long way toward maintaining a good relationship with your neighbors. This approach will also ensure keeping your amateur station on the air at all times.

ham radio

the ham notebook

600 kHz offset for frequency synthesizers

The circuit shown in **fig. 1** is designed for use with any frequency synthesizer which uses a programmable divider with outputs available from each flip-flop. The schematic shows its implementation in the GLB 400B synthesizer; only two ICs are needed to accomplish the function. The output of the divider at the end of a count sequence pulses the phase comparator U1, and reloads the counters U7, U8, and U9 to the number determined by the frequency set switches. (All of this circuitry is not shown in the partial schematic of the GLB 400B.) When enabled, the offset circuit gates off this pulse until the 100 kHz counter U8 reaches a count of 6 which corresponds to 600 kHz. At this time, a pulse is gated through to U1 and the counter load circuits. Other numbers could be decoded from U7, U8, and U9 by using a similar gating arrangement for any desired offset, for application in synthesizers for 220 MHz and 450 MHz.

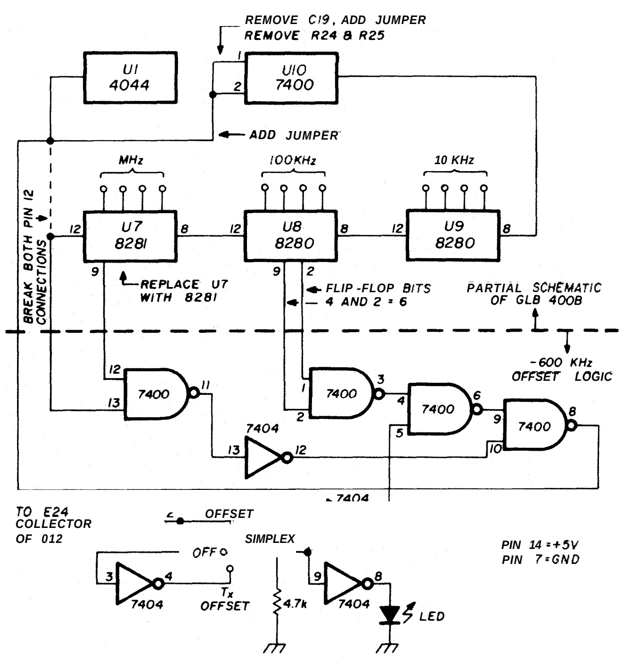


fig. 1. Simple circuit for generating 600 kHz offset with a frequency synthesizer that uses a programmable divider; only two ICs are required. The circuit here shows how the circuit is used with the GLB 400B synthesizer.

In addition to its simplicity, one advantage of this circuit is that it will only offset the synthesizer −600 kHz, and accidental operation above 148 MHz is not possible. When operating, always select the higher frequency of a repeater pair on the frequency set switches. In the 146-147 MHz segment offset Tx to transmit on the repeaters input; in the 147-148 MHz segment select offset Rx to receive the repeater output. To operate reverse simply flip the switch the other way. With the center-off position of the switch, transmit and receive will be on the selected frequency. The offset function is disabled in this position. The LED indicator will only light while offset is actually taking place, so it will go on and off between receive and transmit, always indicating the operating condition.

Only two ICs are required for the actual offset circuit, and U7 must be changed to an 8281, which is simply a plug-in substitution. Also, remove the Ik, 1.5k resistors, and the 150 pF capacitor from pins 1 and 2 of U10 (R24, R25, C19). They are no longer required because the added circuitry always presents the proper TTL signal level to this gate. Be sure to put a jumper in place of C19.

Dave Sargent, K6KLO

illumination for lever action switch

Having problems reading thumbwheel or lever action switches in the dark? The new Heath HW2036 is a perfect example of a fine synthesized receiver at a price anyone can afford. One of its shortcomings is the lack of illumination on the lever action switches. Material used to make the light bar in **fig. 2** came from the junk box of my model railroad. However, the brass and lamps can be obtained at any hobby shop for less than $2.00.

approximately 2 mm (1/16 inch) or so. Bend brackets, drill holes for mounting screws and for the lamp wires in sleeving to pass through the right hand bracket; file smooth and paint.

The brackets should hold the tube high enough above the top of the switch to clear the upper most position of the lever switch. One screw is sufficient to mount, and the wires in sleeving are passed through the hole in the panel and connected to ground and the meter lamp wire. Use grain or wheat lamps and carefully insulate the wires at the base with five-minute

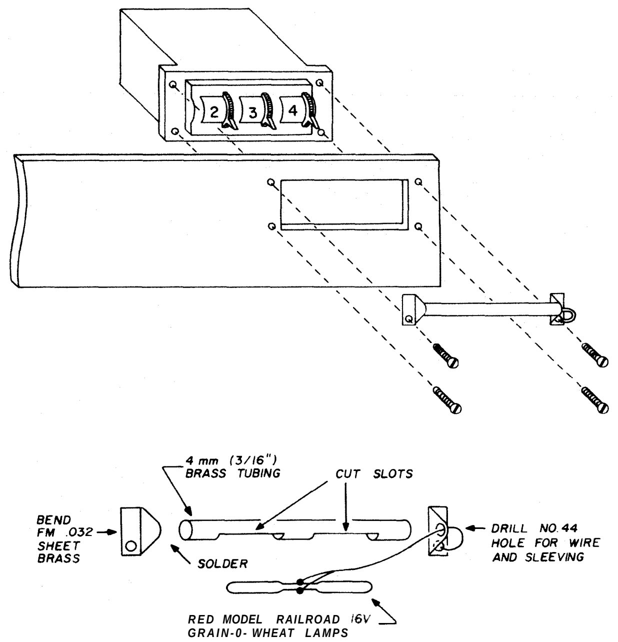


fig. 2. Adding illumination so you can see the HW2036 thumbwheel switches in the dark. Brass stock is available in most hobby shops.

Use a 3116-inch (4mm) diameter brass tube cut to the width of the switch assembly so that the brackets, when soldered to the tube, will fit snugly against the side of the switch bezel and mount under the existing screw. Cut two slots for the lamps, large enough for the lamps to pass through after soldering and painting. Rough trim the brackets, then solder so that the slots aim down and the tube is spaced away from the panel the thickness of the protruding bezel, epoxy. I used red lamps to cut the reflected glare from the switch.

Another modification for the HW2036 which improves the operation is to change the back panel and replace the RCA phone jack with an SO239 connector. Enlarge the hole in the printed-circuit board to take the stub of the SO239 and replace the remote speaker RCA phone connector with a miniature phone jack.

Fred W. Snow, W2IFR

re-forming the oxide layer in electrolytic capacitors

Electrolytic capacitors, including computer grades, which have not been used for any length of time should not be subjected to full voltage without first re-forming the internal oxide layer. If this is not done, they may have high leakage which will result in rapid failure due to internal heating.

The oxide layer may be re-formed in the following manner: Connect the capacitor to a power supply set to the dc voltage rating of the capacitor with a series resistor to limit the short circuit current to around 10 mA. For example, for a 200-volt capacitor, the power supply would be set to 200 volts and the series resistor would be 20,000 ohms. If several capacitors are being re-formed simultaneously, they should not be directly paralleled. Instead, each capacitor should have its own current limiting resistor. This allows each capacitor to charge independently, at a rate dependent on its internal leakage. It also allows the voltage on each capacitor to be measured separately as an indication of its condition. It has been my experience when re-forming large numbers of surplus computer grade electrolytics in this manner that most of them will charge to close to full voltage in just a few minutes. A few will stabilize at a considerably lower voltage, indicating that they have a high leakage and really need the re-forming procedure.

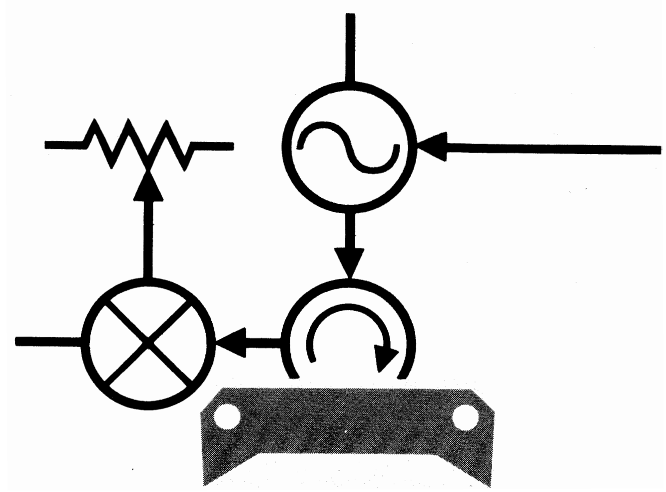
I have not yet enountered a capacitor that would not charge to the same level as the others if left connected long enough, although this has taken as much as several days. When this finally happens, it means that the oxide layer is totally re-formed, and the leakage has dropped to a normal level.

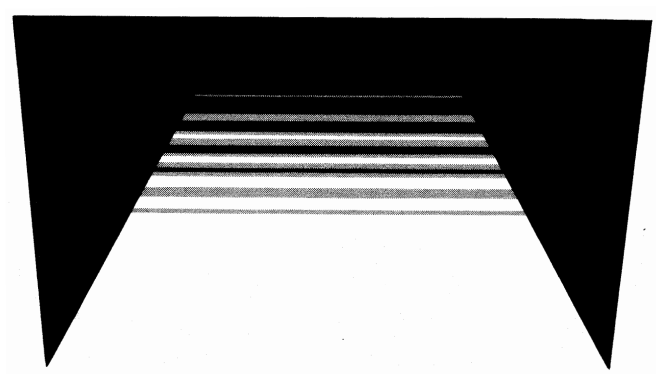
John Becker, K9MM

$1.50

radio

AUGUST 1978





for amateur microwave communications

ham radio magazine

AUGUST 1978

volume 11, number 8

T. H. Tenney, Jr., W1NLB
publisher

James R. Fisk, W1HR
editor-in-chief

editorial staff

Martin Hanft, WB1CHQ
administrative editor

Charles J. Carroll, K1XX
Patricia A. Hawes, WA1WPM
Alfred Wilson, W6NIF
assistant editors

Thomas F. McMullen, Jr., W1SL
Joseph J. Schroeder, W9JUV
associate editors

Wayne T. Pierce, K3SUK
cover

publishing staff

C. Edward Buffington, WB1AMU
assistant publisher

Fred D. Moller, Jr., WA1USO
advertising manager

James H. Gray, W1XU
assistant advertising manager

Therese R. Bourgault
circulation manager

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

United States: one year, $12.00
two years, $22.00; three years, $30.00

Canada: one year, $14.00
two years, $26.00; three years, $36.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

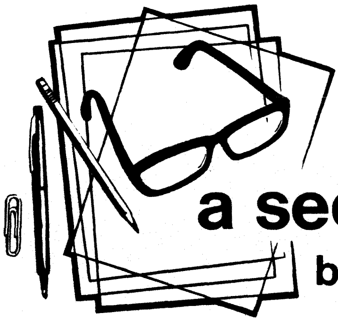
Foreign subscription agents are
listed on page 101

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

In recent months there has been rising concern about the possible harmful effects to living tissue due to heating by radio-frequency energy at 10 MHz and above. The weekly CBS TV news magazine, 60 Minutes, devoted a segment to this topic several months ago, numerous "rf radiation" stories have been published in newspapers and magazines, and now there is a best-selling book on the subject: *The Zapping* of America, by Paul Brodeur. Although much of Brodeur's book is devoted to what he calls the "deadly risk of microwave radiation" and its "cover-up" by the government, he apparently doesn't know the difference between high-power radar or TV transmitters and high-frequency amateur and CB equipment. He would have you believe that little or no research has been done on the dangers of electromagnetic radiation; if your neighbors believe him, you may find your radio activities squelched by local citizens who are afraid of being "zapped" by your amateur transmitting equipment.

Contrary to what Brodeur says, microwave engineers have been aware of rf radiation hazards for 30 years or more, and the scientific community has spent thousands of man hours investigating its effects and establishing safety standards. It is known, for example, that the internal body organs are susceptible to damage from heating caused by high-power radio energy in the range from 150 to 1200 MHz, and that the eye is especially prone to damage from radiation above 1000 MHz. More importantly, it is known that power levels which cause damage are much higher than those found in the average ham shack. Kilowatt transmitters on the amateur uhf bands (432 MHz and above) are potentially hazardous, but if they are completely shielded they are not dangerous to your health. On the lower frequencies there is practically no danger, even if you're running 2000 watts PEP.

Based on present knowledge, which is extensive, various government agencies have established rf radiation safety standards with recommended exposure limits referred to as Radiation Protection Guide Numbers (RPGN). The accepted RPGN value is 10 milliwatts per square centimeter of body area, the standard set by the Occupational Safety and Health Administration (OSHA). Although there are some scientists who disagree with this standard, most agree that rf power levels one-half the OSHA standard (5 mW/cm^2) have little effect on the human body, and practically no one objects to a standard of 1 mW/cm^2. Note that this is based on continuous exposure.

If your transmitter is well shielded, and you use coaxial transmission line, the only possible danger is radiation from your antenna. Assuming a kilowatt linear with 65% efficiency and no feedline loss places about 650 watts at the antenna; what is the minimum safe distance? This depends on the directivity of your antenna, but for a half-wavelength dipole it equates to a distance of about 3 meters (10 feet) for a power density of 5 mW/cm^2. If you're running less than a kilowatt, of course, the safe distance is less. Since most amateur dipoles are installed at least 8 meters (25 feet) above the ground, they obviously pose no radiation threat.

What about multi-element Yagi beams and stacked arrays? Since most of the power is concentrated in front of the beam, there is little danger above or below the antenna. Even with 650 watts input, the beam must have at least 15 dBd gain before the power density reaches 5 mW/cm^2 in the center of the forward lobe, 10 meters (30 feet) in front of the antenna, Few amateur antennas have this much gain, and those that do are used on uhf where it's impossible to generate 650 watts into the antenna and stay within the legal power limit.

On the high-frequency bands, if your beam is on a tower at least 10 meters (30 feet) high and not pointed into a building less than 10 meters away, there is absolutely no hazard at legal amateur power levels. Keep this in mind if you start getting grief from your neighbors.

Jim Fisk, W1HR
editor-in-chief

comments

microstripline impedance

Dear HR:

The formula W1HR deduced for microstrip impedance in the December, 1977, issue is interesting because it can be rewritten in the following way:

$$Z = \frac{376.7}{\sqrt{E_r}} \cdot \frac{h}{w+h} \text{ ohms} \qquad (1)$$

where Z = stripline impedance (ohms)
h = height of stripline
w = width of stripline (in same units as h)
E_r = relative permittivity of dielectric

The number 376.7 ohms (per square) is the intrinsic impedance of free space which by coincidence is nearly equal to 120π.

If there were no fringing of the electric field at the edges of the stripline the characteristic impedance of the line would be given ***exactly*** by

$$Z = \frac{376.7}{\sqrt{E_r}} \cdot \frac{h}{w} \text{ ohms} \qquad (2)$$

In **eq. 1** the $w+h$ in the denominator takes account of the fringing effect by considering that the stripline is ***effectively*** wider than its actual width w by the amount of the height h. As the ratio of the width to height becomes larger, the effect of the fringing becomes less significant and for a very wide stripline its characteristic impedance would approach that of **eq.** 2.

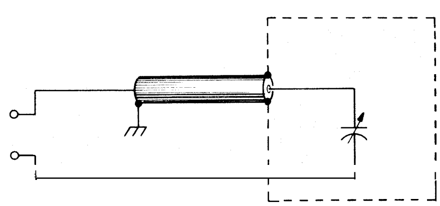
The above discussion is derived directly from a consideration of the field cell concept and of field maps for the transmission lines. It also follows that a line of ***any shape*** can be either calculated from a map or measured very simply with an ohmmeter and resistance paper as described on page 492 of ***Electromagnetic***~by J. D. Kraus and K. R. Carver (McGraw-Hill, New York, 1973).

John Kraus, W8JK
Director, The Ohio State
University Radio Observatory

bandspreading techniques

Dear HR:

I read with interest Mr. Leonard Anderson's excellent article on bandspreading techniques in February, 1977, ***ham radio.*** I would like to propose an alternate to his standard capacitor. By using a 3-wire ***guarded*** circuit, as shown in **fig. 1,** the cable



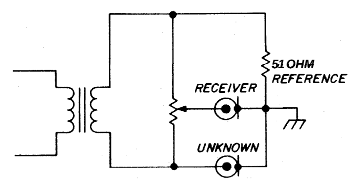
length will not cause the standard to read in error. This is due to the shield of the coax acting as a shield between the two leads from the capacitor, This method is used quite frequently by GenRad and other companies when measuring very accurate capacitance values. The main disadvantage of the guard circuit is that the capacitor must be isolated from ground.

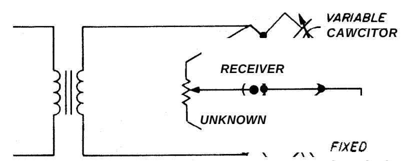
Robert Heider, WØEJO
Glendale, Missouri

antenna noise bridges

Dear HR:

I found the recent article on RX noise bridge measurements very interesting. As the developer of the original antenna noise bridge I would like to point out that two basic models were developed. The TE701 used a similar output circuit to the one shown in the article and worked well to over 100 MHz. The Model TE702 used a variation and worked to over 250 MHz. The bridge circuit was as follows:



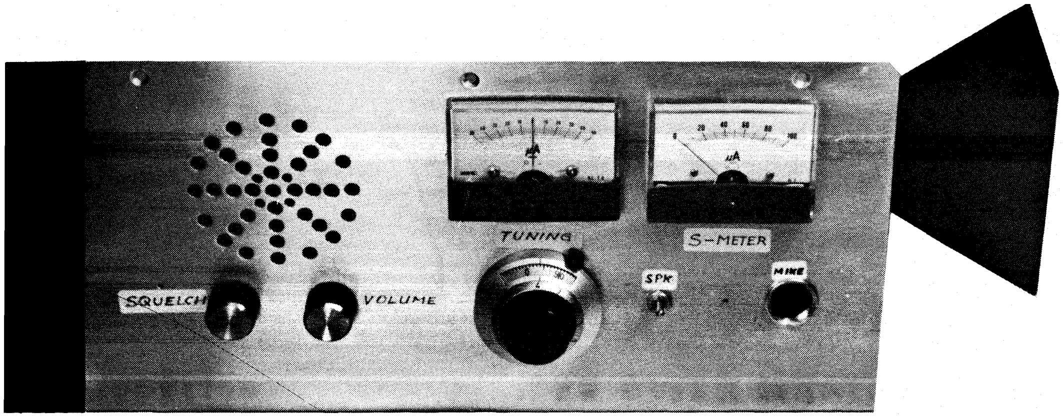


Note that the transformer does not need to be accurately center tapped and that it can be bifilar wound. Also, with a 100-ohm variable pot the calibration range is zero to infinity. To make a reactance bridge, place a fixed capacitor across the unknown terminal and a variable capacitor across the reference resistor. With less effort a lot more accuracy is available with this network over a wider frequency range.

Ted Hart, W5QJR*
Richardson. Texas

*W5QJR is the inventor of the Antenna Noise Bridge, and holds the patent on this very useful device. Readers who are interested can obtain copies of the patent (number 3,531,717, dated September 29, 1970) for 50 cents from the Commissioner of Patents, Washington, DC 20231. **Editor**

(Continued on page 82.)



10-GHz transceiver

for amateur microwave communications

Construction of a complete 10-GHz Gunnplexer transceiver with 30-MHz i-f and automatic frequency control

A little over a year ago Microwave Associates introduced a new component for amateurs which greatly simplifies the construction of a 10-GHz transceiver for operators who are interested in microwave communications but don't have experience with microwave construction techniques. Without special knowledge or an extensive test setup amateurs can now use a Microwave Associates MA-87127 Gunnplexer to operate on the 3 cm (10 GHz) amateur band. No special mechanical work is required. The Gunnplexer is a complete transceiver which consists of a varactor-tuned Gunn diode rf source, a ferrite circulator which decouples the transmit and receive functions, and a Schottky mixer diode for the receiver signal.[1] A diagram of the basic Gunnplexer system is shown in **fig. 1;** a block diagram of the complete transceiver is shown in **fig. 2.**

The Gunn diode oscillator requires a regulated 10 Vdc source which is capable of supplying 200 mA. The rf output power is approximately 20 mW;* a 17 dB gain horn antenna is available from Microwave Associates. The frequency of the Gunn diode can be tuned with the built-in varactor diode over a frequency range of 60 MHz minimum (100 MHz typical). The required varactor bias is +1 volt to +20 volts and should be controlled by a good quality multi-turn potentiometer.

The Gunnplexer can be easily frequency modulated with a small modulating voltage (mV range) which is superimposed on the varactor's dc bias supply. Since a very small modulating voltage is required, the

*Three models are available: the 15-mW MA-87127-1, the 25-mW MA-87127-2, and the 40-mW MA-87127-3. Units are stocked by Glen Whitehouse, Newbury Drive, Amherst, New Hampshire 03031, and in Europe by Microwave Associates, Munich.

This article was translated from German by Konrad Benz, Microwave Associates, Inc., Burlington, Massachusetts 01803

By Klaus H. Hirschelmann, DJ7OO, Reger Strasse 4,6500 Mainz 31, West Germany

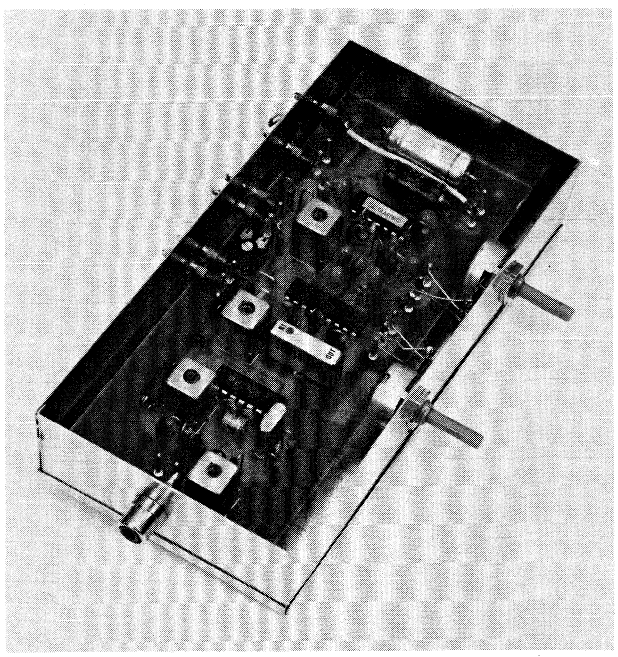
amplification factor of a single-transistor microphone amplifier is sufficient.

i-f amplifier

To complete the 10-GHz transceiver, an i-f amplifier is required. Because the antenna and Gunnplexer and its antennas are normally physically separated from the operating position (for roof or tower mounting), an i-f amplifier with a low noise figure should be connected directly to the Gunnplexer's mixer diode. A noise figure of 1.5 dB or less and a good impedance match ($Z = 200$ *ohms* at 30 MHz) is requried to obtain an overall system noise figure of 12 dB or better. With careful design, a system noise figure of less than 10 dB can be achieved.

The coaxial connection between the i-f preamplifier and the post amplifier/receiver at the operating position is not critical; a proven design is presented later in this article. When considering the noise figure of a Gunnplexer system it's important to remember that the receiver has no preselection so the two receiver sidebands (carrier plus *and* minus the i-f) contribute equally to the overall noise figure.

Standardization of a single i-f system is essential for the operation of a 10-GHz system among a large group of amateur microwave enthusiasts. A 100-



Construction of the 30-MHz receiver designed by DJ7OO. At the bottom left is the mosfet input stage, followed by the 40.7 MHz local oscillator and mixer, TDA1047 fm i-f strip, and TAA611 audio power amplifier. The two potentiometers are for squelch and audio gain.

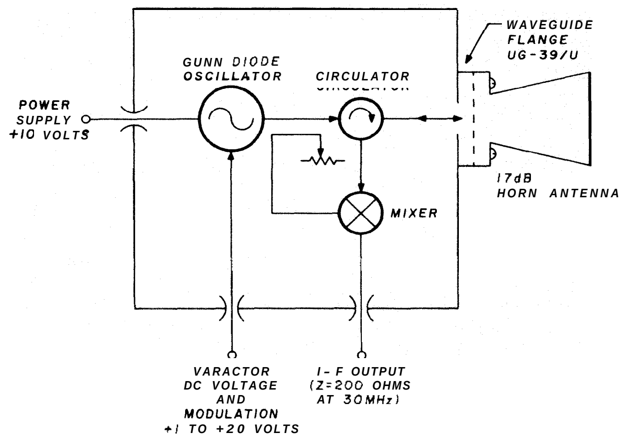


fig. 1. Basic Gunnplexer system showing the varactor-tuned Gunn-diode oscillator, ferrite circulator, and Schottky mixer diode. A portion of the rf power from the oscillator is coupled to the mixer through the circulator. The i-f output impedance at 30 MHz is 200 ohms; a 4:1 transformer is required to provide a good match to 50 ohms (see fig. 2).

MHz i-f has been recommended by several German amateurs,[2] but this is useful only if communications between two fixed stations is all that you want. The result is a full duplex system without transmit-receive switching where the Gunn oscillator operates simultaneously as a receiver local oscillator and frequency-modulated transmitter. Each partner operates at a different frequency, which results in the intermediate frequency as shown in **fig.** 3.

In most cases, however, amateurs want to contact as many other 10-GHz stations as possible. This requires that each station must be able to transmit and receive on either frequency. Since the varactor diode provides a maximum frequency tuning range of only 60 MHz, the use of a 100-MHz i-f would require mechanical tuning of the Gunn oscillator. Mechanical tuning of the Gunnplexer provides a tuning range of ± 100 MHz minimum, but this would unduly complicate a two-way communications set-up. By choosing a 30-MHz i-f, however, you can switch frequencies with a simple voltage change on the varactor diode.

In the Rhein-Main area in West Germany various Gunnplexers are operated at 10350 MHz (transmit) with +4 volts of varactor bias; with +10 volts on the varactor the transmit frequency is 30 MHz higher at 10380 MHz. If an operator knows whether the other station is using the lower (10350 MHz) or higher (10380 MHz) frequency, it is only necessary to tune the receiver over a small range of frequencies.

The instability of the self-oscillating Gunn diode requires wideband frequency modulation; a transmit bandwidth of 75 kHz and an i-f bandwidth of 200 kHz gives satisfactory results.

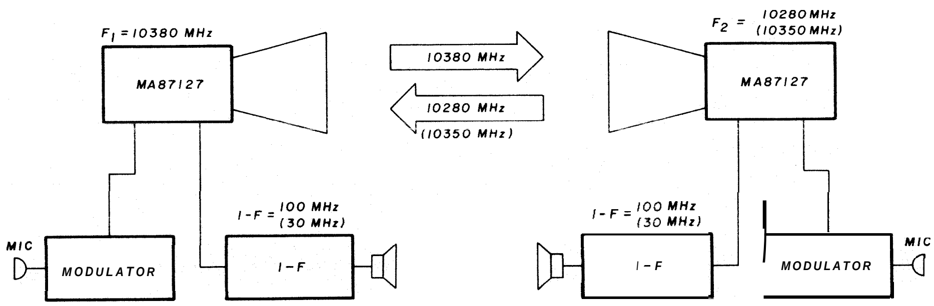


fig. 3. Duplex operation of the **10-GHz** Gunnplexer system, showing the oscillator frequencies for **100-**MHz and 30-MHz intermediate frequencies. As discussed in the text, a 30-MHz i-f is preferred because of the 60-MHz tuning range provided by the varactor: the use of a 100-MHz i-f would require mechanical tuning of the Gunnplexer.

i-f post-amplifier

The 30-MHz i-f post-amplifier and receiver shown in fig. 4 was developed by the Zweite Deutsches Fernsehen amateur group. More than fifty of these receivers have been built and used on the air, and all operate well."

The first 30-MHz amplifier stage uses a dual-gate BF900 MOSFET transistor (similar to the RCA 40673). The self-oscillating mixer is based on a Siemens SO42P IC and translates the 30-MHz input signal down to the 10.7-MHz i-f. The parallel tuned circuit (L1-C1) resonates at 40.7 MHz, the frequency of the third-overtone crystal. Without inductor L1 in the circuit the oscillator has a tendency to run at the crystal's fundamental at approximately 13.56 MHz; this can result in unwanted modulation products ($13.56 + 10.7 = 24.26$ *MHz*).

The Murata SFW10.7MA ceramic filter determines the i-f response characteristics of the receiver; the 3 dB bandwidth is 220 ± 40 kHz. The Siemens TDA1047 IC, which was developed for fm broadcast radios, is used as an amplifier and fm demodulator; it has excellent limiter capabilities and includes a built-in squelch circuit — its symmetry guarantees trouble-free operation.

An S-meter is connected to pin 14 of the TDA1047

"Kits to build your own 30-MHz post-amplifier are available from Elektronik Laden, Wilhelm-Mellies-Strasse 88, D4930 Detmold 18, West Germany; the price is 89 DM ($45) postpaid.

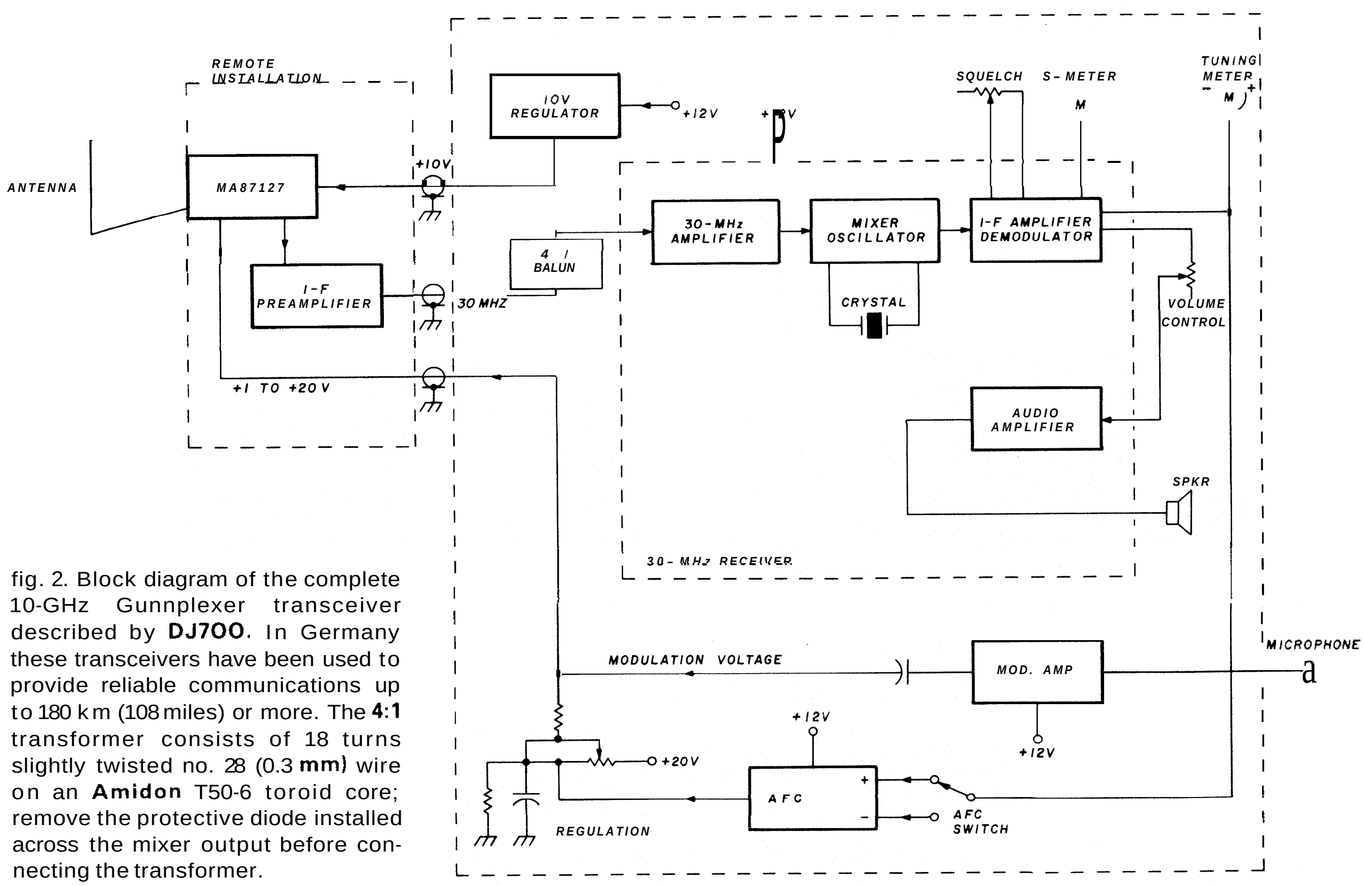
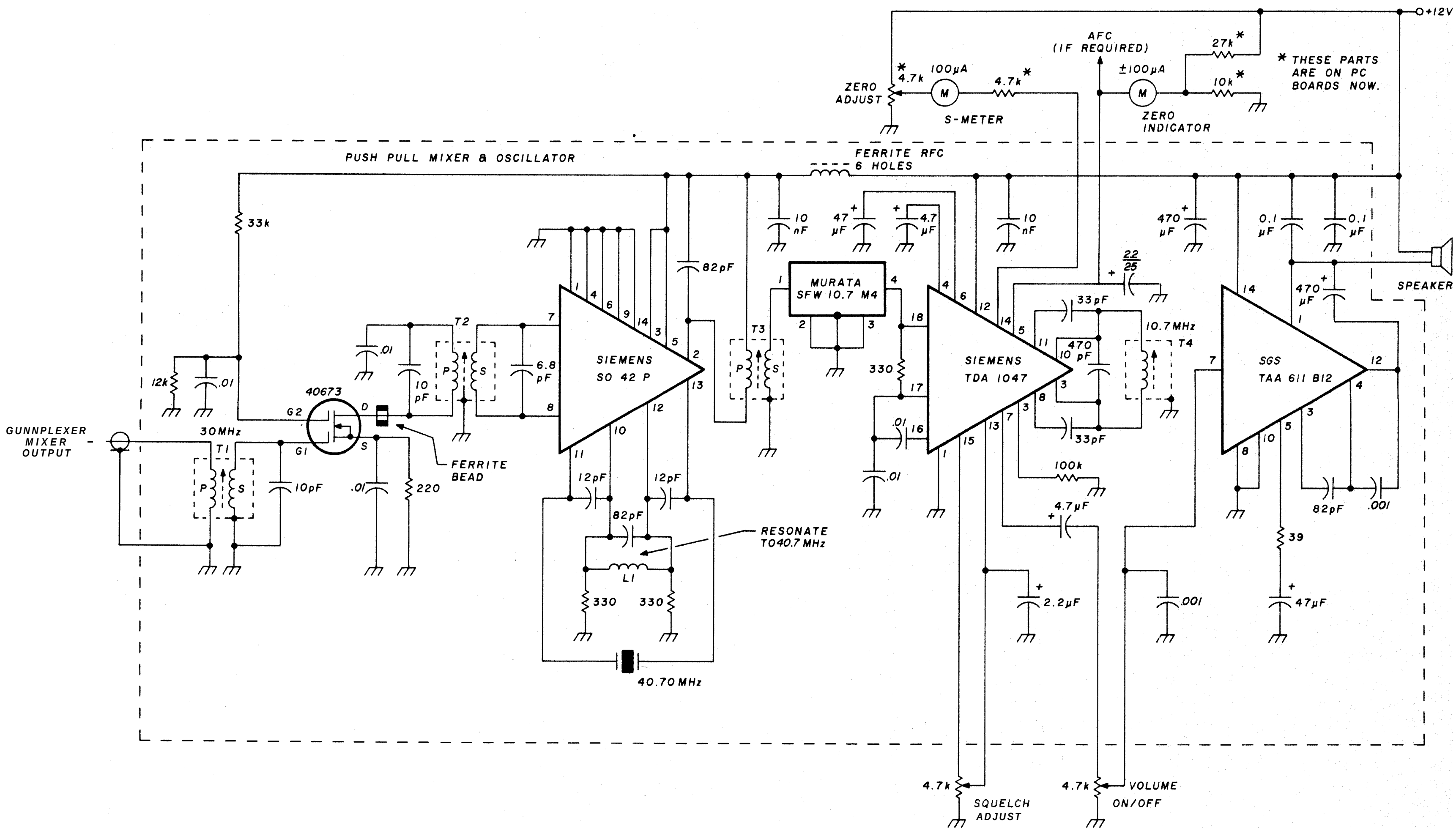


fig. 2. Block diagram of the complete 10-GHz Gunnplexer transceiver described by **DJ7OO**. In Germany these transceivers have been used to provide reliable communications up to 180 km (108 miles) or more. The **4:1** transformer consists of 18 turns slightly twisted no. 28 (0.3 **mm**) wire on an **Amidon** T50-6 toroid core; remove the protective diode installed across the mixer output before connecting the transformer.



L1 15 turns no. 28 AWG (0.3mm) unsupported, closewound on 4 mm (5/32 inch) diameter mandrel

T1 Primary, 3 turns; secondary, 18 turns

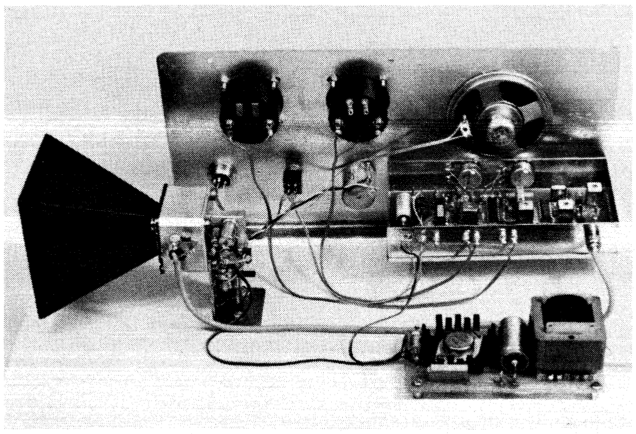
T2 Primary, 18 turns; secondary, 5 turns

T3 Primary, 18 turns; secondary, 5 turns

T4 8 turns

All transformers wound with no. 32 AWG (0.2mm) wire on Vogt D41-2520 forms.

fig. 4. Schematic diagram of a broadband 30-MHz i-f post-amplifierlreceiver which features a MOSFET input stage, SO42P self-oscillating mixer. 10.7-MHz ceramic filter. TDA1047 arnplifierldemodulator, and TAA611 audio power amplifier. The complete receiver is built intoa package measuring 14.7 cm long, 7.4 cm deep, and 2.9 cm high (5.8x2.9 x 1.1 inches). A kit is available.



Layout of DJ3KM's 10-GHz Gunnplexer system, as set up for display at a German club meeting. The 30-MHz receiver is mounted on the front panel, under the speaker; the avc circuitry is built on a small board mounted next to the Gunnplexer. An ac power supply for the system is in the right foreground (photo by DB3PR).

amplifier/demodulator. This is a big help when aligning antennas for maximum received signal. The inherent noise of the TDA1047 produces a small current through the S-meter which can be nulled out by adjustment of the 4700-ohm ZERO ADJUST potentiometer. The output at pin 5 of the TDA1047 is a frequency-dependent dc voltage which can be connected to a carrier meter and/or an **AFC** circuit for the Gunnplexer (fig. **5**). The Fairchild SGS TAA611B12 (or Texas Instruments 76001) serves as an audio power amplifier.

The frequency stability of the Gunnplexer is important for successful two-way communication; the manufacturer specifies a drift of −350 kHz per °C maximum. When the Gunnplexer is first turned on, the oscillator will drift a few MHz as the Gunn diode warms up, so the 220-kHz i-f bandwidth requires continuous tuning of the oscillator. The Gunnpiexer also continues to drift slightly after the initial warm-up period. **A** simple solution to this problem is to compensate for the drift of the free-running oscillator by changing the operating frequency of the station at the other end of the link.

The **AFC** circuit shown in fig. 5 uses the frequency-dependent voltage available from the i-f post-amplifier, as discussed previously. During two-way communications only one operator has his **AFC** circuit switched on; the Gunnplexer at the other end of the link is allowed to run free. **A** three-position switch is used because the frequency change might be up or down (center position is AFC OFF). The coupling between the **AFC** circuit and the Gunnplexer determines the system's holding range.

performance

The successful operation of various 10-GHz amateur stations in the Rhein-Main area, operating with the equipment described here, has proved the system's feasibility and reliability. The use of 17-dB horn antennas at both ends of the link allows communications up to 60 km (35 miles) or more. The 3-dB beamwidth of the horn antenna is approximately 30 degrees, so antenna alignment is not particularly critical.

Some stations are using home-built 23 dB horn antennas or 2 meter (6 foot) parabolic reflectors, so there have been many 10-GHz contacts in the range

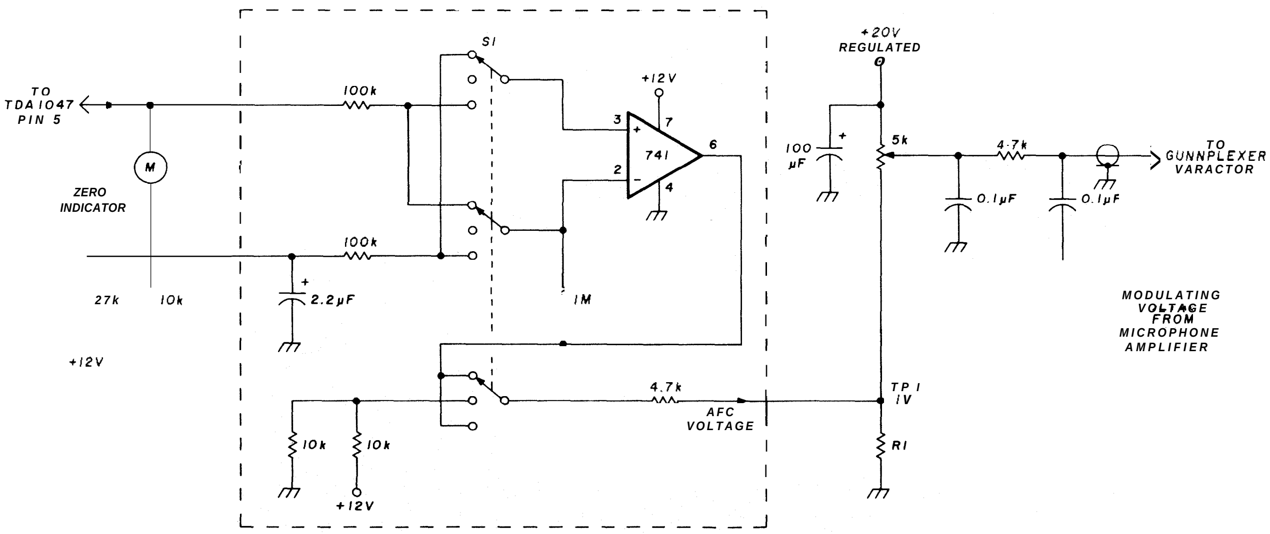
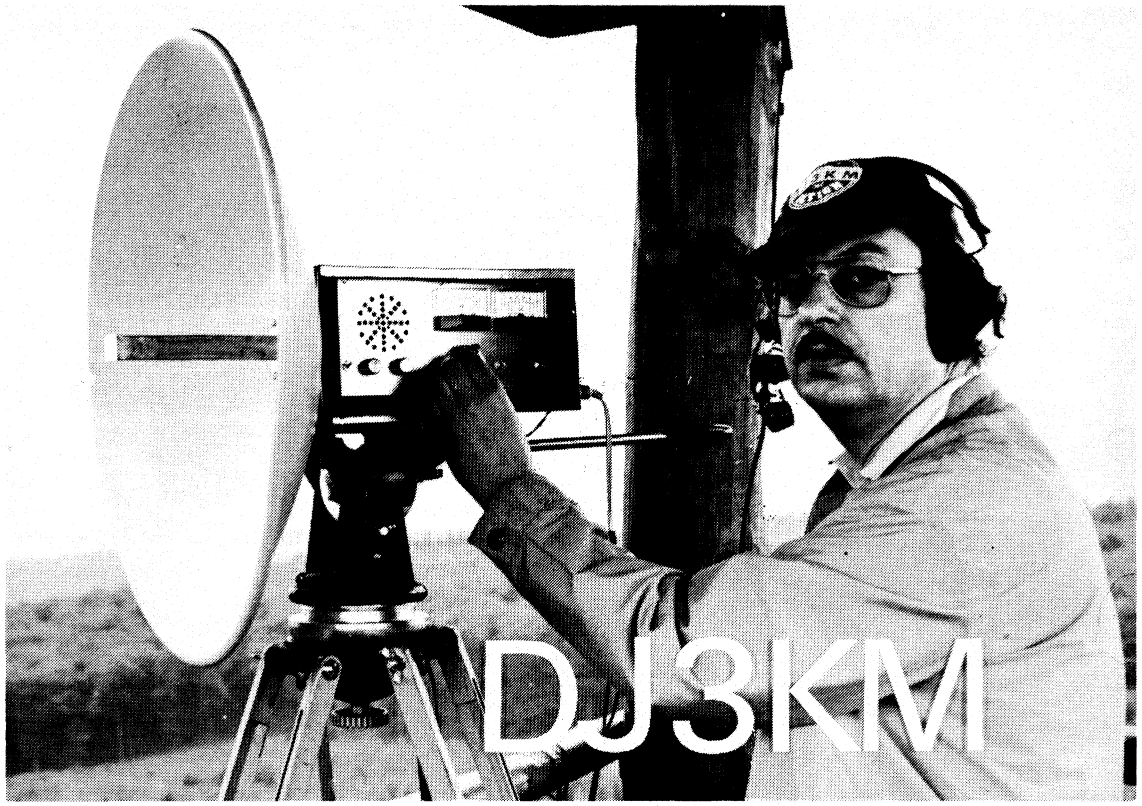


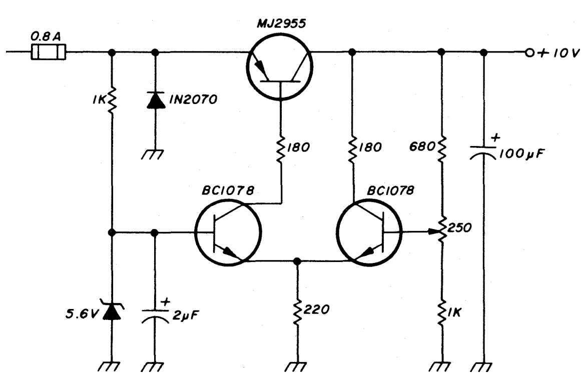
fig. 5. AFC voltage for the 10-GHz Gunnplexer transceiver is derived from the frequency-dependent voltage available from the 30-MHz receiver (fig. 4). The value of resistor R1 (approximately 330 ohms) must be determined experimentally so that 1 volt is measured at TP1.



QSL card used by DJ3KM showing his 10-GHz Gunnplexer and 30-MHz i-f receiver.

of 100 to 200 kilometers (60-120 miles). Since a pair of Gunnplexers with these high-gain antennas has a calculated systems range of at least 400 km (240 miles), we could work over distances greater than 200 kilometers (120 miles) if we could find a non-obstructed path that long.

When setting up the Gunnplexers it's helpful to have a secondary link on 144 or 432 MHz, but many contacts have been achieved without it. The operation of a microwave transceiver with the aid of a map and compass is a new challenge and hobby for many amateurs in Germany.



A +10 volt regulated power supply recommended for use with the 10 GHz Gunnplexer transceiver. The BC107B transistors may be replaced by any small-signal NPN silicon transistors such as the 2N4124. The MJ2955 may be replaced by a 2N3789 or similar 10 amp PNP device.

Activity on 10 GHz in Europe has now reached the point that a 10-GHz bandplan has been approved by amateur groups in Germany, Holland, and Switzerland. In addition to providing space for communications between individual amateurs, the bandplan accommodates beacons, repeaters, and narrowband modes (CW, RTTY, SSTV, and single sideband).

Trial runs with higher gain antennas, narrower i-f bandwidths, and phase-locked loop circuitry for frequency stability are presently going on (reference 3, which describes a phase-locked Gunnplexer system devised by WA6EXV, is available from Microwave Associates).

I would especially like to thank DJ6RW, DJ3KM, DK2DRX, DJ8QL, and DJ8CY for their help in the construction and planning of this equipment.

references

1. J. R. Fisk, W1HR, "Solid-State Microwave RF Generators," *ham radio,* April, 1977, page 10.
2. B. Heubush, DC5CS, Dr. Ing. A. Hock, DCØMT, and H. Knauf, DC5CY, "Ein Sende-Empfanger fur das 10-GHz Band," UKW *Berichte,* Autumn, 1976, page 184; Winter, 1976, page 245; and Spring, 1977, page 47.
3. C. Swedblom, WA6EXV, "ROCLOC Gunnplexer Stabilization System," available from Microwave Associates, Inc., South Avenue, Burlington, Massachusetts 01803.

ham radio

frequency-lock loop

Oscillator stability can be improved by applying this simple but effective frequency-lock loop

One of the main considerations in the design of radio communications systems is frequency stability. The objectives in the amateur radio service, however, are often quite different from those of other hf services. Amateurs have band allocations, while most other users have spot frequencies to work on, and consequently the vfo is usually our preferred primary frequency source. There are three basic frequency generation techniques in common use at the present time. The vfo is the oldest, offering simplicity and the very real asset of continuous tuning, but it is difficult to achieve high stability, especially in the long term. The crystal-controlled oscillator is also simple and very stable, but offers little flexibility, although such variations as the vxo and the "Rock-Mixer" have offered some help in this direction. Finally, there is the synthesizer, based on the phase-locked loop. At the expense of some complexity, this method offers excellent stability and can be very flexible. However, it is inherently a noncontinuously-tuned device, and, therefore, not as well suited to amateur applications — especially on the hf bands.

The vfo, in all respects except stability, offers what we need. It seems a pity to throw away all the results of the continuing development which have made the vfo as good a piece of equipment as it is, and start all over again with the synthesizer. On the other hand, the approach I have taken with the frequency-lock loop (FLL) takes advantage of the positive points of the vfo and adds to it the stability of the crystal oscillator. Moreover, you can readily add an FLL as an outboard unit to an existing vfo without major modification to your equipment.

basic principles

If you have a good frequency counter with a readout down to 1 Hz, you can, by manual tuning adjustments made suitably often, keep the vfo on the required frequency indefinitely. The stability in the medium to long term is that of the counter's clock. The function of the FLL is to automate this operation.

The frequency-lock loop consists of a simplified counter with a crystal derived clock, an error detector and latch circuit, a filter section, and a controlled reactance to compensate for drift in the vfo tank circuit. The error detector may be compared with the operator's recognition of a significant change in frequency, the filter his decision on the magnitude of the correction, and the controlled reactance the action of his hand on the vfo tuning knob.

counter

The purpose of the counter in the FLL system is not to display frequency, but to control it. And, as there is no reason to operate in the decimal or BCD modes, the simple binary counter is used. Comparing the FLL with the manual control, it should be obvious that there is no need to consider the most significant digits of the count. It is hoped that the vfo will not drift so much that the tens and hundreds of kHz would ever change, and surely not the MHz! So, for compensation of drift instabilities, only a small portion of a counter is required, and that can be in binary form.

The gate period is also of fundamental importance.

By Crawford MacKeand, WA3ZKZ, 115 South Spring Valley Road, Greenville, Delaware 19807

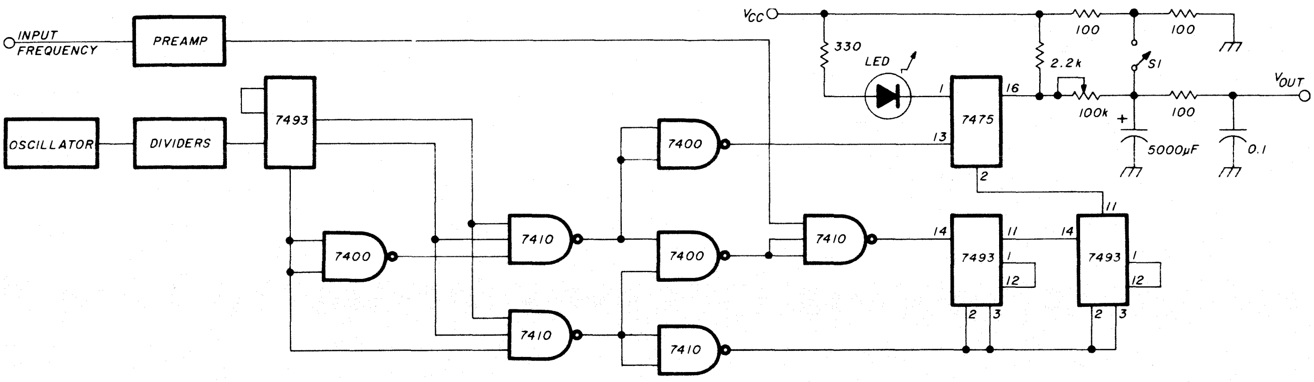


fig. 1. Partiai schematic diagram of the basic frequency lock loop circuit. At this point, interpolation within the 256 cycle groups has not been taken into account.

I originally decided on an updating frequency, based on my feelings for drift rate, of once every 3 seconds, (clock 4.2 Hz), arguing that no significant drift would occur in a gate period of 2.8 seconds. Although this is true, I have changed to a higher clocking frequency of about 420 Hz and a gate of 28 mS. The longer period works fine, but the device takes so long to decide what to do next that the user rapidly loses patience with it.

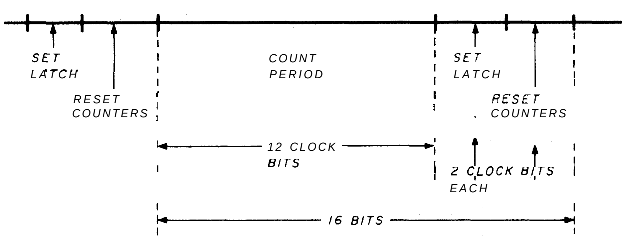


fig. 2. Timing cycle of the basic FLL system.

The counter gate logic is a modification of that presented by MacLeish.[1] The crystal oscillator and dividers can be any arrangement that supplies the correct clock frequency, provided that it has the requisite stability. The counter preamplifier is also a standard circuit for sampling the output of the controlled oscillator.[2]

error detector and latch

At the end of each count period the counter will be in a state which is dependent on the frequency of the controlled oscillator. If the frequency does not vary, neither will the counter's state at that instant. I initially felt that I would need to devise a circuit which would provide an output indicating whether the controlled oscillator was too high or too low in frequency. The obvious way to do this was by the use of a binary logic comparator such as the 7485. However, this would entail the use of switched inputs to cover all the 256 possible states of the counter. Of course, one point of the 256 is available without any comparator at all: when the final stage of the 8-bit counter makes a transition, either 1 to 0 or 0 to 1. This means that during the period the gate was open some multiple of 128 cycles of the input frequency has been counted (256 cycies if you are oniy iooking at the 1 to 0 transition). Therefore, without any further circuitry, the basic FLL shown in **fig. 1** would indicate whenever the input frequency would satisfy these conditions. Assuming that we consider only 1 to 0 transitions, two successive frequency groups are related by:

$$f_n - f_{n-1} = \frac{k}{12} \cdot f_t \qquad \textbf{(1)}$$

where
k = counter total
f_t = clock frequency in Hz

To complete the error detector, I used a latch to hold the output from one count to the next. The output of the latch is a TTL signal; one state indicates that the input frequency is too high and the other state indicates that the input frequency is too low.

filter

If the latch output were applied directly to the controlled reactance, the output frequency of the vfo would constantly be pulled one way and then the

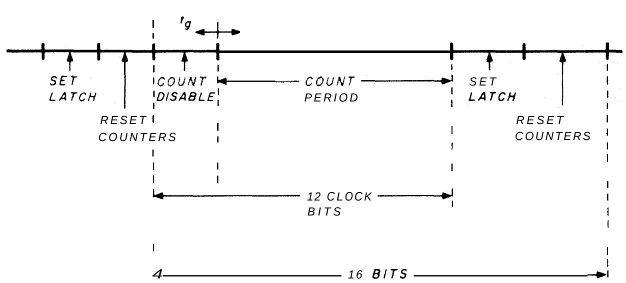


fig. 3. Timing cycle of the frequency lock loop system with interpolation. The 74121 is used to shorten the count period. permitting resolution within a 256 cycle group.

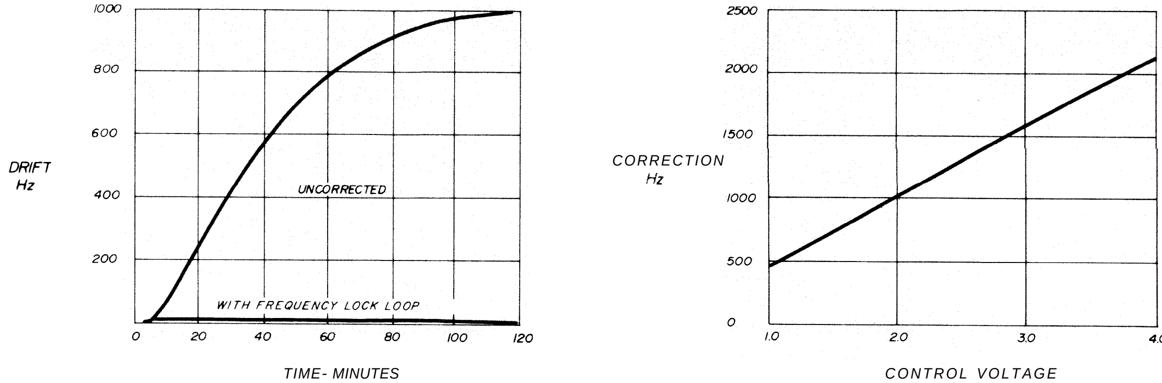


fig. 5. Oscillator drift with, and without, the frequency lock loop system. The range of the correction voltage is shown at the right.

other. However, the mean frequency would be correct. Intuitively, it seems that some smoothing is required. The FLL is very similar to a "bang-bang" servo, and can be readily stabilized by a first order filter or integrator composed of a single RC stage. The optimum filter is probably worth some investigation; nonlinear circuitry may also offer some advantages (a possible approach is described in reference 5).

The filter time constant t_f should be long enough to reduce the fm on the vfo to an acceptable amount, and yet not so long as to make the balancing time excessive. My experiments in this area seem to indicate that somewhere in the region of 50 to 100 seconds is a good starting point.

voltage-controlled reactance

The obvious choice for the controlled tuning reactance is a voltage-variable capacitor diode (varactor

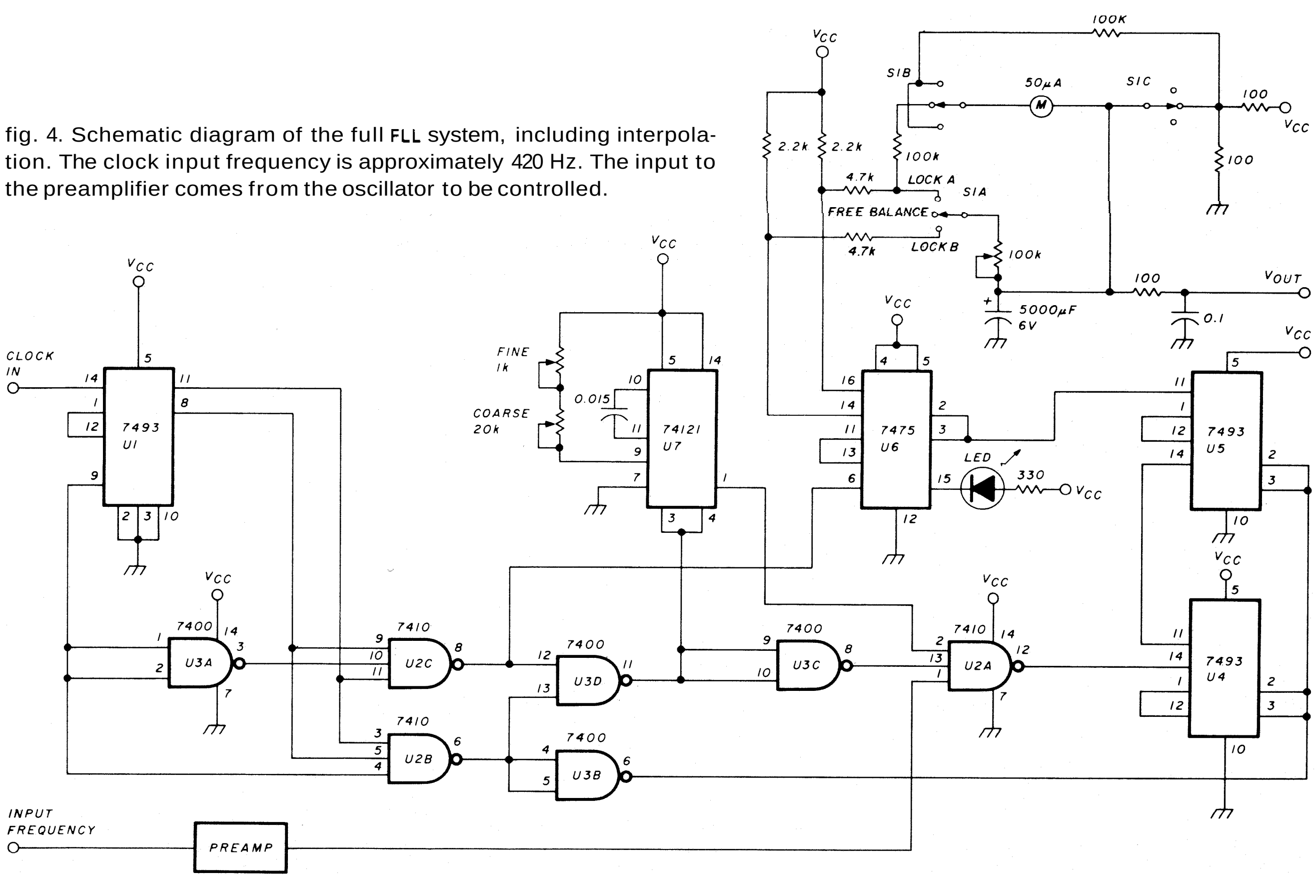


fig. 4. Schematic diagram of the full FLL system, including interpolation. The clock input frequency is approximately 420 Hz. The input to the preamplifier comes from the oscillator to be controlled.

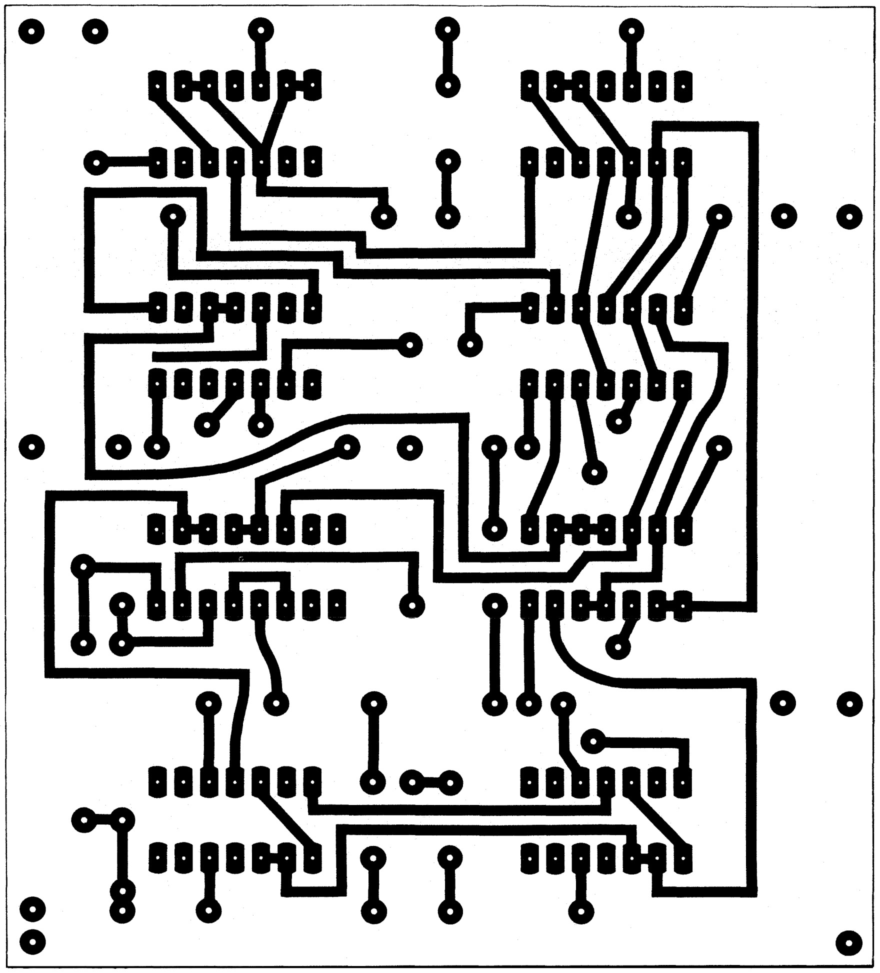


fig. 6. Circuit board layout for the frequency lock loop. Shown above is the back side of the board, with most of the interconnecting wiring; drawing on next page shows the top side of the board and the parts placement diagram. Although not included in fig. 5, this board contains an additional 7490 which is one of the input dividers from the oscillator. Also not shown in fig. 5 are the numerous **0.1-μF** bypass capacitors included on the board.

or varicap). Its application is dependent on the design of the vfo which is to be stabilized. The filter output has a useful range of about +1.5 to **3.5V** dc, although it would be a simple matter to include an op amp if a greater swing were required. The varicap should be connected to the oscillator tank so that it produces, with this voltage range, a frequency variation greater than the drift which is to be corrected.

In mv Hammarlund HQ215 receiver I have been able to stabilize the high-frequency oscillator by coupling into a diode frequency shifter, which is provided for resetting the calibration when changing modes from USB to CW to LSB. Many transceivers have RIT circuits which provide similar access to the oscillator tank, while most transmitters and vfos can easily be modified as if you were providing for FSK operation.

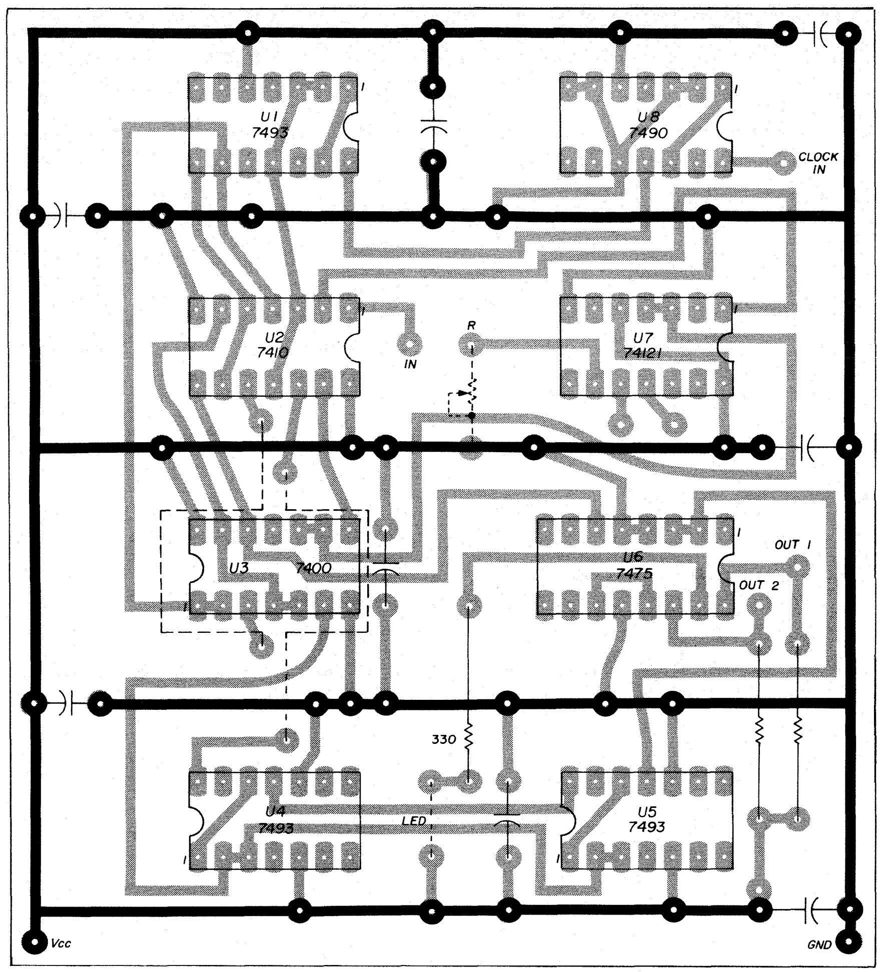
interpolation

The basic FLL of ***fig.* 1** will stabilize a vfo at discrete fixed frequencies, based on a fixed count period determined by the counter clock. The first method of interpolation I considered was that of varying the clock-oscillator frequency, using a vxo as the clock oscillator. With this arrangement I found that

$$\Delta f_a/f_a = \Delta f_x/f_x \qquad \textbf{(2)}$$

where

- f_a is the basic clock oscillator frequency
- Δf_a is the change produced by pulling the vxo
- Δf_x is the resulting change
- f_x is the controlled frequency



This places another constraint on the design, in that Δf_x must be at least as large as $f_n - f_{n-1}$, the difference between successive discrete stabilizing frequencies. But Δf_a is limited by the design of the vxo. Because of this factor, and also the decreased stability of a vxo compared with a regular crystal oscillator, this method was set aside for future consideration in favor of an alternative which permitted the use of a fixed-clock frequency.

The basic timing cycle is shown in **fig.** 2. It should be obvious that if the total counting period could be varied, by at least the time required to count one group of 256 cycles, then the problem of interpolation would be solved. A non-retriggerable one-shot multivibrator is used to create a noncounting period.

The new timing diagram incorporating the interpolating one-shot is shown in **fig.** 3; the schematic diagram shown in **fig. 4.**

operation

The lock switch, S1, is initially set to FREE. In this position the oscillator will be at its nominal calibrated frequency, because R1 and R2 have forced the tuning voltage to its center value. The LED indicator will show the latch's output state. As the oscillator is tuned across its operating range, the LED will cycle on and off every time the frequency changes by $f_n - f_{n-1}$.

If we now choose an operating frequency, the interpolation control is adjusted until the LED flickers, showing that the FLL is ready to lock. The lock point may be either at a 1 to 0 or a 0 to 1 transition as the freqency increases. At this point S1 is moved to either LOCK A or LOCK B. You will know if you've selected the wrong one because the oscillator will rapidly drive off frequency. Initially it is useful to establish a rule such as: clockwise rotation of pot, lights the LED, S1 to LOCK A. After this is established, when you select S1, you're on frequency to stay. Minor frequency adjustments can be made with the potentiometer.

A steady flashing of the LED is a good indication of continuing operation. Meter M1 is valuable in the lock mode to show how far you have drifted and how much corrective capacity you have left. While in the FREE position, it can be used to show which lock position to use and also which way to move the interpolation pot.

performance

In this system almost all of the stability is derived from the crystal clock, with the remainder determined by the RC product in the interpolator. Using the constants discussed, on 80 meters, this amounts to one group out of about 400. In other words, during the total gate period, about 400 groups of 256 cycles are passed, and therefore, only one four-hundredth of the period is dependent on the one-shot's stability. If this is as good as 0.1 per cent, the overall stability is close to one part in 400 000. There is, however, an interesting series of trade-offs between the various constants and values selected. A short-gate period makes the job of the filter easier and reduces the fm effect caused by ripple on the control voltage. A long-gate period, on the other hand, makes the unit difficult to use, but reduces the dependence of the overall stability on the one-shot. Having decided on the gate period, the frequency difference $f_n - f_{n-1}$ is a function of the total count k. If $f_n - f_{n-1}$ is too small, jumping from one stable point to another could presumably occur.

There are a number of points which can be further refined if greater stability were required, but I have found, for instance, that the present design has made it possible to operate unattended on 3600 kHz RTTY autostart, where a stability of ± 10 Hz is desired. My actual achieved stability, as shown in **fig. 5,** is closer to ± 5 Hz, which seems to indicate little drift in the one-shot.

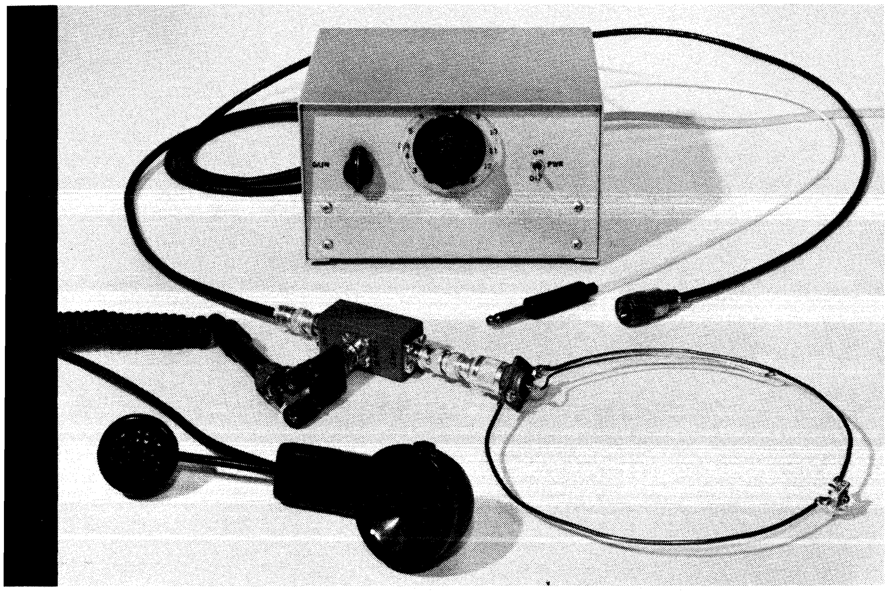
conclusion

The frequency-lock loop provides a simple and effective way of improving the stability of a vfo, effectively competing with a crystal oscillator. Equipment modifications are minimal and can be largely outboard. The components of the FLL itself are all TTL, readily available and inexpensive, while the control system is easy to use and has no tricky components or adjustments. Construction follows normal TTL practice and the simple double-sided layout shown in **fig. 6** is suggested for the main board.

references

1. Kenneth MacLeish, W1EO/7, "A Frequency Counter for the Amateur Station," QST. October, 1970, page 20.
2. Jerry Hall, K1PLP and Charles Watts, WA6GVC, "Learning to Work with Integrated Circuits, Part 8," QST, October, 1976, page 22.
3. Dr. Arthur D. DeLagrange, "Lock onto Frequency." Electronic Design. June 21, 1977.
4. Ulrich L. Rohde, DJ2LR, "The Frequency Counter as a Synthesizer," ham radio, September, 1977. page 44.
5. John C. West, "Servomechanisms," Textbook of Servomechanisms. English Universities Press, London, England, 1953.

ham radio



TVI locator

Locating and correcting the source of TVI is perhaps one of the most difficult tasks facing a radio amateur, one which must be performed methodically if satsifactory results are to be obtained. Much has been learned and written about transmitter harmonic radiation and TV receiver overload, but often very little is said about another prevalent and frustrating source of TV trouble, nonlinear rectification TVI.

Rectification TVI is caused by poor or intermittent contact between two conductors in the radiation field of a transmitting antenna. No amount of filtering or shielding at either the transmitter or TV set will correct the problem, since the interference is generated in the TV spectrum as direct harmonics of the transmitter's fundamental frequency.

In January, 1953, a fine article by Mack Seybold, W2RYI, was published in QST,[1] but I have seen nothing of a concrete nature on this particular problem since that time.

how do I know I have it

Rectification TVI can be suspected when suddenly there is TVI on one or more channels where there was none before, and no changes have been made in transmitter operation. Any metallic discontinuity can cause rectification TVI. In 1947, when I was living in a small town and in the days before the blessings of TV, my next-door neighbor said he heard voices coming from his bathtub drain. Another neighbor heard voices coming from her electric kitchen range. Both voices were caused by detection of my 75-meter a-m kilowatt rig. These two phenomena, no doubt, were caused by rectification.

The strength of the TVI will depend on the efficiency of the rectifier, the length of the "antenna" connected to the nonlinearity, the distance from the transmitting antenna, and the transmitter output power. Two signals on widely separated frequencies can also combine to produce a signal at a third frequency – the faithful $2A \pm B$, or intermodulation products. For example, if two hams live near each other, and one is on 21 MHz and the other on 28 MHz, interference can be caused on channel 4 ($2x21 + 28 = 70$ MHz) or channel 5 ($2x28 + 21 = 77$ MHz), or both, if a nonlinear discontinuity exists in the area. These two signals, of course, will exist only when both stations are transmitting. Also, each signal alone can cause TVI on channel 2 ($28x2$), channel 3 ($21x3$), and channel 6 ($28x3$ and $21x4$).

Visible TVI can be caused by an interfering signal as weak as 40 dB below the video carrier, depending on the frequency of the interference. A 1000 μV video signal, which is an adequate signal, can be interfered with by a 10 μV harmonic. If the amateur transmitter is running one-kW input, this does not leave much margin for harmonic generation.

All 14-MHz harmonics through the sixth can cause trouble, but the greatest problem is caused by the odd harmonics, the third and fifth. **Table 1** shows the harmonic relationships of the 14, 21, and 28 MHz amateur bands with respect to the TV channels. The worst interference is caused at or near the video carrier, 1.25 MHz above the lower TV channel edge. With all stations using color, however, a particularly vicious interference is caused by a harmonic falling on or near the color subcarrier frequency, 4.8 MHz above the lower TV channel edge.

By John E. Pitts, W6BD, 1068 Eden Bower Lane, Redwood City, California 94061

This effect was noted at W6BD on channel 4 when operating near 14.2 MHz. The interference appeared as wide diagonal color (rainbow type) bars on the screen. The fifth harmonic of the fundamental fell within 200 kHz of the color subcarrier at 70.8 MHz. Operation in the CW portion below 14.1 MHz caused no interference. Substitution measurements with a calibrated signal generator showed that color bars were caused by an interfering signal at 71.0 MHz with a signal strength of less than 300 μV. The desired channel 4 signal was 1500 μV. The cause was eventually traced to rectification TVI and was located by the methods presented here.

the fix for the hex

The harmonic chaser used in this hex-pedition (an expedition to find the hex) is simple to construct, easy to use, and will rapidly locate the source of the harmonic radiation. It is also, by today's standards at least, inexpensive. In this instance, the whole system was constructed and tested and the TVI source found in one weekend, so the work involved in the project is not great.

table 1. Amateur-band harmonic relationships to low-frequency TV channels. All frequencies are in MHz.

fundamental harmonics 28	21	14	harmonic frequency	TV frequency band	TV channel number
2		4	56	51-60	2
	3		63	60 - 66	3
		5	70	66 - 72	4
				76 - 82	5
3	4	6	34	82 - 88	6

Since the harmonic strength will be a relative measurement, a narrow-band receiver, tuned to the harmonic frequency, will be used. The easiest approach is to use a TV tuner whose i-f output is in an amateur band, This allows the selective station receiver to become the i-f amplifier and detector.

There are generally two types of tuners used for replacement purposes, the turret type and the wafer type. Due to the coil arrangement of the wafer-type tuner, it is unsuited for this purpose because the tuner's oscillator frequency must be changed. The most easily modified is the turret type, because the coils

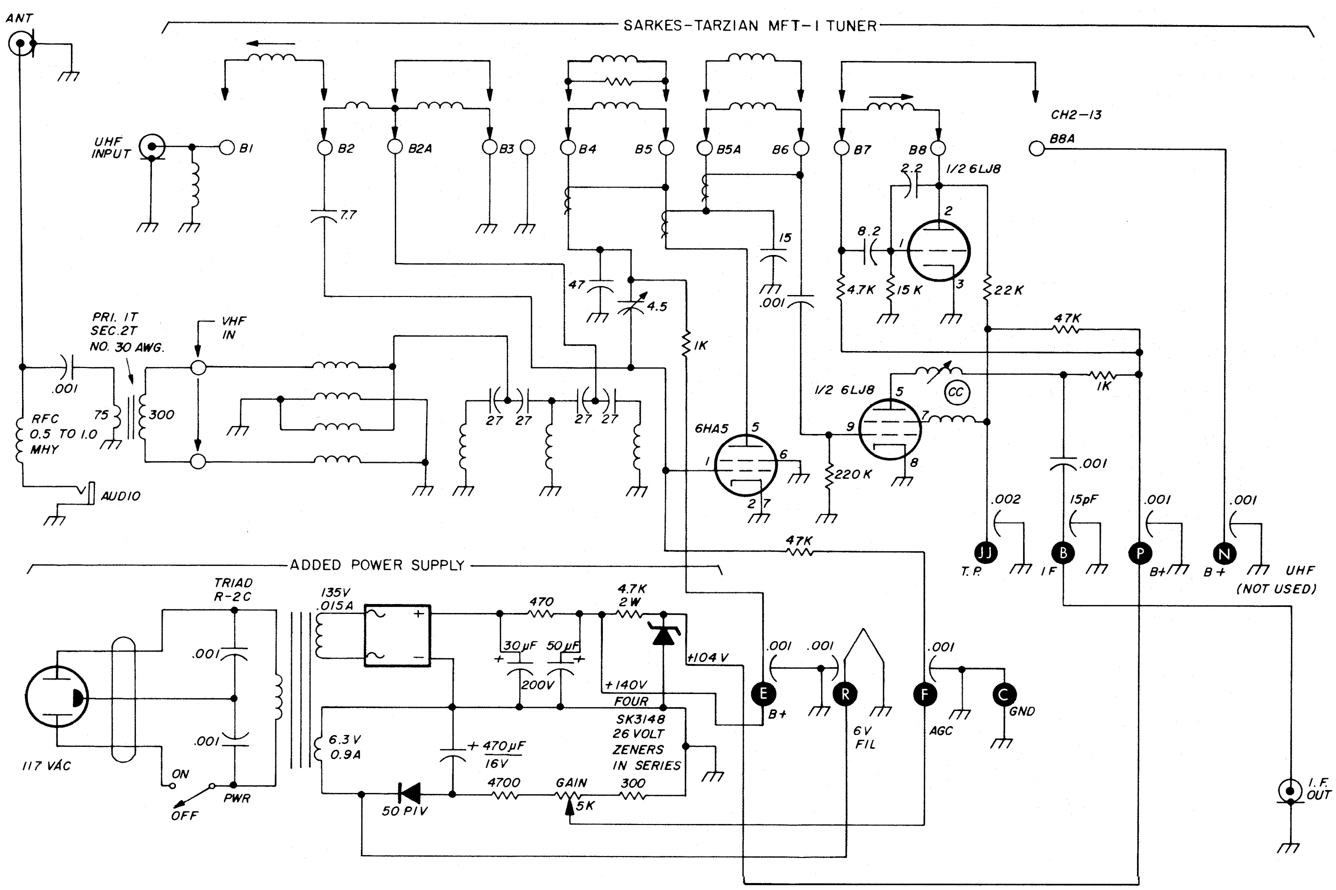


fig. 1. Schematic diagram of the Sarkes-Tarzian tuner and power supply. The coil marked CC is tuned for maximum signal into the receiver. You should not use more than about 60 cm (2 feet) of cable between the tuner and the receiver, otherwise the tuner may not cover the desired output frequency range.

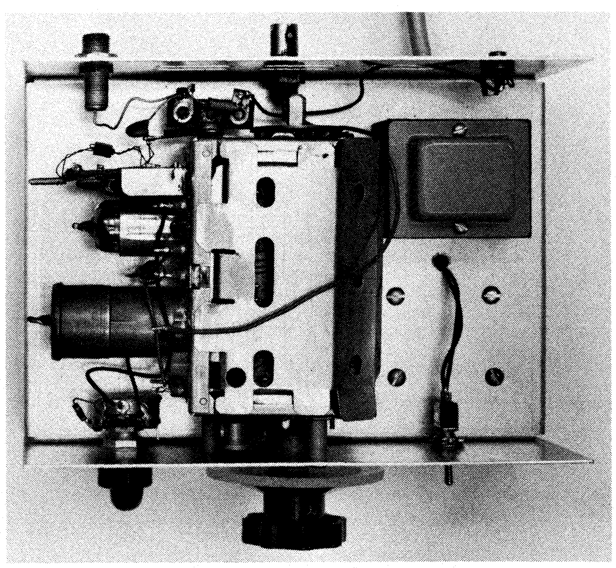
for each channel are mounted on an easily removable bar. Present-day tuners have an i-f of 41.25 to 47.25 MHz, out of the range of most ham receivers. The widest high-frequency amateur band is 28 to 29.7 MHz. Therefore, the oscillator frequency needs to be lowered only about 10 to 12 MHz to produce an i-f output at 29.0 MHz.

The tuner used is a replacement type, Sarkes-Tarzian MFT-1 preset replacement tuner (see **fig. 1**). It is housed, with a small power supply, in an LMB 12.7 x 11.4 x 19.1 cm (5 x 4-1/2 x 7-1/2 inch) W-2F chassis cabinet. Except for the tuner and cabinet, which cost about $27, all parts came from the junk box. Purchasing everything, and with a little horse trading and typical ham ingenuity, the entire cost should not exceed $40.

construction

The original cut-and-try coil modification was performed using a frequency counter. A counter is not absolutely necessary, but if one is available, the job is much easier. If not, a reasonably accurate grid-dip oscillator (GDO) can help set the tuner's oscillator to the required frequencies. The oscillator was tuned to the high side of the desired signal because it did not want to oscillate on the low side. Therefore, as shown in **table** 2, the 10-meter receiver tunes backwards.

The only coil to be rewound is in the oscillator, the coil with the fine-tuning screw slug. Remove the snap-off shield from the tuner chassis. The channels to be modified are 2 through 6, since 7 and above are



Interior view of the tuner section. Loop-antenna input connector is at the left rear, i-f output jack to the receiver in the center, and audio from the receiver is at the extreme right. The 4:1 balun, to match the 75-ohm line to the 300-ohm input, can be seen just below the type-F connector.

not normally subject to rectification TVI. Channel 5 doesn't have to be modified, since no discrete amateur-band harmonic normally falls in this channel. Citizens band harmonics, however, do fall in channel 5.

Rotate the shaft until the bar with the greatest number of coil turns (channels 2 through 6), starting with the bar adjacent to the uhf strip, can be pulled out with the long-nose pliers. The uhf strip has no oscillator coil. The bars are easily removable, but use caution, as they can be broken. Pull at the pressure-finger point, the end with the tuning screw.

Remove all turns from the oscillator coil and clean the soldered portion of the contacts. Use care not to get solder on the switch contact portion of the terminals. Rewind the coils as shown in **table** 2; number 28 (0.32mm) AWG or number 30 (0.25mm) AWG enameled wire can be used. Wind on the number of turns indicated for each channel, observing the same winding direction as used on the other coils on the bar. Wind the turns close-wound, starting at the slug end. If necessary, the turns can be spaced later for the proper frequency range. Unscrew the fine-tuning screw about five turns out from full in. This will provide adjustment range later for the oscillator. Screwing the slug into the coil raises the oscillator frequency, and therefore raises the intermediate frequency to which the receiver is tuned. After each coil is rewound, return the bar to its original position in the turret to prevent mixing their positions.

Install and wire the power supply, jacks, and splitting filter as shown in **fig. 1.** Jacks and power supply may be whatever you have on hand in the junk box. Plate voltage for the tuner may be anything between 110 and 140 volts dc. The bias voltage is obtained from a rectifier on the 6-volt ac filament winding. The values shown for the resistors give a minimum of – 0.8 volt and a maximum of about – 4 volts. Normal operation is at full negative, but, if desired, the bias may be permanently set at – 3 volts by selection of appropriate resistor values.

Install the tuner in the cabinet, mounting it with screws and spacers to the panel. Three of the front holes (near the shaft) will conveniently accept a 6-32 (M3.5) tap or a number 6 sheet-metal screw. For ease of fine-tuning adjustment, a piece of lucite (Plexiglas) — cut to 5.7 cm (2-1/4 inches) in diameter by a circle cutter — forms a good control wheel, similar to the fine-tuning control on a TV set. The center hole is sized for a force fit on the fine-tuning shaft, which is 9.5 mm (3/8 inch) in diameter. Mark the plastic shaft and then cut it to length with a hacksaw, after which the fine-tuning wheel may be forced onto the shaft.

Mark the length required on the selector shaft, cut it with a hacksaw, and smooth with a file. Rotate the

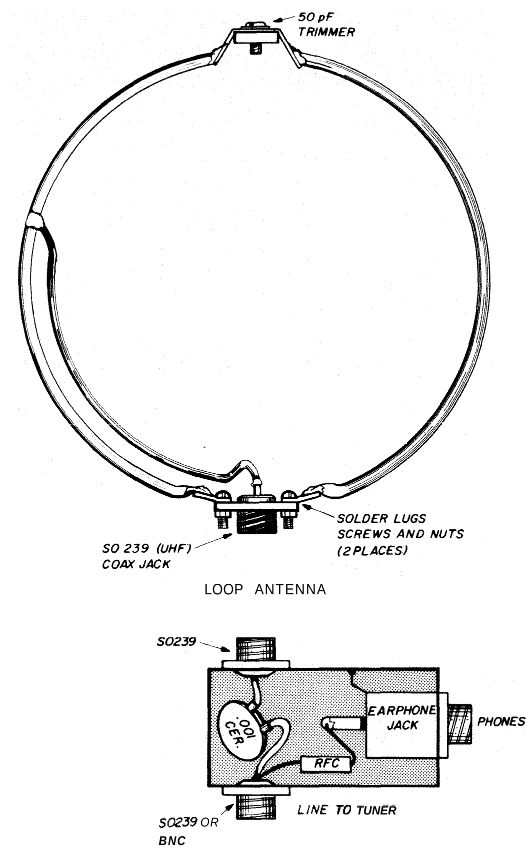


fig. 2. Diagram of the loop antenna and splitting filter. The filter is constructed in a small box, such as an LMB-M-00. The capacitor at the top of the loop is a 50-pF compression trimmer and is used to tune the loop to the desired frequency.

shaft with pliers so the uhf strip is in its operating position, then mount a skirt-type knob with the indicator mark toward the bottom of the panel. Channel 2 will be at the first position to the left of bottom as the turret knob is rotated clockwise.

loop antenna

The loop is constructed of two 25-cm (10-inch) lengths of number 10 (2.6 mm) AWG wire formed into a loop about 18.5 cm (7-1/4 inches) in diameter (see fig. **2).** The base of the loop is fastened to the shell of an SO239 uhf jack, with screws and nuts holding two soldering lugs onto which the loop wires are soldered.

A 50-pF trimmer is soldered to the wires at the top of the loop. A piece of number 12 (2 mm) AWG or number 14 (1.6 mm) AWG copper wire is soldered to the inner terminal of the SO239 jack, formed to the contour of the loop with 6- to 9-mm (1/4-to-3/8-inch) separation, and soldered to the loop 13 cm (5 inches) up its circumference.

Using appropriate connectors and a very small metal box, the splitting filter is constructed for the earphone or telephone connection at the base of the loop. When connecting the filter to the antenna, observe the connections shown in the figure. If connected backwards, the loop will work, but no sound will be heard in the phones.

tuning

Connect the tuner and station receiver together as shown in fig. 3. Temporarily connect the harmonic-producing network (fig. **4**) between the tuner and transmitter output. Place the tuner on channel *2;* tune the receiver to 29 MHz and the transmitter to 28.000 MHz or 14.000 MHz. Only very low output is necessary, just enough to make the diode conduct, producing harmonics. Turn the transmitter on, and also the receiver bfo. Very ***slowly,*** rotate the fine-tuning control until the transmitter harmonic at 56 MHz is heard. Verify this frequency by using the GDO as a signal generator. If no signal is heard, tune the receiver between 28 and 30 MHz and adjust the fine-tuning control until the 56 MHz harmonic is received. Do not confuse the desired signal with the fundamental or second harmonic of the transmitter output, bypassed around the tuner. Then jockey the receiver tuning and fine-tuning control on the tuner until the second harmonic of 28 MHz or the fourth harmonic' of 14 MHz (56 MHz) is at 29 MHz on the receiver. Look up the signal frequencies for the various TV channel video and sound carriers in table **2.** If channel 2 exists in your area, it can easily be heard when an antenna is connected to the tuner input and the

table 2. LO coil winding and i-f frequency output data for TV tuner modification. All frequencies are in MHz.

channel	LO coil number of turns	LO frequency	TV video receiver i-f	TV sound receiver i-f	receiver dial frequency 31	30	29	28	27	26	25
2	16	85	55.25 29.75	59.75 25.25	54	55	56	57	58	59	60
3	14	92	61.25 30.75	65.75 26.25	61	62	63	64	65	66	67
4	14	99	67.25 31.75	71.75 27.75	68	69	70	71	72	73	74
6	11	113	83.25 29.75	87.75 25.25	82	83	84	85	86	87	88

receiver is tuned to the indicated i-f frequency. In my test set-up, a 3 μV signal on any of the converted TV channels could easily be heard in the receiver.

Repeat the tuning procedure for the other channels and amateur bands according to the table. Note that the video or sound carrier can be used as check points if they're within the tuning range of the station receiver. I use my old Hammarlund HQ129X. The video carrier is a strong signal with 15.75 kHz sidebands extending several hundred kHz each side. The sound carrier has distorted modulation, since it is fm.

The loop is connected to the tuner via a convenient length of RG-58 or RG-59 cable equipped with suitable connectors. The most inexpensive connectors are F-type, used for TV cable connections. In my case, in order to reach the source of the rectification, 60 meters (200 feet) of cable was required. If you use F-type connectors, note that they are designed for coax with a solid center conductor.

The loop antenna operates as a radio direction finder to locate the source of signal rectification causing generation of harmonics. In order to hear the effect of loop rotation on the signal, the audio output of the receiver is sent via the coax cable to headphones or a telephone carried by the ioop-antenna operator. While slowly rotating the loop about its vertical axis, a distinct null, about 2 or 3 degrees wide, is easily heard.

Although a loop is normally bidirectional, in this case, due to the tapped feed point, it exhibits about 10 dB of front-to-back ratio when properly tuned. With the operator looking through the loop, he is facing the signal when the deepest null is heard with the feed tap on the ***left*** side of the loop. Rotating the loop about its horizontal axis will indicate, by a deeper null, the angle of elevation of the incoming signal. For maximum directivity, the trimmer capacitor must be tunned for maximum signal at the frequency of interest.

Loop operation can be verified by tuning it and the

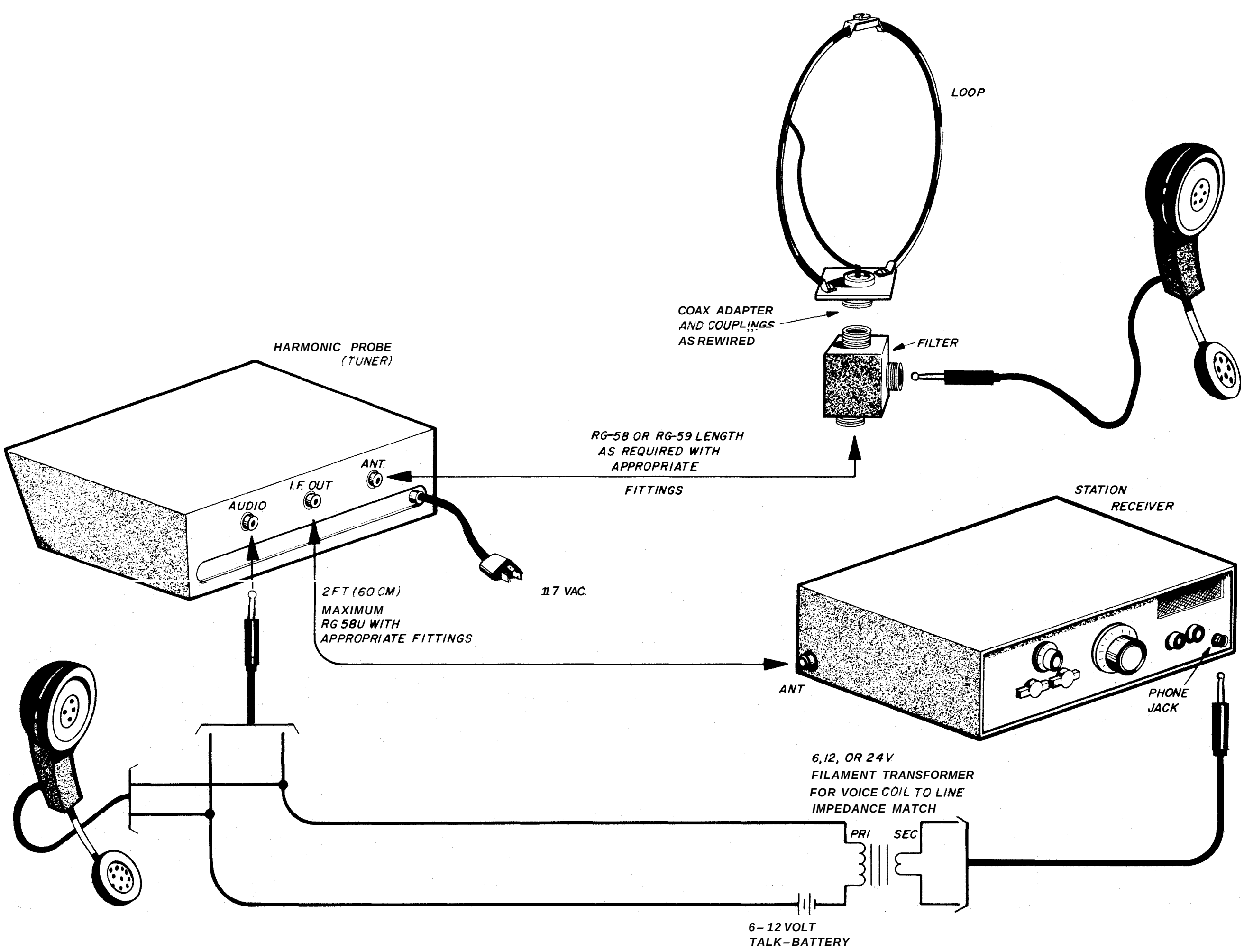


fig. 3. Interconnection diagram of the loop, tuner, and receiver. The earphone and microphone of each handset are connected in series. The battery is not required if the earphones alone are used.

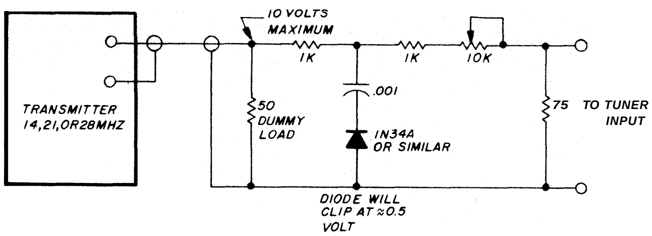


fig. 4. Schematic diagram of the harmonic-producing network. This circuit is used to produce harmonics to calibrate the tuner and receiver. A maximum of 2 watts should be applied.

receiver to a TV station and observing the effect of rotation. This test may be invalid if many echoes exist from pipes, ducts, or other large metal surfaces. This same effect must be considered when looking for TVI.

finding the hex

Set up the equipment as shown in the block diagram, **fig.** 3. Tune the TV set to the channel having interference. Turn on the transmitter and verify that it is causing interference, then turn the tuner to the same channel. Use only sufficient transmitter power to cause TVI. Tune the receiver to 29 MHz and find the harmonic. Note that **table 2** is based on the lower edge of the 14-, 21-, and 28-MHz bands. Also note that the receiver, used as an i-f amplifier, tunes backwards. For example, if 21 MHz interferes with channel 3, the third harmonic is at 63 MHz and is tuned on the receiver at 29 MHz. If the transmitter is tuned to 21.3 MHz, the third harmonic is at 63.9 MHz and will fall at an i-f frequency of 28.1 MHz on the receiver dial.

Set the receiver controls for CW operation, and tune the harmonic so that its detected audio frequency is about 1 kHz. It will be necessary to retune the receiver from time to time since the oscillator will drift slightly. Therefore, an operator should be at the receiver for periodic tuning, and to key the transmitter on request. If two telephone handsets or operator's headsets are available, constant communication between the antenna and receiver operators is possible.

Go outside the house and take a preliminary bearing on the interference source. Note the direction (a rough sketch or map may be helpful). Go to a second location and take a second bearing. In all but the most elusive cases of interference, two or three bearings will suffice. Rotating the loop axis vertically, rather than horizontally, will indicate the elevation of the source above ground level.

Under certain conditions, it may be advantageous to turn off the receiver agc and have the receiver operator control the signal level with the receiver's rf gain control. When nearing the interference source, or when using the probe as a "sniffer" for harmonics radiating from equipment, a coax plug fitted with a few centimeters of stiff wire will serve as a probe antenna.

Due to the attenuated response of the loop antenna at the normal amateur frequencies, a highpass filter of the TV type was not found necessary. If one is used, it must be located after the splitting filter in the tuner, or the telephone extension will not work.

where to look

Many things can cause a rectification-harmonic problem. Some of these are rain gutters, downspouts, roof flashing (the metal under shingles), corroded TV antennas, rusty TV masts, poor (unsoldered) splices in TV feedlines (or in the station antenna system both transmitting and receiving), poor electrical conduit joints and other metal junctions of this nature, all transistorized equipment, intercoms, pipes, telephones, concrete reinforcing bars – the list is almost endless. Any two touching pieces of metal more than a few centimeters long in the field of the transmitting antenna are suspect. The obvious solution to the problem is to permanently bond the two pieces, or, if no electrical continuity is necessary, to permanently insulate them.

Three cases have been found and corrected at my location, galvanized-tin roof flashing and corroded TV antennas on two adjacent houses being the cause. In the latter case, good relations have always been maintained with the neighbors, so no problem existed in correcting the situation. In fact, one case resulted in a very nice Christmas gift as an expression of gratitude. The tin flashing problem was fixed by permanently connecting the two pieces with sheet metal screws and anti-corrosive grease, permanent separation being impractical. The corroded TV antennas were scraped clean at the connection points and then painted with an anti-corrosive grease.

About eight years ago, long before this equipment was built, I found a source of rectification in my own TV antenna so severe that a 75-watt transmitter on 3.5 MHz feeding a dummy load caused TVI. A friend and I found the cause, wholly by accident, after a prolonged search. With the equipment described here, it would have been found in minutes. Now that you have the tools, good hunting, and may all your hexes be easy ones.

reference

1. Mack Seybold, "Harmonic Radiation from External Nonlinear Systems," QST, January, 1953, page 11.

ham radio

a dream realized:
the ultimate antenna array

A 7-element quad on 40 meters? You'd better believe it! Here's an account of how one DXer solved the problem of big antennas

Most of us at one time or another have fantasized about having the ultimate array: the antenna to make you king of the band; the supergain bone crusher. Usually these dreams are dashed away by the reality of circumstances, but sometimes someone will succeed in getting one of these monsters up. Although, generally, this supreme achievement will go unnoticed by most, the rewards of the labor are still collected in abundance by the ambitious amateur who undertakes the challenge and succeeds.

The following account isn't meant to be a construction project but is presented with the hope that some of the ideas will convince others that, first, you don't need a lot of money to build a large array; and, second, some dreams can come true with a little applied ingenuity.

how it all began

Having been one of those few fortunates who've had the pleasure of operating at a large multi-multi station during DX contests, I've become appreciative of the merits of high-gain antennas. One day in early 1973 I was discussing various antennas with Jerry, WA7KYZ, when the subject of 40-meter arrays came up. Since 40 meters is generally considered to be the transition between wire dipoles and rotatable beams, we decided to experiment with some high-gain fixed-wire antennas on that band. Fortunately, we had a sizable piece of land on which to work. This property was dotted with 46-meter (150-foot)-high Douglas fir trees.

initial attempts

The first antenna we tried was a full-size four-section 8JK beam. On paper it looked really simple, but it turned out to be a real monster. We had to resort to using 2.6 mm (no. 10) copper-plated steel wire for the elements and 17-foot-long 1x6s (5 meters x 25 x

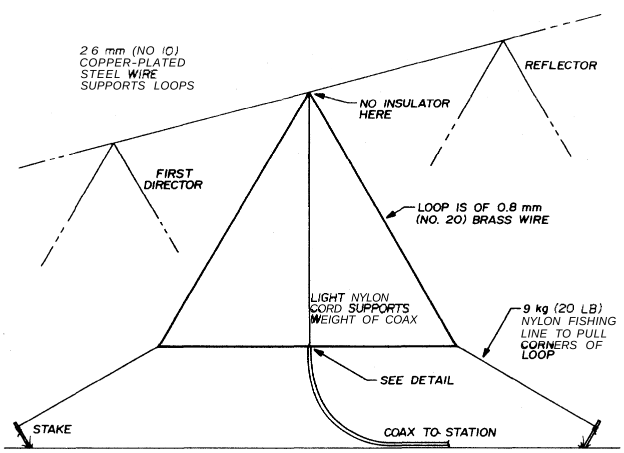
By Paul Kiesel, K7CW, 3522 N.E. 115th Street, Seattle, Washington 98125

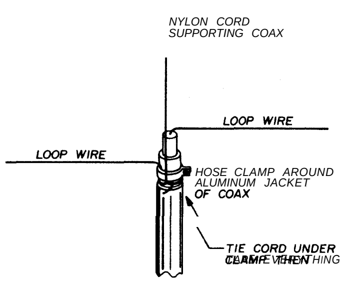
152 mm) for the spreaders. We finally got the thing up in the air by pulling up the ends with a pickup truck.

The antenna worked reasonably well. It seemed to have a low angle of radiation, as it was supposed to, and it definitely had gain. But it petformed well in only two directions. It had a very narrow beamwidth. Additionally, we had a problem that we hadn't contemplated: we had to keep untangling the open-wire feedlines. Also, we had to use a transmatch, which made things even more cumbersome. The antenna eventually came down when an ambitious ten-year-old neighbor untied one of the support ropes at the base of the tree. We had mixed feelings about the array's demise.

the grand experiment

We fiddle-fumbled around for some time before we came up with the ultimate solution, the utopian array. It was to be a multi-element delta-loop quad. We decided to go with seven elements aimed at Europe. Every amateur in Washington state who works DX knows that the European path is the toughest nut to crack, because we have to battle the northern auroral zone. The east-coast guys have the same problem working into Japan. So we had to have a lot of gain; however, we didn't want to narrow the pattern of the array too much. We could have put three times as many elements on the thing,





DETAIL OF FEED POINT

fig. 1. Details of the driven element for the 7-element, 40-meter delta-loop quad antenna. High Douglas fir trees provided the supports. Handbook data were used for loop dimensions and element spacing, which was 0.2 wavelength. The vswr was measured at 1.5.

since the supporting wire was 137 meters (450 feet) long! We couldn't use the 2.6-mm (no. 10) wire for all the elements and support because it would have made the antenna much too heavy, so we used 0.8-mm (no. 20) brass wire for the elements. Sounds

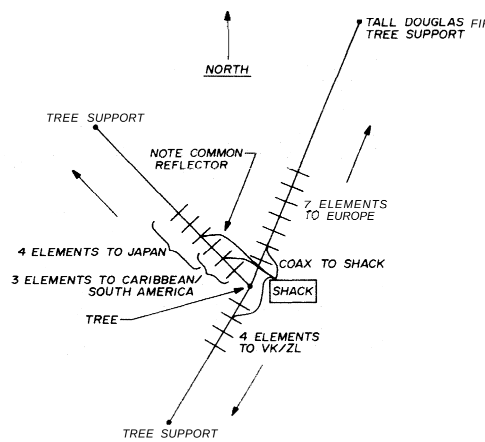


fig. 2. The ultimate antenna farm, which includes three switchable monster arrays covering Oceania, Europe, and South America, including the long path to the Orient.

flimsy, all right, but it works fine for a temporary effort. We really lucked out on the element wire. I picked up 1220 meters (4000 feet) of it at a surplus place for $4.00. We used 9 kg (20-lb) nylon fishing line to pull out the corners of the loops. Jerry picked up some 75-ohm coax remnants from the local cable television company, so we used some of this for the feedline.

up she goes

Now, back to the support wire. Our two support trees were 137 meters (450 feet) apart. We couldn't afford polypropylene or similar rope, so we decided to use some of the 2.6-mm (no. 10) wire left over from the 8JK project. We used no insulators to separate the loops from the support wire as they were unnecessary. As soon as the array was secured, we checked the swr to find that it was only about 1.5:1. We used 0.2-wavelength spacing between elements and the customary formula for loop dimensions. We kept all directors the same size. We found that the array would sway freely but not excessively with the breeze.

I wanted to photograph our antenna for posterity, but this proved impossible. All that was visible in the photos was a 137-meter-long (450 feet) wire, two trees, and a piece of coax reaching up into the air and seemingly terminating into nothing. There are definite possibilities here for those who need to display as little antenna as possible.

It took Jerry, my brother, John (K7TU, ex-WA7OTT), and me about a day and a half to build the antenna from scratch. As we secured each element, we pulled up the support wire a bit. We used a small butane torch to solder the loops to the supporting wire. To support the coax so it wouldn't pull the loop out of shape, we ran a nylon cord down from the supporting wire to the feed point (fig. 1).

performance

I only use published gain figures as a ballpark method for deciding what kinds of array to consider, and I never tried to measure the gain of this antenna with respect to another antenna because of the many variables involved. My method of evaluation is to just get on the air and see how well the array works by communicating.

With the antenna connected to the rig, we tuned through the CW band and came across several UA1s and UA3s !Remember – this was on the 40-meter band.) Their signals were so weak the S-meter didn't move, but all were solid copy. The only other signals on the band were some rag-chewing W6s. It doesn't sound too impressive until I mention that this first check was carried out at 12 o'clock noon.

A couple hours later we were able to work into Europe with consistent S4-S6 signals using only a barefoot T4X exciter. Later on that evening we were getting consistent 599 reports from all over Europe. The Europeans we worked were all solid copy. Our antenna was turning out to be a great performer.

further experiments

When we found out how well our antenna was performing, we decided to erect three more of similar type. We had enough material to erect three elements centered on the Caribbean/South America region, four elements on Japan, and four elements on the VK/ZL/Europe long path (fig. 2). By now we knew how to go about constructing the antennas, and the three new ones only took another day to erect. Upon trying them out we found that all three new antennas performed very well also.

sidelights

At this point a small anecdote relative to our antennas is appropriate. Soon after we got the four arrays up I was stringing JAs at about 9 o'clock in the evening when Gordy, W7SFA (now W7FU), broke in. He wanted to know how well I was hearing Japan. I told him that signals were moderate. This revelation must have surprised him, because he asked me what kind of antenna I had. I told him I was using a 4-element quad and that it was working quite well. Upon being asked by him if I could rotate my quad, I replied that I could and told him to stand by. At that point I paused and flipped the antenna switch over to the European 7-element job, which just happened to be pointing directly at him. I hit the key again and asked him how it sounded. I won't repeat his reply here. And, quite frankly, his signals came up so much with the switch of antennas that their strength almost blasted me right out of my chair. Whew!

As flimsy as the antennas appeared, they proved to be very durable. They remained erect through several storms and during both weekends of the CQ WW DX Contest. In fact they were still in the air when the location had to be vacated a couple of months later.

Although we couldn't use the antennas on a permanent basis, they were still well worth the effort. I'll never forget how much fun it was to tell the Europeans on 40 meters; "The antenna here is a 7-element quad."

ham radio

"Since you're afraid of heights I was sort of hoping for a brother to help with my antenna."

higher frequency resolution for an hf synthesizer

A unique method for obtaining 10-Hz increments from an hf synthesizer using a dual vco system

The advent of low cost phase-locked-loop frequency synthesizers has had considerable impact on vhf amateur radio equipment. Unfortunately, the frequency synthesizer is most adaptable to channelized systems, consisting of a finite number of discrete operating frequencies. Typical high-frequency amateur activity consists of tuning an analog-oscillator controlled radio to a clear frequency and making a call. To answer such a call, with single-sideband equipment, you must tune to within 50 Hz of the originating station's frequency for near natural voice reproduction. This means that a practical, synthesized, hf ssb transceiver must be capable of continuously tuning in 100 Hz steps, and requires an internal, loop-reference frequency of 100 Hz in a conventional configuration. The loop filter cut-off frequency required to effectively eliminate reference frequency sidebands from the synthesizer's output increases the loop lock-up time to several seconds after each frequency change. The ideal amateur CW receiver, equipped with narrowband i-f or audio filters, must be continuously tunable in 10-Hz steps, and this would have ten times longer lock-up time between frequency changes.

This article briefly describes a less common approach to an hf synthesizer which offers 10-Hz frequency steps from a 10000 Hz reference, and therefore offers 1000 times faster recovery after frequency excursions. In addition, less rugged mechanical construction of system oscillators is possible because of loop correction of low-frequency fm due to vibration.

Fig. 1 is a functional block diagram of a conventional PLL frequency synthesizer. The phase detector supplies correction pulses to the vco, through a low-pass filter, at a frequency equal to the reference frequency, until the vco is locked at a frequency equal to N times the reference frequency. A lower reference frequency requires a lower filter cutoff for a given attenuation of the ac component of the reference frequency. The filter is part of the closed loop and its response determines the maximum rate of vco frequency correction.

Fig. 2 is a functional block diagram of a two-part synthesizer, offering 10-Hz steps with the advantages of a 10 kHz reference frequency. Note the use of two reference frequencies, 10.000 kHz and 9.990 kHz. The output frequency is actually the difference frequency between two phase-locked oscillators. To change the output frequency by 10 Hz, we move the first oscillator 10.00 kHz; next, we move oscillator number two 9.99 kHz in the same direction. The difference between the oscillators' frequencies has only changed 10 Hz.

In the following example, I have made provisions for high-side LO injection in a super-heterodyne application, employing a 9.0 MHz i-f system. Also, sample calculations are shown for bfo/carrier frequencies of 8998.5 kHz for lower sideband and 9001.5 kHz for upper sideband (remember the sideband inversion with high-side local oscillator injection). The actual operating carrier frequency is programmed in BCD into the circuit's adders. Functionally, the circuit is divided into two major sections. Section two covers a range of approximately 50 to 60 MHz, with section one covering approximately 59 to 99 MHz.

To run section two at approximately 50 MHz, a divide ratio (programmed vco 2 offset) of 5005 is *initially* chosen for divider 2. To the resulting frequency

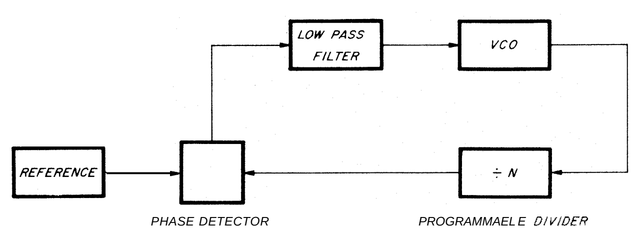


fig. 1. Block diagram of a conventional PLL synthesizer.

By William E. Coleman, N4ES, E-Systems/ECI Division, Box 12248, MS 11, St. Petersburg, Florida 33733

(49999.95 kHz) is added the bfo frequencies of 8998.5 kHz and 9001.5 kHz, producing the initial vco 1 frequency. This procedure allows us to calculate the final required programmed offsets for dividers 1 and 2 (fig. 3).

Therefore, for a programmed operating frequency of 00000.00 kHz, divider 1 must divide by 6744 for lower sideband, and 6045 for upper sideband as shown in fig. 1. Vco I will then operate at 67440 kHz for LSB and 60450 kHz for USB (fig. 4).

And, for a programmed operating frequency of 00000.00 kHz, divider 2 wiii divide by 5850 for LSB and 5150 for USB. Vco 2 will operate at 58441.5 kHz for LSB, and 51448.5 kHz for USB (fig. 5).

With a sample operating frequency of 14307.96 kHz USB programmed into the synthesizer's data input, divider 1 would divide by 8271. Vco 1 would operate at 82710 kHz (fig. 6A). Divider 2 would divide by 5946, and vco 2 would operate at 59400.54 kHz (figs. 6B and 6C).

The i-f offset is implemented by separating the divide cycle of each programmable divide chain into

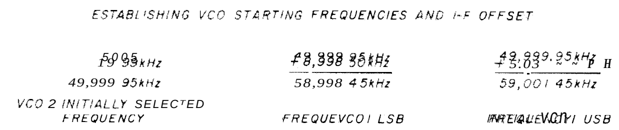


fig. 3. The initial frequency for vco 2 is determined by assuming there is no programmed input frequency (00,000.00 kHz). For the vco to be at its proper lower frequency limit, the divider would have to have an initial divide factor of 5005, producing a 49999.95 kHz output. Vco number 1 will also start with the same frequency, except with the added offset required for either LSB or USB.

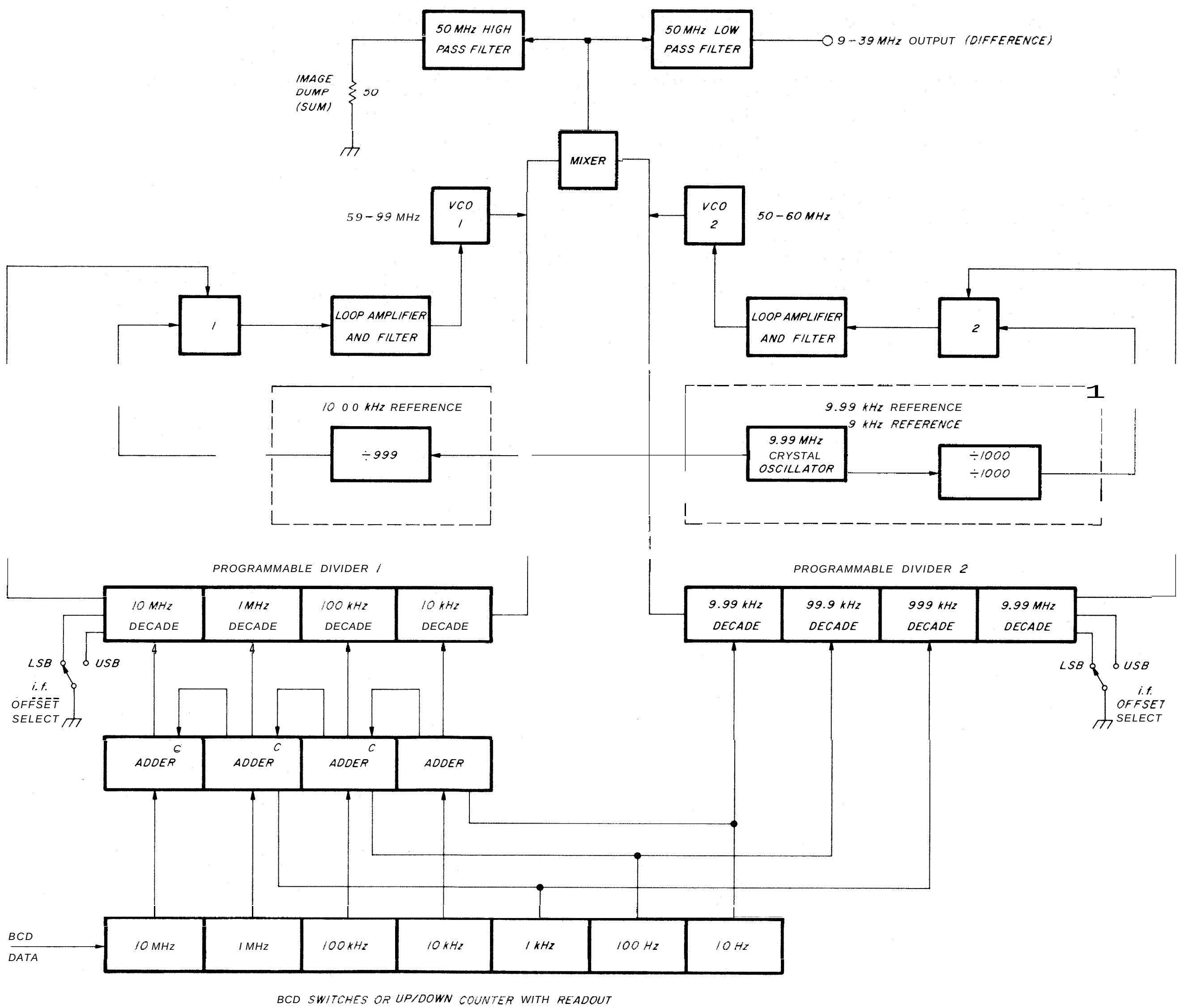


fig. 2. Functional block diagram of a high frequency synthesizer capable of 10-Hz steps, with a 10-kHz reference.

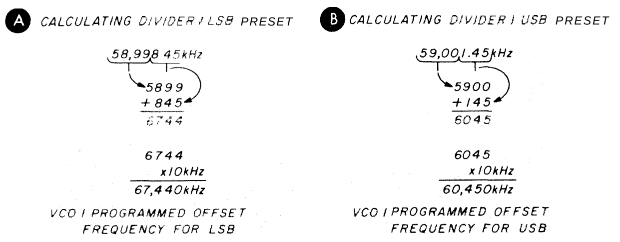


fig. **4.** As seen in fig. 2, the data from the three most significant digits is added to the data from the fourth through the sixth significant digits. This number is then the final number which is used to control the programmable divider number one. **(B)** shows the same divider calculation. except for **USB** instead of **LSB**. The divider then controls vco 1 in 10 kHz steps.

two subcycles. During thé first subcycle, the chain is loaded from and divides by the number programmed in a diode **ROM** (i-f offset data for the sideband in use). For the second subcycie, the chain is loaded with and divides by the data presented to the BCD frequency-select input lines. After both subcycles are completed, one pulse is sent to the phase comparator, and the entire cycle repeats. This approach may also be applied to single-divider PLL systems.

The reference frequencies are derived by dividing a 9.99-MHz crystal oscillator's output by 1000 and 999 to obtain outputs at 9.99 kHz and 10.00 kHz, respectively. An oven is recommended if the synthesizer's 10-Hz resolution is to be used to full advantage.

Vco 1 and vco 2 are combined in a mixer, which supplies the difference-frequency output through a 50-MHz lowpass filter. The mixer is image terminated through a highpass filter to reduce any unwanted intermodulation products which would result from an impedance mismatch at the vco sum frequencies.

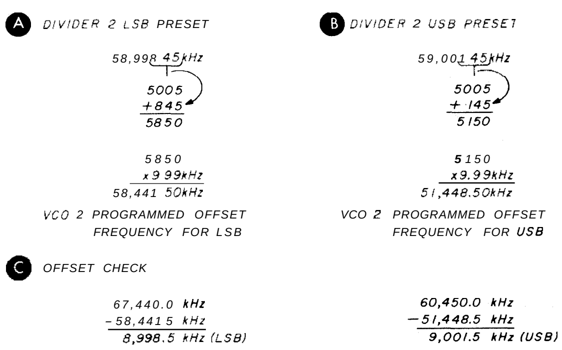


fig. 5. For vco 2, the same initial frequency, as determined in fig. 3, is preset into the counters, plus the three most significant digits from the data switches. The final divider number, times 9.99 **kHz,** produces the required vco frequency, except now in 9.99 kHz steps. As a check, subtracting the value of the two vco frequencies will equal the sideband offset value **(C).**

Both filters must employ T-input sections so they appear invisible to the mixer in their cutoff regions.

Because the high-reference frequencies used in this synthesizer allow loop correction of vco microphonics below about 1 kHz, this circuit is ideal for mobile applications where vibration would otherwise require the use of extremely rugged oscillator enclosures.

This approach may be expanded to yield 1 Hz steps, or to utilize a 100 kHz reference, but either change would require a ten times greater frequency

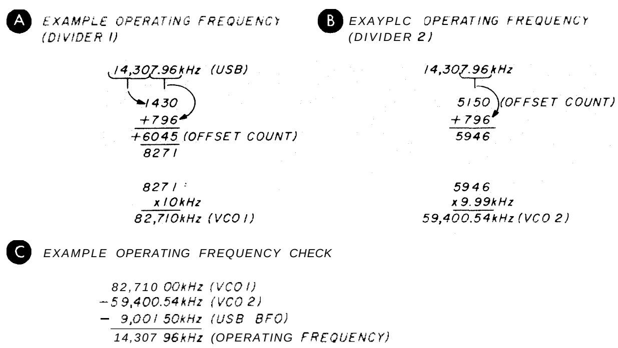


fig. 6. With a sample operating frequency of **14,307.96** kHz **(USB),** the programmable divider, for vco 1, would have a value of 8271. The offset is that calculated in fig. **4B.** This final value determines the frequency of vco 1. The programmable divider for vco 2 uses the input from the data switches plus the offset from fig. 5B to determine the vco's frequency. Since this synthesizer is actually a local oscillator in a transmitter or receiver, its output will differ from the actual transmitted (or receiver) frequency by the value of the i-f frequency. Therefore, the difference between vco 1 and vco 2 will equal the operating frequency plus the i-f offset. Or, subtracting both the i-f offset and vco 2's frequency from vco 1, will give the actual operating frequency **(C).**

range of vco 2, and that much additional range added to the upper frequency limit of vco 1.

A recommended frequency programming scheme would incorporate thumbwheel switches to set the MHz range, with smaller frequency divisions being continuously tunable by an optically coupled, tuning-knob-controlled, up-down counter with readout. The high speed of this synthesizer makes it an ideal candidate for use in an automated, microprocessor controlled station, in which case tuning could be implemented with the computer's ASCII keyboard, or even remote controlled via a modem. Wouldn't it be nice to have a remote 20-meter transceiver atop a 150-meter (500-foot) building?

I would like to thank Fred Studenberg, W4CK, and Chuck Jackson for their technical advice and constructive criticism of this article.

ham radio

repair bench

Bob Stein, W6NBI

automatic noise-figure measurements — fact and fancy

One of the major attractions at the regional vhf/uhf conferences is the noise figure measurement session. Everyone with a new converter or preamp, and especially anyone who does not possess or have access to calibrated noise-figure measuring equipment, anxiously awaits the measurement results. The session becomes a contest of sorts, in which you find out how your latest endeavor or purchase compares with the best of show.

Invariably these measurements are made using automatic noise-figure equipment because of the simplicity and speed with which measurements may be accomplished. Given any one test setup, the comparisons among the various units under test may be valid, in that the relative noise figures for each unit can be determined, as can the difference in noise figure between any of the units. But what about a preamp whose measured noise figure was 1.5 dB at the West Coast Conference and another which yielded 1.7 dB at the East Coast Conference? Is this a valid comparison, or going one step further, is either measurement really accurate? The discussion which follows will, I hope, provide these answers.

In addition, this article becomes another in the series devoted to the use and application of available surplus test equipment. Automatic noise-figure meters, such as the Hewlett-Packard models 340A, 340B, and 342A, and the AILTECH* types 74A and 75, are typical of those in general use. Noise sources which are usable from 10 MHz to 5 GHz, and which may be used with one or more of these instruments, are the Hewlett-Packard models 343A, 345B, and 349A, and the AIL types 7006,7010, 7011, and 7012. **Tables 1** and **2** list the pertinent specifications for the noise-figure meters and noise sources, respectively.

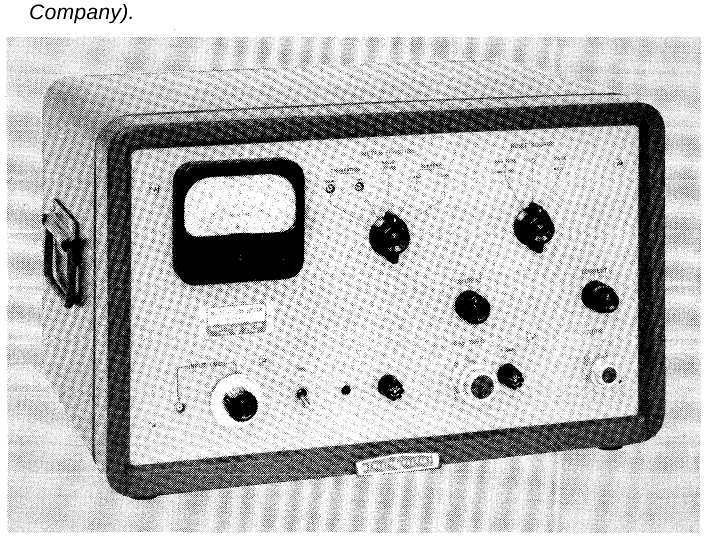
With the exception of the AIL 75, which is probably too new to show up on the surplus market, all of these noise-figure meters and noise sources are available from time to time. Oftentimes a solid-state noise source also becomes available. If it is one of the AIL series 76 it is [illegible] the AIL 75 noise-figure meter; if it is manufactured [illegible] it may be adaptable to one of the aforementioned noise-figure meters, but detailed information must be obtained from the manufacturer. However, because of their relative scarcity and the fact that they may not be truly compatible with most of the available noise-figure meters, no coverage can be provided in this article for any solid-state noise source.

As noted in **table 1,** many of the Hewlett-Packard instruments have beer! specia! versions which accept one or more input frequencies different from those in the standard models. This should present no problem to the potential user, even if none of the instrument input frequencies corresponds to the output frequency from his converter. Either the converter output can be heterodyned to the noise-figure meter input frequency, as explained later, or the noise-figure meter can be modified. This modification involves nothing more than retuning or replacing coils in the instrument's tuned amplifier, and is described in the appendix.

noise-figure equations

Before we can analyze the operation of automatic

The Hewlett-Packard model 340B Noise Figure Meter can be used with either a gas-discharge noise source or a temperature-limited diode source. Models 340A and 342A are similar in appearance *(courtesy Hewlett-Packard Company).*



*AILTECH, a Cutler-hammer Company, was formerly Airborne instruments Laboratory. Throughout this article, AIL will be used to identify equipment manufactured under either name.

noise-figure meters, it is necessary to understand the mathematical definition of noise figure insofar as it applies to the noise-figure meter. I will skip the basic mathematical derivation, since this has been covered extensively in references 1 through 5. Instead, I will start with a mathematical definition of noise figure and proceed to derive the relationships which permit the automatic noise-figure meter to function.

To eliminate any confusion between noise figure expressed as a numerical ratio and the logarithmic equivalent expressed in dB, I will limit the use of the term *noise figure* (NF) to the logarithmic version and call the numerical ratio *noise factor* (F).

The noise factor of a receiver is defined as the input signal-to-noise power ratio divided by the output signal-to-noise power ratio, and is expressed as

$$F = \frac{S_i/N_i}{S_o/N_o} \quad (1)$$

where S_i is the input signal power, N_i is the noise power at the input, S_o is the output signal power, and N_o is the output noise power.

If a broadband noise source is used at the receiver input

$$S_i/N_i = EN = \frac{T_2 - T_o}{T_o} \quad (2)$$

where

EN = excess noise power (generally expressed in dB, but the equivalent numeric ratio in this case)

T_2 = equivalent absolute temperature of noise source when on, in °K

T_o = °K

If we designate the output power from the receiver when the noise source is off as N_1, and the receiver output power when the noise source is on as N_2, the output signal power, S_o, can be expressed as

$$S_o = N_2 - N_1 \quad (3)$$

and since

$$N_o = N_1 \quad (4)$$

$$S_o/N_o = \frac{N_2 - N_1}{N_1} = \frac{N_2}{N_1} - 1 \quad (5)$$

Rewriting **eq. 1** by substituting **eqs.** 2 and **5,** we get

$$F = \frac{\dfrac{T_2 - T_o}{T_o}}{\dfrac{N_2}{N_1} - 1} \quad (6)$$

Converting **eq. 6** to the equivalent noise figure, where $NF = 10 \log F$,

$$NF = \left[10 \log \frac{T_2 - T_o}{T_o} \right] - \left[10 \log \left(\frac{N_2}{N_1} - 1 \right) \right] \quad (7)$$

Since the first term in **eq. 7** is the logarithmic equivalent of **eq.** 2, it is equal to the excess noise ratio (ENR), in dB, of the noise source. Therefore **ea. 7** can be rewritten

$$NF = ENR - 10 \log \left(\frac{N_2}{N_1} - 1 \right) \quad (8)$$

The ratio N_2/N_1 is often designated the Y-factor, so that an equivalent expression for **eq. 8** is

$$NF = ENR - 10 \log (Y - 1) \quad (9)$$

Thus we have arrived at an expression for noise figure in which the only variable is Y when a noise

table **1.** Automatic noise-figure meter specifications.

model	NF ranges and accuracy	input frequencies (MHz)	bandwidth (MHz)	sensitivity (dBm)	agc range (dB)
Hewlett-Packard 340A, B	0-15 dB: ±0.5 dB 3-30 dB: ±0.5 dB from 10-25 dB ±1.0 dB from 3-10 dB ±1.0 dB from 25-30 dB	*30, 60	1 (min.)	−60	50
Hewlett-Packard 342A	Same as 340A, B	*30, 60, 75, 105,200	1 (min.)	−60	50
AIL 74A	0-25 dB: ±0.5 dB 23-36 dB: ±1.0 dB	**One of the following:			65
		10, tunable 20-60	2	−67	
		30	6	−73	
		30, tunable 40-180	2	−67	
		60	6	−67	
AIL 75	0-15 dB: ±0.15 dB from 0-3 dB ±0.25 dB from 3-6 dB 3-18dB: ±0.15 dB from 3-6 dB ±0.25 dB from 6-9 dB 6-21 dB: ±0.15 dB from 6-9 dB ±0.25 dB from 9-12 dB 12-27 dB: ±0.5 dB from 12-18 dB 18-33 dB: ±1.0 dB	30 (other options available)	5 (min.)	−73	65

*Standard input frequencies are listed; many special models were manufactured with one or more different input frequencies.
**Depends on i-f amplifier installed in noise-figure meter.

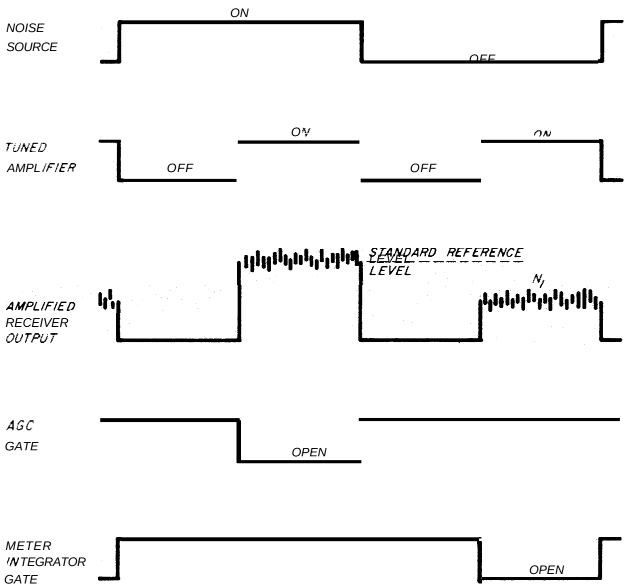


fig. 2. Operational effect of the switching circuit in the automatic noise-figure meter.

source of known ENR is used. Furthermore, since Y is the ratio of N_2/N_1, we need only measure this ratio, without regard to absolute power levels, in order to obtain the receiver noise figure.

automatic noise-figure meters

Now that we have a simplified noise-figure equation, let us investigate the actual method by which noise figure is measured automatically. Fig. **1** shows a simplified block diagram of a typical Hewlett-Packard noise-figure meter. The external noise source and receiver under test are included in the diagram so as to close the loop.

The noise source, powered by the noise-figure meter, is connected to the antenna input of the receiver under test, and the i-f output from the receiver is connected to the input of the noise-figure meter. A switching circuit in the noise-figure meter gates the tuned amplifier, meter integrator, agc, and noise source (via its power supply) on and off at a low audio-frequency rate. As shown in fig. **2,** the noise source is gated on and approximately one-quarter cycle thereafter the tuned amplifier is gated on and the agc gate is opened. During this "on" time the receiver output, N_2, consists of the amplified power from the noise source plus the amplified receiver noise. This is used to establish a standard reference level through agc action.

After the noise source is gated off, the amplifier is again gated on, as is the meter integrator. During this time the receiver output, N_1, consists only of the amplified receiver and input termination noise, which is detected and integrated, and applied to the meter. Since the noise-source ENR is known, and N_2 is always amplified to a standard reference level, the only variable in **eq. 8** is the metered value of N_1. The meter can therefore be calibrated to display noise figure directly, as a function of N_1.

The tuned amplifier is gated on for only one-half of the noise-source on and off periods in order to be certain that the output from the noise source has reached its maximum level or has fallen to its quiescent level.

The AIL noise-figure meters function in a similar manner, although there are differences in the actual circuits used to achieve the resulting indication.

noise sources

Although both coaxial and waveguide noise sources are available for use with automatic noise-figure meters, this discussion will be limited to the coaxial types, since waveguides presently have limited amateur use.

The Hewlett-Packard model 343A VHF Noise Source provides a wideband noise spectrum from 10 to 600 MHz and has a source impedance of 50 ohms. It generates an excess noise ratio of 5.2 to 6.6 dB, depending on the measurement frequency, at its specified current of 3.31 mA; refer to **table 2.** The noise source employs a special temperature-limited

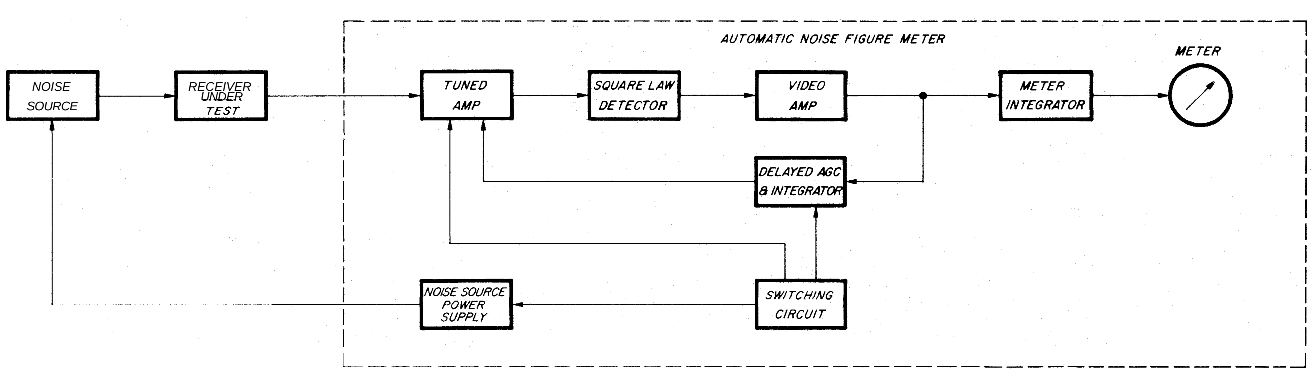


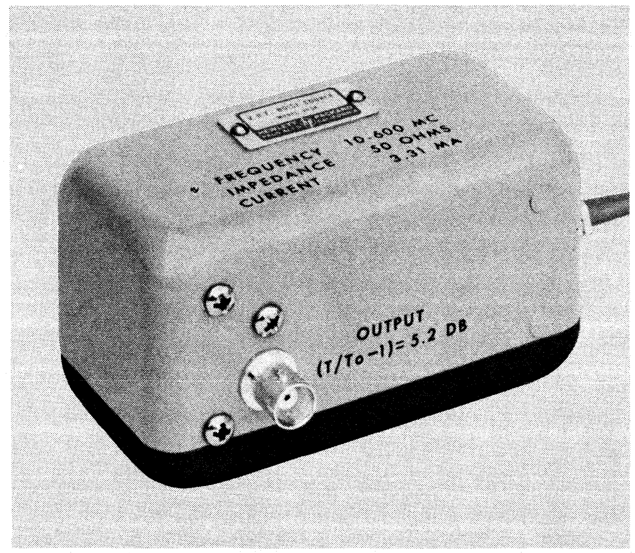
fig. 1. Simplified block diagram of the Hewlett-Packard model 340A or 340B Noise Figure Meter.

vacuum-tube diode (similar to the Sylvania 5722) whose plate and filament voltages are supplied by the Hewlett-Packard 340A, 340B, or 342A noise-figure meter. The noise output of a saturated temperature-limited diode is a function of plate current, which is set to an established value by means of a control on the noise-figure meter.

The Hewlett-Packard model 345B IF Noise Source produces an excess noise ratio of 5.2 dB from any one of four switch-selected source impedances: 50, 100, 200, or 400 ohms. Since tuned circuits within the noise source limit its spectrum to either 30 or 60 MHz, also switch-selected, it has limited amateur application and can be dismissed without further discussion.

Also usable with any of the Hewlett-Packard noise-figure meters are several 5722 noise generators whose construction has been described in references 6 through 8. It should be noted that most, if not all, of these homebrew noise sources suffer from impedance mismatch problems when used above 400 MHz. However, it is possible to build a noise source which compares favorably with the Hewlett-Packard 343A to at least 450 MHz; its construction will be described in a future article.

For receiver frequencies above 450 MHz, a noise source using a gas-discharge tube is generally used. Coaxial noise sources of this type may also be used at frequencies as low as 200 MHz. The Hewlett-Packard model 349A UHF Noise Source and the AIL Type 7010, 7011, and 7012 Coaxial Noise Generators all employ an argon-filled gas-discharge tube coaxially mounted in a helical transmission line. These noise



The Hewlett-Packard model 343A VHF Noise Source utilizes a temperature-limited diode, and can be used from **10** to ***600*** **MHz** (*courtesy Hewlett-Packard Company*).

sources produce excess noise ratios of 15.2 to 15.7 dB over their specified frequency ranges, as shown in **table** 2.

A simplified diagram of a coaxial gas-discharge tube noise generator is shown in **fig.** 3. The tube is alternately turned on and off by the switched power supply in the noise-figure meter. When a high-voltage pulse is applied between the anode and cathode of the tube, the argon gas ionizes and noise is generated. The noise is coupled into the helix and appears at the two coaxial connectors. One connector is used

table **2.** Coaxial noise-source specifications.

model	frequency range (MHz)	excess noise ratio (dB)	operating current (mA)	output vswr
Hewlett-Packard 343A	10-600	5.2 ± 0.2 from 10-30 MHz 5.5 ± 0.25 at 100 MHz 5.8 ± 0.3 at 200 MHz 6.05 ± 0.3 at 300 MHz 6.3 ± 0.3 at 400 MHz 6.5 ± 0.5 at 500 MHz 6.6 ± 0.5 at 600 MHz	3.31	10-400 MHz: 1.2 max. 400-600 MHz: 1.3 max.
Hewlett-Packard 345B	30 or 60, switch selected	5.2	0.41 to 3.31	50, 100, 200 or 400 ohm (± 4%) source impedance, switch selected
Hewlett-Packard 349A	400-4000 (usable to 200)	14.6 ± 0.6 at 220 MHz 15.6 ± 0.6 from 400-1000 MHz 15.7 ± 0.5 from 1-4 GHz	150	200-2600 MHz: 1.35 max. fired 1.5 max. unfired 2.6-3.0 GHz: 1.5 max. fired 1.5 max. unfired 3.0-4.0 GHz: 2.0 max. fired 3.0 max. unfired
AIL 7006	10-250	15.2 ± 0.5	33.1	1.2 max.
AIL 7010, 7011	200-2600	15.2 ± 0.3	175	1.15 (nominal) fired 1.3 (nominal) unfired
AIL 7012	2-5 GHz	15.5 ± 0.2	175	1.5 (nominal) fired 2.0 (nominal) unfired

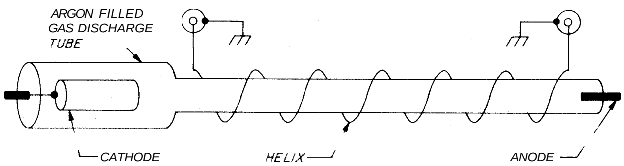


fig. 3. Simplified diagram of a gas-discharge tube noise source. One connector is used for the noise output; the other must be terminated by a precision 50-ohm load.

for the noise output, while the other must be terminated by a precision 50-ohm load.

connecting the noise source to the noise-figure meter

Obviously each manufacturer supplies noise sources which are compatible with his own noise-figure meters. There is nothing, however, other than the problem of physically interconnecting the source and noise-figure meter, which precludes using one manufacturer's gas-discharge source with another's noise-figure meter.

The same interchangeability is not true for diode noise sources. The AIL type 7006 noise diode can be used only with the AIL 74A noise-figure meter, and the Hewlett-Packard diode noise sources can only be used with the Hewlett-Packard noise-figure meters. But, as previously mentioned, there are several 5722 noise generators which can also be used with the Hewlett-Packard noise-figure meters.

All of the Hewlett-Packard noise-figure meters discussed in this article originally included an interconnecting cable for that manufacturer's gas-discharge noise sources. Unfortunately, the cable is invariably missing when the instrument reaches your local surplus emporium. On the other hand, the interconnections are rather simple, so that with the information which follows, you should be able to solve all of your interface problems.

To connect the Hewlett-Packard 349A coaxial (or any of the Hewlett-Packard 347-series waveguide) gas-discharge noise source to any Hewlett-Packard noise-figure meter, a 6-foot (1.8-meter) type 340-16A cable was originally furnished. You can buy a replacement directly from Hewlett-Packard, but the price was $100 the last time I checked. A good alternative is to make your own cable, as shown in **fig. 4.**

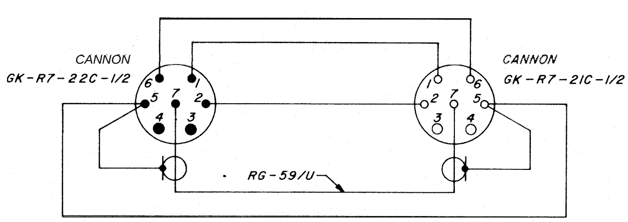
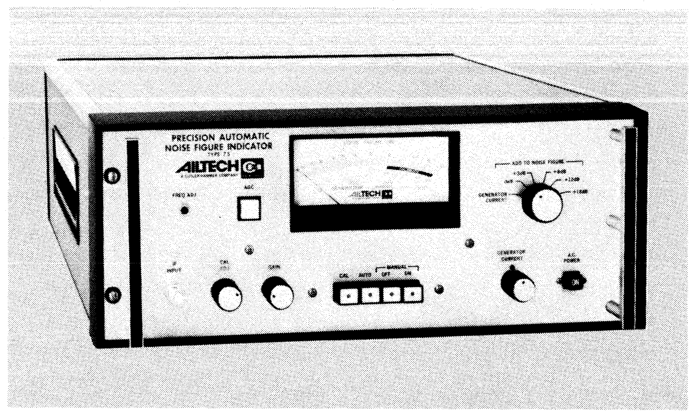


fig. 4. Interconnecting cable for use between a Hewlett-Packard 349A or 347-series noise source and a Hewlett-Packard 340A, 340B, or 342A noise-figure meter. The cable length should be limited to about 2 meters (6 feet).

If you want to connect a Hewlett-Packard gas-discharge source to either the AIL 74A or 75 noise-figure meter, it will be necessary to modify the captive cables on the AIL instrument, as follows."

1. Cut the high- and low-voltage cables at a point at least 20 inches (51 cm) back from the connector ends.



The AILTECH type 75 Precision Automatic Noise Figure Indicator is the latest and most versatile instrument of its kind on the market. It can be supplied for use with both gas-discharge and solid-state noise sources, or for use with the latter type only *(courtesy* AILTECH).

2. Wire the cables from the noise-figure meter as shown in **fig. 5A.**

3. Wire the cable ends with the connectors attached as shown in **fig. 5B.**

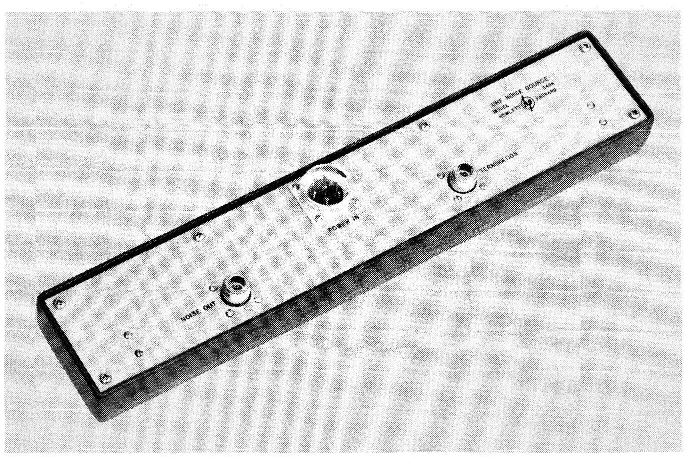
The instrument cables, now terminated in a single multipin connector, will plug into the connector on the Hewlett-Packard noise source. An AIL noise source can still be used by connecting the multipin connector on the instrument cables to the mating connector on the now separate cable assembly shown in **fig. 5B,** thereby restoring the original configuration.

Fig. 6 shows the cabling needed to connect an AIL 7010, 7011, or 7012 noise source to any of the Hewlett-Packard noise-figure meters. The high-voltage connector required to mate with the anode connector on the AIL 7010 or 7012 is not available except as a part of the AIL 7003 cable set. Therefore it must be fabricated by using brass tubing of the appropriate size and suitable high-voltage insulating material

"Taken from AIL Application Engineering Bulletin 70-15, dated August, 1961.

(teflon, polystyrene, etc.) between the inner high-voltage and outer ground conductors. WARNING: the anode lead and connector must be insulated for up to 5 kV; USE EXTREME CARE.

The Hewlett-Packard 343A and 345B noise sources each have a captive cable and connector to mate with the Hewlett-Packard 340B and 342A noise-figure meters. If you intend to use any of the previously referenced 5722 noise generators, the appropriate connector wiring is shown in **fig.** 7. The 5722 filament draws between 1.5 and 2 amperes, so



The Hewlett-Packard model 349A UHF Noise Source is a gas-discharge type using an argon-filled tube. Its usable range is 200 MHz to 4 GHz *(courtesy Hewlett-Packard Company).*

be sure to use wire that is large enough to minimize the voltage drop between the tube and the noise-figure meter.

Connecting any of the commercial or homebuilt diode noise sources to the Hewlett-Packard 340A noise-figure meter requires somewhat more discussion. The *diode* panel connector on the 340A was originally a 3-pin connector, rather than the 5-pin Cannon WK-5-31S connector used on the later 340B and 342A models. (It was intended to mate with the long obsolete 345A noise source.) However, Hewlett-Packard Service Note 340A-1A* described how to replace the old 3-pin connector with the later 5-pin connector, and virtually everyone who had a 340A made the change. In fact, I have never seen an instrument with the old connector still in place.

If a new connector has been installed and you are going to use a homebuilt 5722-type noise source, the wiring information in **fig. 7** applies. If the old connector is still on the instrument, you will have to be lucky enough to find a mating connector or else change

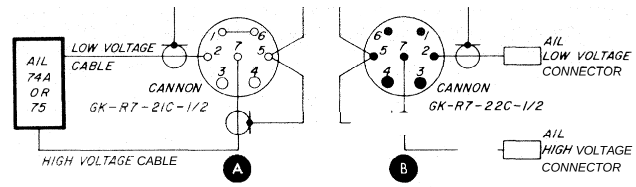


fig. 5. Modification of the AIL 74A or 75 noise-source cables which will permit use of Hewlett-Packard 349A or 347-series noise sources, (After AIL Application Bulletin 70-15, August 1961).

the connector on the instrument. The filament connections on the old 3-pin connector are pins 1 and 2; pin 3 connects to ground,

The Hewlett-Packard 343A noise source can only be used with the 340A if the 5-pin connector has been installed. Before trying to use it, however, check the internal wiring to the connector to be sure that there is a connection to pin 4. If there is no wire on pin 4, run one from that pin to the black center-tap lead of transformer T101. (The transformer may be identified by tracing the leads from pins 1 and 2 of the connector, which are wired to the green and yellow leads, respectively, of T101.)

measurement techniques

We can now discuss the procedures and techniques of measuring noise figure with an automatic noise-figure meter. Because the actual operating procedures vary, depending on the instrument being used, no attempt will be made to describe the detailed operational steps. These appear in the instruction manual for the specific instrument, available from the manufacturer at nominal cost. In general, operation of the noise-figure meter consists of the following steps:

1. Connecting the receiver under test as shown in **fig. 8.** For the purposes of this discussion, the term "receiver" means any receiver or portion thereof (such as a converter, mixer, etc.) which provides an output signal at the input frequency of the noise-figure meter.

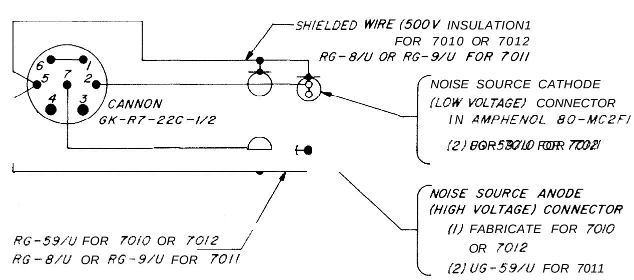


fig. 6. Interconnecting cable for use between an AIL 7010. 7011, or 7012 noise source and a Hewlett-Packard 340A, 340B, or 342A noise-figure meter.

*Available from Hewlett-Packard Company, 1820 Embarcadero Road, Palo Alto. California 94303.

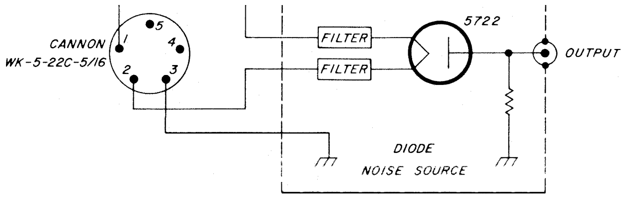


fig. 7. Wiring a homebuilt 5722-tube noise source to permit its use with a Hewlett-Packard 340B or 342A noise-figure meter. See the text for additional information applicable to the Hewlett-Packard 340A.

2. Energizing the noise-figure meter and establishing the fixed agc reference level.

3. Setting the noise-source operating current.

4. Calibrating the noise-figure meter to the noise-source *ENR*. (This step is not applicable to Hewlett-Packard noise-figure meters.)

5. Reading the receiver noise figure on the meter, and correcting for noise-source characteristics and/or loss-pad attenuation.

Although this generalized procedure sounds almost too simple to be true, it is just about as easy in practice. Any adjustments on the receiver can be made while it is under test by merely tuning for the best noise figure. Compare that to measuring noise figure by the manual twice-power method!

Let us now examine **fig. 8** block by block. Most important, there should be a direct connection between the noise-source output connector and the loss pad, and between the pad and the receiver. This means no *cables,* and a minimum of adapters.

The use of a loss pad between the noise source and the receiver is an absolute necessity when using a gas-discharge source, and is highly recommended for diode noise sources. One of the disadvantages of gas-discharge noise sources is that they require a high-voltage ionizing pulse, typically several kilovolts. Because of capacitive coupling between the tube and the helix, this pulse appears as a spike in the noise output. Although the pulse is attenuated, the amplitude of the spike may still be several volts, which may be enough to destroy your expensive low-noise transistor. Therefore, a 6- to 10-dB pad must always be used with a solid-state receiver to attenuate the spike. Even with a 10-dB pad, there may be sufficient spike voltage to destroy certain GaAs fets. A solid-state noise source should be used to check devices of this type.

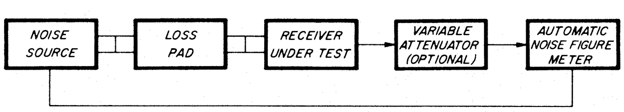


fig. 8. Measuring noise figure with the automatic noise-figure meter. The use of the loss pad and variable attenuator are discussed in the text.

The use of a loss pad with a diode noise source is not absoiuteiy essentiai, but can minimize severai problems. First of all, most receivers (remember the general use of this term) do not present a 50-ohm input when optimized for a noise match. Because the rated *ENR* of a noise source is based on a 50-ohm load, there will be an indeterminate mismatch loss if the receiver vswr is greater than 1.0:1. A 3- to 6-dB pad will not eliminate the mismatch loss, but may reduce it somewhat.

A second reason for using a loss pad is to ensure a 50-ohm source impedance for the receiver, since the vswr of the noise source also is not a perfect 1.0:1.

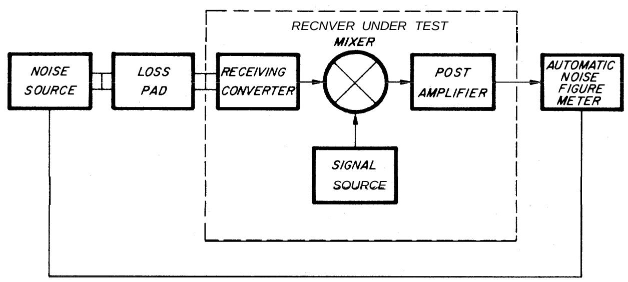


fig. 9. Using an external mixer and signal source to convert the receiver output frequency to that of the noise-figure meter input. The "receiver under test" corresponds to the equivalent block in fig. 8.

Any tendency of the receiver to "take off" when looking into an impedance of other than 50 ohms will be reduced by use of a pad.

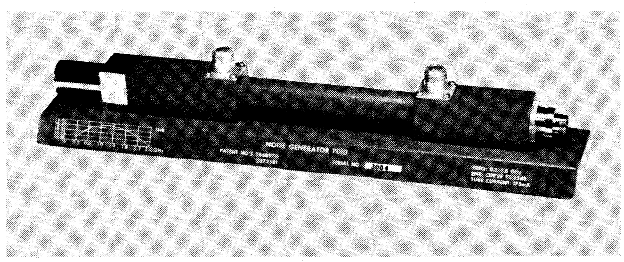
Both of these reasons apply equally to gas-discharge noise sources, but are automatically realized by the absolute necessity of using a loss pad to protect the receiver. In either case, the attenuation of the pad must be known to a fair degree of accuracy; remember that the loss, in dB, must be subtracted from the reading obtained on the noise-figure meter, since any loss ahead of a receiver adds algebraically to the receiver noise figure.

In addition to providing an output at a frequency which is compatible with the noise-figure meter, the receiver must also provide enough gain to amplify the noise input to a level which exceeds the sensitivity of the noise-figure meter. However, the difference in the receiver output levels with the noise source on and off must not exceed the agc range of the instrument and overload it. This means that the output from the receiver, when the noise source is off, shouid be only slightly above the sensitivity level of the noise-figure meter. This can be determined by the use of the variable attenuator shown in fig. **8.** The amount of attenuation which is introduced should be 6 to 10 dB less than the value which causes the receiver output to fall below the threshold required to establish the reference level.

Conversely, if the receiver has insufficient gain, it will not be possible to establish a signal reference level in the noise-figure meter. In that case, a post-amplifier must be substituted for the variable attenuator to provide sufficient overall gain.

receiver configuration

Thus far we have imposed two conditions which the receiver must satisfy – its gain must be compatible with the sensitivity requirement and agc range of the noise-figure meter, and its output frequency must match that of the noise-figure meter input. We have already discussed the first of these, and can proceed to the second.



The AILTECH type 7010 Noise Generator covers the range from 200 MHz to 2.6 GHz. The similar appearing type 7012 is used between 2 and 5 GHz. Both noise sources utilize gas-discharge tubes *(courtesy* **AIL TECH).**

The typical vhf or uhf receiving converter provides an output which is applied to a communications receiver used as a tunable i-f amplifier, detector, and audio amplifier. This output can fall anywhere in the entire hf, vhf, or even uhf spectrum. On the other hand, the input to the noise-figure meter must be at a specific frequency which may not correspond to the output from the converter. The combination of a 30-MHz amplifier in most of the noise-figure meters and most receiving converters with a nominal 28-MHz output has sufficient bandwidth to permit meaningful measurements.

For other converter output frequencies which do not correspond to the noise-figure meter input, an external mixer and signal source can be used to convert the output frequency of the receiving converter to the input frequency of the noise-figure meter. Fig. **9** shows the connections for such a frequency conversion. Any type of mixer can be used, although a packaged double-balanced mixer with coaxial connectors will prove to be the most convenient and versatile. These are available commercially;* a less expensive mixer can be packaged as described by Paul Shuch.[10]

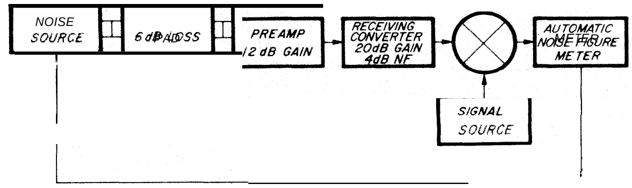


fig. 10. Test setup to determine receiving system and preamplifier noise figures, using a receiving converter following the preamplifier.

The signal source functions as a local oscillator for the mixer, and may be anything which generates a relatively stable unmodulated output, such as a signal generator or vfo. The sum of the converter and signal-source frequencies, or the difference between them, must result in the input frequency required by the noise-figure meter. It makes no difference whether the sum or difference frequency is used; however, the signal source must not be set to a frequency which is close to or is a subharmonic of the noise-figure input frequency.

For example, if the receiving converter output is 14 MHz and the noise-figure meter input is 30 MHz, a signal-source frequency of either 16 or 44 MHz would provide the proper output frequency from the mixer. In this case, the 16-MHz choice should be avoided because its second harmonic at 32 MHz may appear at the mixer output and fall within the passband of the noise-figure meter. This, of course, will render any measurements meaningless, since the noise-figure meter cannot distinguish between the receiver signal and the spurious mixer output.

If a signal generator is used as the signal source, it can function indirectly as the variable attenuator shown in fig. **8.** Varying the signal-generator output will vary the conversion loss of the mixer and permit

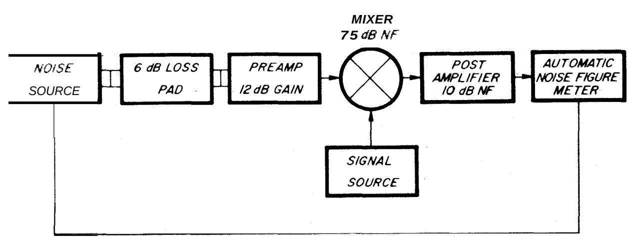


fig. 11. Test setup to determine receiving system and preamplifier noise figures, using a mixer following the preamplifier. In this case, a poor post-amplifier contributes significantly to the overall noise figure.

*Available from Mini-Circuits Laboratory, Merrimac Industries, Anzac Electronics, Hewlett-Packard, Cimmarron, and others.

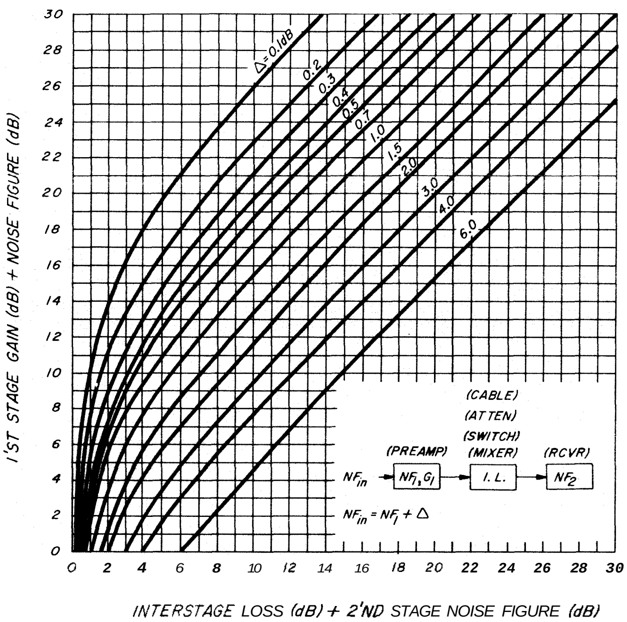


fig. 12. The system noise figure (NF_{IN}) can be found from these curves when the first- and second-stage noise figures, first-stage gain, and interstage losses are known *(courtesy Daico Industries, Inc.)*.

control of the mixer output level. A post-amplifier has been shown in **fig. 9** because the receiving converter gain, less the mixer conversion loss, will generally be insufficient to provide the necessary signal level into the noise-figure meter. This post-amplifier can be any type, such as a surplus i-f strip or a broadband amplifier, which will amplify the desired mixer output frequency. The mixer image frequency will be rejected by the tuned amplifier in the noise-figure meter.

The configuration shown in **fig. 9** will result in valid noise figures only if the gain of the receiving converter is 20 dB or more. If the converter gain is less than 20 dB, the indicated noise figure will be worse than the actual noise figure because of the noise contribution of the external mixer. The actual noise figure must then be calculated, as explained later.

preamplifiers

In order to measure the noise figure of a preamplifier, the test setup of either **fig. 8** or **fig. 9** is used, with the preamp inserted between the loss pad and the receiver. To prevent any possible instability of either the preamp or the converter, in the event that either is not unconditionally stable, it is wise to use an additional loss pad between the preamp and the converter, directly at the converter input. This pad should provide about 3 dB loss; the exact value is not critical nor need it be known accurately. Its effect will be incorporated into the measured noise figure of the receiver (without the preamp), since this measurement is necessary to *calculate* the true noise figure of the preamp, as explained in the following section. If only an overall system noise figure is of interest, then the second loss pad should not be used.

A major consideration in the measurement of preamp noise figure is that of image response and the effect of filters on that response. It is mentioned here as a preliminary precaution only, and will be covered in detail under the heading *image response error.*

cascaded stages

The overall noise figure of any receiver depends primarily on the noise figure of the first stage. However, the noise generated by succeeding stages will contribute to the overall noise figure and, if the gain of the first stage is not over 20 dB or so, have a significant effect. Since low-noise preamps seldom have gains in excess of 15 dB, taking the noise contribution of the following stages into account is of major importance in determining the preamp noise figure.

The general equation for the noise *factor* of networks in cascade is

$$F_s = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots \qquad \textbf{(10)}$$

where

F_s = system noise figure

F_1, F_2, F_3 = noise factor of first, second, third stages

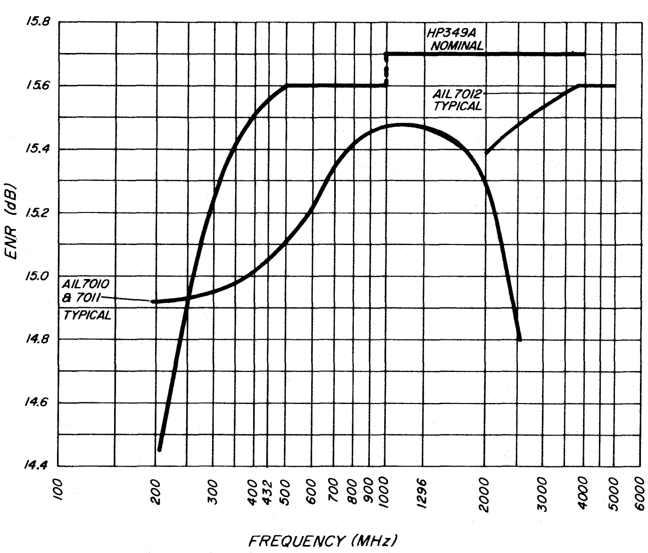


fig. 13. Excess noise ratios for Hewlett-Packard and AIL coaxial noise sources which utilize gas-discharge tubes. The curve for the Hewlett-Packard 349A is subject to the *ENR* tolerances listed in table 2; those for the AIL noise sources are typical and take into account the gas-tube-variations and hot-cold insertion losses.

G_1, G_2 = power-gain ratio of first, second stages

The system noise figure, NF_s, is derived from the expression

$$NF_s = 10 \log F_s \tag{11}$$

It can be seen from **eq. 10** that the numerical value of the second and succeeding terms is a function of the first-stage gain; the higher the gain, the lower will be the values of those terms. If all stages are active amplifiers, only the second-stage noise will be of importance, and only the first two terms of **eq. 10** need be used in the calculation. However, if any of the early stages in the receiver has a gain of 1 or less (*e.g.*, mixers and loss pads), three or even four terms of the equation may have to be taken into account.

As an example, refer to the test setup shown in **fig. 10.** Assume that the preamp gain (A_1) and the converter gain (A_2) have been measured at 12 dB and 20 dB respectively; the measured converter noise figure (NF_2) is 4 dB; and the measured mixer noise figure (NF_3) is 7.5 dB. The system noise figure indicated on the noise-figure meter is 8.65 dB. What is the preamp noise figure (NF_1)?

First, we must subtract 6 dB from the noise-figure meter reading, since that is the increase in noise figure caused by the 6-dB pad. Therefore the actual system noise figure (NF_s) is 2.65 dB. Second, we must convert the noise figures and gains (in dB) to noise factors and numerical gain ratios, as follows:

$$F_s = antilog \frac{NF_s}{10} = antilog \frac{2.65}{10} = 1.841$$

$$F_2 = antilog \frac{NF_2}{10} = antilog \frac{4}{10} = 2.512$$

$$F_3 = antilog \frac{NF_3}{10} = antilog \frac{7.5}{10} - 5.623$$

$$G_1 = antilog \frac{A_1}{10} = antilog \frac{12}{10} = 15.849$$

$$G_2 = antilog \frac{A_2}{10} = antilog \frac{20}{10} = 100$$

By rearranging **eqs. 10** and **11,** and substituting the above values,

$$NF_1 = 10 \log \left(F_s - \frac{F_2 - 1}{G_1} - \frac{F_3 - 1}{G_1 G_2} \right)$$

$$= 10 \log \left(1.841 - \frac{2.512 - 1}{15.849} - \frac{5.623 - 1}{15.849 \times 100} \right)$$

$$= 10 \log (1.841 - .095 - .003)$$

$$= 10 \log 1.743 = 2.41 \text{ dB} \tag{12}$$

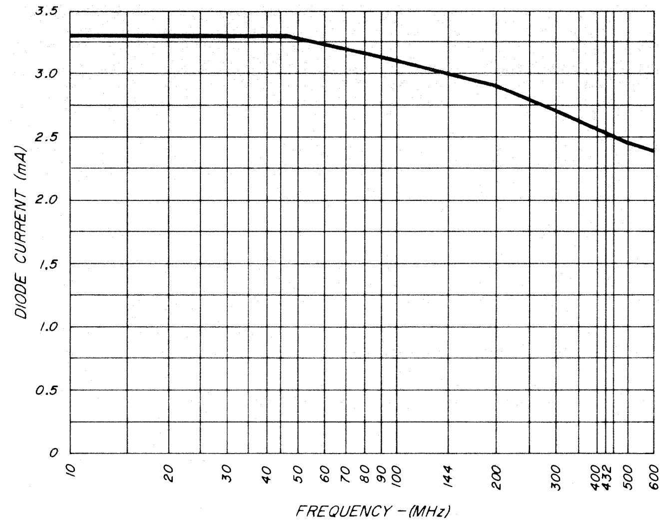


fig. 14. Diode current required by the Hewlett-Packard 343A VHF Noise Source to maintain a nominally constant 5.2-dB excess noise ratio ***(curve*** **by J. R.** *Reisert, W1JR***).**

It can be seen from inspection of the above example that the third term is insignificant and can be ignored. However, suppose that the configuration shown in **fig. 11** is being used to check out the same preamp, using the same mixer and a post-amplifier with a measured noise figure of 10 dB. Now the mixer is the second stage, so that $F_2 = 5.623$. Since, in a stage having a gain of less than 1, the noise figure is equal to the loss, the converse must be true. Therefore, the mixer loss ratio is also 5.623, making the gain (G_2) the reciprocal of 5.623, or 0.178. The post-amp noise figure of 10 dB results in F_3 being equal to 10 (the antilog of 10 divided by 10). In this case, the receiver noise figure reading on the noise-figure meter is 13.2 dB. Subtracting 6 dB for the loss pad, we obtain an actual noise figure (NF_s) of 7.2 dB, or a noise factor (F_s) of 5.248.

Again substituting the measured noise factors and gain ratios in **eq. 12,** we arrive at the following:

$$NF_1 = 10 \log \left(5.248 - \frac{5.623 - 1}{15.849} - \frac{10 - 1}{15.849 \times .178} \right)$$

$$= 10 \log (5.248 - ,292 - 3.190)$$

$$= 10 \log 1.766 = 2.47 \text{ dB}$$

This is approximately the same noise figure as was calculated in the previous example, taking into account rounded-off numbers. Note that in this case the third term of the equation contributes significantly to the overall system noise figure, which leads to the conclusion that a low-noise post-amplifier should always follow a mixer when there is only one stage of amplification ahead of the mixer. If the post-amplifier noise figure were reduced to 1.0 dB, the system

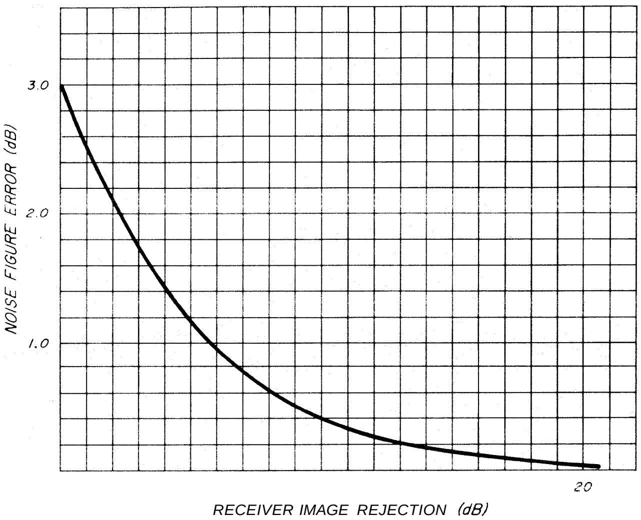


fig. 15. Noise-figure error as a function of receiver image rejection.

noise figure would improve from 7.2 to 3.3 dB. To paraphrase many textbooks, the calculation of this improvement is left as an exercise to the reader.

An interesting graphical representation of the relationships expressed in **eq. 12** is shown in **fig. 12.** If the various stage noise figures, gains, and losses are known in dB, the difference between the system and first-stage noise figures can be read as 4 directly from the curves, which is added to the known first-stage noise figure to obtain the system noise figure.

For instance, let's use the same preamp and mixer, and the l-dB post-amplifier from the preceding example. The first-stage gain plus noise figure is 14.4 dB, and the interstage (mixer) loss plus second-stage (post-amplifier) gain is 8.5 dB. The curves of **fig. 12** show that Ais 0.9 dB at the intersection of these two values. Thus, the system noise figure is equal to the preamp noise figure plus 0.9 dB, or 3.3 dB.

noise-source ENR corrections

In our discussion of the noise-figure equations which govern the operation of automatic noise-figure meters, we estabiished that the only variable in **eq. 8** was the ratio N_2/N_1, and that the ENR term in that equation was a fixed value determined by the noise source. It follows, therefore, that the noise-figure meter calibration must be based on a known or predetermined ENR.

In the case of the AIL noise-figure meters, the operating procedures entail calibrating the instrument to the ENR of the noise source. For the AIL 7006 diode noise source used with the AIL 74A noise-figure meter, the ENR is 15.2 ± 0.5 dB when the diode current is set to 33.1 mA. If a gas-discharge noise source is to be used with either the AIL 74A or 75 noise-figure meter, the instrument is calibrated for the ENR of the source at the frequency of measurement. **Fig. 13** shows typical values for the AIL noise sources and the nominal ENR for the Hewlett-Packard 349A for frequenciesbetween 200 and 5000 MHz.

The meter scales of the Hewlett-Packard 340A, 340B, and 342A noise-figure meters are based on a fixed ENR of 5.2 dB from a diode noise source and a fixed ENR of 15.2 dB from a gas-discharge noise source. There is no adjustment for other values of ENR; if the noise-source ENR is different from the value upon which the calibration is based, a correction factor must be applied to the noise-figure reading. If the ENR exceeds the noise-figure meter calibration design value, the "excess" ENR must be added to the noise-figure readings. If the ENR is less than the calibration design value, the difference be-

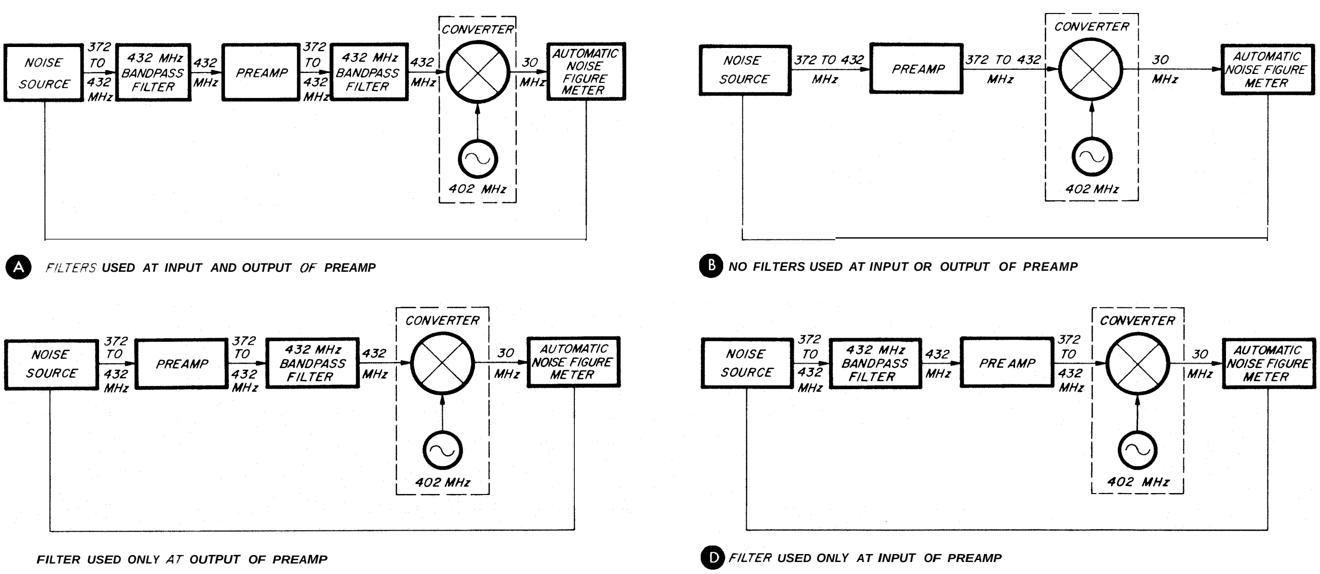


fig. 16. Simplified test setups for noise-figure measurements, using an untuned pre-amp and bandpass filters. The positions of the filters in each test setup are different. All configurations, except that at D, yield valid system noise figures.

tween the two values must be subtracted from the noise-figure readings.

An examination of **table** 2 reveals that the only noise sources which match either of the Hewlett-Packard design values are the Hewlett-Packard 343A, only when used below 30 MHz, and the Hewlett-Packard 345B, which has little amateur application. This means that whenever the 343A is used above 30 MHz, the indicated noise figure must be increased by the difference between 5.2 dB and the noise-source *ENR* at the frequency of use. If the measurements were being made at 400 MHz, for instance, the indicated noise figure would be too low, requiring a corrective addition of 1.1 dB to the meter reading.

Fortunately, in the case of the 343A, there is a simpler method which allows use of the noise-figure meter readings without correction. This involves reducing the diode current from its specified 3.31-mA value to one which maintains a nominally constant 5.2-dB *ENR*. **Fig. 14** is a plot of the diode current versus frequency, which provides this compensation.

Varying the current in a gas-discharge noise source has little effect on its *ENR*, typically only 0.005 dB/mA. Therefore, the Hewlett-Packard noise-figure meter readings must be corrected if the noise-source *ENR* is not 15.2 dB at the frequency of measurement. See **fig. 13** to determine the difference between the *ENR* and 15.2 dB; remember that the difference is added to the noise-figure readings if the *ENR* exceeds 15.2 dB, and is subtracted if the *ENR* is less than 15.2 dB.

If measurements are being made with a Hewlett-Packard noise-figure meter at one or only a few frequencies, as is the usual case for amateur receivers, the need for making corrections to the noise-figure readings can be eliminated by the use of a selected loss pad. The graduations on the two scales of the Hewlett-Packard instrument meters correspond, differing by exactly 10 dB. Therefore, if the noise-source *ENR* were exactly 15.2 dB, and a precision 10-dB pad were used between it and the receiver under test, the noise figure could be read directly on the meter *diode* scale, which is based on an *ENR* of 5.2 dB. Since the noise-source *ENR* is not 15.2 dB, a pad must be selected so that its loss is algebraically equal to the difference between 5.2 dB and the *ENR* shown in **fig. 13.** Thus for the 349A noise source, the use of a 10.35-dB pad at 432 MHz, or a 9.5-dB pad at 220 MHz, will allow direct readout of noise figure on the *diode* scale of the noise-figure meter.

image-response error

Noise figure, as we have been using the term, is more correctly designated as the single-sideband noise figure. This has nothing to do with ssb as a method of modulation; it refers to the fact that any superheterodyne receiver has both a signal and an image sideband (channel), and that we are interested in measuring the noise figure in only a single (the signal) sideband.

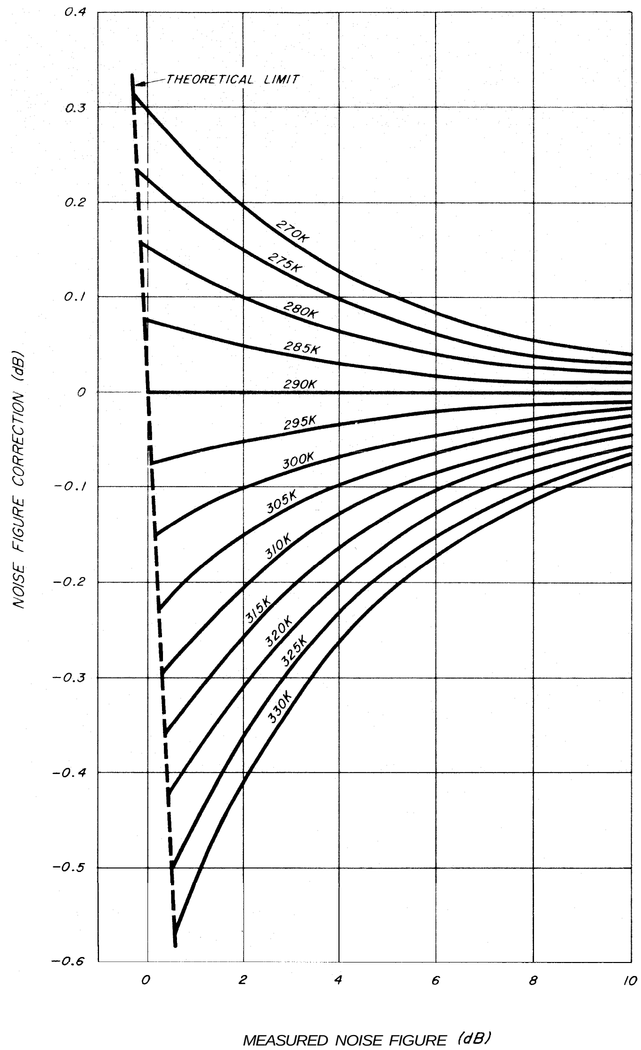


fig. 17. Noise-figure corrections, as a function of termination temperature, for a typical gas-discharge noise source *(courtesy* AIL *TECH).*

The image response of a system consisting of wideband amplifiers followed by a heterodyne receiver (converter) can be an important factor when measuring noise figure. The image-channel signal adds to the signal-channel signal and results in a measured noise figure which is lower than the true noise figure. The noise-figure error is a function of the receiver gain at the signal and image frequencies, as expressed by the following equation:

$$Error_{NF}\,(dB) = 10 \log \left(1 + \frac{G_i}{G_s}\right) \qquad \textbf{(13)}$$

where G_i is the image-channel gain ratio and G_s is the signal-channel gain ratio. It can be seen that, for

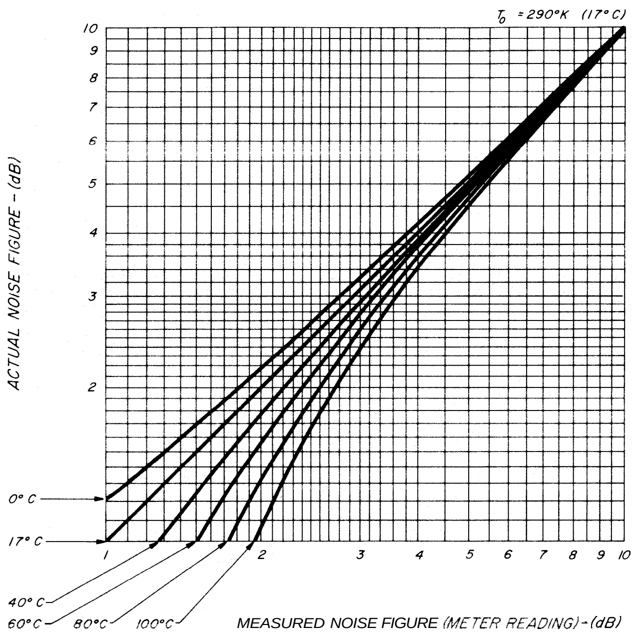


fig. 18. Temperature corrections for the Hewlett-Packard model 343A VHF Noise Source *(courtesy Hewlett-Packard Company).*

equal signal- and image-channel gains, the error will be equal to 10 log 2, or 3 dB.

Fig. 15 is a graphical representation of **eq. 13,** and shows the amount by which the measured noise figure must be increased. It is apparent that the noise-figure error is negligible if the receiver image rejection is greater than 20 dB. However, the corrections, especially when measuring low noise figures, can be significant at image rejections under 20 dB.

Which brings us back to the measurement of low-noise preamp noise figure. The prevailing concept in the design of low-noise amplifiers is to have untuned input and output circuits, and to rely on a separate high-Q filter ahead of the preamp to discriminate against unwanted signals. This is a valid design concept, but may lead to serious errors in the measurement of noise figure, especially at the aforementioned vhf/uhf conferences.

Fig. 16 illustrates, in simplified form, the four possible configurations of measuring preamp noise figure, using a receiving converter and a noise-figure meter with a 30-MHz input. In the examples shown, measurements are to be made at 432 MHz, and both the noise source and the preamp are assumed to be flat from 372 to 432 MHz. The bandpass filters can be of any type — tuned circuits, cavities, stub tuners, etc. — and may actually be part of the input or output circuits of any of the individual blocks. However, the absence of a filter indicates that it does not exist, either separateiy or as a part of any circuit; thus, in **fig. 16B**, the preamp has no filter in either its input or output circuit, and the converter has no filter at its input. The converter output circuit has no effect on this discussion, and is assumed to be tuned to 30 MHz.

In **fig, 16A**, bandpass filters at the input and output of the preamp limit the overall system response to 432 MHz. This is the normal communications configuration, and yields valid noise-figure measurements. In **fig. 16B**, both filters have been eliminated, so that the preamp accepts noise inputs from 372 to 432 MHz and, in addition, contributes its internal noise over the same frequency range. Since the converter produces a 30-MHz output from the noise power at both 372 and 432 MHz equal to $2N_2/2N_1$, the ratio remains the same as from one channel only, resulting in a valid noise figure.

Fig. 16C shows a test configuration in which a filter is used only between the preamp and the converter. It can be seen from the frequencies indicated on that diagram that only the 432-MHz noise signal reaches the converter, so that it too is a valid method of measuring noise figure. (It is equivalent to using an untuned preamp with a normal, tuned-input converter.)

In **fig. 16D**, the filter position has been changed from the output to the input of the preamp. Notice that in this arrangement the noise input to the preamp is limited to 432 MHz. However, this amplified noise plus the internal noise generated by the preamp at *both* 372 and 432 MHz are converted to the 30-MHz input of the noise-figure meter. Therefore, the noise-figure reading will be erroneous. The conclusions to be drawn from this discussion are that meaningful noise figures will be obtained if no filters are used, or if a filter is used following the preamp, but in no case should a filter be used ahead of the preamp without an equivalent filter at the output.

temperature error

In the derivation of the noise-figure equations, T_o is equal, by definition, to 290°K and represents the noise temperature of the noise source when de-energized. It is therefore apparent that if the noise source is at a temperature other than 290°K (17°C or 62.6°F), an error will be introduced. Although such errors are of relatively small magnitude, especially over a limited ambient temperature range, it may be necessary to take them into account when measuring very low noise figures. **Fig. 17** shows the noise-figure correction, as a function of measured noise figure and temperature, for a typical gas-discharge noise source. It can be seen that if a noise-figure measurement of 2.0 dB were made in a room where the iemperature was 80°F (approximately 27°C or 300°K), a correction of −0.1 dB should be made, making the corrected noise figure 1.9 dB. Extended use of a gas-discharge noise source also raises the cold temperature, so that the source should be allowed to cool down to room ambient when accurate measurements are required.

The temperature-correction curves for the Hewlett-Packard 343A diode noise source appear in **fig. 18.** Since the diode heat will cause the termination temperature inside the noise-source housing to be somewhat higher than the ambient, this factor should be taken into account when establishing a corrected noise figure.

Noise power obeys all power-transfer laws, but because noise is random in phase, mismatches cause ambiguous errors rather than known power losses. Because an automatic noise-figure meter measures the ratio N_2/N_1, a mismatch affecting both of these powers has no effect on accuracy, since the ratio remains unchanged.

The major consideration in matching involves the excess noise power available from the noise source. The *ENR* of any of the noise sources discussed in this article, excluding the Hewlett-Packard 345B, is based on developing the noise power in a 50-ohm resistive load. If the load is not 50 ohms, the *ENR* will be indeterminate.

The mismatch error must be calculated for each system by considering that noise power follows the power-transfer laws. A typical plot of error limits for a receiver swr of 2:1 is shown in **fig. 19;** the actual error can fall anywhere between the limits indicated.

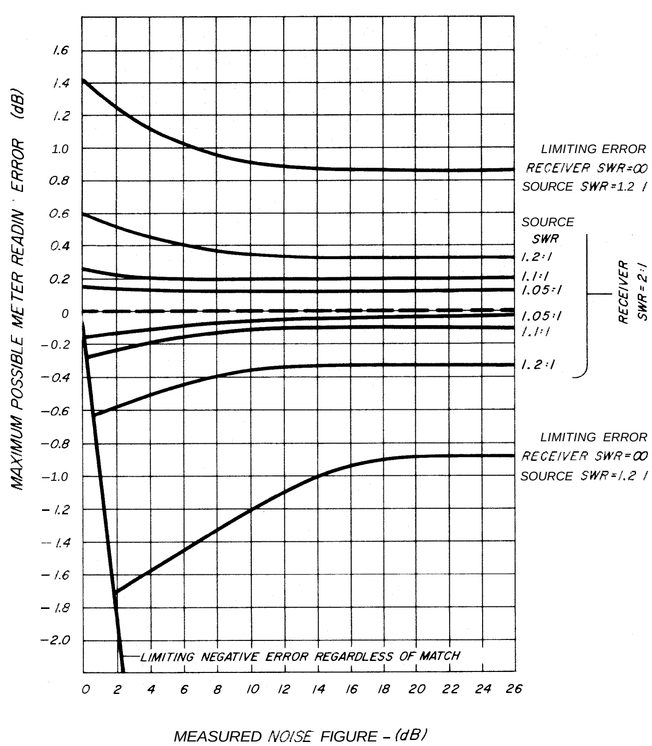


fig. 19. Typical errors for several possible conditions of mismatch between the noise source and receiver *(courtesy Hewlett-Packard Company).*

transmission-line (insertion loss) error

This error is a function of the coupling of the gas-discharge tube to the helical transmission line (see **fig. 3**) and the attenuation of the transmission line, and always reduces the noise-source *ENR*. In **fig.** *13*, this error has been taken into account in determining the excess noise ratios of the AIL noise sources. It has not been considered in the *ENR* curve for the Hewlett-Packard 349A, but according to the published data, the error is less than 0.25 dB. This is well within the *ENR* tolerances specified in **table** 2.

cumulative errors

If all errors are in decibels, they accumulate additively. Therefore, the total possible measurement error will be the sum of the following:

1. Noise-figure meter accuracy
2. Noise-source accuracy, corrected for frequency
3. Receiver image-response error
4. Temperature error
5. Mismatch error
6. Transmission-line (insertion-loss) error, if a gas-discharge noise source is used.

This is an imposing list, and could total well over 1.5 dB if all errors were of the same algebraic sign. However, many of these errors will cancel because of opposite signs, and generally the accuracy of the test equipment is far better than the limits of its specifications. Nevertheless, these possibilities of error are very real and cannot be ignored except for *comparative* measurements using the same equipment at one particular time. And even then, the mismatch errors between the noise source and the different receivers under test still exist.

summary

We have seen how automatic noise-figure measurements are accomplished, and have discussed their limitations and areas of inaccuracy. We can therefore answer the questions posed at the beginning of this article. A difference of 0.2 dB, or even 0.5 dB, in the noise figures of two different preamps measured at two different times on two different test setups is meaningless. Furthermore, the accuracies of such measurements, especially outside of a laboratory environment, are subject to rigorous examination. This is not a criticism of those amateurs who spend many hours at vhf/uhf conferences performing the tests; they know the limitations of their methods and equipment. Rather it is meant to warn those who accept noise figures as gospel that such is not the case.

The same warning applies equally to the "typical noise figure" specifications on some transistor data sheets. Unless an accurately calibrated hot-cold noise source is used to determine the noise figure of the device (and I personally know of cases where this was not done), the published figures may be no more accurate than you or I could obtain with surplus equipment — *caveat emptor.*

The automatic noise-figure meter can also be used to measure noise figure by the twice-power or Y-factor methods. However, since this article was intended to cover only the automatic mode of measurement, no attempt has been made to discuss manual operation. Gain measurements may also be made, in a clever application described by Paul Shuch.[11]

I sincerely hope that this article has taken some of the mystery out of automatic noise-figure measurements. More important, I hope that it will dispel any idea that you know your receiver or preamp noise figure to within a tenth of a dB.

appendix

To change the input frequency of a Hewlett-Packard noise-figure meter, only the frequency-determining circuits in the instrument's tuned amplifier need be changed to the desired frequency. If the new frequency is relatively close to one already established in the instrument, it may be possible to make the change merely by retuning the variable inductors in the tuned circuits. Otherwise, the coils will have to be modified or replaced.

The Hewlett-Packard models 340A and 3408 employ a four-stage tuned amplifier. The front-panel *input* frequency switch selects one of two coils in each stage of the amplifier. In the standard models, these coils and their inductance ranges are as follows:

input frequency (MHz)	model 340A	model 340B	inductance (μH)
30	L1, L3. L5, L7	L6, L8, L10, L12	1.2- 1.75
60	L2, L4, L6, L8	L5, L7, L9, L11	0.27-0.41

Regardless of whether the standard model, having the above input frequencies, or a special model having different input frequencies, is involved, the coil designations will be the same.

The coils resonate with approximately 20 pF. If the existing coils have sufficient inductance range to permit retuning, the procedure is simple. Set the METER FUNCTION switch to NOISE FIGURE and turn the INF CALIBRATION potentiometer fully clockwise. Then, using a signal generator, feed an unmodulated signal at the new frequency into the input of the noise-figure meter. Keep the signal level low enough so that the meter is never at full scale, and adjust the appropriate coils for maximum meter reading. If the meter reaches full scale, reduce the input signal before continuing the adjustments.

If new coils are required for the desired input frequency, select four adjustable coils whose mid-range inductance will resonate with 20 pF at the new frequency. Coils in the Miller 4300 series or Cambion 556-3338 series will replace those in the instrument without any mechanical modifications. After replacing the existing coils in the noise-figure meter with the new ones, retune the amplifier as described above. If the new frequency is 30 MHz or lower, a 3300-ohm composition resistor can be connected across each coil to broaden the amplifier response.

The Hewlett-Packard model 342A has provisions for five input frequencies. Instead of switched coils throughout the tuned amplifier, a fixed-tuned 30-MHz i-f amplifier is used in conjunction with a mixer and local oscillator circuit, which heterodynes four of the five input frequencies to 30 MHz. The fifth input frequency is 30 MHz, which is fed through the mixer with the oscillator disabled.

In the standard model 342A, the following coils and frequencies are used in the mixer-oscillator circuit:

input frequency (MHz)	mixer coil	oscillator coil	oscillator frequency (MHz)
30	—	1100-ohm resistor)	—
60	L10 (0.32 – 0 55μH)	L14 (0.156 – 0.228μH)	90
70	L9 10.32– 0.465μH)	L13 (0.156 – 0.228μH)	100
105	L8 (0.156 – 0.228μH)	L12 (0.32 – 0.465μH)	75
200	L7 (0.0392 – 0.0412μH)	L11 (0.057 – 0.063μH)	170

Special versions of the 342A will have one or more input frequencies which differ from those listed above. Therefore the coil inductances may also be different, but the designations will be the same.

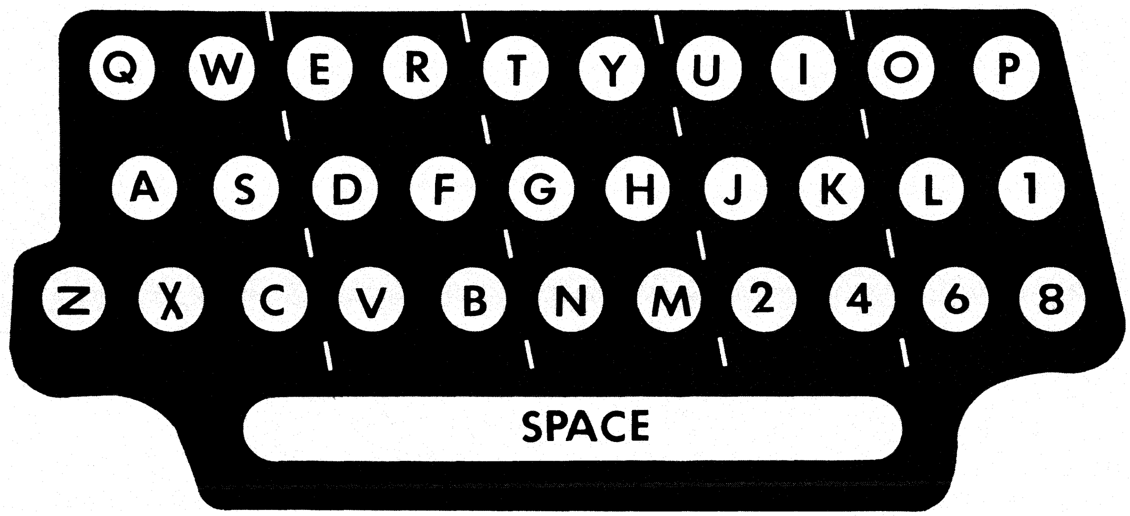
To change any of the input frequencies except the 30-MHz input, either retune the coils associated with the *input* switch position to be changed, or use new adjustable coils whose mid-range inductance will resonate with 14 pF at the required frequencies. The mixer coil obviously must be tuned to the input frequency. The oscillator coil must be tuned to a frequency which is 30 MHz above or below the input frequency; be sure that the oscillator frequency and its harmonics are well removed from the 30-MHz intermediate frequency.

Set the local oscillator on frequency by means of a frequency counter or a dip oscillator. Then adjust the oscillator and mixer coils (but not the coils in the i-f stages) as previously described for the model 340A or 3408, using a signal generator at the new input frequency.

references

1. "An Automatic Noise Figure Meter For Improving Microwave Device Performance," Hewlett-Packard Journal, Vol. 9, No. 5, Hewlett-Packard Company, Palo Alto, California, January, 1958.
2. "Noise Figure Primer," Application Note 57, Hewlett-Packard Company, Palo Alto, California, January, 1965.
3. James R. Fisk, W1DTY, "Receiver Noise Figure, Sensitivity and Dynamic Range – What the Numbers Mean," ham radio, October, 1975, page 8.
4. Benjamin L. Lowe, K4VOW/WA5UVM, "Hot and Cold Resistors As UHF Noise Sources," *QST,* September, 1976, page 32.
5. Charles E. Swedblom, WAGEXV, "Receiver Noise – Theory & Mesurement," unpublished.
6. Joseph H. Huie. K2PEY, "A VHF Noise Generator," *QST,* February, 1964, page 23.
7. Ronald E. Guentzler, W8BBB, "Noise Generators," *QST,* March, 1972. page 44.
8. Louis N. Anciaux, WBGNMT, "Accurate Noise-Figure Measurements for VHF," ham radio, June, 1972, page 36.
9. Hank Olson, WGGXN, and Gene Lehman, WAGJZN, "Noise Generators for 420 Mc. and Up," *QST,* February, 1964, page 33.
10. H. Paul Shuch, WAGUAM, "Circuit Packaging for UHF Double-Balanced Mixers," ham radio, September, 1977, page 41.
11. H. Paul Shuch, NGTX, "Calculating Preamplifier Gain from Noise-Figure Measurements," ham radio, November, 1977, page 30.

ham radio



electronic teleprinter keyboard

This novel electronic keyboard uses toroids and a pulse generator to create the correct marklspace coding

This project grew out of a desire to design an electronic Teletype keyboard which would permit handicapped people to communicate with others. I took an initial design suggested by Ed Brown, WØEPV, determined that it was feasible, and developed the circuit shown in this article. The keyboard itself would generally have to be custom designed for each particular individual; for amateur use any surplus keyboard with a simple, closing-type contact will be sufficient.

circuit description

The heart of the circuit is a set of eight shift registers, composed of four 7474 ICs (see fig. 1). During static conditions, the $\overline{Q}$ outputs of the shift registers are all high, forcing the 7430 NAND gate low. This low disables the clock, while also enabling the pulse generator. When a particular key is actuated (key line taken low), the pulse generator is discharged through the respective cores. For example, if the Y line went low, the pulse generator would be discharged through cores T1, T3, T5, T6 and T7. The pulse from each core sets its respective shift register, causing the $\overline{Q}$ output to go low. Now that one of the inputs of the 7430 has been driven low, its output will change states, both enabling the clock and disabling the pulse generator.

Prior to the change in the 7430, the magnet driver was held in a mark condition by the action of U2.

By L. A. Stapp, WØPHY, 2903 Ash Street, Hays, Kansas 67601

With both inputs to U2A now low, U2 pin **8** will go high, forcing the magnet driver into the space condition. Since the clock has now been enabled, it will put out a pulse after 21 ms, moving the contents of the shift register down one increment. Each low state, on the Q_0 line, will create a ***mark,*** while a high forms a ***space.***

With the clock free running, as long as there is a low into U1, the shift register will receive pulses and move the data one increment for each clock pulse. After **8** shift operations, the registers will be clear and their high outputs will cause U1 to change states again, completing the character.

construction

Building this electronic keyboard is extremely straight-forward with the use of TTL circuitry.* Before operation, R5 should be adjusted with U1 pin 8 high, to provide **1.6** volts on pin **6** of the NE555. If the keyboard is to be used at 60 words per minute, R1 should be adjusted so the 555 oscillates at **45** Hz; for 75 words per minute, use **61** Hz.

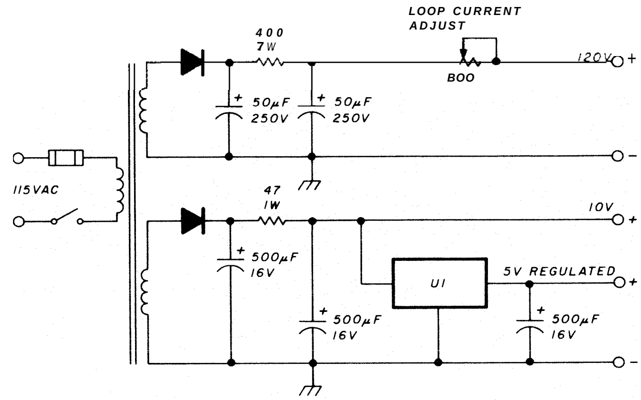


fig. 2. This power supply will provide voltages for both the keyboard plus the Teletype loop. The transformer is a Stancor 8626.

*A copy of the circuit board layout and parts placement diagram isavailable by sending a self-addressed, stamped envelope to ham radio, Greenville, New Hampshire 03048.

ham radio

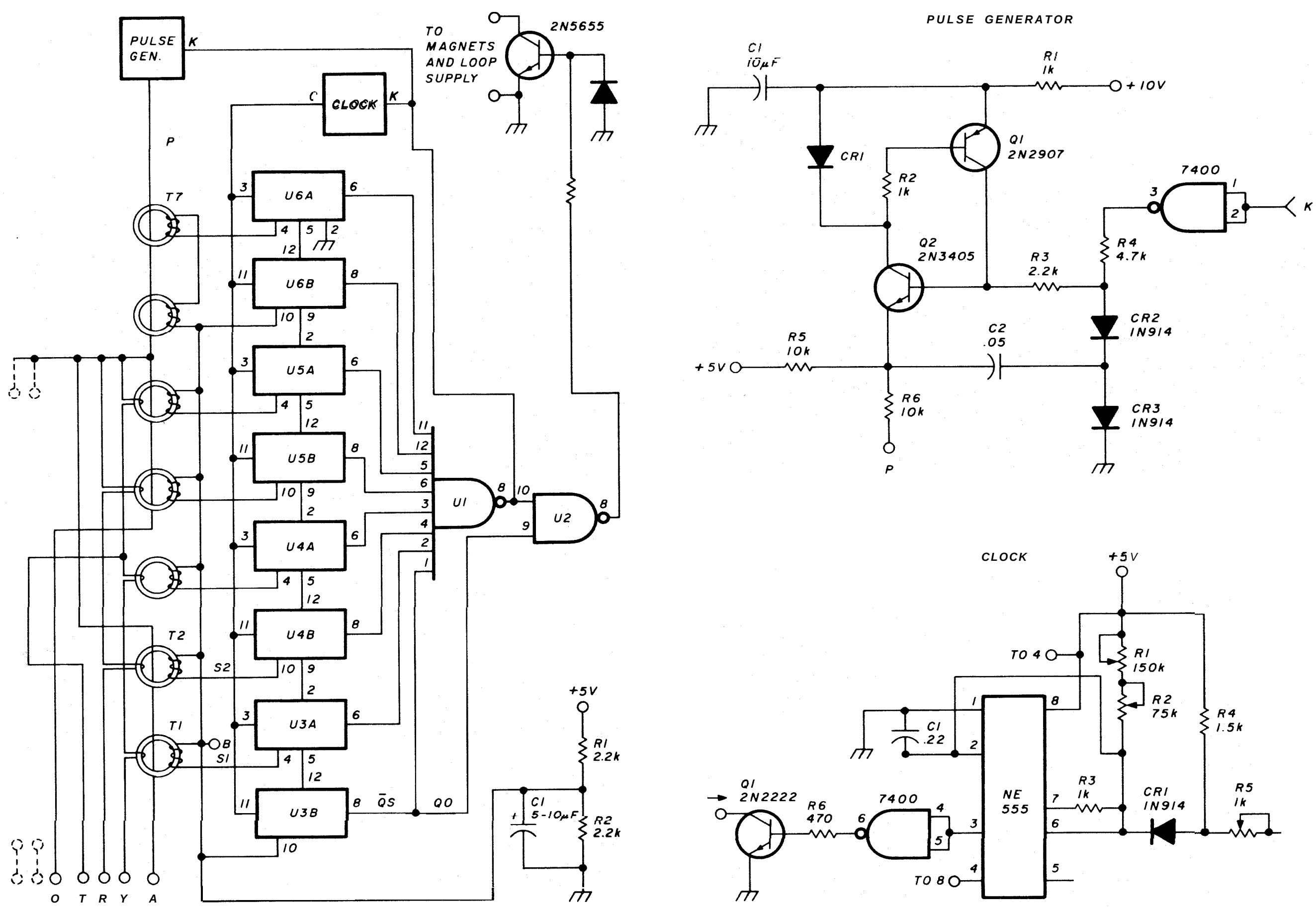
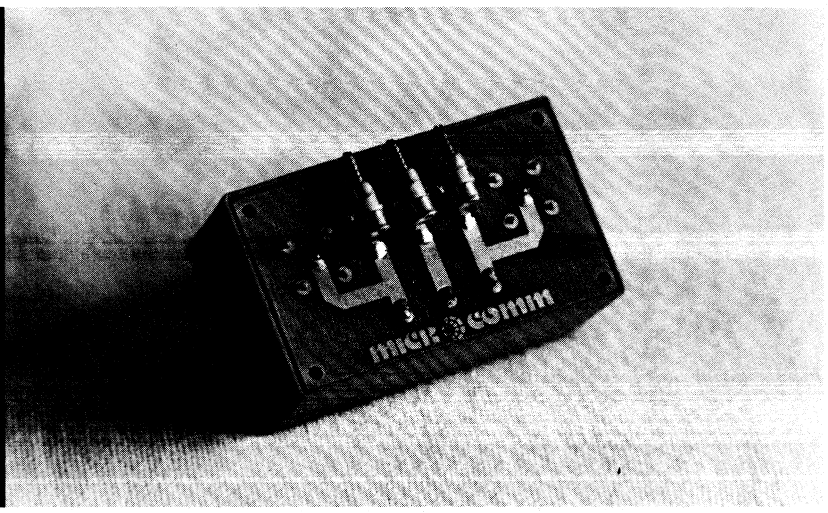


fig. 1. Schematic diagram of the electronic keyboard. Each toroid is an Indiana General CF-102, 0-6, wound with a 10-turn primary. The wires from the pulse generator are fed through the appropriate core to form the mark periods for each character.



improved grounding for the 1296-MHz microstrip filter

Construction techniques for improving the performance of the three-pole 1296-MHz bandpass microstripline filter

The 1296-MHz microstripline bandpass filter I discussed in a previous article has allowed dozens of uhf experimenters to "clean up their act" on the 23-cm band.[1] As shown in **fig. 1,** the filter consisted of three top-coupled parallel-resonant circuits with grounded microstripline inductors. The filter is easy to assemble and tune, but several amateurs who have built it experienced difficulties caused by erratic stripline grounding. The new design presented here eliminates those difficulties and provides lower insertion loss and steeper skirts that will not tend to degrade as the filter is used.

The grounding of the microstriplines in the original design was accomplished by wrapping a thin brass or copper strap around the edge of the PC board, soldering it to the stripline on one side of the board and the groundplane on the other. Although this method of grounding worked well in the prototypes, the stripline inductance is a function of the placement of the grounding strap. Furthermore, the strap's placement on the edge of the board makes it extremely susceptible to physical damage, especially when installing or removing the filter board from its box. A third difficulty with wraparound grounding is that it forces the end of the microstripline inductor to extend to the edge of the board, where stray coupling can cause the tuning of the filter to change when the unit is placed inside an enclosure.

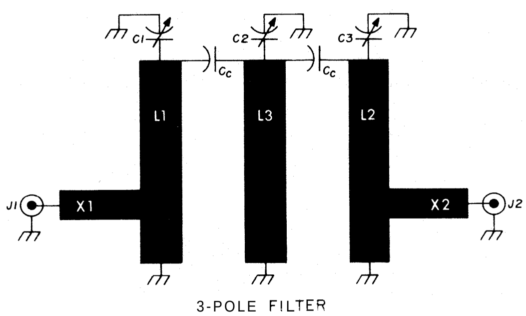
All of these difficulties can be easily eliminated by placing the grounded end of the microstripline inductors somewhat away from the edge of the board and drilling through the board for the installation of a grounding wire or post. With the ground connection running through (rather than around) the board, its mechanical integrity is assured, and the groundstrap inductance is more nearly constant from one filter to the next, especially if the diameter of the grounding wire or post is specified.

By H. Paul Shuch, N6TX, Microcomm, 14908 Sandy Lane, San Jose, California 95124

Although a short length of number 18 (1.0 mm) tinned copper wire makes an acceptable through-board ground connection, maximum grounding effectiveness and mechanical integrity, I have found, can be achieved by installing small electronic eyelets through the board, setting them with a press, and soldering both sides. The eyelets I use are made of thin brass, measuring 0.47 inch (1.2 mm) diameter by 0.093 inch (2.5 mm) long. They look something like tiny rivets. The eyelets are available from a number of vendors* and can be easily set using a center-punch (or sharp nail) and hammer.

grounding the trimmer capacitors

Another area of difficulty encountered by several readers is the grounding of the piston trimmer capacitors. The capacitors I originally used were designed for chassis mounting, so it was necessary to modify



C1 - C3 1.5-pF ceramic piston trimmer (Triko 202-08M or equivalent)

C_c Stray coupling capacitance between stator ends of trimmer capacitors

J1, J2 SMA or equivalent microstripline launchers (E.F. Johnson 142-0298-001 or similar)

L1, L2, L3 Microstripline inductor, 0.5″ (13 mm) long, 0.1″ (2.5 mm) wide, spaced 0.3" (7.5 mm) center to center. Bottom ends strapped to ground plane with thin copper strap

X1, X2 50-ohm microstripline, 0.1" (2.5 mm) wide, any length. Centerline tapped to L1 and L2 0.2" (5 mm) from grounded end

fig. 1. Three-pole microstripline bandpass filter, which will tune the range from 1100 to 1500 MHz. Full-size printed-circuit layout for this filter is shown in fig. 2.

them for circuit-board use by adding a bus-wire loop around the terminal nearest the adjusting screw (see **fig. 4** of reference 1). It would have been better to use a trimmer specifically designed for PC use, with legs installed for grounding the rotor end through the circuit board. One such capacitor is the R-Triko 202-

*One acceptable eyelet is part number F-4793-B, available from International Eyelets, Inc., 528 Santa Barbara Street, Santa Barbara, CA 93101.

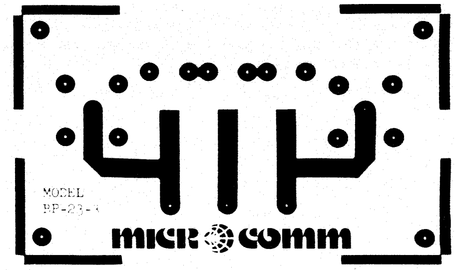


fig. 2. Full-size printed-circuit layout for the three-pole 1296-MHz bandpass filter. Etched on double-clad 1/16″ (1.5 mm) fiberglass-epoxy circuit board; the unetched side serves as a ground plane.

08M, a German ceramic piston trimmer available in the required 1-to-5-pF range." I find filters using this capacitor easier to tune up, although I caution the builder against repeated adjustments because the tuning mechanism loses spring tension and becomes erratic after a couple dozen adjustments. The best procedure is to set the filter on frequency ***once,*** and then place a dot of nail polish, epoxy paint, or *Loctite* on the tuning screw as a reminder to leave it alone!

assembling the modified filter

Fig. 2 is a full-size printed circuit layout for the 3-pole bandpass filter, modified for through-board grounding. The board should be etched from double-clad 1/16-inch (1.5-mm) fiberglass-epoxy printed-circuit stock, with one side left unetched to serve as a ground plane. The board should be drilled in the same manner as the template in **fig.** 3 and the three eyelets installed at the bottom end of the microstriplines.‡ Don't forget to remove a bit of ground plane

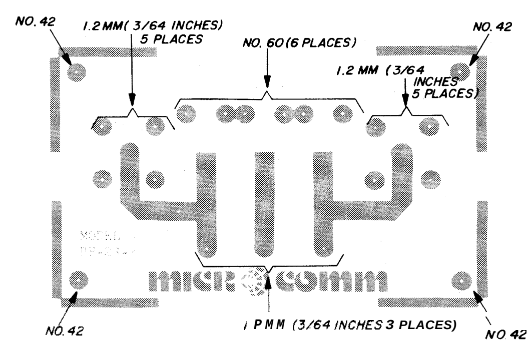
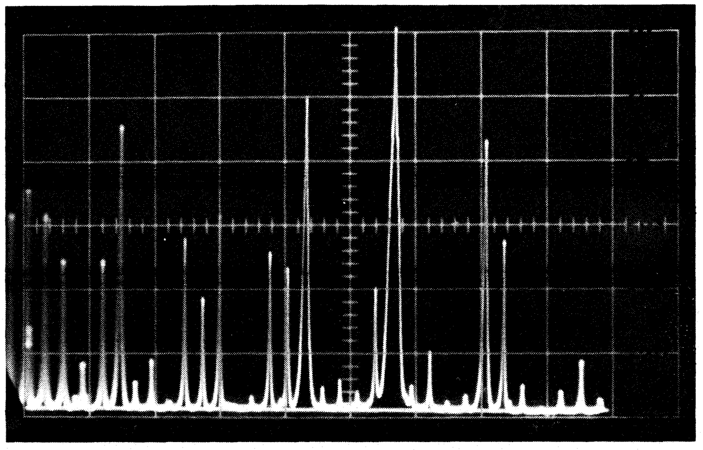


fig. 3. Full-size drilling template for the bandpass filter board.

"These capacitors are available in the United States through Stettner-Trush, Inc., 67 Albany Street, Cazenovia, NY 13035.

‡Completely etched, drilled, and plated printed circuit boards, with the three eyelets installed, are available for $4.50 postpaid within the U.S. and Canada, $5.00 elsewhere, from Microcomm, 14908 Sandy Lane, San Jose, CA 95124. Completely assembled, tuned, and tested filters are also available. Send a stamped, self-addressed envelope for details.



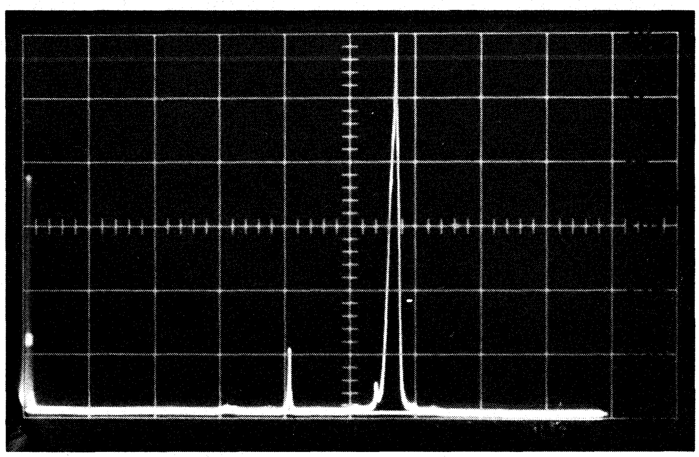


fig. 4. Effect of the bandpass filter on a 1296-MHz local oscillator chain. Spectrum display at top shows the various spurious outputs of a poorly designed LO. Spectrum at bottom is the result of passing the LO signal through a three-pole bandpass filter of fig. 1. The worst remaining spurious component is suppressed 25 dB. (Both displays: dc - 1.8 GHz sweep; horizontal scale, 200 MHz/division; vertical scale, 5 dB/division.)

metallization from around the center-pin holes for the input and output coaxial connectors so the signal isn't grounded out. A 1/8-inch (3-mm) twist drill, used as a deburring tool, works well for this operation.

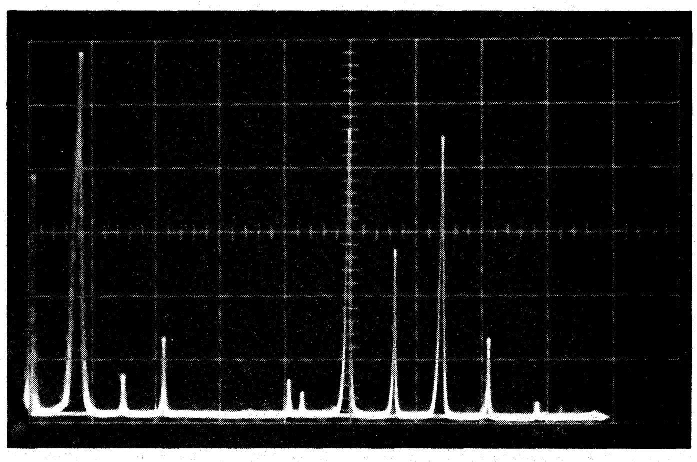
Connectors J1 and J2 are installed next, soldering the center pin to the input and output microstriplines, and running a bead of solder around the connector body on the ground-plane side of the board. The trimmer capacitors are installed last. If you use the recommended Triko trimmer, be sure to bend the two mounting legs nearest the adjusting screw down against the ground plane before soldering. The photograph of the completed filter will assist you in assembly.

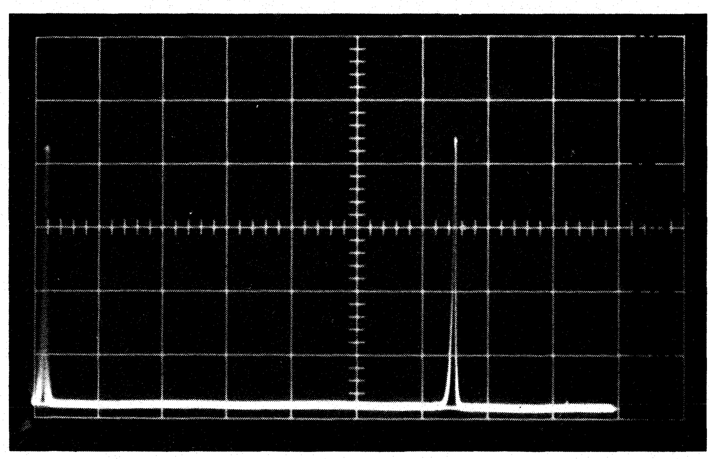
filter performance

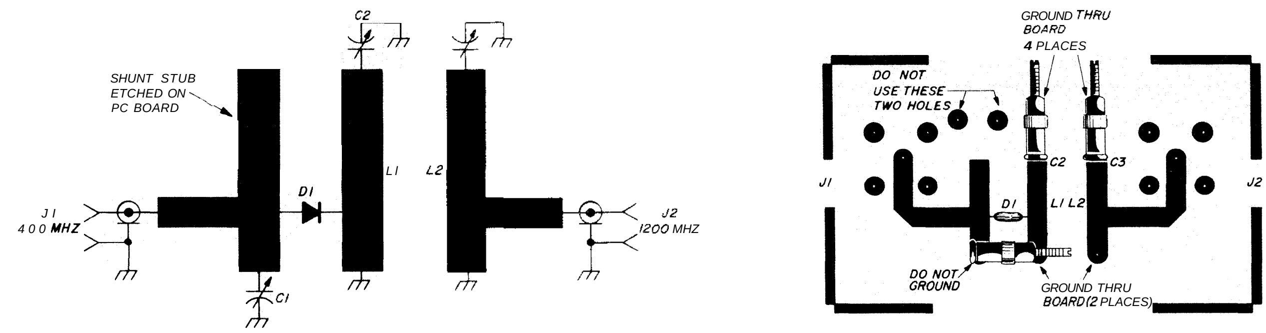
Reference 1 included a swept response curve for the original bandpass filter, as measured on a network analyzer. The response curve for the modified filter design shows slightly reduced insertion loss (on the order of 0.5 dB) and slightly steeper skirts. Perhaps the most realistic indication of filter performance is not its swept response, but the filter's behavior in an actual system. **Fig. 4** shows the effect of installing the bandpass filter behind an extremely spurious local oscillator chain. Note that the numerous spurious components are all significantly suppressed, with the worst remaining spur reduced from −5 dB to about −25 dB, relative to the desired output.

Fig. 5 shows the results when the filter is used to clean up the output of a previously published transmit balanced mixer.[2] Notice that the i-f feedthrough signal and its harmonics, the LO feedthrough, the

fig. 5. Effect of the three-pole bandpass filter on the output of a 1296-MHz transmit mixer. Spectrum display at top shows the output from a singly balanced diode mixer; visible spurious components include the desired signal and image, some LO feedthrough, a very strong component of the i-f injection, and its second and third harmonics, and transmit intermods (resulting from these harmonics mixing with the LO signal). With the three-pole bandpass filter installed in the system, the spectrum (bottom photo) shows that all spurious outputs have been attenuated by more than 25 dB. (Both displays: dc - 1.8 GHz sweep; horizontal scale, 200 MHz/division; vertical scale, 5 dB/division.)







C1	1-9 pF **piston trimmer** (Triko 203-09M **or equivalent)**	J1, **J2**	**SMA coaxial connector** (E.F. **Johnson 142-0298-001 or similar)**
C2, **C3**	1-5 pF pistion **trimmer** (Triko **201-01M or equivalent)**	L1, L2	**Microstripline inductors (see fig. 2)**
CR1	**Step** recovery **diode** (H-P 5082-0180)		

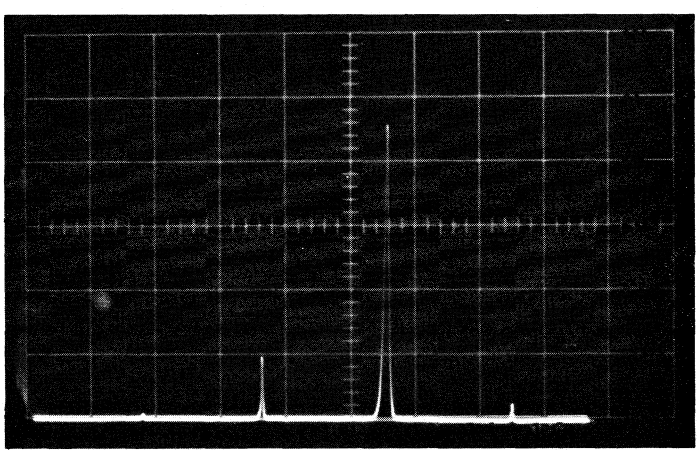
fig. 6. Circuit and construction details for a local-oscillator multiplier which provides 0.5 mW at 1200 MHz from a 5-mW 400-MHz drive signal. This circuit uses the same printed-circuit layout as the bandpass filter (see fig. 2). Output spectrum of this circuit is shown in fig. 7.

image signal, and the intermodulation products are all suppressed below the dynamic range of the spectrum analyzer.

local-oscillator multiplier

In a previous article I described a diode multiplier for developing local-oscillator injection for a 1296-MHz converter.3 As this multiplier used a microstripline output filter, it seemed reasonable to assemble a similar multiplier on the bandpass filter PC board, thus allowing one PC artwork to do the job of two. The circuit, which makes a rather nice low-level tripler, is shown in **fig. 6.** Note that the microstripline previously associated with the first filter pole is now used to support the multiplier diode and its input matching circuit. Do *not* install a grounding eyelet on this first stripline if you are building the multiplier! The other two filtering poles help reject the many other harmonic components generated by the step-recovery diode, as shown in the spectrum analyzer display of **fig. 7.** When driven by the 5-to-10 mW signal from my uhf LO chain, this multiplier provides about 0.5 mW of third-harmonic output. This power level can be easily buffered in a 1296-MHz preamp,[4] applied to a 3-pole filter for additional spurious rejection, and used to drive the LO port of a transmit or receive balanced mixer.

fig. 7. Output spectrum of the 400-MHz to 7200-MHz LO multiplier. Note that the tripler circuit provides some degree of filtering of the 400-, 800-, and 1600-MHz components from the step recovery diode.



I should point out that the circuit of **fig. 6** provides no dc return for the anode side of the multiplier diode. A dc return is necessary for the diode to properly develop self-bias; in my system this dc path is provided at the output of the uhf LO. If the driving stage does not offer dc continuity to ground, it will be necessary to install a dc return circuit on the multiplier board. This can be most readily accomplished by adding a small (0.33-μH) rf choke to ground at the location normally occupied by the first trimmer capacitor when this board is used as a filter.

summary

The printed-circuit layout can be used to fabricate high-quality bandpass filters and diode multipliers for the amateur 23-cm band. The designs are based upon previous articles, but the addition of through-board eyelet grounding significantly improves performance and reliability. Further details on construction, tune-up and testing, and system application are discussed in reference 1.

references

1. H. Paul Shuch, WA6UAM, "Microstripline Bandpass Filters for 1296 MHz," *ham radio,* December, 1975, page 46.
2. H. Paul Shuch, WA6UAM, "Rat-race Balanced Mixer for 1296 MHz," *ham radio,* July, 1977, page 33.
3. H. Paul Shuch, WA6UAM, "Easy-to-build SSB Transceiver for 1296 MHz," *hamradio,* September, 1974, page 8.
4. H. Paul Shuch, WA6UAM, "Low-cost 1296-MHz Preamplifier," *ham radio,* October, 1975, page 42.

ham radio

simple scope monitor

for vhf fm

In the dark
about those signal reports
on two meters?
This easy monitoring system
pays off in
increased versatility

Adding an inexpensive oscilloscope to your fm transceiver will make your base station much more versatile. You'll have a signal monitor capable of measuring (+0–) carrier frequency, peak-deviation modulation, per cent noise quieting, and *Touch-Tone* signal levels. You'll be able to tweak transmit crystals in other rigs (great for netting CD-MARS members on frequency) and much more. Now blow the dust off your old scope and put it to good use — or give the test scope a dual function, as I did.

fm signal reports

How often have you heard, "You're 40 per cent quieting but Q5 copy," and wondered how the operator on the other end arrived at the 40 per cent figure? He must have been good at guessing, had a super trained ear, or maybe trying to be friendly.

When I became interested in two-meter fm I was very much confused with per cent quieting signal reports (especially if I had to give one). The transceiver I used was an ICOM IC20 and it did have a relative strength S-meter. While working simplex, whatever the meter indicated would be the report I gave. However, working through repeaters presented a whole new ball game. Unlike simplex, repeaters may come slamming into your location, but the signal on the input side may be quite weak. Under such circumstances the S-meter reading is invalid for a report, but a monitor scope would immediately show the per cent noise quieting into the machine. One day while working on a circuit and using my old EICO model 460 scope, I saw the light for a per cent quieting indicator.

operation

After examining the IC20 schematic for a good signal pickoff point, I decided the best place would be at the output of :he discriminator stage (at TP3) and ahead of the audio control circuit (see fig. 1).

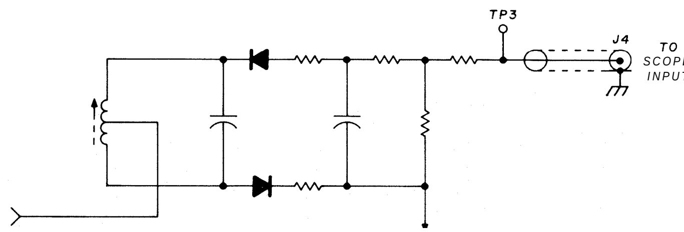


fig. 1. Pickoff point in a typical amateur 2-meter transceiver for connecting a monitor scope. In this case it's at TP3 in the popular ICOM model IC20. The pickoff point should be made at the receiver discriminator output, but ahead of the audio control circuit.

Several advantages would be obtained by doing this: 1) I'd see the noise figure as it appeared, unaffected by audio and squelch-control settings; 2) I'd see peak deviation (modulation); 3) by connecting directly from the discriminator output to my scope's dc vertical input I'd be able to measure the dc voltage produced by the received-signal carrier. Most fm transceivers (or receivers) have a test point at the location mentioned for alignment purposes. Your manual or schematic should indicate if this point is available.

connection

Modification and wiring of the transceiver was simple (fig. 1). Connection to the outside world was made using a small-diameter, single-conductor, shielded cable from the discriminator circuit to a

By Ed Spadoni, W1RHN, 91 Tower Street, Dedham, Massachusetts 02026

rear-panel-mounted phono connector (Radio Shack No. 274-346) designated J4 in **fig. 1**. The shield was wired to J4 ground lug only. Next my scope was connected through a suitable length of shielded cable, with a male phono plug on the end, to mate with J4.

calibration

Scope control settings will vary depending on the types of scope used. My EICO 460 control settings were as follows: Vertical attenuator in the DC-X1 position, horizontal sweep at 100 Hz, and sync on INTERNAL. With the transceiver tuned to an inactive channel, adjust the vertical gain for 51 mm (2 in.) p-p noise level **(fig. 2)**. The noise as seen on the scope indicates the total passband of the rig's receiving section. My IC20 has a passband of better than 16 kHz, which allows me to see signals ±8 kHz from the passband center with reasonable accuracy. At present I can read 1 kHz = 2.5-mm or 0.1-in. division on the scope with good linearity to ±8 kHz.

It's essential to have your receiver properly aligned and your receiving crystals adjusted to the passband center. If you're synthesized, be sure the receive oscillator is also properly adjusted. I've found that a very convenient frequency reference source is a repeater output signal. By tuning in several machines you can see if things are properly adjusted by noting if the repeater signals are located in the passband center. If not, a little trimming must be done.

Working someone with a synthesized or vfo rig can prove very helpful in determining the linear portion of your receiver and its frequency limits. After determining the useful reading range of your transceiver-scope combination, you should be able to read signals to ±1 kHz of your receiver passband center.

using your scope

Here are some samples of the more important waveforms observed while monitoring both simplex and repeater signals.

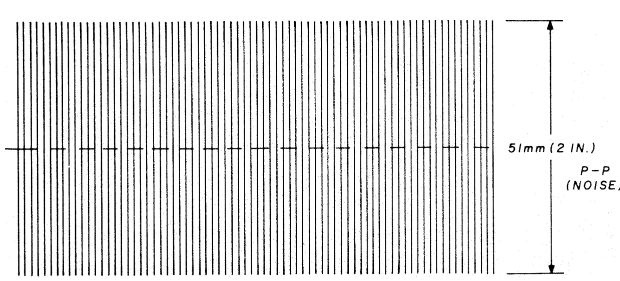
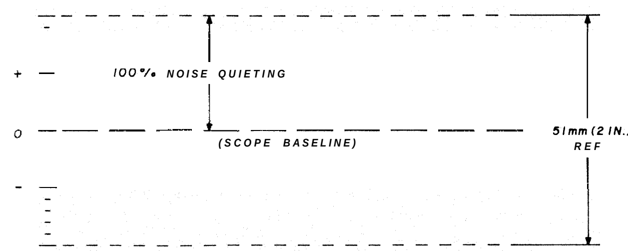
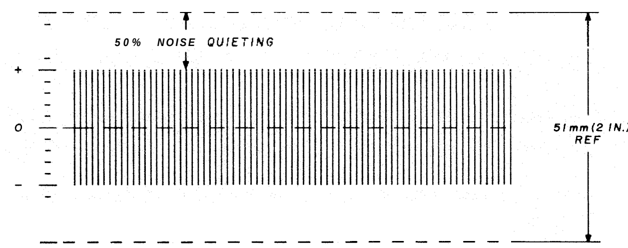


fig. 2. Scope presentation for calibration. peak-to-peak noise should occupy the entire scope face, which represents the total passband of the receiver on noise.

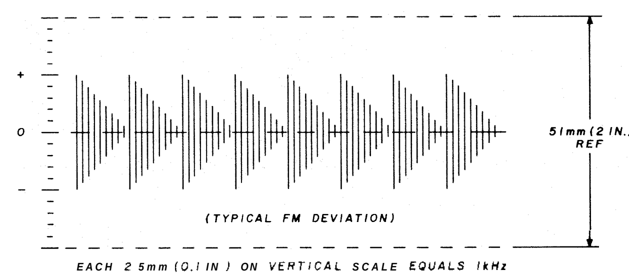
1. 100 per cent noise quieting signal, either simplex or repeater, with a pause in modulation:



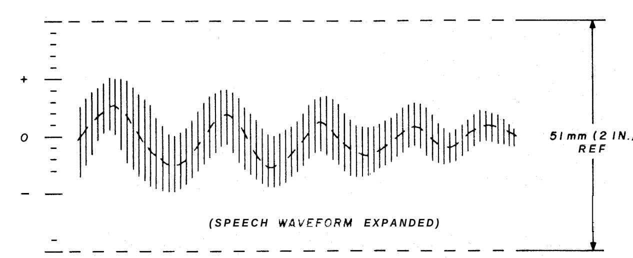
2. 50 per cent noise quieting signal, either simplex or repeater, with a pause in modulation:



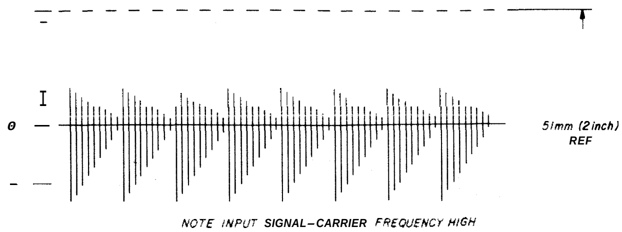
3. Normal signal showing deviation (modulation). Most transceivers and repeaters have their deviation set for a nominal peak of 5 kHz:



4. 100 per cent quieting repeater signal received at your location with a weak and noisy input signal. Horizontal sweep adjusted for expanded one or two speech waveforms:

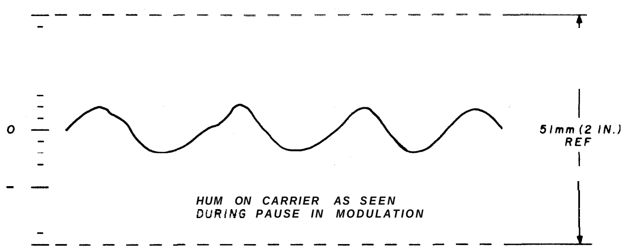


5. Repeaters normally produce equal swings of deviation, in both up and down directions, with most input signals. If a new station comes into the machine and produces the pattern shown below, this means that the new signal's carrier frequency is off and higher than the repeater passband center. The opposite would be true with a low-frequency carrier signal into the machine:

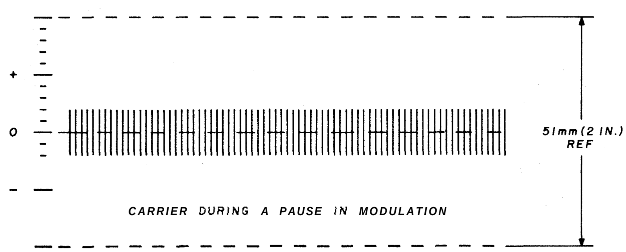


6. Hum accompanying signal. Frequency and ratio of hum to modulation can be determined. Hum frequency is referenced to 60 Hz.

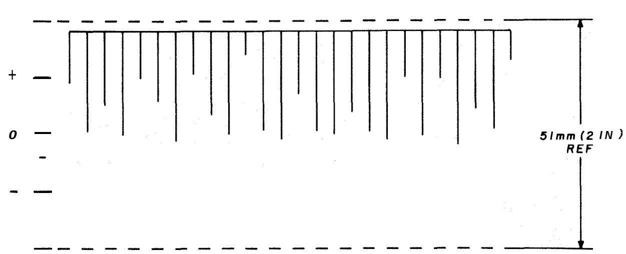
$$\textit{Per cent of hum} = \frac{hum\ p\text{-}p}{deviation\ p\text{-}p} \times 100:$$



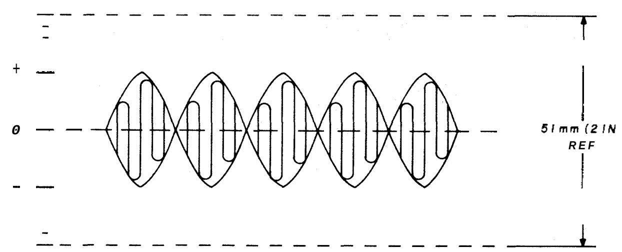
7. Power supply whine accompanying the received-signal carrier has a constant amplitude and frequency. Mobile alternator whine is similar, except the amplitude and frequency will change as the vehicle is accelerated or decelerated. Whine appears as a high-frequency modulated carrier with or without voice modulation:



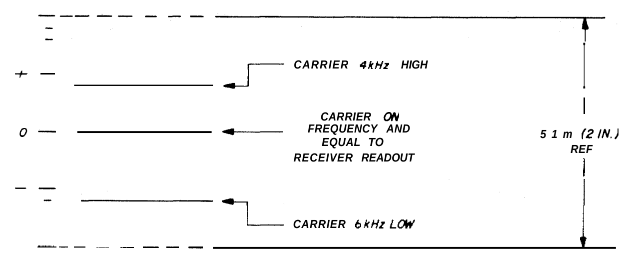
8. Upper-channel interference from a strong simplex or repeater station. Many hams have experienced this condition and termed it (in error) intermodulation. Lower-channel interference is similar, except that the baseline would appear on the lower side of the passband with audio transients shooting upward:



9. Checking and adjusting *Touch-Tone* signals. Preferably, use a simplex channel and have the operator close-talk into the microphone at a normal voice level. (Note the peak deviation.) Next, have the operator switch in the TTP, push buttons **1** and 2 or 1 and 4, and observe the peak amplitude. Peak deviation and amplitude should be the same for both; if not, adjust TTP output level. Then have the operator push all buttons to see if all levels are the same. When a single button is pressed a dual frequency is generated and is a normal function for a TTP, as shown below. Each button produces its own set of different frequencies:



10. Frequency measuring. Using a scope with your transceiver gives you a limited-range frequency meter, allowing you to read up to ±**8** kHz per channel on 2 meters. The range is restricted only by the passband capability of your transceiver.



final comments

You can leave the scope in the circuit without affecting transceiver performance. If you have separate receive-transmit capability, the scope can also be used to monitor your own transmitted signal. The waveforms illustrated are those most often encountered and are therefore the most important.

Scopes aren't difficult to come by. Try surplus houses, auctions, or build one from a kit. A monitor scope beats the cost of frequency counters and you see much more. Making this simple modification to your radio and adding a scope will allow you to keep watch over other rigs (your hand-held; repeater output), especially the transmitted signals. The combination becomes unbeatable when used with synthesized transceivers or those with ± 5-kHz offset or with transceivers having a VFO and 1-kHz readout.

ham radio

single-tone decoder for vhf fm

Design and construction of a false-free device using a single tone to alert you on two-meter fm

A single-tone decoder is adequate for many applications on vhf fm, but very little practical information on these devices has been published. Several years ago K2OAW described the use of a 741 op amp as a carrier-operated relay (COR) and tone decoder,' but I didn't know what a COR was at that time. So I read the part about the tone decoder, and within 15 minutes I was at the junk box struggling to put a reasonable fascimile together to see whether or not all that was claimed for this circuit was true. The claims were that the decoder would not trigger on noise, speech, or even singing but would activate immediately in response to the chosen tone. Beyond that, the decoder bandwidth was not supposed to get any broader as the input amplitude increased. Also, the decoder was not supposed to trigger on the selected tone if that tone were accompanied by any other tone or noise.

This article describes a single-tone decoder with hints on how to set it up. It has the advantage of being free from falsing while coming on quickly enough to stop a scanner on the frequency where the tone was transmitted. An appropriate encoder, small enough to fit in most hand-held transceivers, is also described.

In those days I had two CB hand-held transceivers with built-in tone encoders for mutual noise making on 11 meters. I wanted to use the decoder (if it worked) with these units. Fortunately I had all I needed on hand to make the circuit on a perf board. When I was finished, lo and behold, it worked! It was so selective that it wouldn't false, even on channel 11 with the band open; when my tone came through, the decoder came right on.

I put it all together in a box with a speaker and some jacks and used it that way for several years until I got on 2-meter fm. I then discovered some problems and shortcomings that needed solutions. Over the years I came up with a modified circuit that filled my needs and has been working well ever since. Before going any further, let me explain why I used a single-tone decoder rather than *Touch Tones,* sequential tones, or some other type of selective decoding device. It's not really the decoder but the *encoder* that makes the difference. A stable, single-tone encoder can be easily and inexpensively built to fit into a hand-held transceiver, and that's exactly what I did.

operation

The radio is on at all times and tuned to a repeater frequency, but a relay directs the audio to a 10-ohm, 2-watt resistor and the decoder input. When the desired tone is received the audio is directed through a 7-second timer to both a local and an extension speaker. (The extension speaker, in my case, is located in the kitchen.) After seven seconds the unit resets, and the audio is removed from the speakers. My wife (the technician in the house) then goes to the shack and operates a toggle switch that defeats the decoder, disconnects the extension speaker, and supplies the audio at a conversational level to the

By Steve S. Krarnan, MD. WA2UMY, 29016 Candlelight Way, Lexington, Kentucky 40502

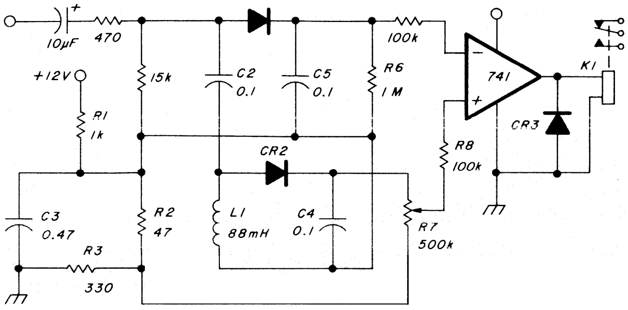


fig. 1. Original circuit of the single-tone decoder, which forms the basis of the modified system. (From *73*, July, 1972.)

local speaker. After two-way communications are completed, the same switch returns the unit to normal operations. Nothing need be disconnected from the rig at any time.

The original circuit is reproduced here (fig. 1) to demonstrate its operation. I'll dispense with a detailed description of how the decoder works. Briefly, the incoming signal is separated into two components — the desired component and all others. The two voltages so derived are rectified and fed respectively to the positive and negative inputs of the 741 op amp. The voltage at the positive input must exceed that at the negative input for the 741 to turn on. The bandwidth is set by R7, which determines the op amp bias. (A complete description of the circuit can be found in the original article.)'

modified decoder

Fig. 2 shows the modified circuit. C1, the frequency-determining capacitor, is replaced by seven capacitors and a seven-pole, single-throw switch on a 14-pin DIP. (This is only a convenience to allow easy frequency change.)

choosing capacitors

Be sure to use only NPO capacitors for C1; that is, capacitors whose values do not change with temperature, otherwise you may find, as I did, that on warm or cool days the decoder will not respond to your tone. A drift of only a few hertz can prevent the decoder from working. If you're not sure of your capacitors but have access to a capacitance bridge, connect the chosen capacitor across the bridge, take a reading, and then coo! the capacitor with freeze mist and take a second reading. If the change is more than slight (say about 10 per cent), don't use it. You'll find that most disk ceramics will change value by as much as 50 per cent under these conditions. Mylar or tantalum capacitors are usually good.

My unit is set up to decode frequencies between 300-600 Hz using a 150-pH toroid; an 88-pH toroid, as originally described, is fine. Experiment with values of C1 to see where you are and work up or down from there. If you have a low-band rig, connect the audio output to the decoder and use the crystal calibrator to generate a test signal. Of course, an audio signal generator may be used, but remember that the amplitude must be similar to that of an audio power amplifier since these are the conditions under which the decoder was designed to work.

adjustment and tests

If the decoder doesn't operate at first, test the frequency determining components by placing a dc voltmeter across the 741 positive and negative inputs. Sweep an audio-frequency generator over as

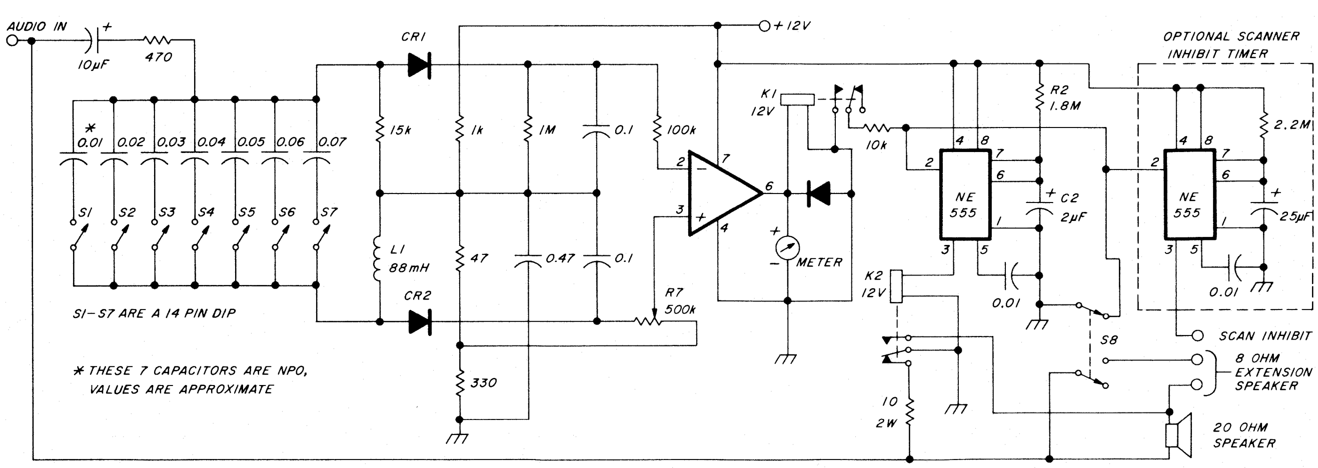


fig. 2. Modified decoder featuring easy frequency change using 7 NPO capacitors and a 7-pole, single-throw switch on a 14-pin dual in-line package. Unit decodes frequencies from 300-600 Hz. The universal 741 op amp does the work; bandwidth is set by resistor R7.

wide a band as possible. The voltage will peak at one frequency and remain negative at all others. Play around with the bandwidth while you are doing this.

The 555 timer is set to keep the speakers on for about seven seconds, but you can substitute a pot for R2 to give different lengths of time.

The meter across K1 is used to adjust the encoder to the decoder frequency. The meter will read highest at the decoder's most sensitive frequency. The meter is also useful for quick checks and adjustments. Almost any sensitive meter can be used with the appropriate series resistor to keep it in range. I used a tape recorder VU meter.

encoder

I tried several circuits as an encoder, but by far the best is the "Twin-T Oscillator" taken from the *Radio Amateur's* VHF *Manual.* While the output amplitude is low and must be fed into the transmitter mike input, the frequency is completely independent of the supply voltage over its operating range. When NPO capacitors are used, a very stable source of oscillation results, which is mandatory if the encoder-decorder pair are to work reliably. It's best to set up the encoder with an oscilloscope to try to achieve a near-perfect sine wave. Other waveforms contain harmonics, which will tend to desensitize the decoder. Keep this in mind when you operate through a repeater. The repeater's audio shaping, or your own overdeviation, may cause tone distortion.

suggested improvements

This unit has worked well for several years, but, for still further usefulness, the following may be done. A scanning board can be added to the receiver and connected so that the decoder, when coming on, will inhibit the scanner and lock the receiver onto the tone frequency. This may be accomplished by a second 555 timer set to lock on for about one minute. The manual defeat switch would also inhibit the scanner. This setup will allow you to use whichever repeater or simplex frequency is most convenient at the time, especially if your favorite repeater happens to be down just when you want to make the call. The decoder can be used in this way because it triggers almost instantly on the appropriate tone. Other single-tone decoders using the 567 chip require a prolonged tone to achieve freedom from falsing. A scanner would pass by too fast to decode the tone in this case. Other modifications will come to mind I'm sure. I hope you find this project useful and fun.

reference

1. Peter Stark, K2OAW, "741 Op Amp COR and Tone Decoder Circuits," 73, July. 1972, page 83.

ham radio

electronic bias switching

for the Henry 2K4 and 3KA linear amplifiers

Easy modifications
you can make
to these popular linears
to increase efficiency
in CW and ssb modes

Two excellent articles have appeared in the amateur literature dealing with electronic bias switching (EBS) for high-power linear amplifiers.[1,2] Why electronic bias switching? It's a great saver of tube life. It reduces tube dissipation, ambient noise, and your power bill. EBS, in general, is a way to make your amplifier operate more efficiently in whatever mode you choose, ssb or CW.

The EBS method described here may be used by those amateurs interested in CW only operation or by those using ssb with or without signal processing. Using the basic circuits described in references 1 and 2, a very efficient EBS circuit can be built into the popular Henry 2K4 or 3KA linear amplifier. The circuit can be adapted to your home-brew linear with a little ingenuity.

Henry rf decks

First of all, for those not knowing it, the rf decks in the 2K4 and the 3KA amplifiers are almost identical. The only difference is the use of wider-spaced loading variable capacitors in the 3KA (two 350 pF instead of two 500 pF, plus three additional 100-pF, 5-kV doorknob fixed capacitors). A 2K4 rf deck can easily be modified to a 3KA rf deck by simply changing these components. Of course, the 3KA uses higher plate voltage (3.6 instead of 2.8 kV). A 2K4 can be driven to 2 kW PEP with 100 watts rf, while the 3KA needs at least 150 watts of rf drive to run at its full rated output.

The EBS circuit will not be dealt with in detail. The referenced articles discuss the principles of operation of the circuit. The circuit components can be mounted on the aluminum panel covering the compartment that houses the swr bridge, zener diode, etc.

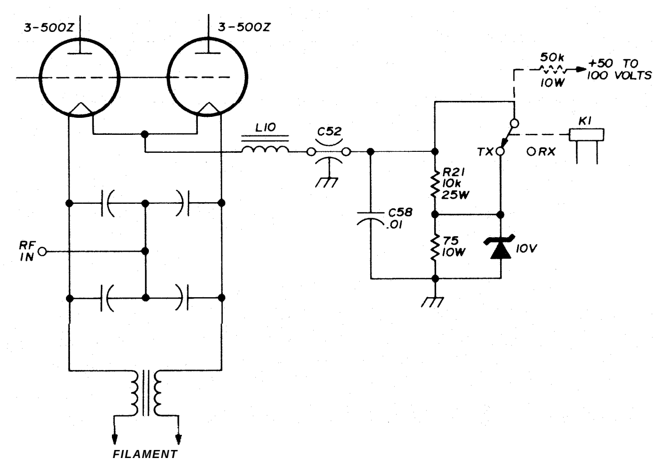


fig. 1. The bias and transmit-receive switching circuits in the Henry 2K4/3KA linear amplifiers. Dashed line (upper right) shows a 50k resistor added to the circuit through a +50-100-volt power supply to effect complete amplifier cutoff during receive mode.

Fig. 1 shows the original bias plus transmit/receive switching circuit as used in the 2K4 and 3KA amplifiers. During receive, R21 is switched into the cathode circuit, whereby the tube is biased to a point very near cutoff. Simply adding a 50-k resistor (**fig. 1**) and connecting it to a +100-volt supply (50-150 V) will improve the circuit; the tube will then be ***completely*** cut off during reception (+100 volts on the cathode).

By Michael James, W1CBY

improved bias circuit

Fig. 2 shows the EBS circuit as developed especially for the 2K4 and 3KA linears. R1 through R4 are well over-dimensioned resistors (in wattage), where a parallel combination is used for added safety (if one resistor opens, the system will still function on the remaining resistor). These resistors can be mounted on the bottom side of the cover plate. The Darlington transistor and the thyristor can be bolted

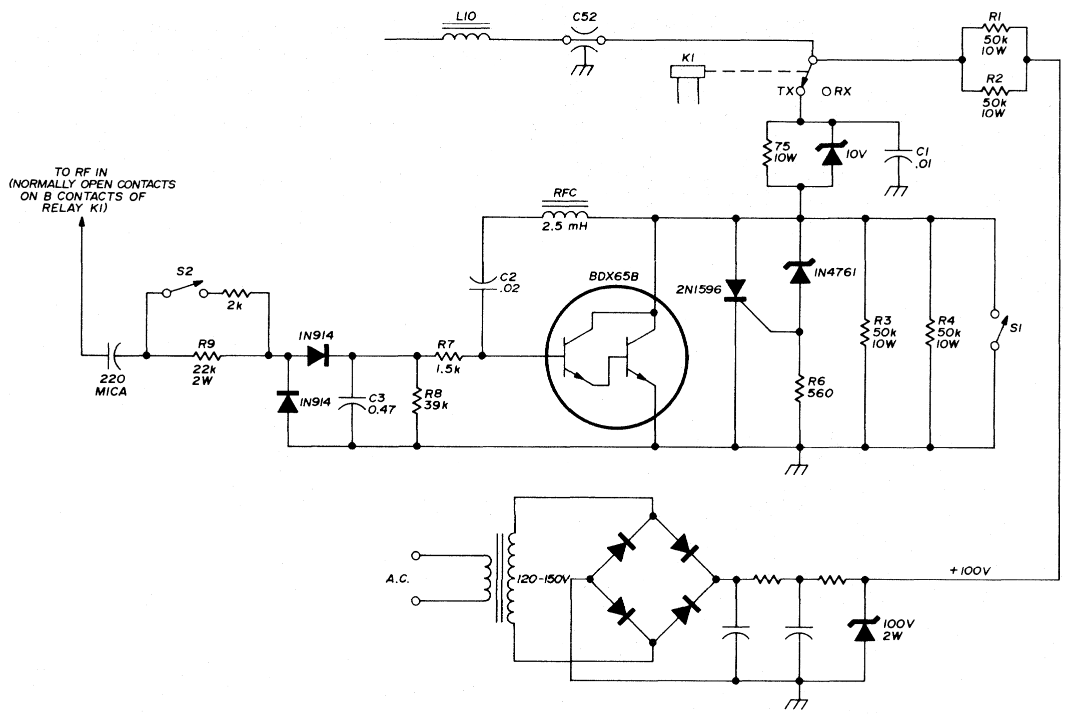


fig. 2. The electronic bias system circuit that can be added to the Henry 2K4/3KA linear amplifiers – an easy way to improve operation on CW or on ssb modes with or without signal processing.

to the plate using mica washers. All other components should be wired point-to-point on wiring strips. The rf choke (2.5 mH) in the integrator circuit between the Darlington transistor collector and base was necessary to prevent rf from getting into the base circuit and being rectified. This was the case with one particular unit in which the circuit functioned even without rf applied to the diode rectifier from the exciter input. Installing the rf choke cured the problem.

operation on cw

If you're a CW-only man, the time constants on the make and break side of the switch can be made much shorter. Change the 0.47-μF capacitor to 0.047-μF and the 20-nF to 2 nF. This will turn the amplifier on and off much faster and still further reduce tube dissipation between dots and dashes. Closing S1 allows you to run the amplifier in its original condition; that is, without the EBS circuit. It also allows you to check tube idling current.

operation on ssb

Using signal processing. One general disadvantage of running a great amount of clipping (10-20 dB) is the objectionable background noise that may occur between words, especially if you have some noisy fans or if room acoustics are not the best.

First it's essential that, when using speech processing, you speak very closely to the microphone (lips almost touching). When doing so the amount of clipping permissible with a processor such as the Magnum Six, Comdel, Vomax, DX engineering, or Datong, in a noisy shack, is determined by the acceptable signal-to-background noise ratio. An acceptable ratio is −25 *dB*. This means that, if your average power output with the processor, on a steady, stretched-out "Aaaa," is 1 kW (on your output meter), the background noise should be no more than 3.2 watts on the same output meter (that is, 25 dB down from 1000 watts). Adjustment procedure:

1. Switch off the EBS circuit by closing S1.

2. Tune up the amplifier in the normal way.

3. Adjust the clipping level in the prescribed way, but certainly no iurther than to the point where the background noise, as indicated on your

wattmeter, is at least 25 dB down from your steady "Aaa. . ." (3.2 watts versus 1000 watts).

4. Open S1, without changing any setting to your transmitter or processor. If the background noise (at – 25 dB or better) trips the EBS circuit (meaning that if you see the background on the scope or output meter, or if your plate meter does not drop to zero) then the EBS-circuit input sensitivity is too high. Increase the value of the 22k resistor until you find a value where the background noise *just* does not trip the EBS circuit. This value should be such that a drive of a little higher than – 25 dB (say 5 watts or – 23 dB in our example) turns on the EBS circuit.

If, when first switching on the EBS circuit, your – 25-dB background noise does not trip the EBS circuit, *reduce* the value of R9 (22k), and determine the value where a – 25-dB signal will not trip the EBS circuit, while a – 23 dB signal does.

Once you've determined the correct value of R9, you've not only installed a good working power saver but have achieved *total elimination of all bothersome background noise,* while running 15 or 20 dB of processing in a noisy environment. Nobody (especially the locals) will hear the blowers and accuse you of running excessive power!

Ssb using no signal processing. Using the EBS circuit with a value of R9 as determined above but driven by a nonprocessed ssb drive signal will result in too-low sensitivity of the input circuit. This will cause the circuit to switch on syllables. The result will be a distorted signal (similar to a vox dropping in and out while you talk because of too-short vox delay).

To work properly, the value of R9 must be decreased until switching does not occur between syllables. The best way to find out is to listen to your own signal using headphones and adjust R9 until the breaking up on syllables disappears. A value of 1.5 - 3.3k seems to be a good starting value.

If you want the EBS circuit to be fully flexible for both ssb modes (processed and nonprocessed ssb drive signals), a small switch (S2) or relay can be installed, which switches a second resistor in parallel with R9 to reduce its value when operating with a nonprocessed drive signal.

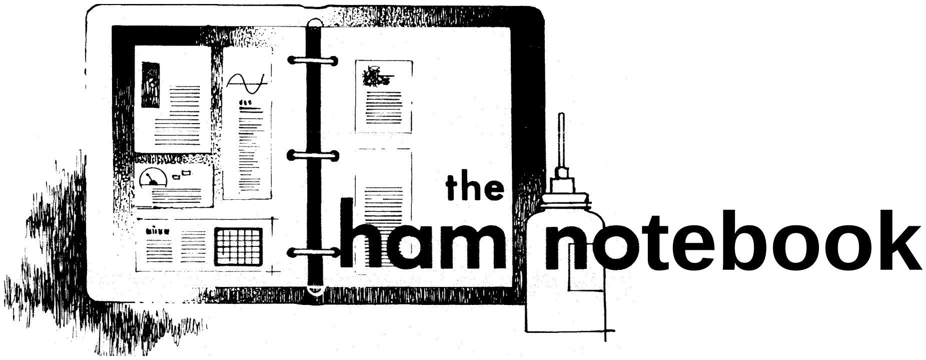
references

1. Marv Gonsior, W6VFR, "Electronic Bias Switching for Linear Amplifiers," *ham radio,* March, 1975, page 50.
2. J. A. Bryant, W4UX, "Electronic Bias Switching for RF Power Amplifiers," QST, May, 1974, page 36.

bibliography

1. John A. Devoldere, ON4UN, "Improved Performance from the Drake R4-B and T4XB," *CQ,* March, 1976.

ham radio



rejuvenating transmitting tubes with thoriated-tungsten filaments

Many amateur high-powered linear amplifiers are designed around the popular Eimac family of tubes, such as the 4-250A, 4-1000A, and the 3-500Z. All these tubes use thoriated-tungsten filaments. All other things being equal, the life of these tubes depends on the filament, which should be treated with care if you expect your tubes to last.

Filament emission is a complex process. As the chemical composition of the filament changes, the electron emission changes. As soon as the tube is turned on, it starts to lose electron emission, which finally drops below a value determined to be the "end-of-life" point. This process generally takes several thousand hours.

Once the end-of-life point is reached, the filament's chemical composition is so changed that nothing can be done by the user to restore the filament emission. The tube is then said to be "decarburized." The ditungsten carbide on the filament surface has thus evaporated or has combined with residual gas, and the carbide surface layer on the filament is gone.

Theoretically, it's estimated that a four per cent increase in filament voltage will result in a 20K increase in temperature, a 20 per cent increase in peak emission, and a 50 per cent decrease in life because of filament carbon loss. This, of course, also works the other way. For a small decrease in temperature and peak emission, life of the filament carbide layer, and hence the tube, can be increased substantially.

For the Eimac 4-1000A and other tubes of this filament voltage, broadcast stations run the tube at 7.2 volts instead of 7.5 volts. The reason is extended life. The 3-500Z filament should be run at 4.8 volts instead of 5.0 volts, and so on. The filament voltage should be checked with a 1 per cent meter to achieve these values.

If the tube filament is contaminated, or if electron emission is otherwise inhibited (perhaps a grid has been overheated and has liberated gas, or filament chemistry has been upset by running the filament at a *very low* voltage), the tube can be rejuvenated by increasing the filament voltage by about 15 per cent and running it for a time at this overvoltage (filament power only; no other voltages on the tube elements). This filament overvoltage action will cause emission material in the filament to "boil" out from the filament interior and form a new emissive surface.

The "cooking" time depends on the filament condition – the time may be only a few minutes or it may be longer. The only way to tell is to test the tube at intervals for emission. If the tube has been cooked properly, and the filament is in the right condition chemically to begin with, normal electron emission will be restored.

If you have a power tetrode or triode tube that has lost filament-emission (evidenced by decreased power output), it's certainly worth a try to get the tube back to near-new condition. Make sure that you meter the "cooking" circuit properly and that adequate cooling for the tube envelope and filament connectors is allowed.

These large tetrodes are expensive to replace, and you haven't anything to lose by cooking the filament of one that's lost emission. However, don't expect miracles. If your linear has used tubes, you probably don't knew the history of the tube's operation and the cause of filament emission loss. If's worth a try, though, and you may be pleasantly surprised.

Alf Wilson, **W6NIF**

audio rolloff

Many people find that their external *Touch-Tone** encoder will not access some systems. Many times, this is not the fault of the radio or the encoder, but actually the interface between the two units. What often occurs is that the signal from the encoder is connected into the audio input. Most radios incorporate a small-value capacitor (0.001 to 0.0033 μF is typical) between the microphone input and the first audio IC. This capacitor rolls off the low frequencies from the *Touch-Tone* encoder, yet passes the high frequencies relatively unattenuated.

One possible solution to this problem is to change the value of the input capacitor to 0.1 μF. If this is not practical, another remedy would be to directly inject the signal from the encoder into the input of the first audio IC. In this case, connect a 0.01 μF capacitor between the encoder and the IC. The capacitor should be mounted as close to the circuit board as possible to preclude any problems with rf getting into the audio stage.

Joe Olivera

"*Touch-Tone* is a registered trademark of the American Telephone and Telegraph Company.

tester for 6146 tubes

Since many popular exciters and transceivers use 6146 tubes, and since it is not easy to find a tube tester to accommodate this tube, there is a need for a simple tester to evaluate the condition of 6146 transmitting tubes. This is particularly important when speech processors are used — they tend to raise the average power input, thus shortening tube life. The circuit shown in **fig. 1** uses junk box parts, but will provide a very acceptable 6146 tube tester.

In this tester an ac bias for the grid is provided from the filament winding. It must be polarized. It must be polarized properly, *i.e.*, the grid must be going positive as the plate is going positive. To check this, reverse the filament connections. Choose the one which yields the greatest plate current. The tester is then ready for use.

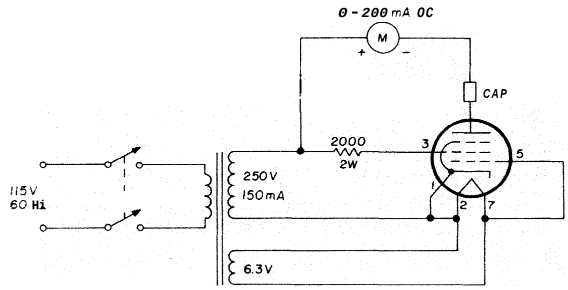


fig. 1. Simple tester for 6146 transmitting tubes is easily built from junk-box parts.

A good tube will draw 115 mA or more as indicated on the meter. Note that this meter indication is the average of half-wave rectified current. Tubes providing 90 mA or less should be discarded or, at most, kept for emergency spares. The tester is also useful for balancing pairs of tubes.

Gary Liegel, W6KNE

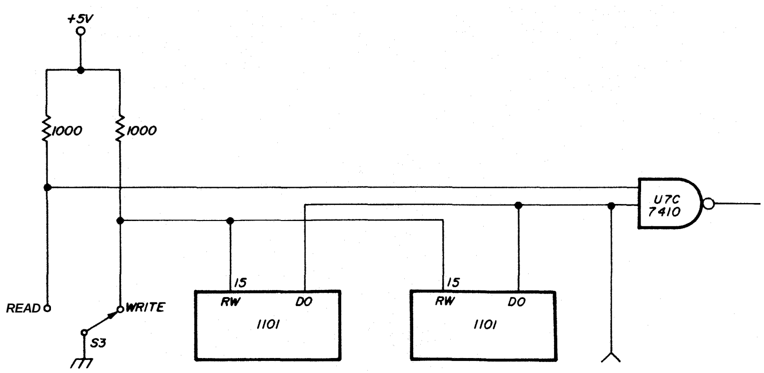
---

programmable accessory for electronic keyers

Since completing the programmable accessory for electronic keyers, August, 1975, *ham radio,* I've struggled to get it operational with my WB4VVF keyer,[1] achieving only intermittent success. The problem has always centered around the memory address and the READ/WRITE control line.

As I've discovered, the READ/WRITE line of the memories does not have to be synchronously pulsed with the address locations, merely taking the R/W line to +5 volts during the READ is sufficient. Therefore, since it turned out that the clock pulse from Q2's collector will directly drive U9A, both U8 and U11A are no longer required. The READ/WRITE switching is still done with S3 as seen in **fig. 2.**

Another problem was that the output pulse from U11B was fast enough to feedthrough the first binary counter in U12 and trigger the second binary counter simultaneously. This prevented full address of the memories. Bypassing pin 4 of U11B with a 1000 pF capacitor cured the problem.



Since programmable memory aa dress was not required, 7493s were substituted for 74193s. Additionally, sockets must be changed from 16 pin to 14 pin. The 7493 is somewhat cheaper and more available from suppliers.

John M. Korns, K9WGN/WØUSL

1. James Garrett, WB4VVF, "The WB4VVF Accu-Keyer," QST, August, 1973, page 19.

---

simple frequency counter

The frequency counter described by K4JIU in February, 1978, *ham radio,* page 30, has proven to be a simple, but useful, design. Unfortunately, after building the counter on the board supplied by Mr. Bordelon, the counter wouldn't operate above about 30 MHz on the 50-MHz range, or above 300 MHz on the 500-MHz range. Discussions with the author indicated that the problem probably revolved around the waveform presented to the 7208. The Intersil data sheet stated that the optimum input waveform should have a 50 per cent duty cycle. This is the case in the 5-MHz range. But, when using the 74196 prescaler, the Q_D output has an 80 per cent duty cycle.

One possible cure is to use the Q_C output from the 74196 to drive the counter. This will give a duty cycle of 60 per cent. This change also requires that the nonscaled 5-MHz input be loaded through Data Input C instead of Data Input D. The change is accomplished by cutting the foil runs at pins 12 and 13 of U3 and using pieces of insulated wire to connect the foils to pins 2 and 3 respectively. After the change, there will be no connections to pins 12 or 13.

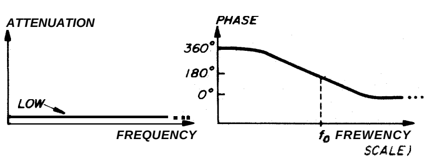
Carroll Hamlet, W2QBR

phasing networks

Dear HR:

In regard to VK2ZTB's article summarizing ssb phase shift networks in the January, 1978, issue, several comments are in order. First, a review of the many existing broadband audio phase-shift networks is fine, but the underlying theory ***common*** to each should also be presented.

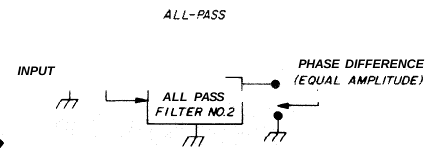
In general, the type of networks employed are called all-pass filters (the attenuation is constant, and the phase changes montoroically with respect to frequency, over the entire frequency band of interest). All-pass filter characteristics:



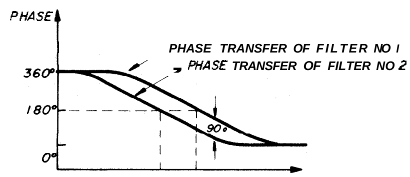
$$\frac{V_{output}}{V_{input}} = \frac{-\omega^2 - j\,\frac{\omega_o}{Q}\,\omega + \omega_o^2}{-\omega^2 + j\,\frac{\omega_o}{Q}\,\omega + \omega_o^2}$$

where $\omega = 2\pi f$ and ω_o and Q are constants.

If two all-pass filters are constructed with the proper Q and ω_o (for each), an audio phase shifter for ssb generation results. This is done as follows:

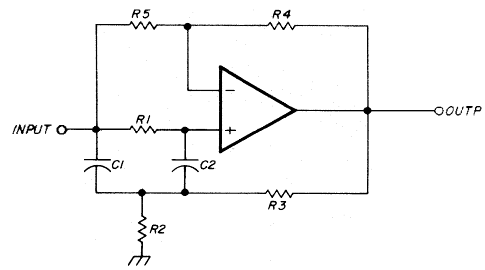


The Q of both filters is chosen to be 0.2252. ω_o of all-pass filter 1 is 2π (428 Hz); ω_o of all-pass filter 2 is 2π (2104Hz).



Note that even though the phase shift of each filter changes with frequency, the phase ***difference*** between the two outputs is 90 degrees over a wide band of frequencies.

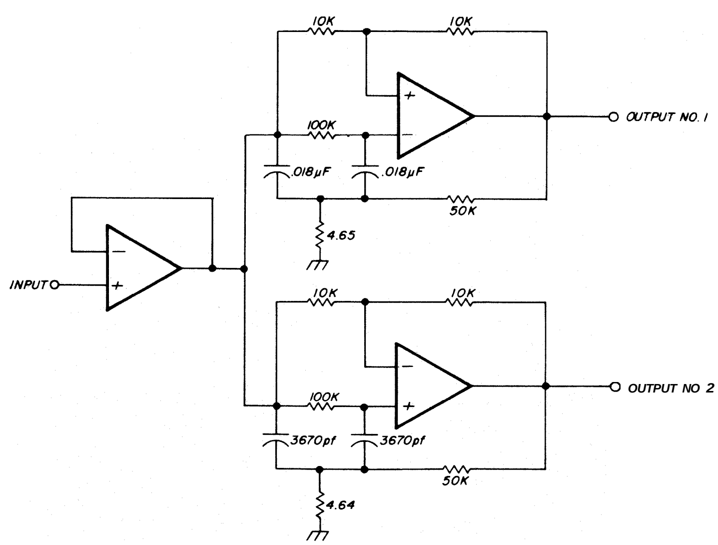
No active filter examples were included in the article (VK2ZTB's **fig. 11** uses op amps, but there are no RC networks in the feedback path). Why not use the more state-of-the-art active filter approach? One realization of this type of circuit uses two Steffen all-pass filters.



The general transfer function for the Steffen circuit is:

$$\frac{V_{out}}{V_{in}} = \frac{-\omega^2 - j\omega\left[\frac{R_4}{R_5}\left(\frac{1}{R_2}+\frac{1}{R_3}\right)\frac{1}{C_1} - \frac{1}{R_1C_1} - \frac{1}{R_1C_2}\right] + \frac{1}{R_1C_1C_2}\left(\frac{1}{R_2}+\frac{1}{R_3}\right)}{-\omega^2 + j\omega\left[\frac{1}{R_1C_2} + \frac{1}{R_1C_1} + \left(\frac{1}{R_2}+\frac{1}{R_3}\right)\frac{1}{C_1} - \frac{1}{R_3C_1}\left(\frac{R_4}{R_5}\right)\right] + \frac{1}{R_1C_1C_2}\left(\frac{1}{R_2}+\frac{1}{R}\right)}$$

The resulting audio phase shift network for ssb generation (300 to 3000 Hz) is as follows:

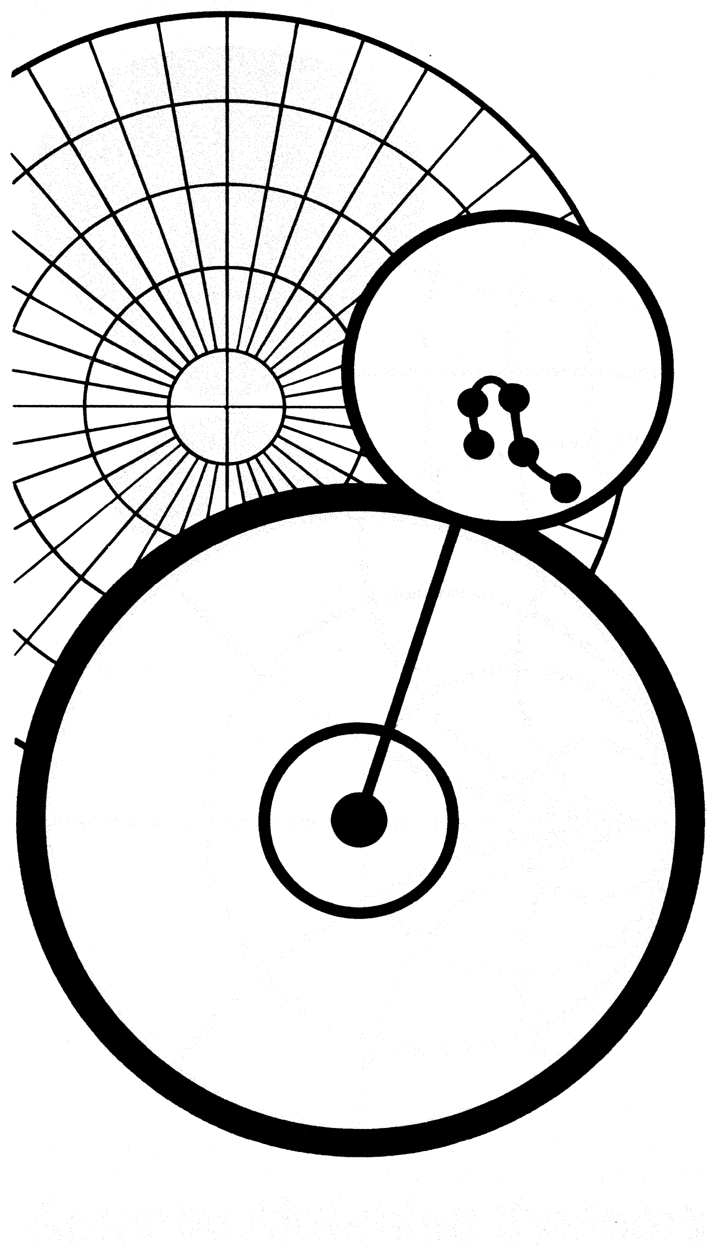


Tom Apel, WB9YEM
Madison, Wisconsin

$1.50

radio

SEPTEMBER 1978



ham radio magazine

SEPTEMBER 1978

volume 11, number 9

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

United States: one year, $12.00
two years, $22.00; three years, $30.00

Canada: one year, $14.00
two years, $26.00; three years, $36.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

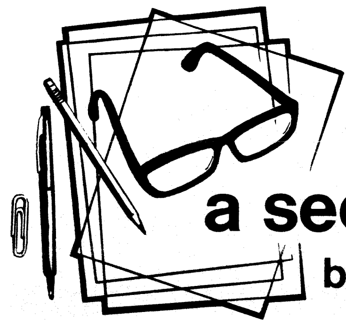
Foreign subscription agents are
listed on page 117

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

The editor of an amateur radio magazine must wear several different and diverse hats. In fact, I could use up several pages describing all the details that need attention to keep the magazine running smoothly. However, I'd like to talk for a moment about one very important task that sets the tone of the magazine: selection of articles.

Most of the articles published in *ham radio* are contributed by readers who want to share an idea or the details of a particularly successful project. Authors range from enthusiastic hams who have never written anything more than a short story for their English professor to fellows with engineering backgrounds who make their livings in front of a typewriter; all want to share an idea. I welcome the output of anyone who is interested in contributing something that will benefit all amateurs.

Prospective authors often ask, "What kind of articles are you looking for?" That's difficult to answer because new manuscripts arrive in the mail every day, so our needs are continually changing. However, I'm always on the lookout for simple construction projects that can be put together in one or two weekends. Larger projects are also welcome, but most of our readers must divide their leisure time between amateur radio and other interests, so they don't have time to build a Chinese copy of a complex piece of equipment. I'm also interested in good technical articles that cover any facet of radio communications.

Although I can accept only technical or construction articles for *ham radio,* articles for our sister publication, *Ham Radio Horizons,* can focus in on any aspect of amateur radio. In addition to simple construction projects for the beginner, I'm looking for DXpedition travelogues, adventure tales, articles on getting started in ham radio, and human interest stories with an amateur radio theme. If you have something you think would be of interest to our *Horizons* readers, I would like to have the opportunity to consider it for publication.

Once a month we set aside one or two days to go over all the manuscripts that have come in during the previous month. Since I seldom use more than a dozen articles in any issue, I don't accept more than that during any one-month period. This is sometimes a nearly hopeless task because there may be three-dozen or more to be considered. The first things I look for are originality and interest value. If the contribution passes this test, the next thing I look for is technical accuracy and attention to detail.

The contributed article doesn't have to be a literary masterpiece to be accepted. If you have a good idea and it's well documented, if the illustrations and technical discussion are clear and accurate – you may have a winner! On the other hand, if the article rambles from one topic to another, or presents inaccurate or misleading information, you will receive a rejection slip.

If your article has been accepted for publication, don't expect to see it published in the very next issue. The production times for a monthly magazine are probably much longer than you ever imagined. The articles for this issue, for example, were being prepared for publication during the month of May. As you are reading this we are putting together material for the December issue of *ham radio.*

In addition to writers, I am always on the lookout for new and unusual ways of *looking* at amateur radio, especially for *Ham Radio Horizons.* If you are an illustrator, painter, photographer, or sculptor, I would like to have an opportunity to review drawings, slides, or pictures of artwork for possible future consideration. Media can include airbrush, pen and ink, wash, watercolor, oils, collage, paper sculpture, *ad infinitum;* subject matter includes full-color covers and lead artwork for articles covering every aspect of amateur radio: antennas, hamshacks, satellites, vhf fm, DXing, field days, etc. Obviously, all artwork which is submitted for publication must be of professional caliber. Artwork done on consignment involves preliminary layouts or sketches, and must follow a regimented deadline schedule. If the prospect interests you or any part-time Rembrandts you know who enjoy amateur communications, write directly to our art editor, Jim Wales, for more information.

Jim Fisk, W1HR
editor-in-chief

calculator design

Dear HR:

I just have to tell you how much I've enjoyed and appreciated your articles in the last six months or so on various kinds of design approaches suitable for use with pocket calculators.

The Whyman pi network article in the September, 1977 issue; the Anderson circuit analysis article in the October issue; and the Ball satellite tracking article in the February issue were very good indeed; and you really hit the jackpot with the Anderson, Hoff, McNally-Keen, Fisk, and MacCluer articles, in March, 1978 ***ham radio.*** While I'm still digesting this last batch, you come along with several more juicy items in April and May; so unless you slacken up a bit, I may never catch up.

I acquired a TI-59 calculator just about the time this string started and have really enjoyed the task of digging into all the design methods and programming them for my TI machine. Having to "translate" the HP-25 program listings into the corresponding algorithms only added to the spice. And, with the 480-step/60-memory capacity of the TI-59, there was plenty of room to expand and improve the programs. I recall that one of the articles mentioned that the calculations were awkward to handle on non-RPN calculators, but I found them to be very straightforward and easy on the TI–59.

It is a bit ironic that I no longer have any real need for any of this. I retired from my engineering career about three and a half years ago at age 67, so have no use for it professionally. I got back on the ham bands at about the same time (after being off some forty-four years), but am not particularly interested in doing my own building or design work. There is just a very high degree of simple intellectual pleasure in finding out what really goes on in networks, transmission lines, and the like.

The orders of magnitude improvement in the speed and accuracy with which one can do complicated calculations can really make all those equations in our musty old engineering textbooks come alive in a way they could never do when one had to struggle with manual calculations, slide rules, and log tables. What an incalculable boon the calculators must be to young engineers just starting out and facing the awesome potential range of engineering knowledge and techniques.

These marvelous little calculators can also be used in ways which were never contemplated by their designers. I've come up with one such use for my TI–59, a fairly simple little program that will convert the TI-59 into an ID timer to let me know when it is time to send my call sign during a QSO.

Robert F. White, W6PY
Palo Alto, California

noise bridge construction

Dear HR:

I have just finished constructing and compensating the noise bridge described in the February, 1977, issue of ***ham radio.*** I had not expected that a major source of trouble would be the 180 pF capacitor. First, I used a ceramic capacitor, since I assumed it would be the least inductive. But, after a long and often disappointing search for the source of the erroneous behavior of my bridge, I found that this capacitor had 1 or 2 ohms of effective series resistance. Since the equivalent parallel resistance at 28 MHz, was only 500 ohms, the R_p reading was too low at the higher frequencies. Several different types of capacitor were tried, but most of them gave the same poor results. The only good results were obtained with a Philips 82 pF ceramic capacitor, and later with a 180-pF "micropoco" capacitor of the same brand. The latter type is a miniature, tubular-molded film capacitor with a polystyrene dielectric.

With this capacitor, the compensation of the bridge was completed as described, It certainly pays to give some attention to this detail beforehand, as it took me quite a long time to find out what was wrong with my bridge.

I built the series range expander in a UG-260/U and a UG-1094/U screwed together. The dielectric and the center conductor of the UG–1094/U have to be shortened, and a little hole is drilled into the remaining part of the center conductor. The two are locked with the large nut from the UG-1094/U.

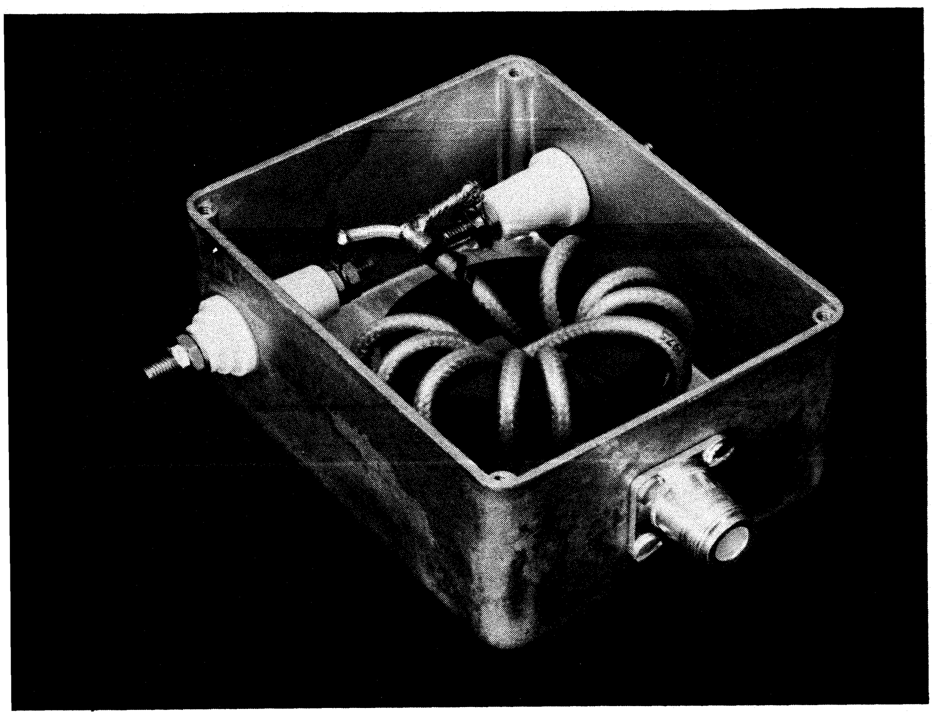
Arjen Raateland, OH2ZAZ
Helsinki, Finland

micoder matching

Dear HR:

In your May, 1978, issue of ***ham radio,*** Wesley Johnson presented a circuit in the ***ham notebook*** for impedance matching of a Heath HD-1982 micoder. It works beautifully. However, he specifies miniature Calectro coupling capacitors. Like many hams, I just couldn't find these parts anywhere. I eventually substituted 0.47 μF and 4.7 μF Radio Shack tantalum caps; they are small enough and can easily be purchased. I made a 2.5 x 2.5 cm (1 x 1 inch) printed circuit board and as he described, it fits perfectly between the top mounting posts and the case. Using the micoder with a Yaesu FT227R yields very good audio reports.

R. A. Stellarini, WB8VUN
Canton, Ohio



simple and efficient broadband balun

Construction of a new, improved balun which introduces no reactive components to the antenna feed system

Balanced versus unbalanced, balun or no balun – how many times have you heard long, philosophical discussions on this subject? I must admit I have had many sessions on the subject myself. Some of the pros and cons on baluns will be discussed in this article; then a new and improved broadband balun design will be described.

Balun vs no balun

It seems obvious that a ground-plane antenna is an example of an unbalanced antenna and therefore can be fed directly with an unbalanced feedline such as coaxial cable. It also seems obvious that a center-fed dipole is a balanced antenna and therefore requires a balanced feed system. This could be twin lead, open-wire line, or a balun fed with coax cable.

Judging from discussions heard on the air, it isn't obvious why a balanced feed system is required or what it "buys" the user. Reading the advertisements of some manufacturers could lead you to believe that a balun is required to prevent TVI or to lower your vswr. This is nonsense, as you will see later.

By Joe Reisert, W1JR, 17 Mansfield Drive, Chelmsford, Massachusetts 01824

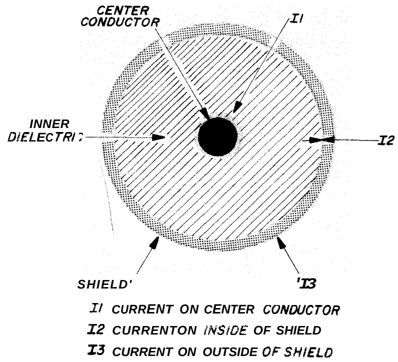


fig. 1. Three-wire representation of a coaxial feedline.

Over the years, coaxial feedline has become very popular while open-wire lines have practically vanished. One erroneous and often-heard story is that open-wire lines radiate, while coax does not; this is not true. The main reasons coax cables are popular are that they do not require special mounting techniques and they are easily monitored for both power and vswr. This has often led to the use of coax cables to directly feed a balanced antenna, especially on the amateur bands below 14 MHz.

There is a simple way to look at this situation;[1,2] a coaxial cable can be viewed as a three-wire feed system since the skin effect will allow one current to flow on the inside of the shield and another, a different current, on the outside (see **fig. 1**). The current on the inside of the shield is equal and opposite to the current flowing on the center conductor. The current on the outside of the shield, however, is a function of the currents induced by the field of the antenna (see **fig.** 2). This current is affected primarily by the geometry; the least current on the outside of the coax occurs when the feedline is at a right angle to the antenna, a rather unlikely situation.

How does feeding a balanced antenna directly with a coaxial cable affect the antenna's performance? The most obvious point is that when a horizontal antenna is used, the feedline re-radiation, if present, will most likely be in the vertical plane. Hence, some of the transmitted signal will also be radiated vertically. Below 14 MHz this probably won't degrade performance because the antenna pattern is probably already distorted — a little vertical radiation may fill in some nulls as well as radiate some power at lower angles. Another effect will be to cause rf feedback or a "hot" rig.

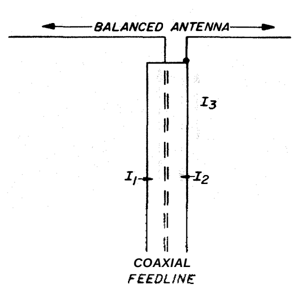


fig. 2. Current on the outside of the coaxial shield, I_3, is affected by relationship of the feedline to the antenna. I_3 is minimum when the feedline is located at right angles to the antenna.

When balanced, directional antennas are fed with coax, however, especially on 14 MHz and above, the feedline re-radiation as described above can be devastating, since it may cause undesirable high-angle signal off the back of the antenna or strong, local, vertically polarized signals which interfere with weak DX signals. Therefore, if coax cable is to be used to feed a balanced antenna, always use a balun.

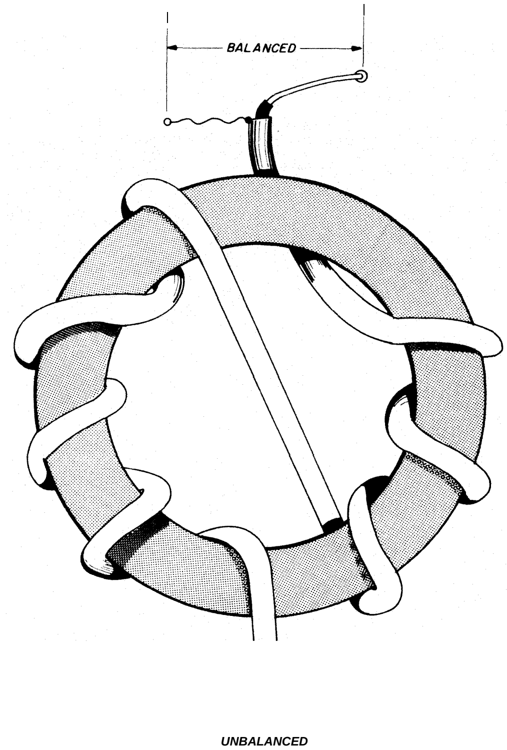


fig. 3. Construction of the improved broad-band balun.

balun types

The most common baluns are the toroid and ferrite-rod types.[1,3] The biggest problem with all these baluns is that they are all frequency sensitive to one degree or another, especially if a wide frequency range (3 to 30 MHz) is desired. Personally, I have had the best luck with the ferrite-rod type.

There are several other problems with toroid and ferrite-rod baluns: they all have some loss and, if not perfectly constructed, can introduce a mismatch; a

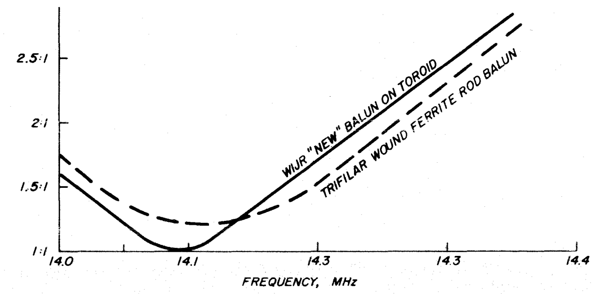


fig. 4. Measured vswr of a TH6DXX antenna on 20 meters with a ferrite-rod balun compared to vswr with the improved coaxial balun. The ferrite-rod balun contributes reactive components to the feed system, as discussed in the text.

less understood problem is the effect of using a toroid or ferrite-rod balun with a mismatched antenna, a common situation since a 1:1 vswr is only present at one discrete frequency on any band. This led me to search for a new type balun which did not suffer from these problems.

new type of balun

In the search for a better balun, I studied the coil of coax approach which is recommended by several antenna manufacturers. This type of balun solves some of the difficulties with mismatch loss, since the impedance is constant, but the thought of using 6-10 meters (20-30 feet) of coax with its loss was not too appealing.

Looking at **fig. 2,** it seems that all we have to do is prevent rf current from flowing on the outside of the coax. In other words, we must devise an rf choke on the outside of the coax shield. This can best be accomplished by wrapping the feedline on a ferrite rod or a toroid core, but some external field will still be present.

Then I saw an article on a super toroid, one that had almost no external field. *Voila,* an answer to my prayer! I quickly wound some RG-58/U coax on a Micrometals T-200-2 toroid, but results were disappointing. I reckoned that an inductive reactance of 500 ohms on the outside shield should be a minimum requirement. But the permeability of this powdered-iron toroid is so low that 14 turns were required on 10 meters — the toroid was barely large enough for that many turns, and 40 turns were required on 3.5 MHz! Hence, a search was conducted for a better core.

Then I looked at Indiana General's *Ferramic* cores and noted that the permeability of their Q1 material was 125 and suitable for operation from 3.5-30 MHz. The F568-1 core* also has a larger inside diameter — 35.5 cm (1.4 inch) versus 28.5 cm (1.25 inch) for the T-200 toroid. With the higher permeability, a suitable balun could be built at 3.5 MHz with only 12 turns of coaxial cable. Also the strays are lower, since the balun is based on coax which has the same impedance as the antenna.

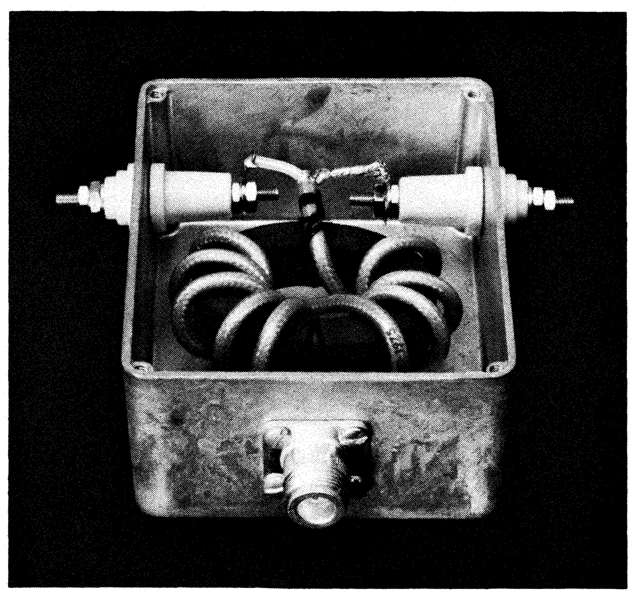
construction

The new toroid balun construction is shown on **fig.** 3. Note that an even number of turns is preferred, with half the turns going on one side and the other half in opposition. Note also that the ends of the coax are on opposite sides of the toroid, which is very desirable mechanically.

For a kilowatt balun for 3.5 to 30 MHz, the Indiana General F568-1 type core is preferred. RG-8/U coax is too large and RG-58/U is undesirable except for levels below 200 watts. A check of available coax cables narrowed the choice to RG-141/U, which is approximately 6.35 cm (0.25 inch) OD with a Teflon dielectric and can easily handle 1000 watts. However, RG-142/U is also acceptable but is quite costly, since it uses a double silver-covered braid. W2DU has also suggested RG-303/U. A 12-turn balun requires approximately 90 to 100 cm (36-40 inches) of coaxial cable.

Table 1 shows the minimum required turns versus frequency. Hence, a 10-turn coil would suffice for 7-30 MHz, but 12 turns are desired for 3.5-30 MHz and should also work at 1.8 MHz with slightly poorer performance. For lower frequencies the type TC9 core material would be a better choice with an appropriate number of turns as discussed above.

Photograph of the improved broad-band balun. The cast aluminum box is a Bud CU234 or similar.



*Indiana General toroids and coaxial cable are available from G. R. Whitehouse, 10 Newbury Drive, Amherst, New Hampshire 03031.

The finished toroid can be placed in an aluminum or plastic box. I mounted mine in a Budd-type CU234 cast-aluminum box with a connector at one end and two insulated, ceramic feedthroughs on the sides. The coax shield should be debraided at each end and then twisted for insertion in the connector and solder lugs as shown in the photograph.

performance

I reasoned that a simple lab test would be to terminate the balun in its characteristic impedance, measure the vswr, and then measure the vswr with a short from the center pin side of the load to ground. If the balun is truly balanced, the short circuit should not affect vswr to a large degree. Indeed, a 10-turn balun tested at 10 MHz showed only negligible change when either of the output leads was grounded. Similar results were noted at frequencies with proper turns. Loss was negligible.

table 1. Minimum number of turns *vs.* lowest frequency of operation for the improved broad-band balun (assumes 500 ohms reactance for 50-ohm system).

frequency	inductance	turns T-200-2	turns F568-1 Q1
3.5 MHz	22.74 H	40	12
7 MHz	11.37 H	28	10
14 MHz	5.69 H	20	6
21 MHz	3.79 H	16	4
28 MHz	2.84 H	14	4

The supreme test was to replace my ferrite-rod balun on a TH6DXX tribander to see if the vswr wouldchange. You will note from **fig. 4** that the original balun had a higher vswr at resonance and a somewhat lower vswr at the high end of the band than the new improved balun. This confirms that the ferrite-rod balun introduces a mismatch at resonance. The lower vswr at the high end of the band is probably due to the increased loss of the ferrite-rod balun, which tends to make the vswr look lower than it really is. Additional baluns are also now in use on a 160-meter dipole and several G5RV slopers on 80 and 40 meters.

other variations

This balun is not restricted to 50 ohms; indeed, any impedance coax could be used if the turns are calculated to yield a reactance at least 10 times the impedance. Smaller or larger cores or coax can also be used if lower or higher power is required. In addition, this type of balun is not frequency-limited to the high frequencies; it also works well at vhf if attention is paid to layout and lead length. The beauty of this type of balun is that it does not introduce any additional reactive components to the feedline.

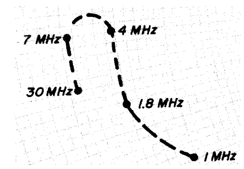


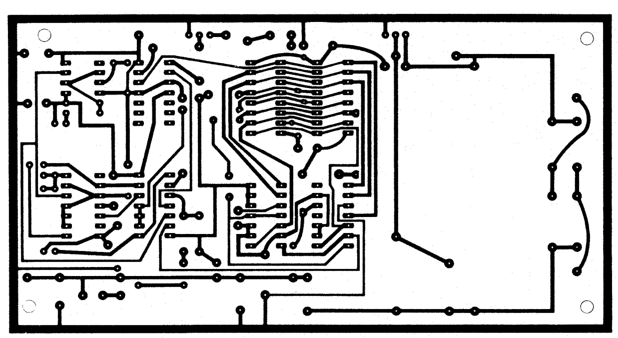
fig. 5. Smith chart plot of a 10-turn broad-band balun built by W1JR, as measured by DJ2LR over the frequency range from 1 to 30 MHz. Vswr is 1.3:1 or less on all high-frequency amateur bands.

I would like to express special thanks to Walt Maxwell, W2DU, and to Ulrich Rohde, DJ2LR, for their encouragement and assistance in preparing this material.

reference

1. Lew McCoy, W1ICP, "Is a Balun Required?" *QST,* January, 1968, page 28.
2. William Orr, W6SAI, "Broadband Antenna Baluns," ham *radio,* June, 1968, page 6.
3. T. A. O. Cross, "Super Toroids with Zero External Field made with Regressive Windings," *Electronic Design,* September 1, 1976.

ham radio



20 meter
delta-loop array

The use
of a scale model
and simple,
but effective,
construction techniques
produces an effective
delta-loop array

The cubical quad type of loop antenna has been a very popular and effective antenna for amateur use between 14 and 30 MHz. The Yagi antenna is probably the most popular of all amateur antennas, and both types have their avid defenders for the most "effective" antenna on a particular band. An excellent comparison of the two antennas is presented in an article by Lindsay, WØHTH.[1] There is another type of loop antenna, however, which is generally overlooked in amateur publications — the delta loop. The delta loop is simply another configuration of a full-wavelength loop antenna like a quad, but the delta loop offers certain advantages: plumber's delight type construction and extremely good vswr bandwidth.

There have been few articles written on either the design or the construction of the multi-element delta-loop antenna,[2,3] and none describe the antenna for use on 20 meters. After reviewing the available information, I concluded that the reasons for neglecting the delta loop on 20 meters were the physical size of the antenna and the resultant problems with successful construction. I hope that my results will bring to light a successful construction technique and a proof of performance that will encourage other amateurs to experiment with and use this antenna.

delta loop vs quad

Consider the current distribution of the familiar quad loop antenna shown in **fig. 1**. A current reversal occurs at the junction of each half wavelength section, so there is a current minimum in each vertical leg and a current maximum 180 degrees from the feedpoint. The electric field polarization of the quad loop is derived from the fact that the vertical components of the current elements produce radiated fields that tend to cancel each other, $y_1+y_2=y_3+y_4=0$, and the horizontal components of the current elements produce radiated fields that are additive, $x_1+x_2+x_3=x_0$, where x_0 is the effective field-producing current of the loop. The polarization of the electric field is a plane perpendicular to the plane of the loop. This same explanation of the properties of the full wavelength loop apply whether the loop is a square, a diamond, a circle, or a triangle.

For the case of the triangular or delta loop antenna, the current elements are shown in **fig.** 2. Again, the current reversal occurs at the junction of each half wavelength. Each current element can be broken into vertical and horizontal components. The vertical components of the current elements produce radiated fields that tend to cancel each other. Unlike the square loop, diamond, or circle, the delta loop produces a horizontal field component (proportional to x_1+x_4) which is 180 degrees out of phase with x_0,

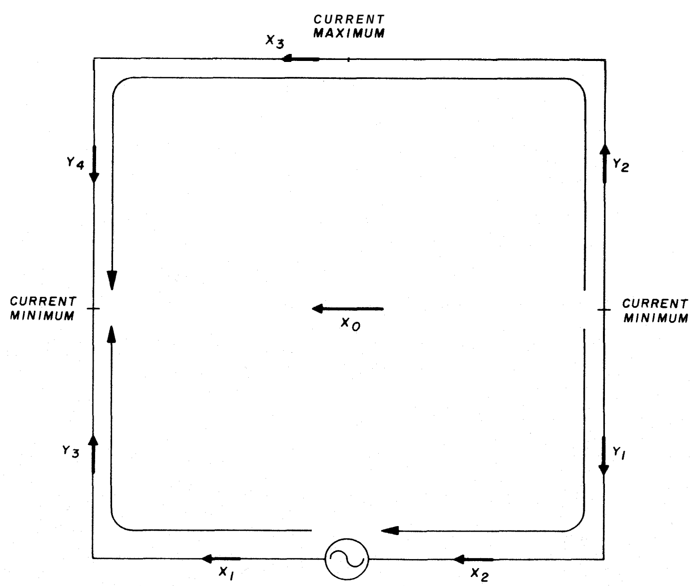


fig. 1. Current flow in a one-wavelength square loop produces two current nulls, one at the center of each vertical side. However, the horizontal current components do not cancel, producing horizontal polarization.

By Glenn Williman, N2GW, 145 Whalepond Road, Oakhurst, New Jersey 07755

the effective field current. Due to the geometry of the delta loop, however, the magnitude of this component, when compared with the magnitudes of all other horizontal components, yields the same overall effective field component as the other types of loops, x_0. Therefore, although the shapes of different full-wavelength loops vary, the effective field produced is of the same magnitude and of the same polarization when fed at corresponding locations. Admittedly, the preceding discussion is a simplistic view of electromagnetic field theory, but, without getting bogged down in the math involved, the concepts are adequate to subjectively describe the radiation of the loop antenna.

It is interesting to note that there is evidence that for a simple full-wavelength loop, the gain is approximately 4 dB above isotropic. In other words, assuming 2.15 dBi gain for a half-wave dipole, the full-wave loop shows a gain of 1.85 dBd (dB above half-wave dipole). This predicted gain differential has been experimentally supported by Lindsay's results.[1] The effect of this differential is to say that, for a given boom length, the loop parasitic array will exhibit a 1.85 dB gain above a Yagi array. Stating it another way, Lindsay's results show the array length of a Yagi must be about 1.8 times as long as the length of the loop parasitic array. This explains why an optimally spaced 2-element loop array is comparable in gain to an optimally spaced 3-element Yagi array.

antenna model

Since I was unfamiliar with the delta loop when I began this project, I decided the easiest way to experiment with a design for the antenna would be to build an operating scale model. A model design frequency of ten times the intended frequency reduced the dimensions by a factor of ten. I chose 147 MHz for the model because the dimensions were manageable and measurements could easily be made on 2 meters.

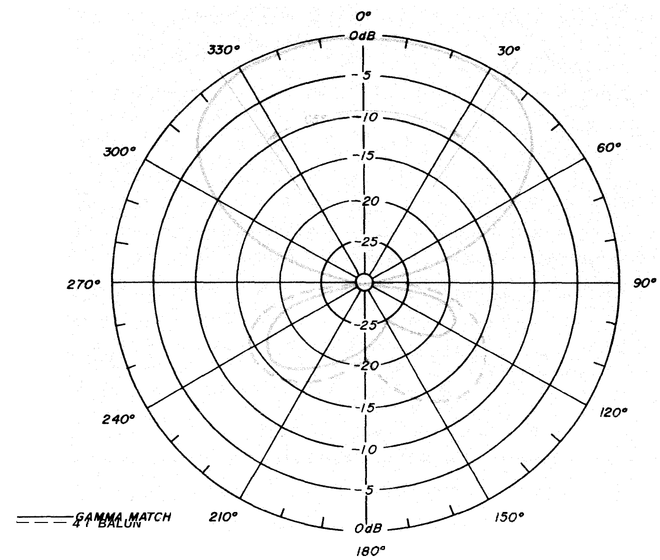


fig. 3. E-field patterns from the scale-model 2-element delta loop. This test was performed at 147 MHz and produced a front-to-back ratio of 22 dB and a front-to-side ratio in excess of 30 dB. The half-power beamwidth measured 65 degrees.

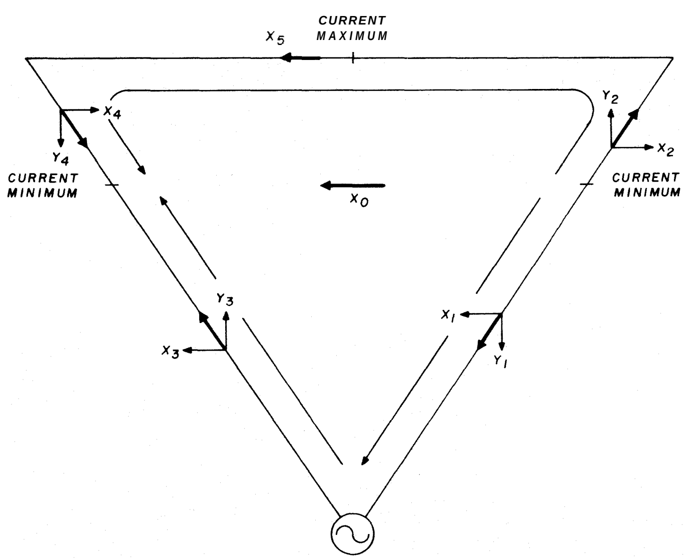


fig. 2. In a full-wavelength delta loop, the electric field is the same as that of the other configurations due to cancellation of the current components.

The generally accepted formulas for element length worked very well.

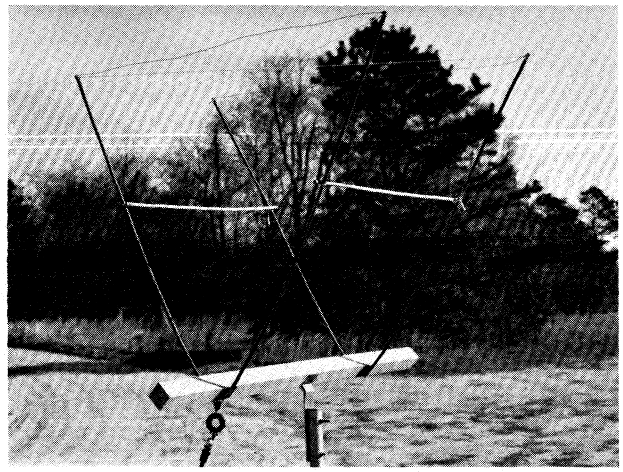
$$\text{driven element} = \frac{306.3}{f_{MHz}} \text{ meters}$$

$$= \frac{1005}{f_{MHz}} \text{ feet}$$

$$\text{reflector} = \frac{314}{f_{MHz}} \text{ meters}$$

$$= \frac{1030}{f_{MHz}} \text{ feet}$$

An element spacing of 0.17λ was chosen, as there was no noticeable difference between that and 0.2λ as seen using the model. Also, 0.17λ translated very closely to a 3.7 meter (12 foot) boom length on the full-size antenna. The model was constructed using 6.5 mm (1/4 inch) copper tubing for the sides of the loops and wire for the tops of the loops. The size of the copper tubing was a poor choice for the model because it presented an unrealistic element-diameter-to-wavelength ratio. But, as it turned out, the



Scale model of the 2-element delta loop. This model was used for measurements at 147 MHz.

performance of the model accurately predicted what I would eventually find with the 20-meter antenna.

Two methods of feeding the antenna were also tried. Measuring a single circular loop of wire showed the resistance to be between 140 and 150 ohms. With a 4:1 balun feeding the driven element, the vswr was about 1.4:1, as would be expected. A better match was obtained with a conventional gamma match, and a vswr of 1.1:1 was easily obtained. Both matching schemes provided an extremelv wide vswr bandwidth. At the design frequency of 147 MHz, the vswr remained below 1.5:1 for 18.4 MHz (– 9.4MHz to +9.0 MHz), or a bandwidth of slightly greater than 12 per cent. E-field patterns were also measured with both versions of the scaled antennas, with the results shown in **fig. 7.** Using laboratory grade test equipment, the gain was measured on both versions and found to be 8 dBd at the design frequency. I was very encouraged with these results, and while I worked on the design of the full-size version, the model was used on 2 meters (vertical polarization is obtained by rotating the array 90 degrees – on its side).

construction

Each delta-loop element consists of two arms of aluminum tubing joined together at one end of the boom, with a third arm of wire completing the loop. As determined from an unsuccessful construction attempt, the critical part of the design is a reliable method of attaching the delta elements to the boom

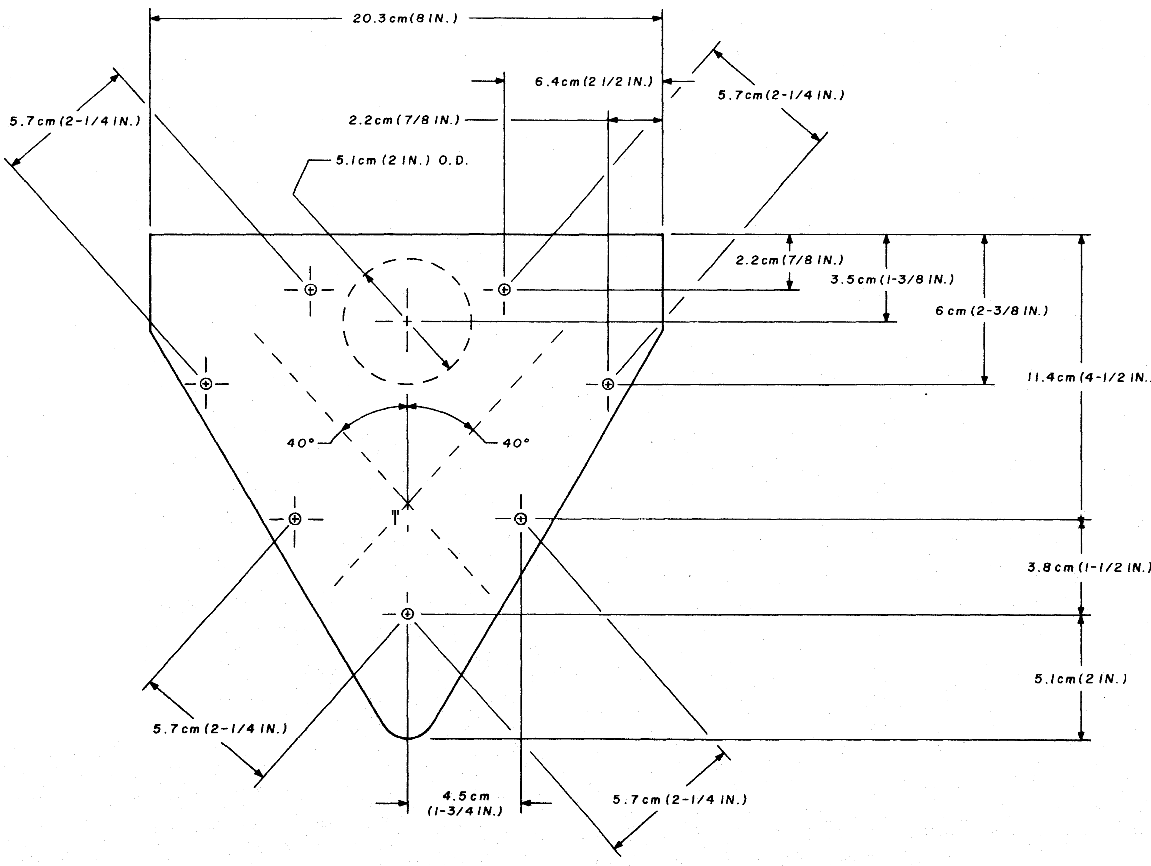


fig. 4. Drilling template for the element mounting plates.

– a method which will be stable in high winds and remain as light in weight as possible. I devised the following solution:

Two triangular aluminum plates are cut from 6.5-mm (114-inch) stock and shaped into the form shown in **fig, 4**. These plates are then welded onto the boom. (Any local welder with the ability to work with aluminum should be able to do the job inexpensively.) Care should be taken to ensure that both end plates are perfectly parallel and are not skewed around the boom. This would tend to make the assembled antenna unbalanced and will also degrade the front-to-back ratio.

Next, using **fig. 4** as a guide, holes for the galvanized pipe clamps are drilled in the end plates. The 19-mm (314-inch) pipe clamps work very nicely, and are a good deal less expensive than 25.5-mm (I-inch) U bolts. With respect to the end plate vertical centerline, each leg is angled 40 degrees, making a total angle of 80 degrees between the two legs. After both end plates are drilled and the pipe clamps are loosely mounted, one pair on each side of the plate, most of the work is done. All that remains is to make up the individual arms.

Each of the four arms is identical, consisting of a 2.7-meter (9-foot) section, a 2.4-meter (8-foot) section, and another 2.4-meter (8-foot) section, with respective diameters of 25.5 mm (1 inch), 22 mm (718 inch), and 19 mm (3/4 inch). Since this tubing is usually supplied in 3.7-meter (12-foot) lengths, the extra 1.2-meter (4-foot) section of 22-mm (718-inch) tubing is inserted into the bottom of the 25.5-mm (?-inch) tubing. One end of the 25.5-mm (I-inch) and 22-mm (718-inch) tubing is slotted, with hose clamps securing the three sections together. In addition, two sheet metal screws are used to secure the piece of 22-mm (718-inch) reinforcing tubing. All the aluminum should be type 6061-T6.

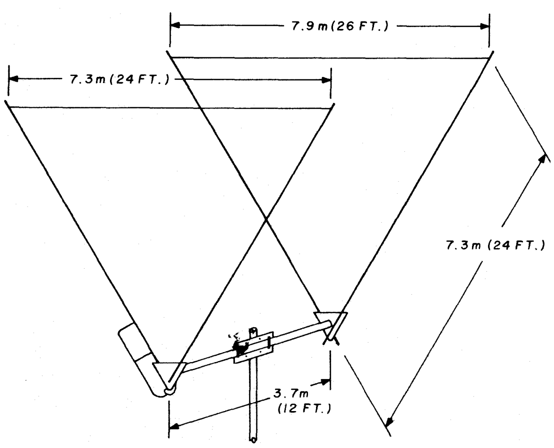
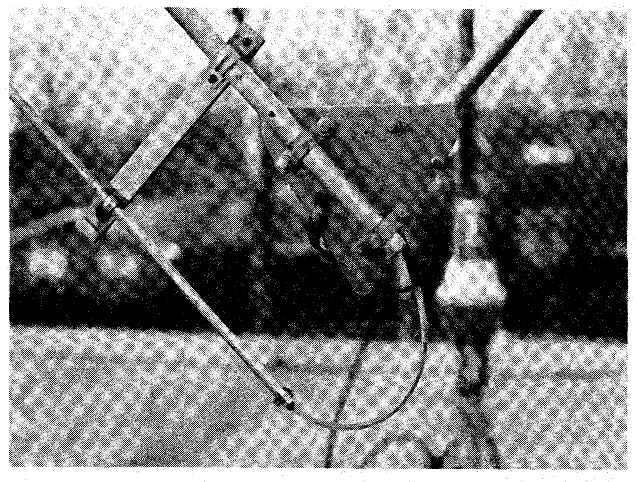


fig. 5. Final dimensions for the two-element delta-loop array.



Details of the gamma match system. Moving the tap point of the gamma arm and changing the amount of coax inserted in the gamma arm will allow you to adjust for minimum vswr.

After the four arms are completed, the top of each is drilled to accept a small eyebolt. The wire legs are then cut to the appropriate lengths and wrapped and soldered to the eyebolts. If the eyebolts are cleaned and well fluxed, a small blow torch will ensure a well-soldered joint.

The last item to be made is the plate that is used to attach the boom to the mast. This should be made of 6.5-mm (114-inch) aluminum or 3-mm (118-inch) steel. The boom is attached to the plate with 5.1-cm (2-inch) muffler clamps and the mast is attached using U bolts of the size required for the mast. The boom is a 3.7-meter (12-foot) length of 5.1-cm (2-inch) OD tubing with a 3-mm (118-inch) wall thickness, also of 6061-T6 material. **Fig. 5** shows the dimensions of the assembled antenna.

final assembly and testing

The antenna should be assembled either on the ground, using a ground-rigged mast, or in a position where it can easily be reached. I assembled mine on a I-meter (3-foot) roof tower, but only after I had proven to myself that everything would fit.

Fix the boom to the mast and tighten all the clamps securely. Then, with someone to help support the boom, insert one leg at a time into the end plate and tighten the securing clamps. After both legs are in place, the completed element will be well-balanced and self-supporting. Finally, move the other end of the boom and complete assembly of the other element. The basic antenna is now complete, and all that remains is to add the gamma match assembly shown in **fig. 6.**

The gamma-match assembly is a length of 9.5-mm (318-inch) aluminum tubing, slotted at one end and bent at the other to the dimensions shown. The outer braid of the coax feedline is stripped away (about 1 meter [3 feet]) and the center conductor, with its insulation, is inserted into the gamma rod. The center conductor of the coax and tubing form the gamma capacitor, with the coax center-conductor insulation forming the capacitor dielectric. The length of coax inside the gamma rod is adjusted for minimum vswr and then secured by a small hose clamp at the bottom of the gamma rod. This is a trial-and-error procedure, and successive lengths of the center conductor can be cut off until the right amount of capacitance is obtained for resonance. Best results will be obtained if the vswr is measured right at the antenna while the adjustments are being made.

One obvious advantage of working with the scale model is the ease with which performance tests can be made. The measurements are more difficult to make with the full-size array, but the agreement in results obtained between the model and the full-size version was very good, and convinced me of the performance of the full-size version. **Fig. 7** shows the vswr and the front-to-back ratio measured at a distance of about 300λ. The reflector is easily tuned by loosening the pipe clamps and sliding the tubing down, thus shortening the overall circumference of the reflector loop. The initial dimensions for the reflector were purposely made long so that tuning could be accomplished.

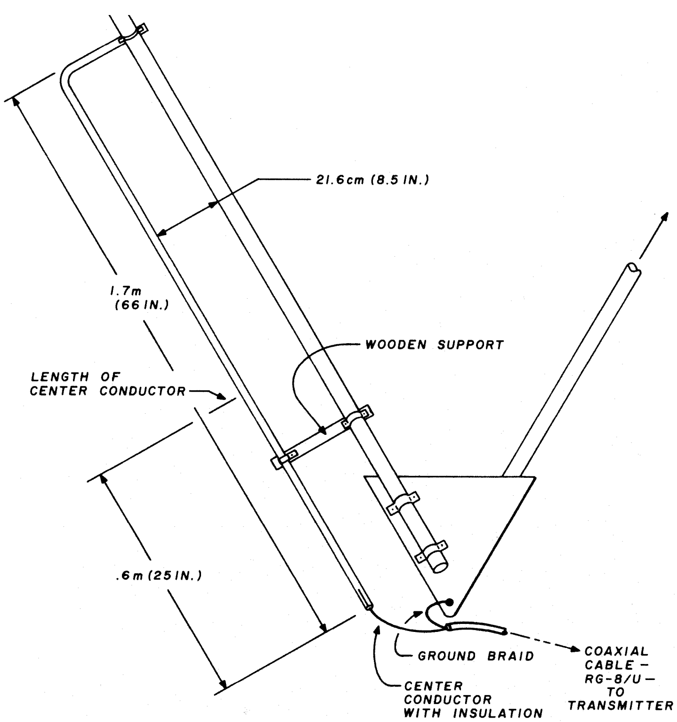
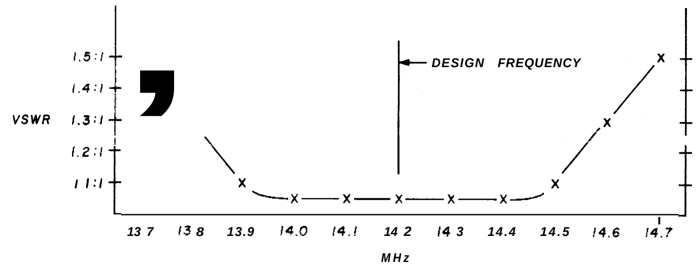
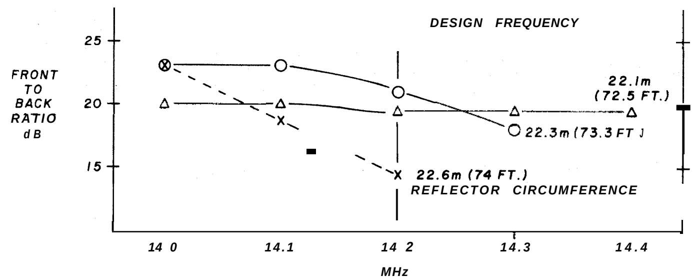


fig. 6. Details of the gamma match for feeding the delta loops.





Boom length	3.7 meters (12 feet)
Turning radius	4.3 meters (14 feet)
Weight	18.1 kg (40 pounds)
Wind load 97 km/h (60 mph)	36.3 kg (80 pounds)
Forward gain	8 dBd
Front-to-back ratio	>20 dB
1.5:1 vswr bandwidth	1 MHz
Impedance	52 ohms with gamma match
Half-power beamwidth	65 degrees horizontal
Cost	$100 price range

fig. 7. Front-to-back ratio and vswr curves for the full-size delta loops.

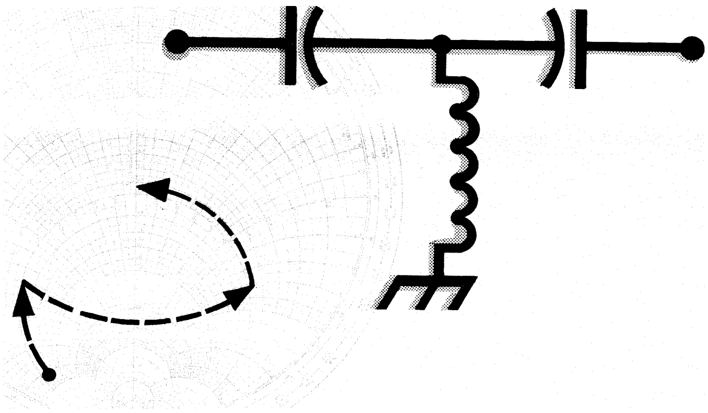
conclusion

The 2-element delta loop has been a real performer for me. It's allowed me to compete in several DX pileups and compare favorably with others using full-size 3-element Yagis and 2-element quads. Even after my linear was recently sidelined, I was still able to work out very nicely just using my exciter. Also, the low vswr allowed me to bypass my antenna tuner and still enjoy a vswr of less than 1.5:1 across the entire band.

references

1. J. E. Lindsay, Jr., WØHTH, "Quads and Yagis." *OST*, May, 1968, page 11.
2. Harry R. Habig, K8ANV, "The HRH Delta-Loop Beam," *OST*, January, 1969, page 26.
3. Lewis G. McCoy, W1ICP, "The Delta-Loop Beam on 15," *QST*, January. 1969, page 29.

ham radio



T-network

impedance matching to coaxial feedlines

Design of a five-band T-network for matching antennas to 50-ohm transmitters

The output circuits of almost all high-frequency transmitters, transceivers, and power amplifiers are designed for use with coaxial transmission lines. Most have a nominal impedance rating of 50 ohms; this means the equipment would like to see $50+j0$ ohms at the output terminals. The transmitter or amplifier is then matched to the antenna for any length of line; maximum power is transferred and loading is easy because the net effect is equivalent to connecting a 50-ohm non-inductive resistor across the output terminals. Reciprocally, this optimum condition also applies to receivers designed for 50 ohms input – maximum power transfer occurs from antenna to receiver.

It is not difficult to achieve the matched condition at a single frequency in a single high-frequency amateur band, but few amateurs limit their operation to a single frequency. Two simple antennas with low feedpoint impedance (when cut to resonance) that have an input resistance close enough to 50 ohms to make a reasonably good match with direct coaxial feed are shown in **fig. 1.** The quarter-wavelength vertical operated against a good ground has an input resistance near 35 ohms, and the half-wavelength center-fed dipole has an input resistance near 70 ohms. In both cases the standing wave ratio is less than 1.5:1 when 50-ohm coaxial cable is used and the tuning circuits in the transmitter can easily compensate for this small amount of mismatch.

By Jack M. Shulman, W6EBY, 789 Garland Drive, Palo Alto, California 94303

In practice there are two considerations which make the simple direct coaxial feed scheme of **fig. 1** unsatisfactory. First, many low feedpoint impedance antennas have an input resistance not equal to or even near 50 ohms. In **fig. 2A** for example, if input resistance R_a is 17 ohms, a 3:1 swr will exist at terminals **JK.** The impedance at the sending end of the line, terminals **AB,** will then be a function of line length,* and, depending on this length, may contain both resistance and reactance. **Fig.** 3 shows impedances at all line lengths away from antennas which have an swr of 2, 3, and 4. These are determined from the constant swr circles which cross 2.0, 3.0, and 4.0 on the resistance axis of a Smith chart.[1]

The second consideration is that operation over an entire amateur band is usually desired, not just on one frequency. When an antenna with a low feed-

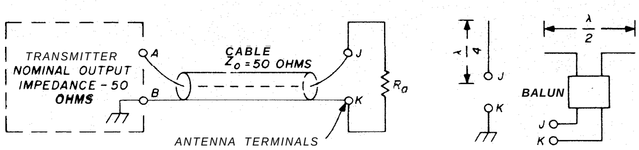


fig. 1. The quarter-wavelength vertical (above good ground) and half-wavelength horizontal dipole are two popular antennas which provide a suitable match to 50-ohm coaxial cable (R_a = 35 to 70 ohms).

point impedance is operated away from its resonant frequency, resistance changes at a slow rate but reactance departs from zero at a rapid rate as shown in **fig. 4.** This is discussed in detail in reference 2. For any given antenna the actual value of R_a and antenna reactance X_a above and below resonance will depend on many variables: the conductor length-to-diameter ratio, antenna height above ground, orientation (horizontal or vertical), number and spacing of elements (if more than one), and proximity to conducting objects. **Figs. 2B** and **2C** show the equivalent circuits of an antenna when operated above and below resonance. A typical set of swr measurements taken at terminals **AB** is shown in **fig. 5.**

Whatever the exact values of R_a and X_a for any low feedpoint impedance antenna at any particular frequency in a band, the important point is they are *low* — probably less than 200 ohms in all cases over the frequency range of interest. It's the job of the matching network to make some low value of anten-

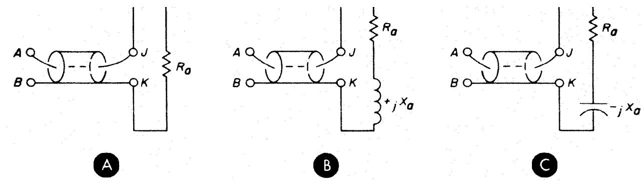


fig. 2. Equivalent circuit of an antenna at resonance (A), above resonance (B), and below resonance (C). With%-ohm coaxial line, if the antenna resistance R_a is not near 50 ohms at resonance, the impedance at AB is a function of line length and the match may be unsatisfactory with direct coaxial feed.

na impedance at terminals **JK** look like $50+j0$ ohms at terminals **AB.** The T network is ideally suited for doing just that. Moreover, the T network can be made extremely simple with only two active components, one coil and one variable capacitor. The operational fundamentals and design of such a network will be described in this article.

impedance matching fundamentals

Assuming that the impedance at the point on the line near **AB** where the matching network will be inserted contains both resistance and reactance, the matching process consists of two steps:

1. Inserting opposite reactance at the output end of the network to result in a net reactance of zero

2. Transforming the remaining resistance value to 50 ohms

The reactance initially present, as represented by

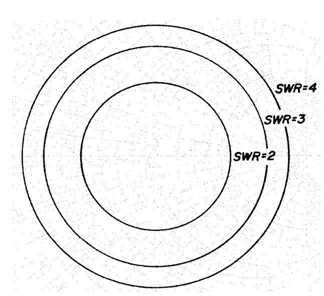


fig. 3. Effect of transmission line length on impedance for swr of 2:1, 3:1, and 4:1. Concentric circles are constant swr values of $R \pm jX$ along the transmission line.

*The impedance at the input of any transmission line is a function of the length of the line and the impedance at the load end. If the load resistance is purely resistive and equal to the characteristic impedance of the line, however, the impedance at the input end is equal to the load impedance, regardless of the length of the line.

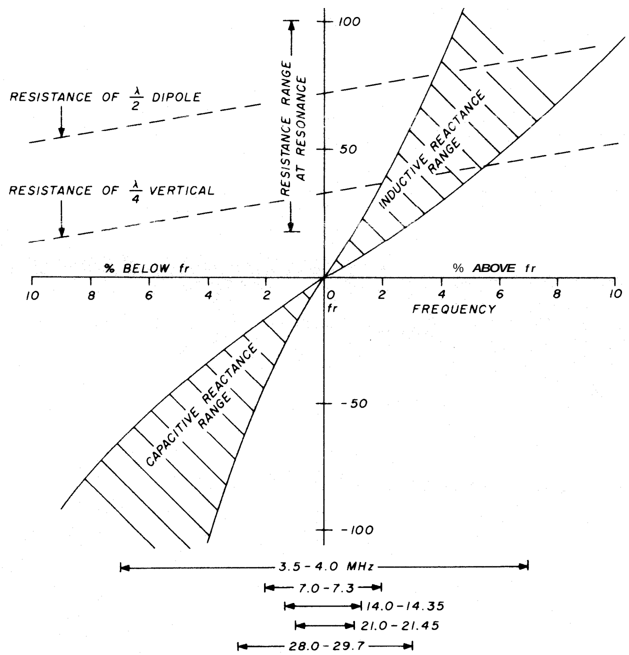


fig. 4. Typical change in resistance and reactance of an antenna as the operating frequency is moved away from resonance. The scales show the corresponding widths of each of the high-frequency amateur bands with resonance at the center of the band.

$+jX$ or $-jX$ in series with R at terminals **AB**, will be low as indicated by **fig.** 3. To compensate for this reactance a low equal value of capacitive or inductive reactance must be inserted either in series or in parallel. If inductive reactance is needed, it can easily be provided with a small coil. If a small amount of capacitive reactance is required, however, the size of the capacitor may be impractically large.

A simple way to take care of the reactance compensation for either case is to incorporate a series inductance leg at the output end of the matching network, and vary that inductance a small amount above and below the design value. This can be done by adjusting a tap on the output end. A T-network inherently provides this series inductance leg.

Having compensated for reactance at the output end of the network, the second step is to transform the remaining resistance to 50 ohms. A network containing inductance and capacitance is capable of doing this because for any series circuit containing R_s and X_s, **fig. 6**, there exists an equivalent parallel circuit containing R_p and X_p where in general R_p is different from R_s.

By definition, equivalence is the case where both series and parallel circuits exhibit the same magnitude and phase angle of impedance to an external circuit. To state this another way, both circuits must have the same external circuit Q where Q is defined as X_s/R_s for the series circuit, and R_p/X_p for the parallel circuit.

The equations relating R_p to R_s and X_p to X_s for any value of Q are as follows:

$$R_p = R_s\,(Q^2 + 1) \qquad \textbf{(1)}$$

$$X_p = X_s \left(\frac{Q^2+1}{Q^2}\right) \qquad \textbf{(2)}$$

If Q is 5 or more, **eqs. 1** and **2** can be simplified to the following with less than **4** per cent error.

$$R_p = Q^2 R_s \qquad \textbf{(3)}$$

$$X_p = X_s \qquad \textbf{(4)}$$

The L network in **fig. 7** is the basic matching network for transforming one resistance at terminals **AB** to another at **CD.** The pi network and T network in **fig. 8** can be thought of as two L networks connected in series. Because of the practical limitations on the size of capacitors for X_p, L networks are best suited for matching a low resistance to a high resistance or vice versa. The pi network is best suited for matching a high resistance to a high resistance, and the T network for matching a low resistance to a low resistance.

If a pi network is best suited for matching one high resistance to another high resistance, you might logically question why it is so universally used in the output stages of transmitters and amplifiers where the output resistance to be matched is 50 ohms. One of the reasons the pi network is popular is that, if a set

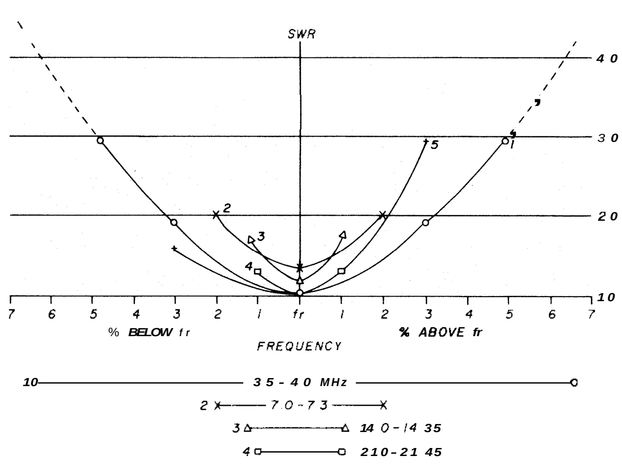


fig. 5. Plots of standing wave ratio as measured at the transmitter output terminals for an 80-meter inverted vee (1), 40-meter inverted vee (2), and three-band Yagi for 14 MHz (3), 21 MHz (4), and 28 MHz (5).

of fixed coil taps is used for all high-frequency bands and if the resistance on the input side is near 50 ohms and there is little or no reactance, adjustment of the capacitor on the output side is noncritical for proper loading.

!f, however, the resistance is not near 50 ohms, or if considerable reactance exists on the output side (as occurs when changing frequency away from resonance), the pi network may not be able to provide satisfactory loading. Adding a series inductance leg on the output side to form a pi-L network is one way to take care of this problem.

T network external to transmitter

A T-network assembled as an individual unit can be inserted in a coaxial feedline at any convenient distance from the transmitter, as shown in **fig.** 9. Any reactance which exists initially at **AB** can be

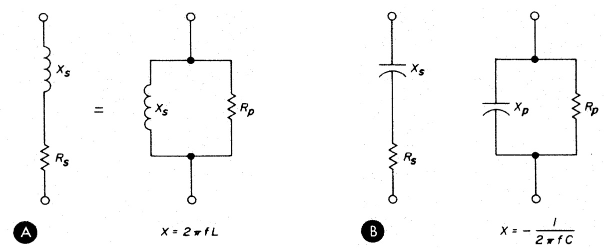


fig. 6. Equivalence of series and parallel circuits containing resistance and reactance. The equivalent element values may be determined from the relationship $X_s/R_s = R_p/X_p$.

tuned out by moving the coil tap on X_{s1}. Elements X_{p1} and X_{p2} of **fig. 7D** are combined into a single value of X_p equal to

$$X_p - \frac{X_{p1}\,X_{p2}}{X_{p1} + X_{p2}}$$

The right-hand side of the T network transforms the low resistance at **AB** to a high resistance at **CD.** The left-hand side then transforms the high resistance at **EF** to 50+ $j0$ ohms. The high resistance at **CD** to **EF** is referred to as the *virtual* or "apparent" resistance, transformed from **AB** and **GH,** respectively. The resultant virtual resistance across X_p is

$$R_p = \frac{R_{p1}\,R_{p2}}{R_{p1} + R_{p2}}$$

If the resistance at **AB** in **fig.** 9 is 50 ohms, and the effective resistance at **GH** is also 50 ohms, the T network becomes symmetrical; X_{s1} is equal to X_{s2}, and

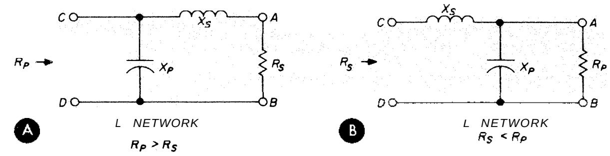


fig 7, L networks are very useful for transforming one resistance value to another. Design formulas for L networks are found in reference 3.

X_p is one half of X_{s1}; R_{p1} is equal to R_{p2}, and their resultant, R_p, is one half of R_{p1}.

T network with single center-tapped coil

The two series inductance legs of the T network, X_{s1} and X_{s2} **(fig. 9)**, don't have to be separate coils. Use of a single center-tapped coil not only simplifies construction but also increases the effectiveness of the network in its operation as a lowpass filter to attenuate harmonics. Each of the two series legs must have a larger inductance than two separate coils would have for a given value of Q, and the extra inductive reactance additionally impedes the flow of harmonic currents through the network.

Fig. 10A shows a center-tapped coil with connections **a** and **c** at its ends, and **b** at the center tap. The equivalent circuit is shown in **fig. 10B** and the coil assembled into a T network in **fig. 10C**. When connected as in **fig. 10C,** the coefficient of coupling k is negative and the effective inductance of each coil half is less than L.

Inductance of the coil alone, measured between **a** and **c,** is greater than twice the inductance between **a** and **b** or **b** and **c** by the factor $2Lk$, where k is defined as the ratio of mutual inductance ***M*** to ***L*.** To obtain a desired value of ***L*** for a T network, a relation between k and coil dimensions is needed. **A** curve of k vs the length-to-diameter ratio of a center-tapped coil is available,' but the same data can be obtained directly for any given set of coil parameters by using an **ARRL** L/C/F calculator Type A.*

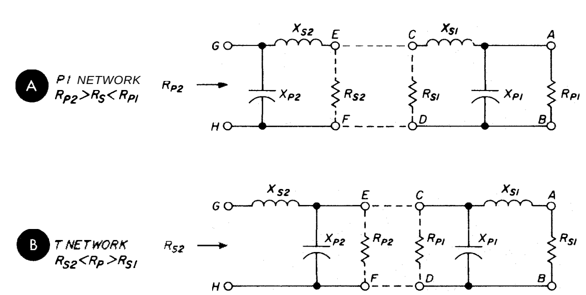


fig. 8. L networks may also be combined into a pi network (A) or a T network (B). The design of pi networks is discussed in references 4, 5, and 6.

*The ARRL L/C/F Calculator Type A is available for $3.00 from the American Radio Relay League, 225 Main Street, Newington, Connecticut 06111.

There are only two active components to be considered in designing a practical single-coil T network: the variable capacitor C1 and the full coil **ac** center-tapped at b. Determination of both depends on the lowest frequency at which the network is to be used,

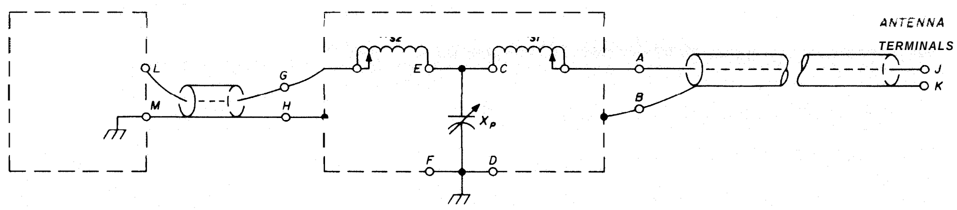


fig. 9. A T-matching network inserted in the line at or near the transmitter output terminals.

so you must decide whether this will be 3.5 or 7.0 MHz. If you choose 3.5 MHz, a second decision must be made as to whether a very large coil should be used to give a network Q of 8 to 10 at that frequency, or a more reasonable sized coil to give that Q at 7 MHz. In the latter case an additional fixed capacitor must be switched across C1 for operation on the 3.5-4.0 MHz band.

An air-wound coil 6.35 cm (2.5 inches) in diameter, 10.2 cm (4 inches) long, and 8 turns per inch (3 turns per cm) is readily available (B&W 3030). It is of reasonable size, and nearly optimum for 7 MHz, so it was chosen with the option of using additional capacitance at 3.5 MHz. A transmitting type variable capacitor should be used for C1 because the rf voltage across it will be Q times that at the input terminals **GH.** The capacitor should be calibrated so that values of X, can be readily determined from the dial settings. A 30-220 pF transmitting unit was selected in my case.

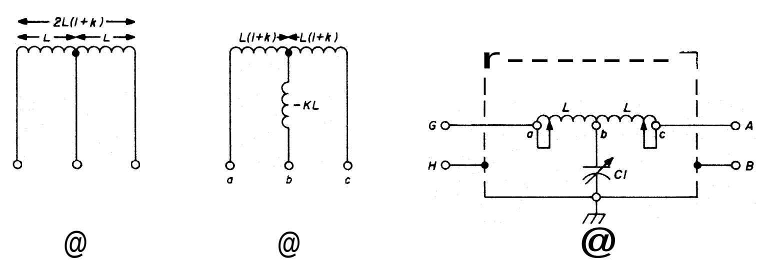


fig. 10. Inductance of a center-tapped coil is shown at (A) with its equivalent circuit at (B). A two-component T network with a center-tapped coil is shown at (C).

Using a Type A calculator the inductance of half coil L is determined as follows:

$$L_{ac} = 31\ \mu H$$

$$L_{ab} = L_{bc} = L = 12.8\ \mu H$$

$$L_{ac} = L_{ab} + L_{bc} - 2M = (2L - 2M)$$

$$M = -2.7$$

$$k = M/L = -0.21$$

The effective inductance of L, L,, is $L(1+k)$.

$$L_e = 12.8\ (1 - 0.21) = 10.1\ \mu H$$

If the resistance at GH and **AB** is 50 ohms, and the full coil is in the circuit at 7.0 MHz, the network components, **fig. 7D,** are as follows:

$$X_{s1} = X_{s2} = 2\pi f L_e = 444 \text{ ohms}$$

$$Q = 444/50 = 8.88$$

$$X_{p1} = X_{p2} = 444 \text{ ohms}$$

$$(X_p = 222 \text{ ohms } (C1 \text{ at } 7 \text{ MHz}$$

$$= 1/2\pi f X_p = 102\ pF)$$

$$R_{p1} = R_{p2} = 78.85 \times 50 = 3942 \text{ ohms}$$

$$R_p = 1971 \text{ ohms}$$

At 3.5 MHz the network Q is 4.44 and X_p is 111 ohms, requiring a total capacitance at C1 of about 400 pF. A fixed transmitting capacitor of 0.0002 μF in parallel with the 220 pF variable may be used.

T network for five-band coverage

Fig. 11 shows the schematic of a tapped-coil network built from the above design data. Construction details are omitted because of the large number of possible options in components and additional convenience features such as tap switches. You may also want to add a coaxial switching scheme to bypass the network so that operation with a given antenna can be quickly compared with and without the matching network. The main construction criterion is to preserve electrical and mechanical symmetry between the input and output sides of the network.

A good method of determining coil tap locations is to connect a transmitter with a nominai 50-ohm output impedance at J1 and a 50-ohm dummy load at J2. An swr indicator should be inserted between the transmitter and J1. With a 50-ohm load the network will be symmetrical. For all bands higher than 7 MHz where the full coil is used, a set of taps can be found which give a 1:1 swr at any desired Q. The calculated values of X_p, X_s, and Q for the coil taps of **fig. 11** are listed in **table 1.**

With the antenna and coaxial feedline combinations which produce resistance at J2 anywhere between 25 and 100 ohms, the capacitor dial settings X_p will be close to the values in the table and the indicated swr will be at or very close to 1.0:1. At frequencies off resonance, where reactance appears at J2, the swr will still be close to 1:1 except at the edges of the widest amateur bands.

Maximum swr readings for antennas 2, 3, 4, and 5 of **fig. 5,** without moving any of the coil taps, were 1.2:1 on the 7 MHz antenna and 1.7:1 on the 28 MHz antenna. It is possible to get a 1.1 swr reading with any of these antennas on any frequency in the band by moving coil taps slightly, but this is not normally necessary.

With a 3.5-MHz antenna you should not attempt to use the antenna and network combination at frequencies where the swr is greater than 3:1 without the network. A broadband or two-frequency antenna should be used if full-band coverage is routinely required.

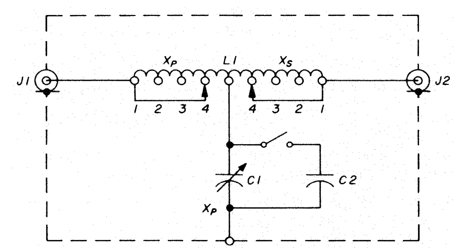
conclusions

Use of a T network in the manner described here is new only in the sense that its advantages have long been overlooked. I have been using a network similar to **fig. 11** at W6EBY for more than 15 years, and it has been used successfully on all frequencies in all five bands with a kilowatt amplifier. In fact, the acquisition of the amplifier and the distressing events which occurred when its load departed too much from 50 + j0 ohms were the motivations which originated the matching network in the first place.

The network can be just as useful, even essential in some cases, with lower power transmitters. Some of the new units with solid-state final stages are not tolerant of an swr greater than 3:1 and they are designed to shut themselves off when confronted with high swr. T networks, although they can't improve the performance of the antenna, will at least allow operation with the antenna as it is.

table 1. Calculated capacitance, inductance, and Q for the network of fig. 11, with 50 ohms at both input and output.

frequency (MHz)	C1, or (C1 + C2) (pF)	X_p (ohms)	X_s (ohms)	Q
3.50	394	115	230	4.60
3.75	342	124	248	4.96
4.00	302	132	264	5.28
7.00	102	222	444	8.88
7.15	98	227	454	9.06
7.30	94	232	464	9.26
14.00	41	277	554	11.08
14.175	40	281	562	11.22
14.35	39	284	568	11.38
21.00	49	155	310	6.18
21.225	48	156	312	6.24
21.45	47	158	316	6.32
28.00	45	126	252	5.06
28.85	43	128	256	5.14
29.70	40	134	268	5.36



C1	30-220 pF transmitting type variable capacitor
C2	200 pF, 5000 Vdc
J1, J2	SO-239 coaxial socket
L1	31 μH air-wound coil, 6.35 cm (2.5") diameter, 10.2 cm (4") long, 3.2 turns per cm (8 turns per inch), B&W 3030 or equivalent.

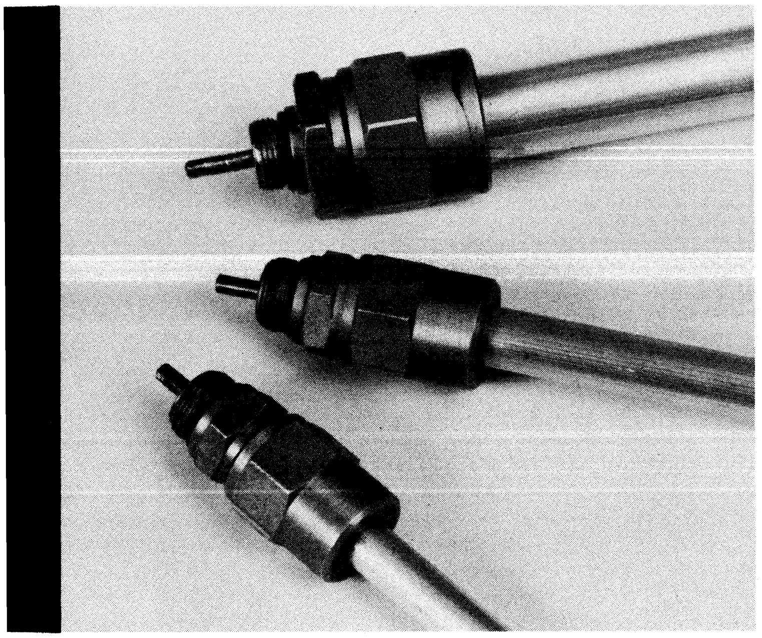
coil taps	1	2	3	4
Band, MHz	3.5, 7	14	21	28
Active turns	32	26	14	10
Turns each leg	16	13	7	5

fig. 11. Tapped-coil T matching network for the five high-frequency amateur bands, 3.5 through 29.7 MHz.

references

1. *J. R. Fisk,* W1HR, *"How to Use the Smith Chart," ham radio, March,* 1978, *page* 92.
2. *H. Jasik, Antenna Engineering Handbook,* McGraw-Hill, *New York,* 1961, *Chapter 3.*
3. R. E. *Leo,* W7LR, *"How to Design L Networks," ham radio, February,* 1974, *page* 26.
4. *E. W.* Whyman, W2HB, *"Pi Network Design and Analysis," ham radio, September,* 1977, *page 30.*
5. *L. H. Anderson, "Pi Network Design," ham radio, March,* 1978, *page* 36.
6. *I. M. Hoff,* W6FFC, *"Pi Matching Networks — Tables of Values," ham radio, June,* 1977, *page*
7. R. *W. Johnson,* W6MUR, *"Multiband Tuning Circuits," QST. July,* 1954, *page* 25.

ham radio



75-ohm cable in amateur installations

Making use of 75-ohm CATV cable results in lower line loss, which means more power to the antenna

Many hams are not aware of one of the best coaxial cables available, the 75-ohm, solid-aluminum sheathed cable made specifically for cable television (CATV). The coax used in these systems is characterized by a minimum attenuation loss, minimum random-signal pickup, excellent weather resistance, and high structural return loss. All this not withstanding the fact that it can usually be obtained as scrap for next to nothing.

"The number assigned to each cable is actually the outside diameter of the aluminum sheath in inches. With cable of primarily U.S. manufacture, this number has become the generic name of the cable.

By Gordon K. Woods. W7VK, 6022 North Second Avenue, Phoenix, Arizona 85013

Leading the list of features is low attenuation loss. As a result, a 100-watt-output, 420-MHz transmitter, feeding power through 30.5 meters (100 feet) of 0.750" hardline, will deliver 75 watts to the antenna; this is quite an improvement over the 40 watts delivered by a comparable length of RG-8/U!

To further illustrate, **fig. 1** shows the loss exhibited by several different types of cable, starting with the relatively lossy RG-58 and ending with 0.750 cable. With CATV cable, even at the lower frequencies, impressive gains are available to people who use long cable runs.

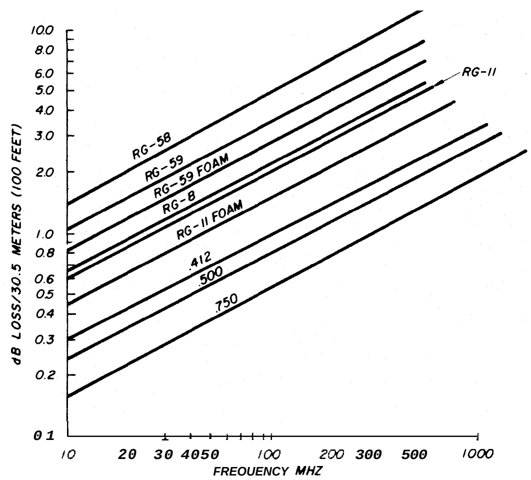
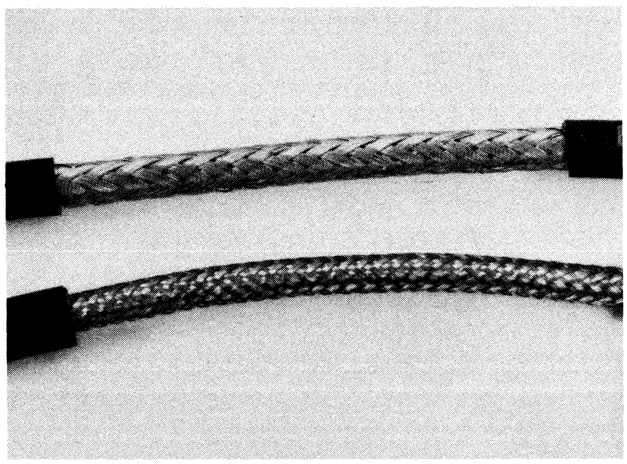


fig. 1. Loss vs. frequency for several types of cable. The CAN cable examples refer to cable manufactured by Systems Wire and Cable, Inc., of Phoenix, Arizona. Cable from other manufacturers may have different attenuation values, but as a rule will be very close to these figures.

In addition to its lower attenuation figures, solid aluminum sheath also reduces random-signal pickup and leakage. The best military braid specifications require only 96 per cent shielding, as compared to the 100 per cent provided by seamless CATV cable. And, as it turns out, most braided cable used by amateurs has a braid coverage in only the 75 to 90 per cent range, and sometimes as low as 60 per cent (see photograph)!

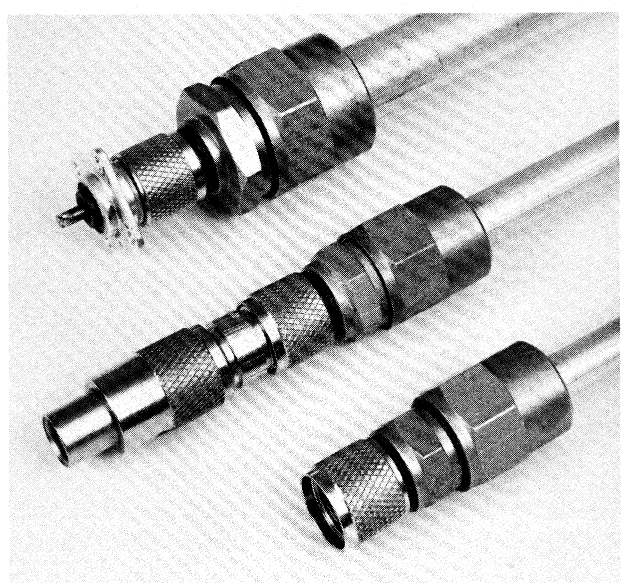
Weather resistance is also greatly improved by using cable with a seamless sheath. After a period of



As seen in this photograph, braid coverage can vary radically. The two examples are both RG-59/U, except that the top cable is normally used in CAN installations. Use of the bottom cable may result in unwanted signal pickup and/or emission.

exposure to sunlight and air, the copper conductors in ordinary braid become corroded; they do not form an electrical bond to one another, but function more as insulated wires, increasing random-signal pickup and radiation losses. This can be easily checked by terminating a well weathered RG-8 line and noting the background noise level. Then, change to CATV cable, equally terminated, and your receiver will be dead.

CATV cable is normally bare, but it is also produced with a black polyethylene jacket if the cable is to be exposed to salt spray, fog, or industrial contaminants. It is also manufactured with a "flooded" polyethylene jacket for underground or underwater installation.



Different examples for using the adapters between 75-ohm CAN hardline and standard UHF connectors.

connectors

The major problem encountered by amateurs using 75-ohm CATV cable has been finding suitable connectors to use between the cable and ordinary UHF fittings. (Special cable connectors which mate with type N and F fittings are available, but they are difficult to locate and buy in small quantities.) There is a practical solution to this problem, however. This is the use of standard CATV "feedthrough" connectors, which, fortunately, end up with 6.5 mm (1/4 inch) of male 518 x 24 (M16-2) thread, the same thread as standard UHF connectors.

Making the adapter begins, as shown in **fig.** 2, with the installation of the appropriate feedthrough connector on the end of the cable. To mate with the UHF connectors, a PL258 female-to-female adapter is slipped over the end of the exposed center conduc-

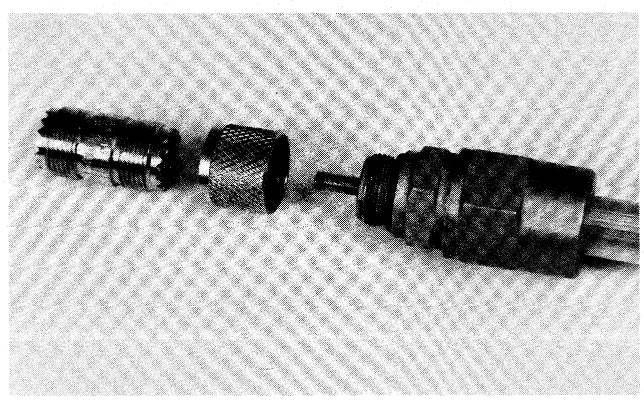


fig. 2. As shown in this photograph, adapting CAN cable to a normal UHF connector requires the use of a CATV feedthrough connector attached to the end of the cable. The 0.500 cable in this illustration has the brass tubing sweated over its center conductor. The threaded coupling cut from a PC259 connector attaches the PC258 adapter to the feedthrouah connector.

tor. Joining the adapter and feedthrough connector is accomplished by using the threaded portion of the barrel from a PL259 connector.

On the 0.412 and 0.500 cable, you will have to sweat solder a piece of 4-mm (5132-inch) OD brass tubing over the center conductor. (The brass tubing is available in short lengths from most hobby stores.) The center conductor of the 0.750 cable is heavily tinned to increase its diameter from 3.7 to **4** mm (0.146to 0.156 inch).

If you want to directly hardwire the CATV cable to an SO-239 chassis connector, prepare the cable end as shown in **fig.** 3. Then, connect the SO-239 to the feedthrough connector, prior to inserting the cable. The final step consists of inserting the cable into the feedthrough connector, making sure that the center conductor mates with the SO-239, and tightening the cable ferrule. Generally, connectors are available from CATV equipment supply sources, although they are not enthusiastic about small-quantity orders. (Try these sources only if you can't con the local CATV system installer out of a few.)

installation

When installing solid-sheath aluminum cable, note that all bends should be made over a grooved form block. Also ensure that all bends are ***never*** made to a radius of less than ten times the cable diameter. Observing this precaution will prevent wrinkling the sheath, which can cause impedance bumps. Too tight a bend may also force the center conductor to one side, since the foamed dielectric is soft and subject to cold flow.

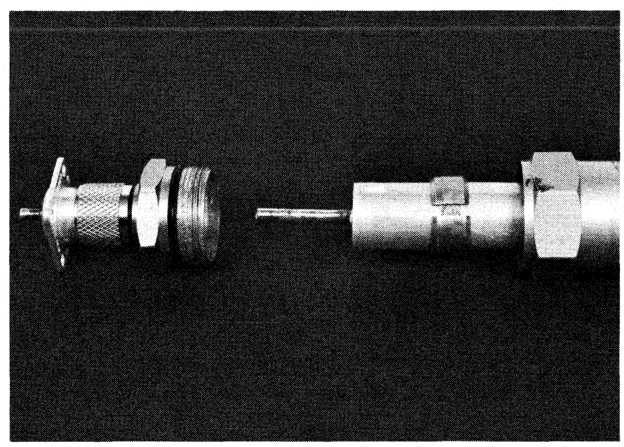


fig. 3. Attaching the hardline to a chassis-mounted SO-239 can be accomplished in the same manner. In this case though, the feedthrough connector and the SO-239 are mated before the cable is attached.

Another item available from equipment supply stores is special heat-shrink tubing that slides over the entire connector assembly. These tubes are usually about 23 cm (9 inches) long and have a special sealant inside that is effective against moisture.

summary

Before changing to 75-ohm transmission line, it is best to ensure that your transmitter and antennas will match the higher impedance. In general, most transmitters with a pi network will match the impedance presented when using 75-ohm cable. Even gamma matches on Yagi arrays can be readjusted to match the new cable. Unfortunately though, it can be an expensive proposition if you try to change your power meters to read correctly in a matched 75-ohm system.

Even with the small problems presented by connectors and, in some cases, matching, the use of 75-ohm CATV cable has one big advantage: more power at the antenna at a highly economical price.

ham radio

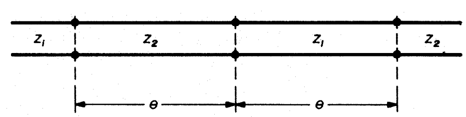
matching 75-ohm CATV hardline to 50-ohm systems

The previous article by W7VK pointed out the significant attenuation differences between the more commonly used RG-8 type coaxial cables and 75-ohm CATV type "hardline." In some amateur installations, changing to hardline could mean large increases in the power delivered to the antenna, especially where long cable runs are being used. As Woods pointed out, switching to this type of cable usually involves only antenna rematching and retuning the transmitter. Unfortunately, in some cases, rematching the antenna to 75 ohms is not possible, and the resultant swr may be intolerable; the ultimate isolation between sections of a repeater duplexer, for example, can be degraded by a high swr on the line. The matchable bandwidth of an antenna can also be reduced, since the output pi network was originally designed for 50-ohm loads. And finally, 75-ohm power meters are not commonly available.

matching

The standard quarter-wavelength transformer or Q section, one of the most popular forms of matching, is unfortunately not readily suited for this task. The impedance of the matching section has to be the geometric mean between the two impedances to be matched, or in this case $\sqrt{Z1 \bullet Z2} = 61.3\ ohms$, not a common coax impedance value.

One little-known matching technique, the nonsynchronous impedance matching transformer, does offer a solution to the problem. W5TRS originally described this method in *ham radio*,[1] though only providing basic design information (see **fig. 1**). In a subsequent letter,[2] W3DVO briefly discussed the bandwidth in relation to a standard quarter-wavelength transformer. Until now, however, nothing has been published on the use of the nonsynchronous transformer. Since the required sections are the same impedance as those to be matched, this method would seem to be an easy solution to the 75-to-50 ohm matching problem, and warrants further examination.



$$\theta = \cot^{-1}\sqrt{\frac{Z_1}{Z_2} + 1 + \frac{Z_2}{Z_1}} \qquad \ell = \frac{\theta\lambda}{360}$$

fig. 1. Diagram of the basic nonsynchronous matching transformer as described by W5TRS. The lengths of the two matching sections will vary according to the impedance ratio.

To evaluate this method, I decided to compare the nonsynchronous transformer to another technique, stub matching. **Fig. 2** illustrates the situations that were considered. (If the load were replaced by an antenna, the system would not be too different from a typical antenna installation.) The main feedline was considered to have a 1 dB insertion loss and be 0.25 wavelength long at the center frequency. Since there will be a perfect match at only one frequency, having the feedline 0.25 wavelength long provided the maxi-

By Charles J. Carroll, K1XX, *ham radio,*
Greenville, New Hampshire 03048

table 1. Impedance values along the lines shown in fig. 1.

frequency, MHz	point A	point B	point C	final swr
	nonsynchronous transformer			
144	74.63 + j0.63	75.34 − j0.50	49.86 − j0.44	1.0093
145	74.82 + j0.31	75.15 − j0.24	49.93 − j0.21	1.0044
146	75.01 − j0.02	74.99 + j0.02	50.00 − jO.O1	1.0002
147	75.20 − j0.36	74.85 + j0.29	50.08 + j0.22	1.0047
148	75.37 − j0.70	74.72 + j0.56	50.16 + j0.41	1.0089
	stub matching			
144	75.60 − j0.12	75.75 − j0.18	50.02 + j0.86	1.0173
145	75.31 − j0.06	75.39 − j0.09	50.00 + j0.42	1.0084
146	75.02 − j0.00	75.03 − j0.01	50.00 − j0.02	1.0004
147	74.74 + j0.05	74.67 + j0.06	49.99 − j0.01	1.0002
148	74.45 + j0.12	74.30 + j0.11	49.97 + j0.04	1.0009

mum impedance change at other than the center frequency. This, along with the low insertion loss, will provide close to the worst-case swr. The matching sections were considered to be lossless lines.

test results

Table **1** shows the different impedance values as the 50-ohm load was rotated back toward the generator. In actuality, the values were determined with the aid of an HP-25 programmable calculator; the use of the Smith chart was precluded since the final differences were extremely small, and beyond the *accurate* resolution of even an expanded chart.

Because the initial results proved so favorable, another set of calculations were performed. This time, instead of the relatively narrow bandwidth afforded by the 2-meter frequencies (approximately 1.5 per cent), calculations were carried out for 80 through 10 meters, with the 80-meter extreme of 6 per cent bandwidth. Table **2** shows the results for the nonsynchronous transformer when applied to **fig. 2A**.

As a final test, the line was terminated with eight different reactive loads. each selected to be on the 2:1 swr circle on a Smith chart (see **fig. 3**). The inner points represent the same impedances, but as seen at the generator (transmitter) end after the different rotations. Table 3 lists the actual computed values.

summary

The nonsynchronous impedance matching transformer can be an extremely valuable tool. With a bandwidth basically comparable to either stub or

table 2. Swr values for the nonsynchronous transformer when used for 80 through 10 meters.

frequency, MHz	final swr	frequency, MHz	final swr
3.5	1.0571	14.3	1.0048
3.6	1.0322	14.4	1.0091
3.7	1.0091	21.0	1.0063
3.8	1.0086	21.1	1.0029
3.9	1.0224	21.2	1.0003
4.0	1.0319	21.3	1.0034
7.0	1.0150	21.4	1.0059
7.1	1.0045	28.0	1.0097
7.2	1.0038	28.2	1.0045
7.3	1.0188	28.4	1.0003
14.0	1.0097	28.6	1.0048
14.1	1.0045	28.8	1.0091
14.2	1.0003		

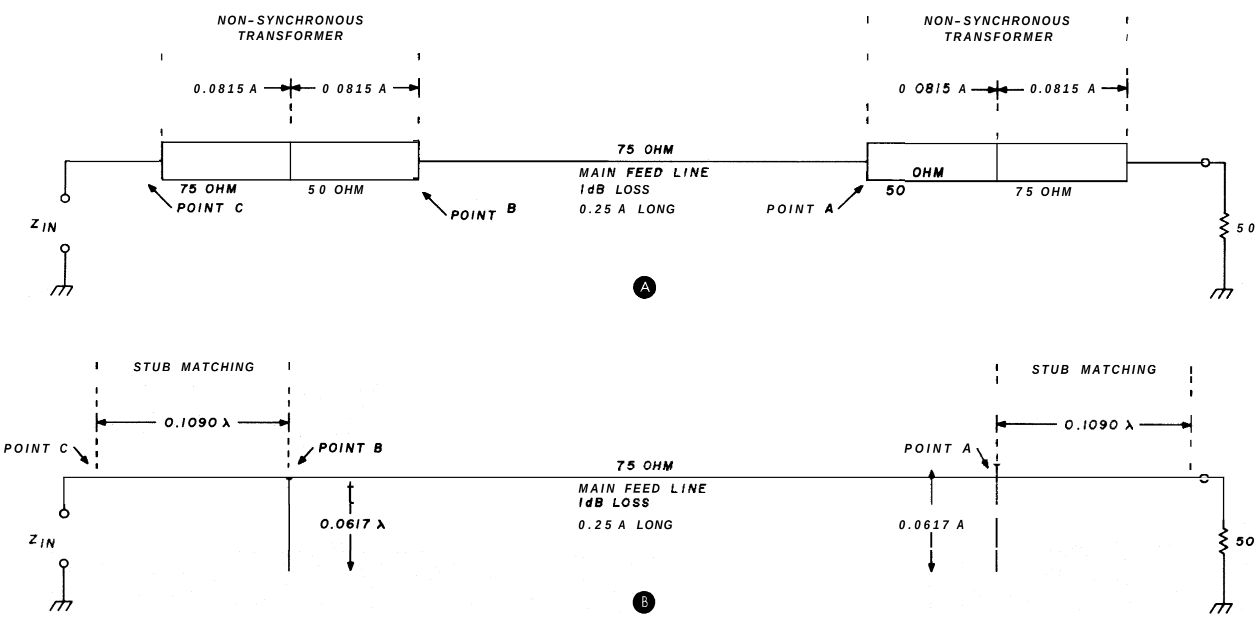


fig. 2. Schematic diagram of the system used to evaluate the bandwidth of the two matching systems. The main feedline, as used in both systems, has 1 dB loss. Points A, B, and Ccorrespond to the impedance values listed in the tables.

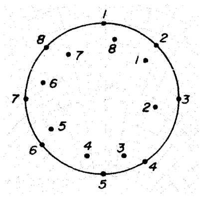


fig. 3. Smith chart presentation of the eight reactive loads used to terminate the line. The inner points represent the final impedances as seen at the transmitter end.

Q-section matching, it has the added advantage of requiring one-third less coax than the quarter-wavelength section and one-half less than a stubbed system. Though these differences may not be significant at vhf, they can save a considerable amount of cable on 80 through 10 meters. In addition, the construction of a nonsynchronous transformer appears to be inherently easier than that of a stub system because of the difficulty in correctly placing a T-type connector. Probably the two biggest disadvantages are that the feedlines have to be dedicated to a particular band (since each transformer length is frequency dependent), and the requirement that the coax be the same impedanceas those to be matched. These factors certainly prevent its qualifying as an all-encompassing matching method, but it more than adequately will handle the problem of matching 75-ohm CATV hardline to a 50-ohm system.

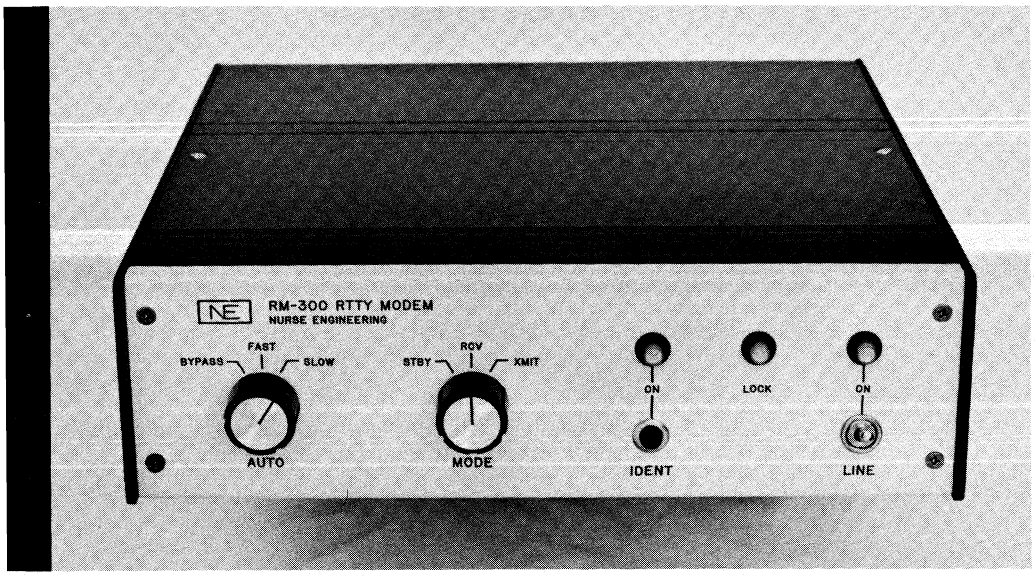
table 3. Computed swr values at 14 MHz with the line terminated in reactive loads.

termination	computed impedance values	final swr
25.00 ± j0	34.17 + j15.90	1.7097
28.09 + j14.90	48.40 + j26.48	1.7040
40.02 + j30.01	74.23 + j22.41	1.7080
69.53 + j36.84	83.89 - j10.66	1.7193
100.0 ± j0	59.09 - j28.80	1.7311
69.39 - j36.86	39.16 - j22.24	1,7367
39.95 - j29.96	30.58 - j9.84	1.7327
28.06 - j14.85	29.21 + j3.11	1.7215

references

1. Henry Keen, W5TRS, *ham notebook, ham radio, September, 1975, page 66.*
2. *Raymond Aylor,* W3VDO, *comments, ham radio, May. 1976, page 63.*

ham radio



RM-300 Modem

RTTY modulator. demodulator for vhf operation

Design and construction details for the hardware to get started in the RTTY mode on the vhf amateur bands

Operating RTTY on vhf fm is a joy. Bothersome fading, static crashes, interfering signals, and drifting VFOs don't stand between you and a QSO. Whether across town on simplex or across the state through a repeater, vhf RTTY is a reliable, trouble-free mode ideally suited for rag chewing or unattended operation.

The ingredients for a vhf RTTY station include a standard fm transceiver, a Teletype machine, a terminal unit (demodulator), an AFSK oscillator (modulator), and a means of sending your call in Morse code as required by the FCC.

The RM-300 Modem (Modulator-Demodulator) was developed to provide a simple way for those interested in vhf RTTY to get on the air. The RM-300 contains a phase-locked loop (PLL) demodulator to convert the 212512295-Hz tones from your transceiver speaker terminals to Teletype keying pulses, a stable AFSK modulator to feed the microphone input of your transceiver, a read-only memory CW identifier, and auto-start logic — all on a single 114 x 152 mm (4.5 x 6 in.) circuit board. A second board, the RP-400, contains loop and low-voltage power supplies and loop keying circuits.

A schematic diagram for the RM-300 is given in **fig. 1.** The three major functions mounted on the cir-

"A complete kit of all parts is available from Eclipse Communications, 5 Westwood Drive, San Rafael, CA 94901. The RM-300 kit less PROM costs $71.25. A PROM programmed with one or two call signs (specify) costs $7. The cost of the RM-300 circuit board alone is $21.25. (The board is double-sided with plated-through holes and a solder mask on both sides. The component side is screened with part numbers and component values.) Add $1 for postage and handling with all orders. California residents add 6 per cent sales tax.

By Howard L. Nurse, **W6LLO**, 665 Maybell Avenue, Palo Alto, California 94306

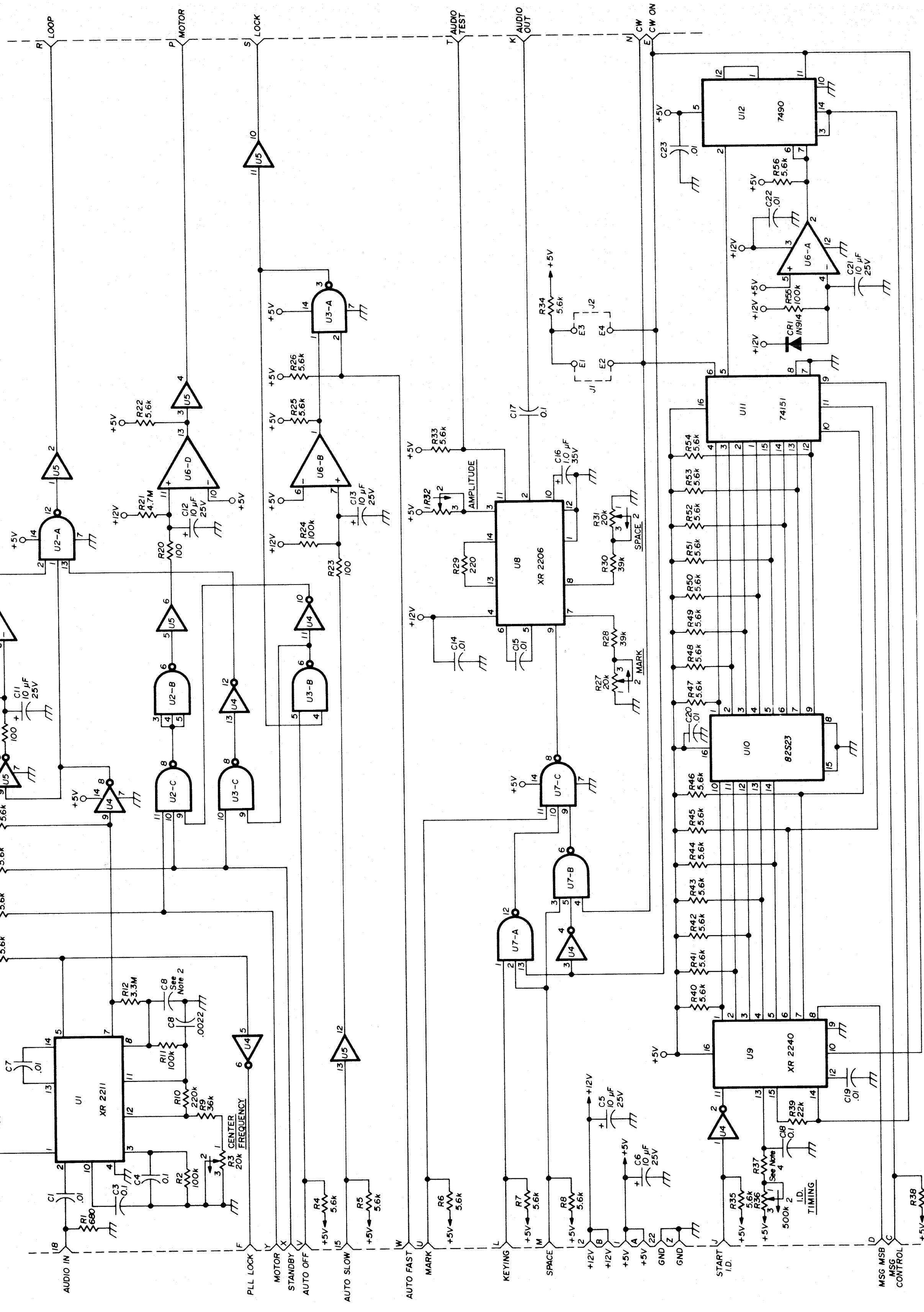


fig. 1. Schematic of the RM-300 Modem (modulator-demodulator) for those interested in vhf RTTY. A complete kit of all parts is available (see text).

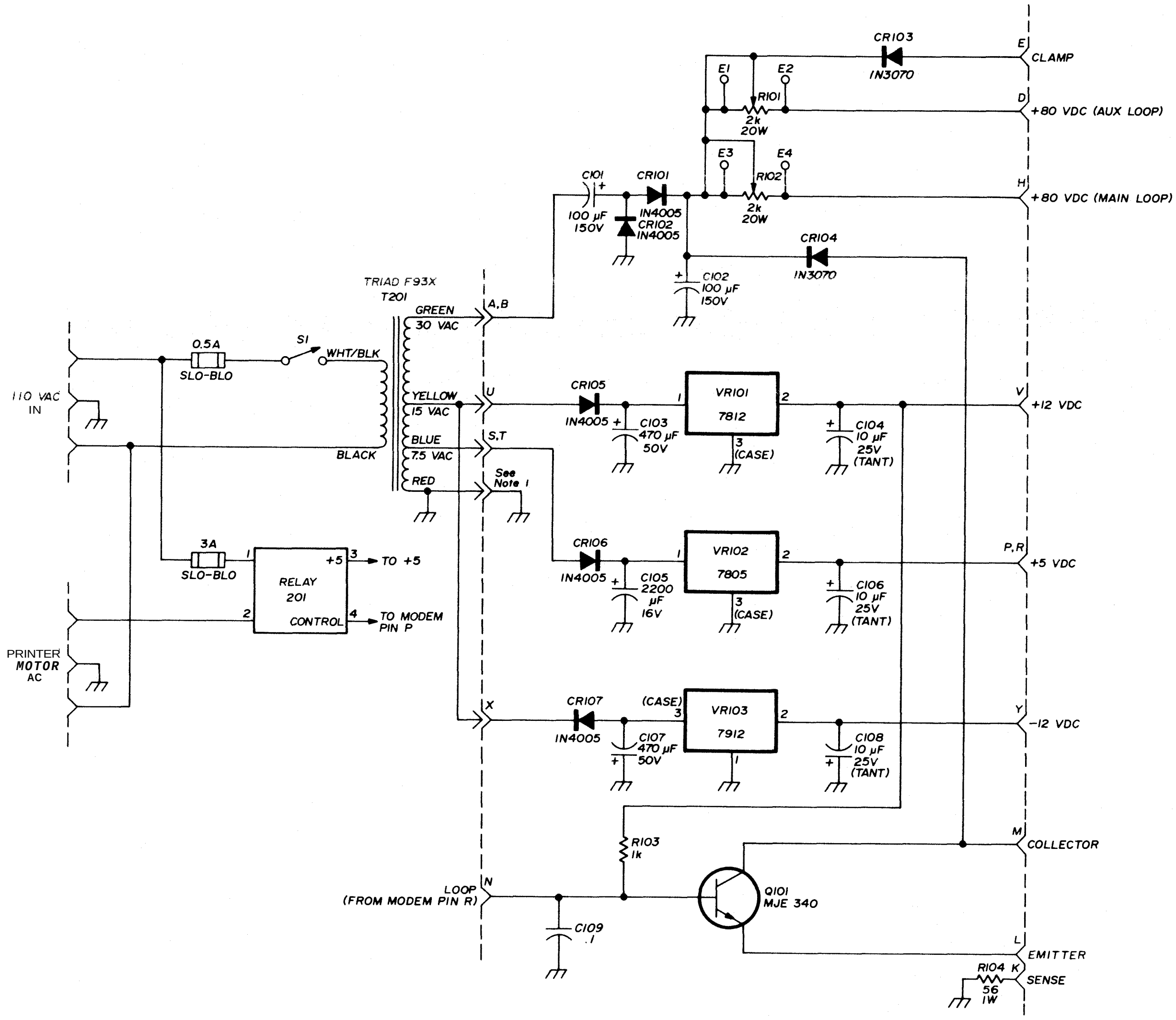


fig. 2. Schematic of the RP-400 power supply. The design features an auxiliary –12 V dc supply in addition to power required for the RM-300 Modem. Kits are available.

cuit board are grouped on the schematic. The demodulator portion is shown in the top third, the AFSK modulator in the middle third, and the CW identification circuitry in the bottom third of the schematic. Detailed descriptions of connections to and from the RM-300 circuit board are given in **table 1.**

demodulator and autostart

Audio from the transceiver is fed to the XR2211 PLL through pin 18 on the board edge connector. The PLL output on U1 pin 7 is at a TTL-compatible high level when the input tone is 2125 Hz (MARK) and low when the input tone is 2295 Hz (SPACE). The frequency at which the PLL switches from high to low is the PLL center frequency as determined by center-frequency potentiometer, R3.

The PLL output signal is inverted by U4 and applied to the antispace circuit (U5 and U6-C) and the keying inhibit gate, U2-A.

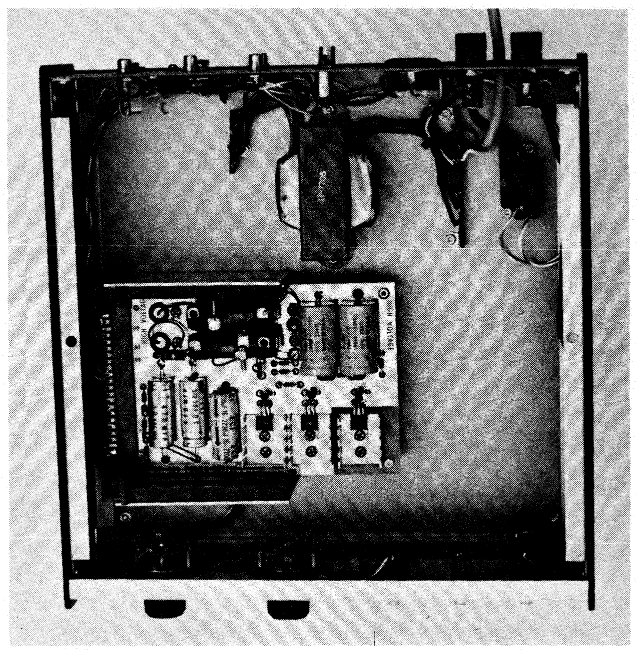
The antispace circuit prevents space tones longer than approximately 200 milliseconds from keying the loop. The output from U5 pin 8 clamps the voltage across C11 near ground as long as the PLL is detecting a mark tone. A space tone allows the voltage across C11 to rise at a rate determined by R18. If the voltage rises above 5 volts, the output of U6-C goes low, which forces the loop into *mark*-hold.

The autostart circuit and standby logic can also be used to force the loop into mark-hold. The PLL contains a carrier detector, which has a TTL-compatible output on pin 5 of U1. This output is low when a signal within the lock range of the PLL is detected. The carrier-detect output, after inversion by U4, can be used to control the fast or slow modes of the autostart. The fast mode allows the Modem to respond immediately to a detected signal, which is desirable when operating fast break-in. The slow mode yields a 0.5-second delay before the Modem responds, which gives sufficient noise immunity on vhf circuits. When a carrier is detected by the PLL, the lock output (edge connector pin S) goes low. This output can be used to control an LED on the Modem front panel.

If the auto off input (edge connector pin V) is grounded or a carrier has been detected, and the standby input (pin X) is high, the loop can be keyed by the PLL because pin 13 on U2-A will be high.

The open-collector loop output from U5 pin 2 is high for mark and low for space. It can be connected directly to a loop keying transistor or to the serial input of a UART.

The motor output (pin P) is low when a signal has been detected, the standby input is grounded, or the motor input (pin Y) is grounded. The motor output will remain low for approximately 25 seconds after a signal has dropped. This timing is determined by the RC time constant of R21 and C12.



Interior of the RM-300 Modem. A solid-state motor-control relay is near the power cable at the rear of chassis. The RM-300 board is below the RP-400 power-supply board.

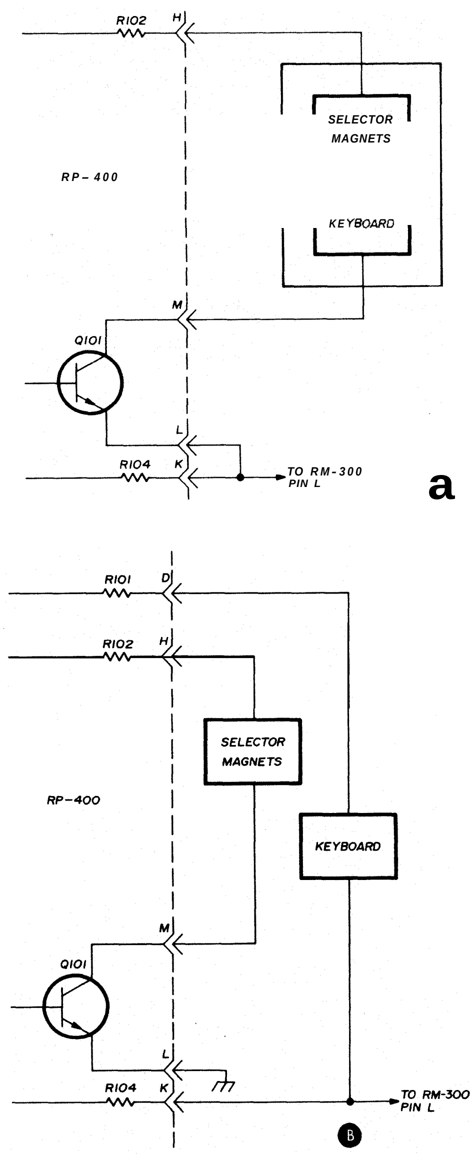


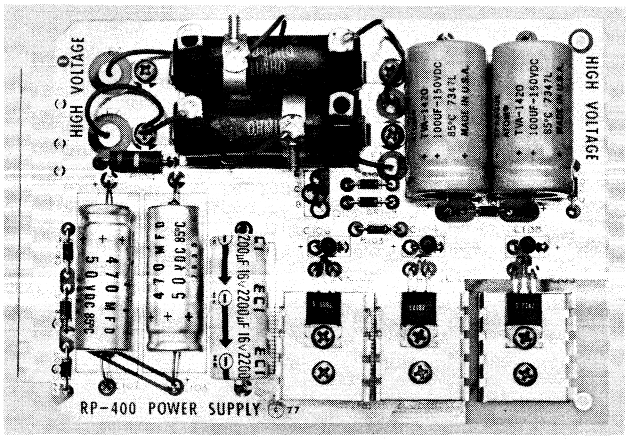
fig. 3. Two suggested ways of connecting your RTTY machine to the loop supplies and keying transistors. Sketch A shows the usual connection. Separate selector magnet and keyboard connections are shown in B.

AFSK modulator

The heart of the AFSK modulator is an XR-2206 function generator integrated circuit. It produces low-distortion sine wave tones at a frequency determined by capacitor C15 and the resistance between pin 7 or 8 and ground, depending on the control input on pin 9. When pin 9 is high (mark), R27 sets the ouput frequency to 2125 Hz; and when pin 9 is low (space), R31 sets the output frequency to 2295 Hz. The output frequency can be monitored at the audio test output (pin T), which provides a TTL-compatible square-wave replica of the function generator output waveform.

The AFSK modulator output level should be adjusted to match the requirements of your transmitter's audio circuit using potentiometer R32. The output amplitude can be adjusted to a maximum level of approximately 5 volts peak-to-peak.

All control inputs to the AFSK modulator are TTL-compatible. The teleprinter keyboard signals are applied to the *keying* input (pin L). Consistent with the design convention used throughout the board, a high input yields a mark tone, while a low input yields a space tone. Mark and space override inputs are provided on pins U and M respectively. These inputs can be used to force the AFSK modulator to either state despite information present on the keying input. The remaining control input to the AFSK modulator is from the CW identifier.



The RP-400 power-supply board. The board is **114** x **152** mm (**4.5** × ***6*** in.). **See** text for **kit** information.

table **1**. RM-300 **RTTY** Modem input-output descriptions

name/pin number	description
AUDIO IN/18	Audio from receiver. 680-ohm impedance. Input level 10 mVrms to 3 Vrms.
PLL LOCK/F	TTL high when PLL locked to input signal.
MOTOR/Y	When grounded causes MOTORIP to go low.
STANDBY/X	When grounded causes MOTORIP to go low and LOOPIR to go high (MARK level).
AUTO OFF/V	When grounded forces Modem into receive mode without control by the PLL LOCK/F function.
AUTO SLOW115	0.5-second Autostart control line.
AUTO FAST/W	Instantaneous Autostart control line.
MARK/U	When grounded forces AFSK modulator to MARK.
KEYING/L	Normal RTTY keying line. MARK (2125 Hz) is high and SPACE (2295 Hz) is low. This input is overridden by CW IDENT cycle.
SPACEIM	When grounded forces AFSK modulator to SPACE.
+1212, B	+12 Vdc at 20 mA maximum.
+5/1, A	+5 Vdc at 200 mA or less.
START ID/J	Momentary ground starts CW IDENT cycle.
MSG MSB/D	Logic low for first 128 cycles of CW IDENT.
MSG CONTROLIC	When grounded enables first half of CW IDENT memory. When high, enables second half.
LOOPIR	Loop control output, high for MARK and low for SPACE. Can sink 40 mA at 15 volts.
MOTORIP	Output goes low to turn on motor. Can sink 40 mA at 15 volts.
LOCK/S	Output goes low when PLL has locked.
AUDIO TEST/T	AFSK oscillator test output, 5 V p-p.
AUDIO OUT/K	AFSK modulator output, 600-Ohms, 6 V p-p maximum.
CW ON/E	Output goes low during CW IDENT.
CW/N	Output goes low during each IDENT key closure.

CW identifier

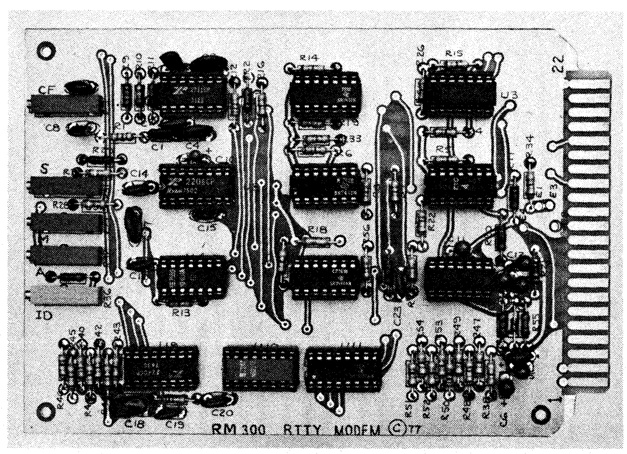
The CW identifier output signals include the CW keying line, which provides full-shift keying, and a CW *on* control line, which precludes all other keying when active. The CW identifier, started by momentarily grounding the *start ID* input (pin J), causes either **O** or **.** characters to be printed by the receiving station. The timing and memory programming required to accomplish this format were described in reference 1.

One new feature, available with the "0" option, is the ability to have two call signs in the read-only memory. One call sign is accessed by grounding the *message control,* pin C, while the other is obtained by leaving pin C open. If your call is too long to fit into half of the memory (128 bits), the whole memory can be used by connecting the *MSG MSB* output, pin *D,* to pin C on the edge connector.

On-off control of the CW identifier is accomplished with U9, an **XR-2240,** which contains an oscillator, an eight-stage divider, and a control flip-flop. The timing of the CW output is established by adjusting the *ID timing* potentiometer, R36. Five of the eight divider outputs from U9 address the **32** bytes of memory in U10, while the remaining three extract the CW characters from the selected bytes with multiplexer U11. Decade counter U12 monitors the CW output for consecutive "blanks;" after it detects eight, it turns off the control flip-flop in U9. Comparator U6-A and its associated circuit resets the decade counter so that the counter will start with the correct count when power is first applied.

adjustments

Proper adjustment of the Modem requires a frequency counter and a voltmeter or oscilloscope. The Modem can be adjusted with help from another sta-



The RM-300 Modem circuit board, which is the same size as the RP-400 power supply board. Kits are available (see text).

tion which has a counter or an AFSK oscillator/demodulator that has a known calibration.

AFSK **generator.** Connect a counter to ***audio test,*** pin T. Ground ***mark,*** pin U. Adjust R27 (**M**) to obtain a counter reading of 2125 Hz. Move the ground lead

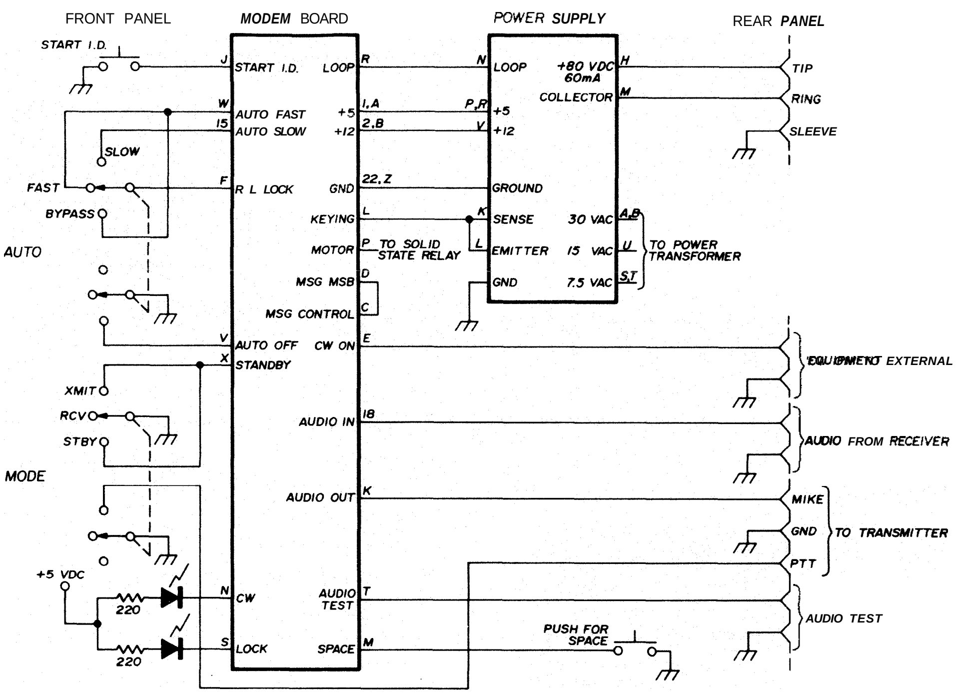


fig. 4. Interface wiring for the PC boards, front-panel controls, and rear-panel connectors.

from pin U to ***space,*** pin M. Adjust R31 (**S**) to obtain a counter reading of 2295 Hz. Use a voltmeter or oscilloscope connected to ***audio out,*** pin K, to adjust the output to a level compatible with your transmitter audio input using R32 (**A**). Remove the ground lead.

Phase-locked loop (U1). Use a jumper to connect ***audio out,*** pin K, to ***audio in,*** pin 18. Connect an oscilloscope or voltmeter to hex inverter U4-8. (This voltage will swing between ground and approximately 3 volts.) Ground ***mark,*** pin U, and adjust R3 (**CF**) until the hex inverter output goes high. Remove the ground from ***mark,*** pin U, and ground ***space,*** pin M. While counting turns, adjust R3 until U4-8 just goes low. (R3 should be adjusted in a ccw direction.) Divide the number of turns by two and adjust R3 in a cw direction by that number.

CW **identifier timing.** This adjustment is most easily made with the aid of another station. Initiate an ident while the other station monitors your signal with his printer. When the timing is adjusted correctly his printer should print 00000000. . . with the number of zeroes determined by your call sign. If you've selected the Baudot ident option, adjust the timing with R36 until the other machine prints the encoded message correctly.

power supply

The RP-400 power supply was designed to furnish the voltages required by the RM-300, in addition to an auxiliary -12 Vdc supply for other circuits you may wish to add, such as a UART. Included on the board are the following:

two 80 Vdc, 60-mA **loop supplies**
one $+12$ Vdc, 100-mA **supply**
one $+5$ Vdc, 300-mA **logic supply**
one -12 Vdc, 100-mA auxiliary **supply**

A high-voltage loop keying transistor is also located on the RP-400 power-supply board. A schematic for the board is given in **fig. 2.***

The loop supplies run from a voltage doubler consisting of C101, CR101, and CR102. The resulting dc is filtered by C102 before being sent to the adjustable wirewound resistors, R101 and R102.

Two suggested ways of connecting your Teletype machine to the loop supplies and keying transistor are shown in **fig.** 3. The usual connection, with the

*A complete kit of all the parts for the RP-400 power supply is available from Eclipse Communications, 5 Westwood Drive, San Rafael, CA 94901. The RP-400 kit costs $71.25. The cost of the RP-400 circuit board alone is $21.25. (The board is double-sided with plated-through holes and a solder mask on both sides. The component side of the board is screened with part numbers and component values.) Add $1 for postage and handling on all orders. California residents add 6 per cent sales tax.

selector magnets and keyboard contacts in series, is shown in **A** of the figure. Resistor R104 provides a TTL-compatible low level when the loop is open. If you desire separate selector magnet and keyboard connections, use the wiring diagram shown in B of **fig.** 3. Split wiring might be used when you wish to use the output of a repeater to provide local copy through a demodulator. By operating in this manner it's possible to tell instantly if you aren't making it through the repeater.

The low-voltage power supplies are similar in design. Each has a half-wave rectifier, filter capacitor, and three-terminal regulator.

construction

The RM-300 Modem board, the RP-400 power supply, the solid-state motor-control relay, and the power supply fit nicely inside a 30.5 x 30.5 x 8 cm (12 x 12 x 3 in.) Moduline enclosure.

Fig. 4 is a schematic showing one way to connect the boards, front panel controls, and rear panel connectors. Variations on this wiring will be determined by your requirements.

acknowledgments

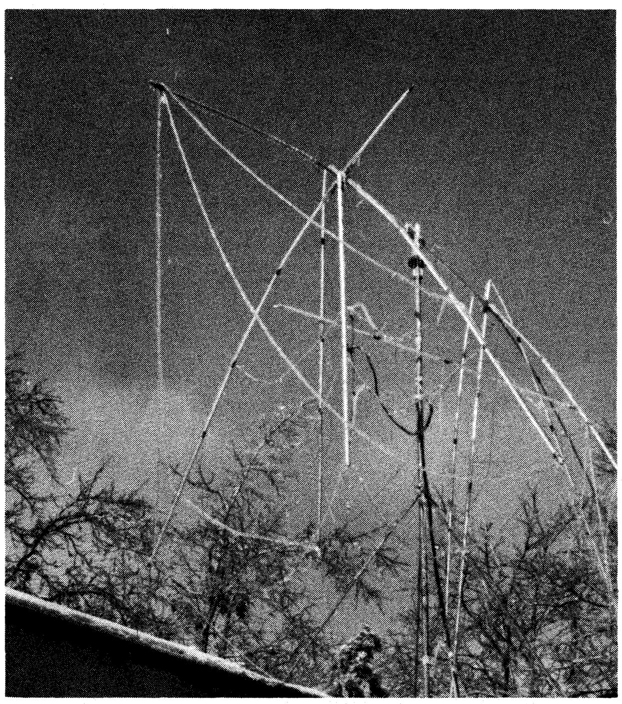
Special thanks are due Herbert Drake, Jr., WB6IMP, Rod Roderique, WA0QII, and Alan Bowker, WA6DNR, for their encouragement and suggestions during the development of the RM-300/RP-400 Modem and power supply.

reference

1. Howard L. Nurse, W6LLO, "CW Memory for RTTY Identification," ham radio, January, 1974, pages 6-12.

ham radio

stressed quad



As you can see by the photo, I have a problem. To eliminate this problem in the future, I have devised a means of stressing a quad (see **fig. 1**). As seen in **fig.** 1, the ends of the spreaders are connected together with 3 meter (10 foot) lengths of 45 kg (200 pound) test mono-filament fishing line. In addition, three 3 meter (10 foot) lengths of conduit are connected together with sleeve couplings and extended through the boom approximately 3 meters (10 feet) beyond the ends of the boom. Additional lengths of fishing line connect the ends of the spreaders to the extended ends of the conduit. With the lines tight, the quad now has the general appearance of a Zeppelin, but is stressed to withstand a large wind load. I also attempted this method with nylon cord, but it had too much stretch; the monofilament fishing line does not. Had I used fishing line first, I would not have this picture.

Ira Hargis, W5TIU

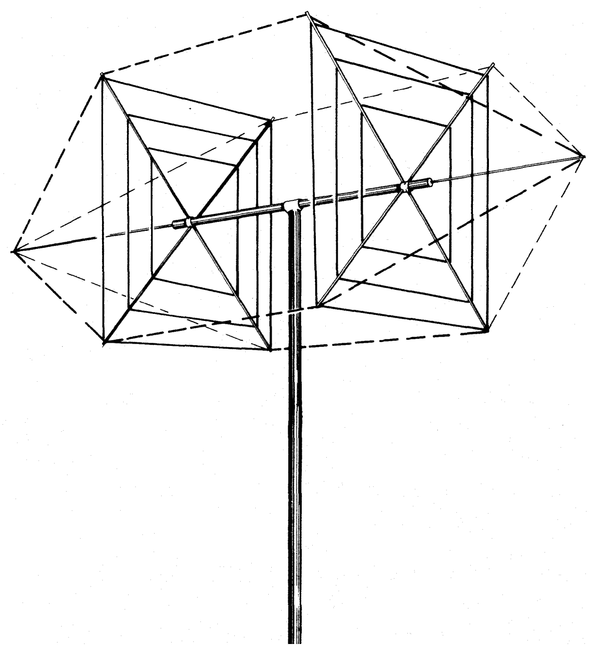


fig. 1. Diagram of the stressed quad as described by the author. The fishing line is almost invisible, yet keeps the spreaders under stress, increasing its survivability during high wind loads.

integrated circuit arrays

Using IC arrays permits the designer to reduce size and component count — without sacrificing performance

Integrated circuits have been around for a decade or so; starting off with just a few transistors on a single chip, inter-connected in amplifier arrangements, they have developed into complex circuits incorporating many transistors, diodes, Zeners, and other components in a single package, often requiring a minimum of external parts to perform the functions of any linear, digital, or logic application. Most hams are familiar with the smaller, linear ICs designed for specific applications. And generally, many would not bother to use discrete transistors for an i-f amplifier or audio power output stage when a single TO-5 can or 14-pin dual inline IC makes construction simpler and more reliable. The majority of ICs are "committed" devices, developed to perform a specific function. Their popularity in amateur construction projects has long been established from the proliferation of articles published on amplifiers, balanced modulators, and phase-lock-loops, in receivers, transmitters, and test equipment.

Much less apparent is the constructor's use of the "uncommitted" IC. These devices are independent diodes and transistors on a single chip, individually connected to the package pins. Known as IC arrays, they are widely used in commercial and industrial equipment for minimizing space and assembly effort. There are also "semi-committed" arrays, in which only two of the transistors have internal connections (such as a differential or darlington pair).

For building equipment, there are several reasons why using IC arrays, instead of discrete transistors or diodes, can be an advantage.

1. In many cases, the IC will cost about the same, or less, than the equivalent discrete components. Due to broad commercial use, many types are fairly inexpensive and easy to find – even at bargain prices from some surplus outlets.

2. The individual transistor and diode parameters are much more closely matched than discrete components of the same type, and also retain matching over wide temperature variations, due to the individual devices being etched on a common substrate. Some arrays have two specially matched transistors, ideal for balanced-circuit applications.

3. Printed circuit board layout may be simplified, using less space than with discrete devices.

4. By using low-profile IC sockets, troubleshooting and repair can be as simple as plugging in a new IC.

Most of the major solid-state device manufacturers

By Peter A. Lovelock, K6JM, Hughes Aircraft Company, P.O. Box 90515, Los Angeles, California 90009

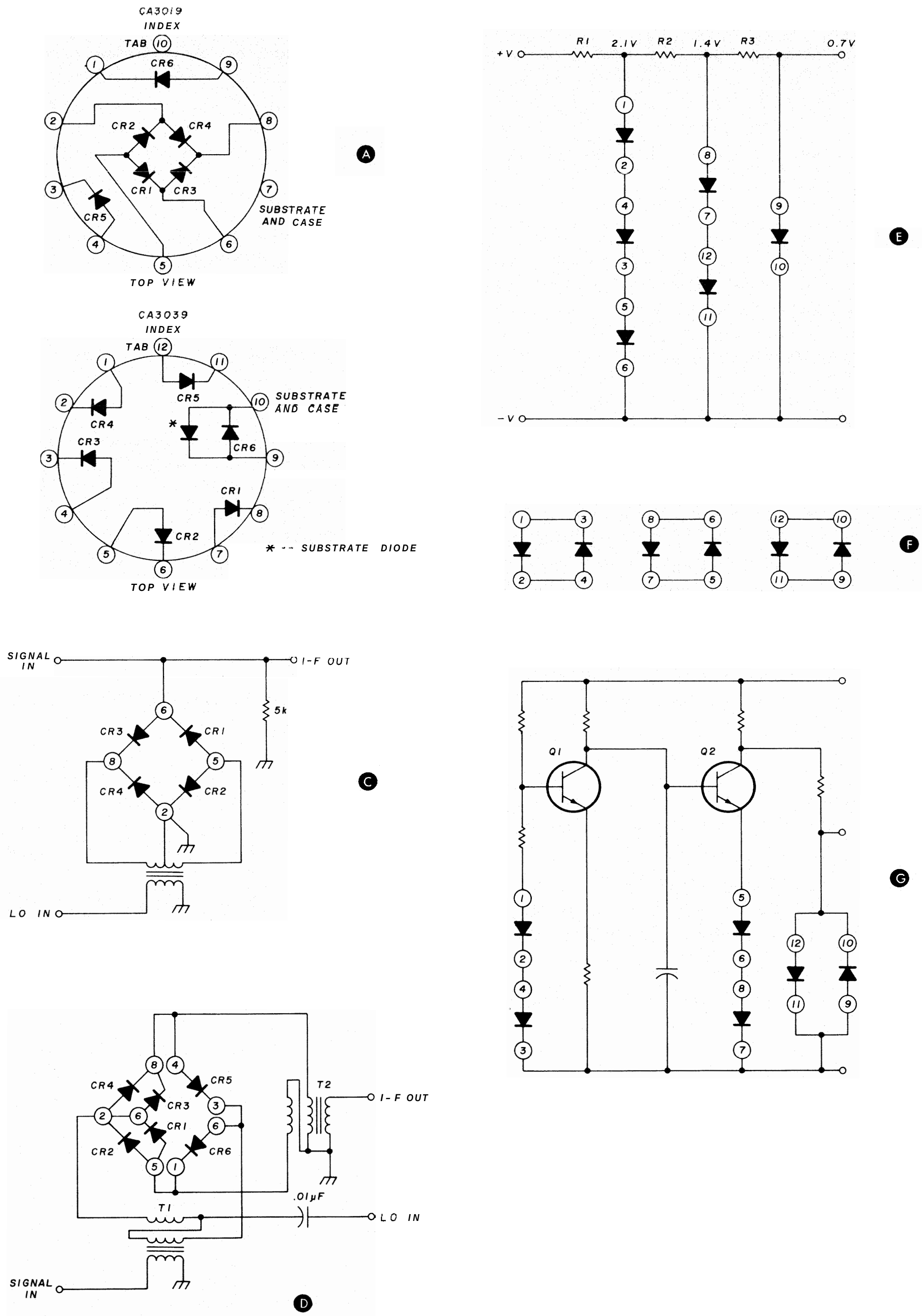


fig. 1. Pinout diagrams for the **CA3019** and **CA3039** are shown in A and **B**, respectively; C and D show two configurations for the **CA3019**, a balanced modulator and a double-balanced ring mixer. A 3-voltage regulator or reference, E, is obtained by using the forward-voltage drop across the diodes. By connecting the diodes back-to-back. they will also function as clippers (**F**). A typical application for almost the entire **CA3039** package, including temperature compensation, emitter bias, and output limiting is shown at G.

produce transistor arrays. For the purpose of reviewing the more commonly available types, reference is made only to the RCA numbers. When a pin compatible equivalent is made by other manufacturers, the RCA number has an asterisk, and the equivalents will be found in table 1. Also, this review is divided into four logical categories, for easy future reference.

1. Diode arrays.
2. Uncommitted-transistor arrays.
3. Semi-committed transistor arrays.
4. Hybrid arrays (incorporating transistors and diodes, both uncommitted and semi-committed).

table 1. Pin compatible IC array equivalents

RCA	National	Fairchild
CA3019	LM3019	μA3019
CA3039	LM3039	μA3039
CA3018	LM3018	μA3018
CA3086	LM3086	μA3086

The review includes only essential information, and some basic applications circuits for each type, to assist device selection. The reader should refer to the manufacturers specification sheets, data manual, and application notes for complete information on any specific device. Some complete IC array circuits, both amateur and commercial, are included after the device review.

diode arrays

1. **CA3019***

Configuration:	Six silicon diodes, four internally connected as a quad, two independent.
Package:	10 pin, TO-5 (fig. **1A**).
Applications:	Modulators, mixers, analog switches (figs. **1C** and **1D**).

2. **CA3039***

Configuration:	Six ultra fast, low capacitance silicon diodes, independently connected.
Package:	12 pin, TO-5 (fig. **1B**).
Applications:	Balanced modulators, demodulators, voltage reference and bias regulators, clipper limiters, (fig. **1D**).

The inherent advantage of the diode arrays (figs. **1A** and **1B**) is the close matching of the diodes when used as single or double-balanced mixers or product detectors. While the diodes in the CA3019 and CA3039 do not have identical characteristics, they are essentially interchangeable in many circuits. Fig. **1C** shows the CA3019 quad as a balanced modulator. For this application, the diode "ring" is more commonly used. Fig. **1D** shows how the CA3019 diodes can be connected as a "ring" in a double-balanced mixer. Since this involves paralleling the quad diodes, (CR1, CR2 and CR3, CR4) connecting pins 2 and 6, and using the independent diodes CR5 CR6 to complete the ring, balance will not be as perfect as with four individual diodes, though adequate for most purposes. If the circuit in fig. **1D** is used as a product detector, it is recommended that the LO be fed into T2 and audio output be taken from the center of T1's secondary.

The CA3039's independent diodes lend themselves to any circuit requiring up to six fast silicon diodes. Figs. **1E**, IF, and **1G** show these diodes used for low-voltage regulation, varistors, or transistor amplifier biasing with temperature compensation and output limiting (clipping). Note that the voltage regulator is for low current, limited to the maximum forward current of the diodes. It is best applied as a voltage reference, rather than high current applications in which reverse-polarity zener diodes are used. By using one to six diodes in series, low current, stable reference voltages of 0.7 V, 1.4 V, 2.1 V, 2.8 V, 3.5 V, and 4.2 V can be derived from one CA3039. As shown, regulated positive voltages are obtained, but by reversing the supply and diode polarity, negative voltages for fet or mosfet device biasing can be achieved.

uncommitted transistor arrays

1. **CA3083/3183/3183A***

Specification:	Five general-purpose silicon NPN transistors, independent substrate connection. Two transistors (Q1, Q2) matched. $I_{,,,,}$ 100 mA, h_{FE} 76 typical.
Package:	16 pin, plastic DIP (fig. **2A**).
Applications:	Signal processing/switching from dc to vhf, lamp and relay driver, differential amplifiers (fig. **2E**).

2. **CA3127**

Specification:	Five high-frequency general-purpose silicon, NPN transistors, with independent substrate connection. DC to 500 MHz. Low noise (3.5 dB at 100 MHz). High power gain (30 dB at 100 MHz).
Package:	16-pin plastic DIP (fig. **2B**).
Applications:	vhf amplifiers, mixers and oscillators. I-f amplifiers, synthesizers, synchronous detectors (fig. **2F**).

3. **CA3096A/3096AE**

Specification:	Three general-purpose high-voltage silicon NPN transistors, and two general-purpose, high-

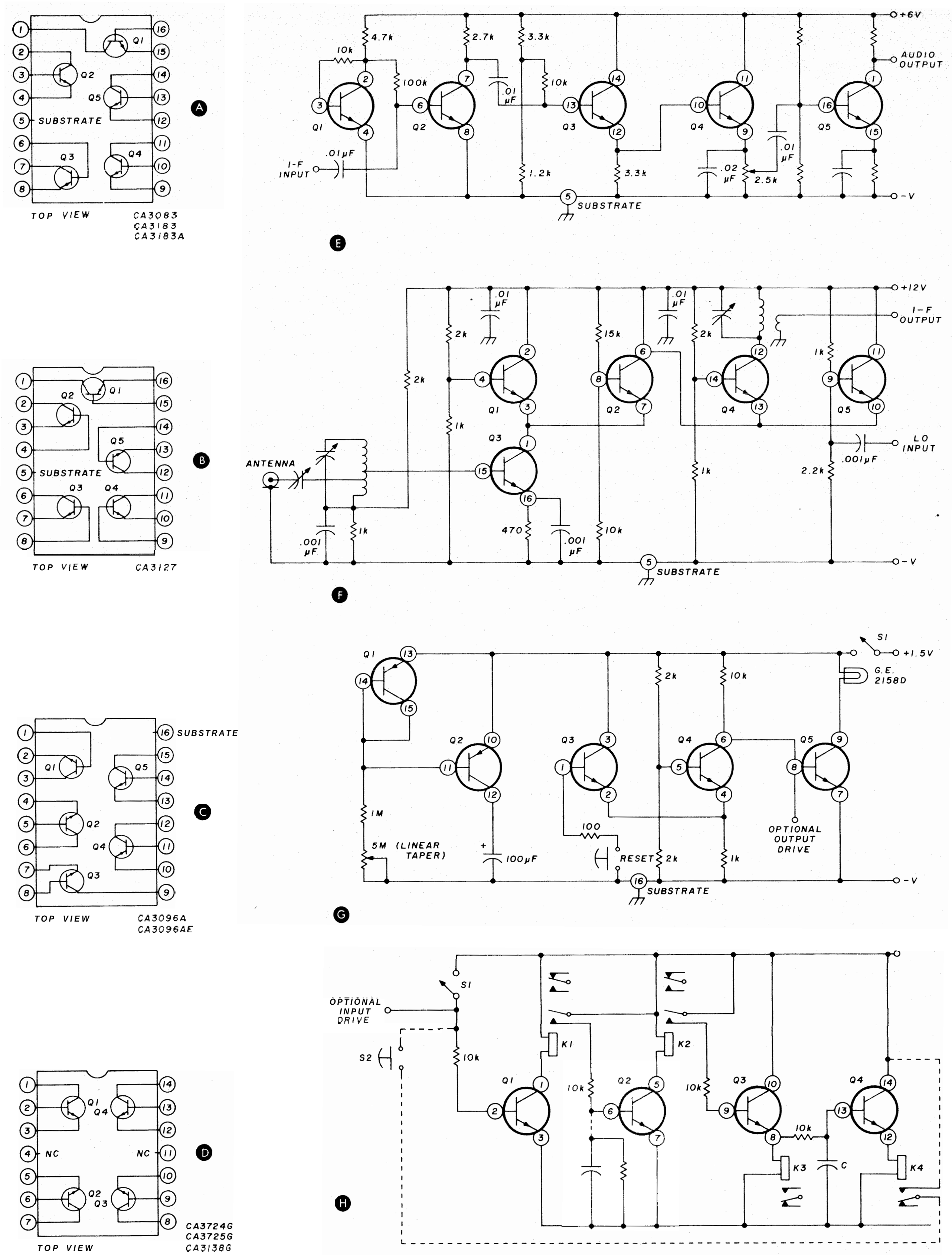


fig. 2. Pinout diagrams for different transistor arrays are shown in A through D; typical applications are shown in E through H. In H, the parallel time constant, from the base of Q2 to ground, causes K2 to open only after **K1** has opened. This delay is adjustable according to the values used for the resistor and capacitor. By using S2 and the drive line from K4, the relays can be activated by **S1** and de-energized by S2.

voltage silicon, PNP transistors, independent substrate connection.

Package: 16 pin, plastic DIP (fig. **2C**).

Applications: Differential amplifiers, dc amplifiers, timers, lamp and relay drivers (fig. **2G**).

4. **CA3724G/3725G/3138G**

Specification: Four high-current silicon NPN transistors. I_c 1 A fast switching (30 ns at 0.5 A).

Package: 14 pin, plastic **DIP** (fig. **2D**).

Applications: High-speed switching, high-voltage switching, high-current LED, lamp and relay drivers (fig. **2H**).

The first two devices in this group, the CA3083 and CA3127, have a lot in common, both having five independent NPN transistors which can be used in a very wide variety of circuits. However, the transistor characteristics in each type are quite different, even though there may be some circuits in which either could be made to work. The CA3083 has higher I_{max} and is suited more for hf and audio circuits which require transistors with high-signal level parameters. Fig. 2E is a typical application for a final i-f amplifier (untuned), second detector, and first audio amplifier, using all five transistors in the CA3083. Q1 is used for bias stabilization of the first i-f amplifier. If Q2 was conventionally biased with two resistors, Q1 could be applied to some other function.

In contrast, the CA3127 incorporates transistors specifically designed for small signal, high power gain characteristics up to 500 MHz, making it ideal for receiver front-end use, including rf amplifier/mixer/oscillator combinations.[1] Fig. 2F is such an application, using all five transistors for the rf amplifier and mixer stages of a front-end suitable for vhf. Needless to say, with the high power gain of the individual transistors, considerable care should be exercised in the layout of such circuits to avoid building only oscillators.

The CA3096 is an interesting IC which contains a mixture of three NPN and two PNP transistors. The CA3096AE has more closely matched parameters between transistors, useful for complementary pair circuits.

Fig. 2G is the circuit of a ten-minute interval timer using all the transistors in the CA3096 – a handy device as an identification reminder for ragchewers. The time constants allow adjustment of the timer from 1.6 to 10 minutes by changing the 5-megohm potentiometer. The maximum time can be changed by increasing or decreasing the value of the 100 μF capacitor; minimum time is varied by changing the value of the 1-megohm fixed resistor in series with the pot. The lamp indicator may be replaced by a relay or audio reminder (such as NE555 oscillator), so long as Q5's collector current does not exceed 50 mA, or a load of less than 30 ohms at 1.5 volts.

Since the CA3724 was meant to be a high-current driver, an appropriate application is shown in fig. **2H**. Many times it is necessary to close and release relays in proper sequence. Closing S1 turns on Q1, which closes K1, but two normally-open contacts on K1 are employed to put forward bias on Q2. Thus, K2 can close only after K1. Opening S1 will cause both relays to release at almost the same time.

If reverse sequencing is required, a parallel RC time constant, from Q2's base to ground, will cause K2 to open after K1. K3 and K4 are driven by the emitters of Q3 and Q4, with the voltage across the coil of K3 used to drive Q4. This technique eliminates the need for a pair of normally open relay contacts as in the Q1, Q2 arrangement. The single capacitor, from base to ground of Q4, provides for a time-delay release of K4; the time constant is dependent on the value of C, the 10-k ohm base resistor, and the resistance of K3's coil.

If a latching relay combination is desired, the normally open contacts of K4 can be used to connect drive through S2 (momentary pushbutton, normally on) to the base resistor of Q1. S1 should be changed to a momentary pushbutton, normally off. Push S1 to activate the relays and push S2 to release.

The circuits in fig. 2G and 2H can be combined by connecting the optional drive output, from pin 8 of the CA3096, to the optional drive input of the CA3724. With this combination, the timer will activate the relay driver circuit after a preset time delay, initiated by the timer ON switch. This arrangement would be useful for a high-voltage time delay turn on in tube-type linears. S1 would, of course, be eliminated in this arrangement, but S2 may be left as an emergency OFF switch.

semi-committed transistor arrays

1. **CA3018/3018A/3118/3118A***

Specification: Four general-purpose silicon NPN transistors, two independent; two internally connected as a Darlington pair. Independent substrate connection.

Package: 12 pin, TO-5 (fig. **3A**).

Applications: General use in signal processing, dc through vhf range. Suitable for use in circuits similar to CA3083.

2. **CA3086/3146/3146A/3045/3046***

Specification: Five general-purpose silicon NPN

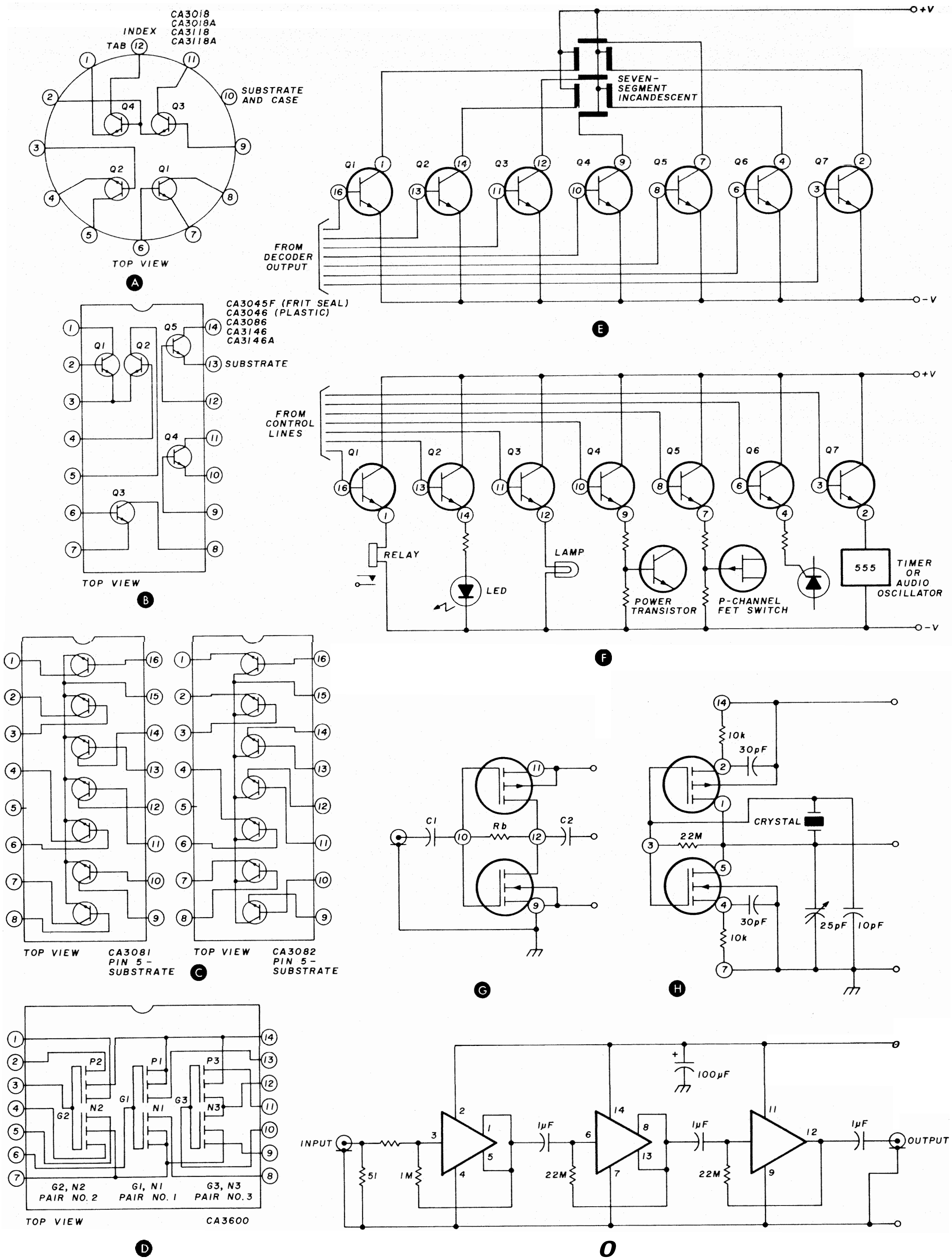


fig. 3. Pinout diagrams for transistor arrays which have internal connections between the transistors (semi-committed arrays). Various applications for the arrays are illustrated in E through **I**. Using the **CA3600** mosfet array in **I** will limit the upper frequency to approximately 5 MHz. Even though the **CA3600** has protected inputs, normal precautions should be exercised while working with the **IC.**

transistors, three independent; two internally connected as a differential pair (common emitters).

Package: 14 pin, DIP package type varies with type number. CA3086/3146/3146A are plastic package (fig. **3B**).

Applications: General purpose use for signal processing in the dc to 120 MHz range.

3. **CA3081 and CA3082**

Specification: Seven high-current silicon NPN transistors. (CA3081) All emitters internally connected in common. (CA3082) All collectors internally connected in common.

Package: 16 pin, plastic DIP (fig. **3C**).

Applications: Lamp, LED, and relay drivers.

4. **CA3600**

Specification: Complementary mosfet array of 3 N-channel and 3 P-channel enhancement mode mosfets, gate protected.

Package: 14 pin, plastic DIP (fig. **3D**).

Applications: High input-impedance linear amplifiers, low-power oscillators.

As mentioned earlier, this group of arrays has two or more of the transistors internally connected in a partial circuit, but this does not eliminate their use as independent transistors, though there are more limitations than the undedicated group.

The characteristics of the individual transistors in the CA3018 and CA3086 families are sufficiently close to the CA3083 so that RCA specifies the same application note for all three. The CA3083 has an I_{cmax} of 100 mA, while the CA3018 and CA3086 have I_{cmax} of 50 mA. Otherwise, the parameters are close enough to make these three devices interchangeable in almost any circuit where the partially committed arrangement of the latter two permits. In both the CA3018 and CA3086, the four transistors may be used independently by allowing the pins of the fifth device to float unconnected. Or, in the CA3086, the pins of the differential pair may be jumpered (5-1 and 4-2) creating a single transistor with an I_{cmax} of 100 mA. Also, the darlington pair in the CA3018 may be used as a single, super-beta transistor by using pins 11 and 12 as the collector, 9 as the base, and 1 as the emitter. Since applications for the CA3018 and CA3086 are similar to the uncommitted CA3083, no basic circuits are shown for these types,

The CA3081 and CA3082 both incorporate seven NPN transistors having identical characteristics, the CA3081 having all collectors connected together, and the CA3082 with all emitters connected. These devices were designed to drive seven-segment displays as shown in fig. 3E. However, these devices can be used for low-current, remote driving of a variety of other components including relays, LEDs, incandescent lamps, power transistor switches, fet switches, SCRs, and complex devices such as an NE555 oscillator. The only constraint is that current supplied to the driven device does not exceed the I_{cmax} of 50 mA for each transistor. If a higher current rating is required, two or more of the transistors may be paralleled.

The CA3081 may also be used as seven independent amplifiers in circuits where grounded emitters are appropriate. Likewise, the CA3082 may be employed as seven emitter followers with independent inputs and outputs. Such circuits may be used as buffer mixers (separate inputs, common output) or distribution isolators (common input, separate outputs).

The CA3600 may be one of the less familiar ICs. This device has three pairs of complementary mosfet transistors, with the input gates of each pair connected. Each pair has one P-channel and one N-channel enhancement-mode mosfet. The N-channel, depletion-mode mosfet is familiar to most, appearing frequently in single- or double-gate versions for rf amplifiers and mixers. The depletion-mode devices have channels which are normally ON in the absence of gate bias. Enhancement mode mosfets have channels normally OFF until bias is applied to the gates, positive bias for P-channel and negative bias for N-channel. Using an ac input signal as bias, the linear amplifier shown in fig. 3G will have the P-channel transistor conducting on positive half cycles, and the N-channel device on negative half cycles. The output from C2 will be a complete, amplified replica of the ac input.

One nice thing about complementary amplifiers of this type is that they maintain a high-linearity output swing, over almost the entire limit of supply voltage. Therefore, you can obtain close to 10 V p-p output with a 10 Vdc supply. In addition, the input impedance is very high, with the circuits requiring a minimum of external components. Fig. 3H shows the basic amplifier expanded into a crystal oscillator, and fig. 3I illustrates the entire IC connected as a complete, cascaded 100 dB amplifier. The high input impedance and high off-resistance of the P and N channel enhancement-mode mosfets make them highly suited for switching circuits where the load current does not exceed 10 mA. For those who are hesitant to use CMOS devices, considering their sensitivity to static voltages, the CA3600 device has resistor/zener diode-protected gates to minimize the chance of static damage.

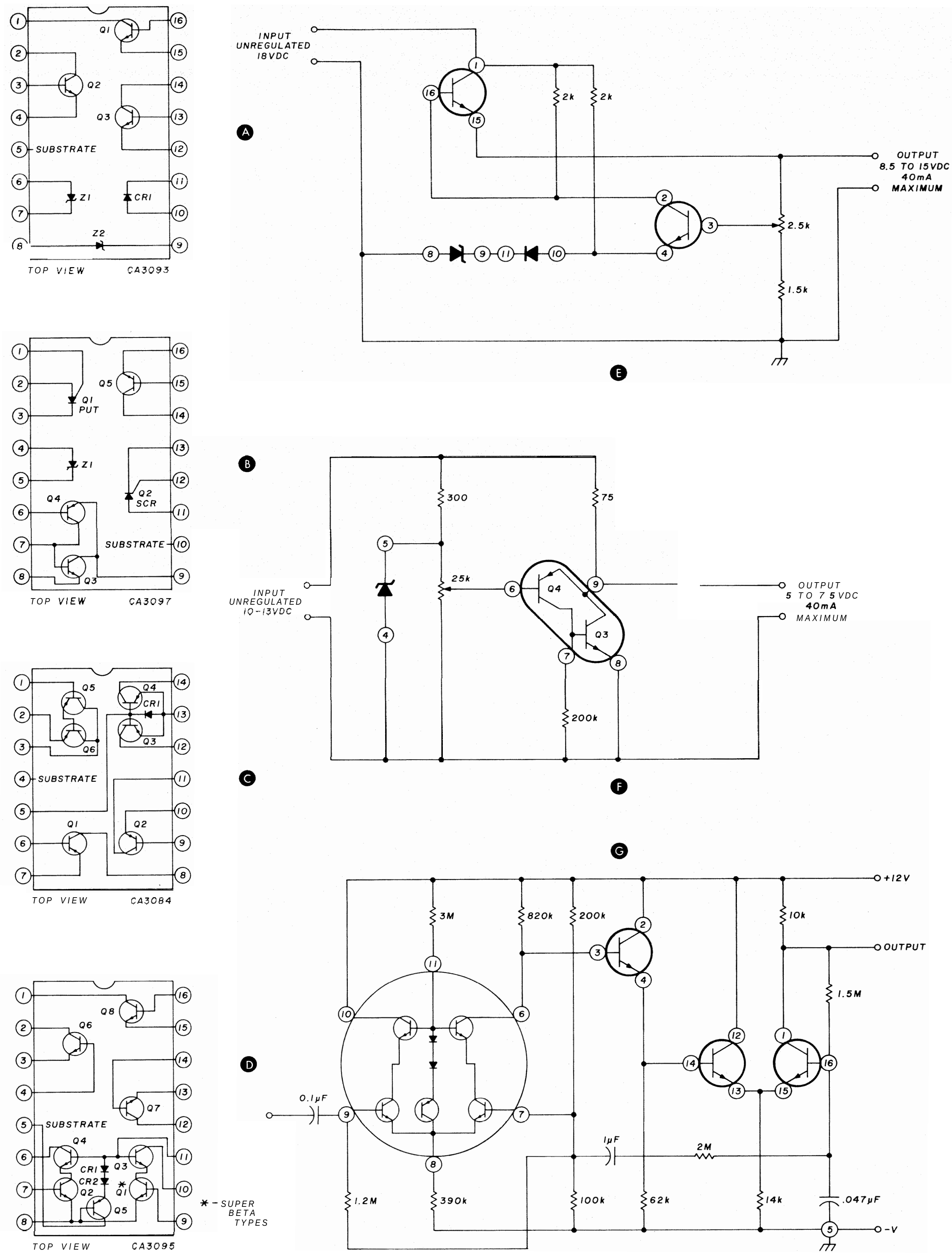


fig. 4. Pin diagrams for the hybrid integrated circuit arrays are shown in A through D. Applications for this group of arrays include voltage regulators and a low-noise preamplifier, as shown in E thorugh G. The amplifier in G has a voltage gain of 30 dB and a noise factor of 2 dB at 10 Hz, decreasing to 0.3 dB at 1 kHz.

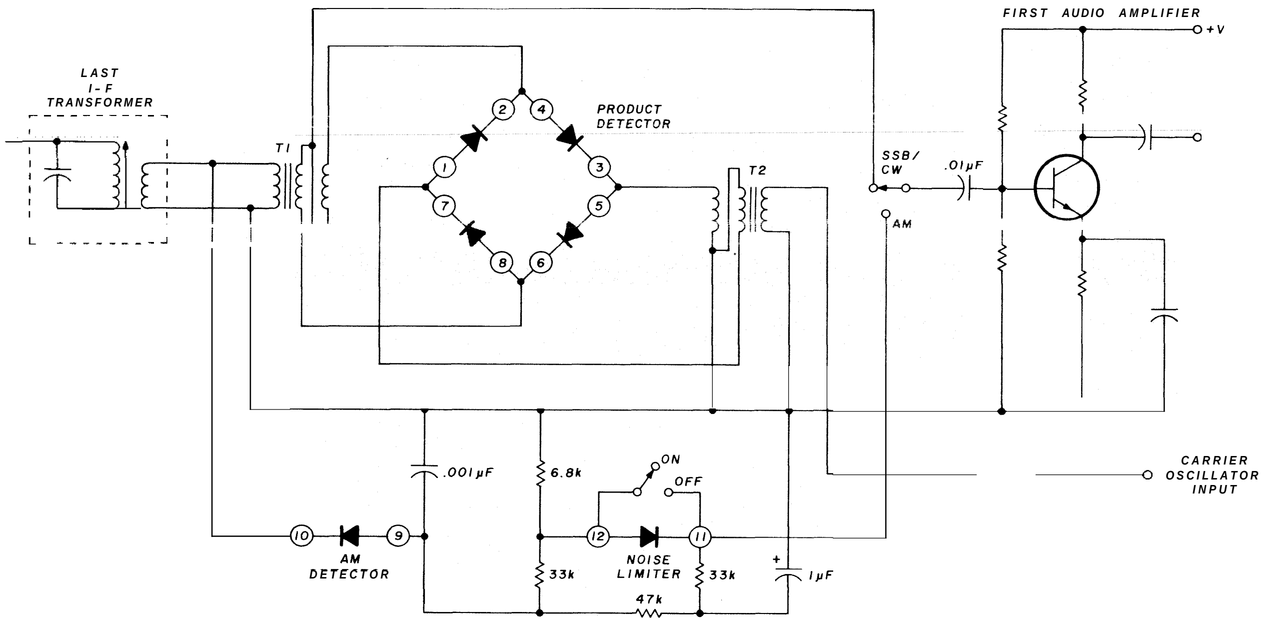


fig. 5. Schematic diagram of a product detector, a-m detector, and noise limiter using the CA3039. T1 and T2 are Q2 ferrite cores wound with 12 turns of no. 26 AWG (0.4 mm) enameled wire. The individual wires should be twisted approximately 1 turn per centimeter (3 turns/inch). The carrier oscillator signal (+7 dBm) must be off when the detector is used for a-m.

hybrid arrays

1. CA3093

Specification: Three independent general-purpose, high-current silicon NPN transistors — two independent 7-volt, 114-watt zener diodes; one general-purpose silicon diode. Two of the transistors, Q1 and Q2, are closely matched at 1 mA.

Package: 16 pin, plastic DIP **(fig. 4A)**.

Applications: Signal processinglswitching from dc to vhf. Temperature compensated voltage and current regulators.

2. CA3097

Specification: One independent silicon NPN transistor. An NPN and PNP transistor pair, one zener diode, one programmable unijunction transistor, and one silicon-controlled rectifier.

Package: 16 pin, plastic DIP **(fig. 4B)**.

Applications: Voltage regulators, timers, constant current source, oscillators, multivibrators.

3. **CA3084**

Specification: Two independent general-purpose silicon PNP transistors, two PNP transistors connected as a Darlington pair, combination of two PNP transistors and one diode connected as a "current mirror."

Package: 14 pin, plastic DIP **(fig. 4C)**.

Applications: General signal processing, low power, low frequency; double-balanced mixers/modulators, product detectors.

4. **CA3095**

Specification: Three independent high-voltage silicon NPN transistors, five NPN transistors, differential amplifier array, two of which (Q1 and Q2) are super-beta types.

Package: 16 pin, plastic DIP **(fig. 4D)**.

Applications: Super-beta preamplifiers, high-impedance dc meter amplifier, low-noise video amplifier. The independent transistors are usable for signal processing from dc to vhf.

This group of arrays will test your imagination for using different solid-state devices. However, the manufacturers were not arbitrary in selecting the different devices.

Fig. 4E, one example of a use for the CA3093, is a temperature compensated, series voltage regulator, adjustable from 8.5 to 15 volts at a maximum of 100

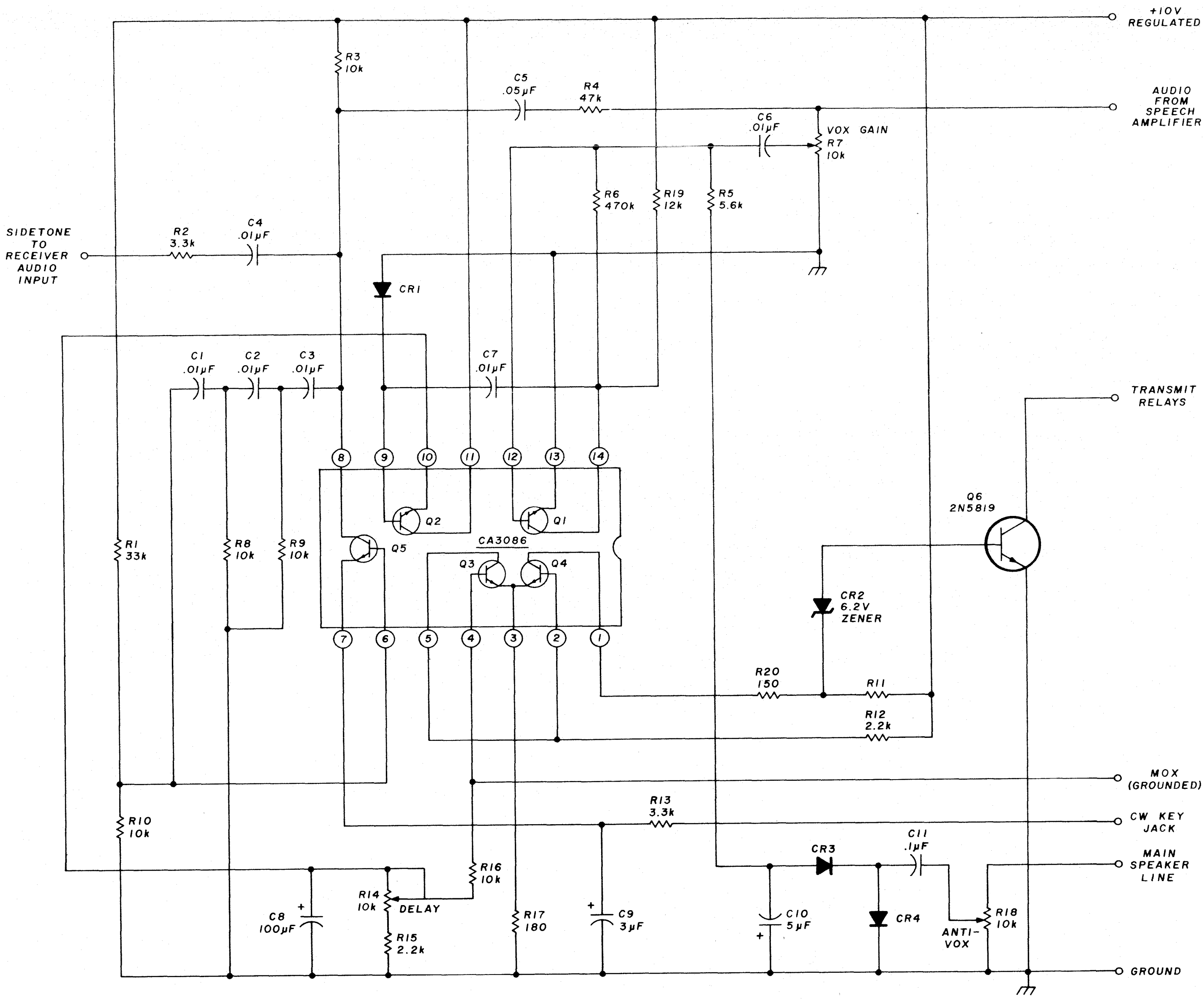


fig. 6. Schematic diagram for a vox/semi-break-in system using the CA3086. Q6 is an external transistor which must have a collector current rating sufficient to control the external relays. CR2 is a 6.2 volt, l-watt Zener diode. CR3 and CR4 must be germanium diodes.

mA. **Fig. 4F** is a shunt voltage regulator using the CA3097, leaving one transistor, SCR, and a programmable UJT for other use.

The CA3084 includes four general-purpose PNP transistors, with Q3 and Q4 connected as a darlington pair. Also included is a "current mirror" arrangement of two PNP transistors and a diode. The two independent PNP transistors are closely matched, and can be employed in separate circuits, or as part of a balanced complimentary circuit with NPN transistors. The darlington pair, having three basic connections (base, collector, emitter), can be considered as a single super-beta transistor with an h_{FE} of 1250. The PNP current mirror is suited as an active load for differential amplifiers which use NPN transistors. In general, the CA3084 device is used to furnish circuit sections for other ICs or discrete devices, rather than providing a complete circuit function.

Three independent general-purpose NPN transistors are contained in the CA3095, in addition to a differential-amplifier array of five NPN transistors and two diodes. Two of the transistors (Q1 and Q2) are super-beta types with an h_{FE} of more than 1000. The differential amplifier and independent transistors may be used separately, or combined **(fig. 4G)** into a very high-input impedance, low-noise amplifier.

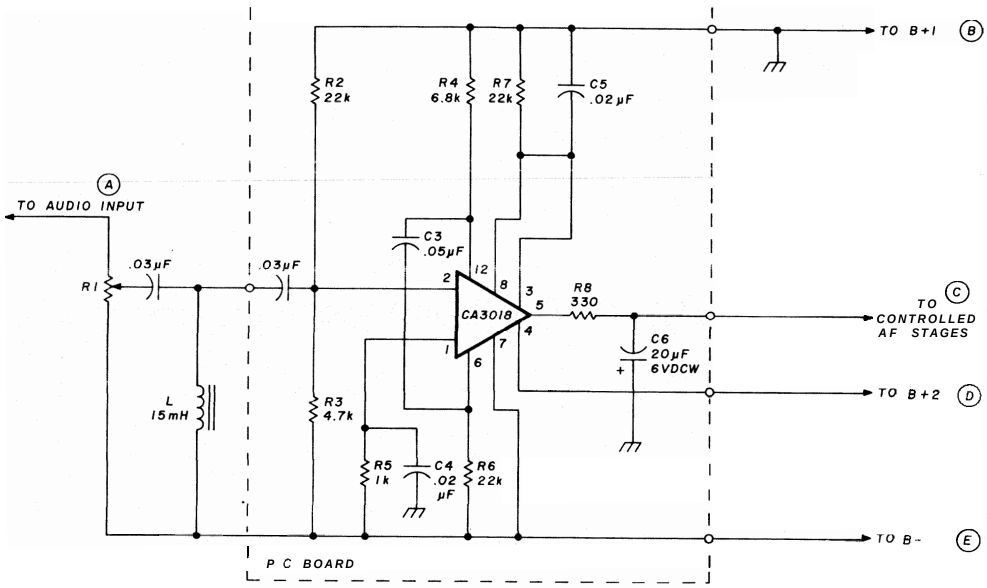


fig. 7. Schematic diagram of a squelch circuit using the CA3018.

getting the most from practical circuits

The basic circuits, illustrating the use of each type array, were intended as examples of the versatility of these devices. Many different uses will be found by the enterprising builder. Some of the array devices, such as the uncommitted transistor arrays, will find many more applications than the more specialized types.

Determining whether to build a project with an IC array or discrete components is a matter of planning and trade-offs. Once you have a circuit to construct, research the available devices to see if one of them incorporates all, or most, of the active components required. After all, it doesn't make sense to use a CA3093 to use only two transistors, leaving the third, plus two zeners and a diode as spares; though on occasions, one unused section can come in handy for later modifications. The real trick is working all of the active IC sections into a construction project, and

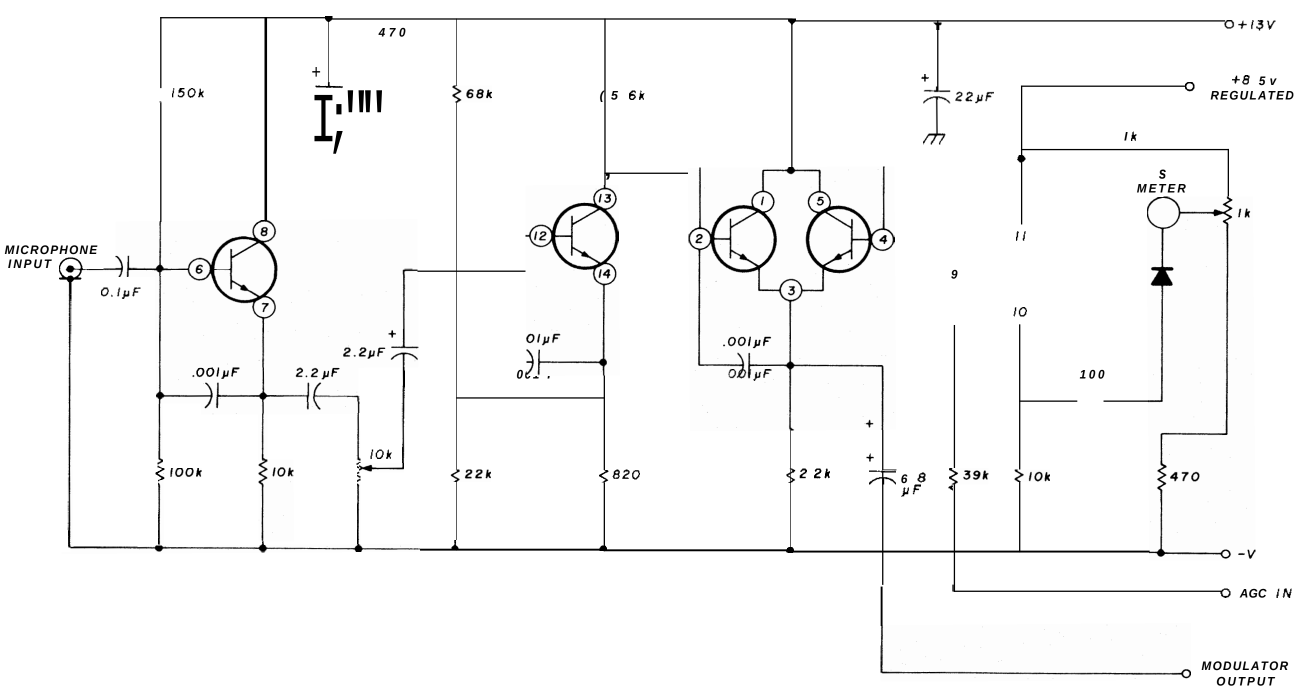


fig. 8. The CA3086 has been used to its fullest advantage in the Atlas series transceivers. This schematic shows the microphone/modulator amplifiers. and S-meter amplifier in the transceivers.

this article would not be complete without including some practical amateur and commercial circuits where this has been achieved to advantage.

Fig. 5 is a double-balanced, diode-ring product detector, a single-diode detector, and a series-gate noise limiter, using all the devices in a CA3039. The TO-5 package occupies less space than individual diodes flat on the board. In addition, the matched diodes give superior performance in the product detector. A module of this type, to replace the triode product detector in a tube receiver, has been built on a printed-circuit board measuring only 4 x 4 cm (1-1/2 x 1-112 inch).

Fig. 6A is the circuit of a complete vox and semi-break-in CW module, including sidetone using a single CA3086 IC. One external transistor, Q6, is used as a solid-state switch to energize the transmitter relays.

The output from Q1, the input audio amplifier, is rectified by CR1, applying bias to Q2. This emitter follower (Q2) with a variable RC time constant in the emitter lead, sets the vox delay. The differential pair 0 3 and 04, are connected as a Schmitt trigger, with Q4 normally conducting, its collector voltage is about +2 volt and the 6.2 volt zener (CR2) cannot conduct. When rectified audio causes the Schmitt trigger to switch, Q4 conducts and the resultant 10-volt collector voltage causes CR2 to conduct, turning on Q6. 0 5 is an RC coupled, phase-shift audio oscillator, with an output from 700 to 800 Hz. When its emitter resistor is grounded through the key jack, the tone output is fed to the transceiver (or receiver) audio output stages, providing a sidetone in the CW mode. The oscillator's output is also fed to Q1, causing the vox circuit to function in the same manner as with a normal microphone input. A manually operated mode is provided hy grounding the base of 03. CR3 and CR4 rectify the speaker output, providing the anti-vox control.

Fig. 7 is the circuit of a squelch module, based on the CA3018,[2] for use with fm receivers. The entire circuit was built on a board measuring only 2 x 3.5 cm (0.8 x 1.4 inches).

The circuit for the microphone amplifier, balanced-modulator audio driver, and S-meter amplifier stages used in the popular Atlas Radio 2101215 series transceivers is shown in **fig. 8.** This is an excellent example of making full use of an IC array. Q1 is a high impedance emitter follower, Q2, a voltage amplifier, Q3 and Q4, a differential pair strapped to form a single low-impedance emitter follower.

references

1. Bill Hoisington, K1CCL, "Two High-Gain RF Stages in One IC for 2-Meter FM," *73,* May, 1974, page 47.
2. Peter Lovelock, K6JM, "The Postage Stamp Squelcher," *73,* May, 1975, page 103.

ham radio

tracking down repeater jammers

Malicious interference is increasing on the amateur bands, particularly on 2 meters — here are some ideas for concerned hams who want to do something about it

Repeaters seem to attract a variety of intentional jammers. Kerchunking is by far the most common offense, and nearly every 2-meter operator does it at one time or another to check that the rig is getting out. But there are others who do it regularly and repeatedly, even to the point of trying to play "Jingle Bells." They have been heard to say, "I guess I'll wear out another relay."

the problem

Almost every time a 2-meter rig is ripped off in the mistaken belief that it can be peddled as CB, there's an epidemic of jamming. It starts with kerchunking, progresses to attempts to raise another CBer, then often proceeds to false calls, obscenity, or the type of operation that's led to the term "Chicken Band."

Another source of repeater jamming is the malcontent who's unhappy because the repeater is operating on *his* frequency. Some of his tricks are blocking signal levels, noise modulation, endless kerchunking, and retransmission of signals from other channels or of other services. A really mean trick is to key the transmitter at a rate that may simulate "picket fencing," or may lead users to believe that a rig or repeater malfunction has occurred.

Amateur clubs (and stations) should have the ability to track down and identify these jamming sources.* For one thing, this can lead to the recovery of a stolen rig and to the word being passed to stay off ham equipment. Also, besides leading to a better repeater operation, the track and identification capability should be of major value in a manmade noise program. This can help all hams, not just repeater users.

Although the operations involved require individual effort, it's probably better to undertake such a program as a club effort. Some reasons for this are discussed at the end of the article.

interference plan

For an anti-jamming or interference program to work, a plan of action is needed. The basic elements are:

1. Detection
2. Alerting
3. Preliminary locating
4. Spotting
5. Identifying
6. Acting

Each involves some effort and some time. Several elements also involve some skill, but the effort per person can be kept low. These factors are among the reasons this anti-jamming effort makes a good club project.

detection

Detection can be left to chance, but it's a good idea to have at least some planned effort. In fact, jamming detection can also serve another purpose — intruder detection: the steps and procedures are the same.

One way to handle the detection problem is to

By R. P. Haviland, W4MB, 2100 South Nova Road, Daytona Beach, Florida 32019

*An example of what's being done to stem the vhf jammer problem appears in reference 1. **Editor.**

combine it with another one. At any time a club probably has one or more members who are shut in for some reason, usually illness. Many of these people are happy to have some light chore to help pass the time. Monitoring can help. The club can help by having a set of loan equipment available, or possibly two sets, one receive-only for nonlicensed members.

While any reasonable antenna and receiver combination is suitable for monitoring, a comprehensive plan should be based on the use of scanners. Very good commercial scanners covering up to ten channels or so are available at reasonable cost. Panoramic vhf receivers can sometimes be found in surplus sources, or a Panadaptor can be added to a vhf receiver. Incidentally, it doesn't make much difference whether an a-m or fm receiver is used. So older rigs can serve the monitoring function. A mechanical dial drive along the lines of the General Radio 1521 drive unit could be worked out for older, continuously tuned rigs.

Owners of synthesized rigs are in a good position to check all or at least a good number of the channels. However, doing this by throwing lever switches becomes very tedious. The procedure can be mechanized by the simple addition of a counter chain that feeds the rig's internal counters. The basic principles are shown in **fig. 1.** The OR gates can be mechanical switches, but a diode or IC OR gate is simpler. Some form of level changing will be needed to allow the squelch to stop scanning; it can be a relay. The same circuit can start a recorder.

A few simple logic circuits can be used to give a no-noise, unattended monitor system, which can also reduce time in checking for interference. The trick is to use two tape recorders. One would be set up to run when a signal is detected. Ideally it would be a two-track unit with time on the second track, say by digital clock readout or WWV/CHU recording. The other recorder would start when a signal is received and run for about thirty seconds. Then it

Public awareness of personal two-way radio communication is keeping pace with the population — both are increasing. Adding fuel to the fire is the increasing availability of low-cost equipment, which means that almost anyone who wants to own a two-way radio can be in business with little effort and a modest outlay of cash, often without bothering to obtain a license. Along with this upsurge of personal radio activity is the inevitable: an increase in malicious interference by those of marginal mentality who seem to enjoy disrupting legitimate communications. The problem exists in the Citizen Band radio service as well as the Amateur service; the problem is particularly severe on 2 meters.

Radio amateurs are noted for policing their operating frequencies. In this article, author Haviland offers some ideas for those interested in ferreting out jammers on two meters. He presents an interference plan and suggestions for implementing it together with suggestions on avoiding legal pitfalls during final action. The theory behind the program is that, once the program is working and the pressure is on, jammers will give up and disappear. **Editor.**

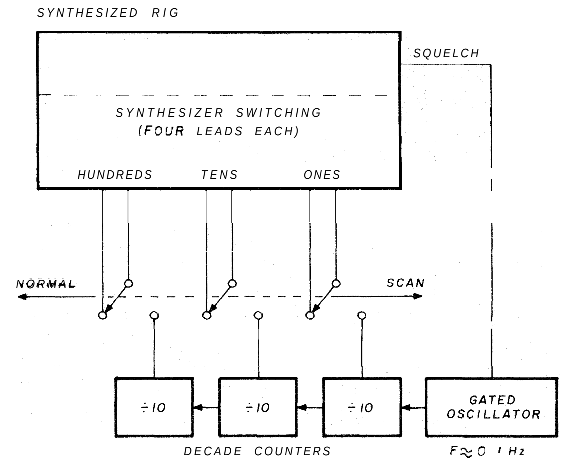


fig. 1. Adding scan-capability to a synthesized radio. Switching can be mechanical or by logic elements.

would not restart until the signal had been off for, say, fifteen seconds. This avoids recording most legitimate contacts and long patches of interference. The compression allows fast review of a day's record. If suspicious signals are found, the full tape can be reviewed.

This type of system doesn't permit immediate action, but it is good for uncovering the operating habits of intruders, or of intentional jammers. It also gives a useful record for analysis and possible action.

alerting

The purpose of alerting is to bring the locating and identifying resources of the interference plan into play. For several reasons it's best to use a small number of alerters. This helps prevent false alerts and callup for marginally traceable interference. Also it helps prevent annoyance to a number of people — if a dedicated, intentional jammer is around it's virtually certain that he'll try to harass all concerned by trying to break up contacts by 3 AM phone calls or even dirtier tricks.

The alerters must be prepared to put up with such tricks and should be ready with appropriate counters — fast frequency change, power increase if necessary, and call monitoring by the phone company. In severe problems it may be necessary to go to a two-tier alerting plan — a fairly large number of people who will relay a detection notice to the actual alerter.

The alert should be in stages. The first step is to determine if further effort will be worthwhile. This can be done by one or two stations who can tell if the signals are strong enough to get a good DF bearing, if the interference source is fixed or moving, if it's distant or local, or if the problem warrants further action,

If further actions seems in order, enough direction-finding stations should be alerted to get a good fix.

One situation needing fast action is that of a ripped-off rig being tested. This is almost always a local fixed-station problem. Chronic and intentional interference may be local or distant, but doesn't require immediate action — it may be better to wait, perhaps

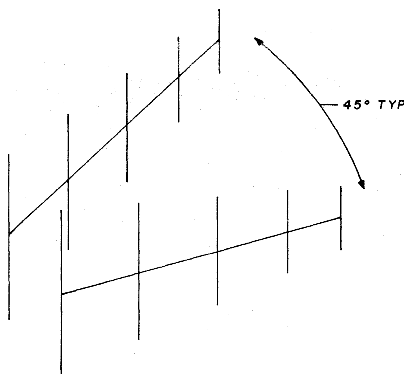


fig. 2. Skewed Yagis for direction finding. Feed may be switched between antennas or at the receiver.

until a key DF station is available or until signal levels increase so that a good DF bearing and good identification are possible.

Overall, the alerting must involve decisions on:

1. Preliminary evaluation
2. DFing
3. Mobile DFing
4. Identification
5. Urgency

Some of these decisions can be handled by landline but others require coordination and passing of data. A "jammer net" is good, with the alerting station acting as net control. Obviously, coordinating operations should not be on the channel subject to interference. Probably the operations should be simplex, although the net alert can be called by a repeater.

direction finding

The DF techniques will vary somewhat for fixed and mobile interference and for local and DX signals. The standard rules apply, of course: a minimum of two bearings are needed. A minimum signal-to-noise ratio is needed for good angle accuracy, and for good geometric resolution the bearings must cross at angles approaching 90°.

For preliminary evaluation, bearings can be obtained from any station with a rotary beam. Two such stations may give sufficient accuracy on local signals to justify dispatching one or more mobiles for pinpointing the source.

Better DF accuracy can be secured with a special DF antenna setup. One simple method, shown in **fig.** 2, is to mount two identical beams at an angle of 40° to 60° using two feed lines to a switch, which may be at the receiver. Rotating the array until the signals from the two beams are the same gives the signal bearing as the bisector of the angle between the arrays.

Somewhat better accuracy can be obtained with phased antennas. The Adcock array of **fig.** 3 is very good. For this antenna the bearing is the angle giving a null signal and lies at right angles to the plane of the elements. Watch the 180° null.

For distant signals working at the null is not too accurate, and it may be necessary to use lobe-switched arrays as shown in **fig. 4.** As for the skewed array, **(fig. 2)**, the bearing is that giving equal signals as the switch is operated.

Various refinements to these simple techniques can be made, such as fast switching, synchronized indicator switching for right-left indication, and visual presentation. See the various handbooks, such as Terman's *Radio Engineers Handbook* for ideas and data.

If the interference source is moving a special problem exists, which is related to the problem of relative motion for ships at sea. See Dutton's *Navigation* for solution to these problems by use of the maneuvering board.

pinpointing

Sometimes bearings from fixed stations will give sufficient accuracy to pinpoint a source, but it's often necessary to call on mobiles.

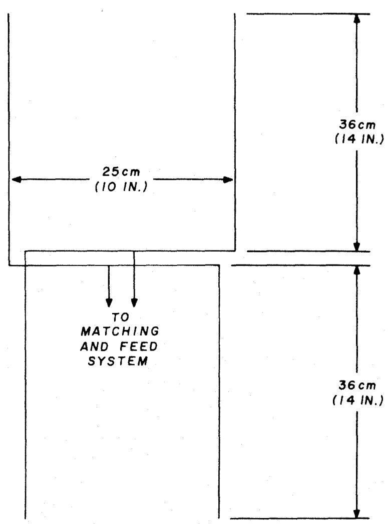


fig. 3. Simplified Adcock array. Matching may be omitted for receive only. The feed impedance is low. System can be made from simple materials.

Fair results can be obtained with mobiles having simple antennas if the receiver has an S-meter.[1] The procedure is to drive in directions that give signal-strength increases. Driving around a suspected area may confirm that the signal source is enclosed, and it may be possible to spot an antenna system. Mobiles doing this should remain on public access roads or areas and probably should not stop for any reason.

Better and faster results can be secured if the mobile has DF capability. Probably the simplest DF unit is made of two half-wavelengths of wire mounted on a wood pole, as shown in **fig.** 3. It's not too difficult to work out a window mount, which will allow DFing while in motion.

With this type of antenna, a very close fix is possible by using the process called "doubling the angle on the bow," a nautical technique. Drive in a straight line past the suspected location. When the angle to the signal is 45° to the line of motion, note the position along the route. Continue driving until the bearing is 90°. The signal is coming from a point at right angles to the route, and its distance from the route is equal to the distance from the first position, as shown in **fig. 5.** See navigation texts for variations in the technique.

This mobile pinpointing can be speeded up if the fixed-location DFing stations track the mobiles with respect to the interference. Angle differences can be measured more accurately.

identification

While locating and pinpointing are going on, steps that will permit identification of the signal can start. One element of this procedure is to show that harm to communications actually results from the interference. Another part is to show that the signal is coming from the pinpointed location. A final goal is to identify the transmitter causing the interference, or even the person responsible. This identification is not just for information — it may be crucial if action is to be taken, as discussed later.

The first element of identification is *complete* and *accurate* logging. This should show as a minimum

1. Date and time
2. Band and frequency
3. Station or communications being interfered with
4. Description of interference

In addition it's desirable to record any analysis of the interference, including bandwidth, center frequency, results of a-m and fm reception, and approximate or exact spectral distribution. Each log entry should be signed, and, if possible, witnessed.

The use of tape recordings for interference detection has been mentioned. These records are valuable in identification, especially if some simple precautions are taken. Each tape should start with an announcement of the fact that it is an interference

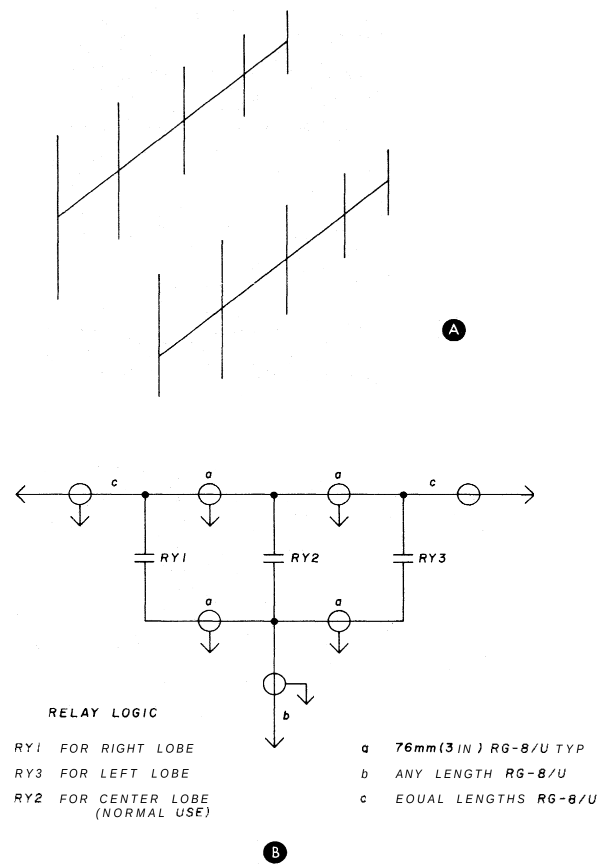


fig. 4. Lobe-switched system for DXing. Sketch A shows the antenna arrangement. Booms are parallel; spacing is typically one wavelength. The connection harness is shown in B.

record, and the date, time, operator, and place given. It is well to state the make and serial number of equipment used in reception and in recording. Before the recorder is shut off, the date and time should be repeated. Strip charts are also useful. These may give the pattern of interference, including times of occurrence and relative levels of wanted signal and interference.

Recordings should also be made by the mobiles used in pinpointing. Signal level at various locations is helpful in showing the extent of interference and the general location of the source. When the source location has been identified, simultaneous recordings at the locating mobile or mobiles and the tracking stations should be made. Each should be identified, of course.

In many interference cases pinpointing the location of a jammer, and showing by records that the interference ***does*** come from that location, is sufficient for action. However, cases may arise where further identification is needed.

Identification of a particular transmitter or person is not easy but possible. Two types of material are needed: the interference records and identified records of suspected sources. The identification rests on "signature" analysis and matching, that is,

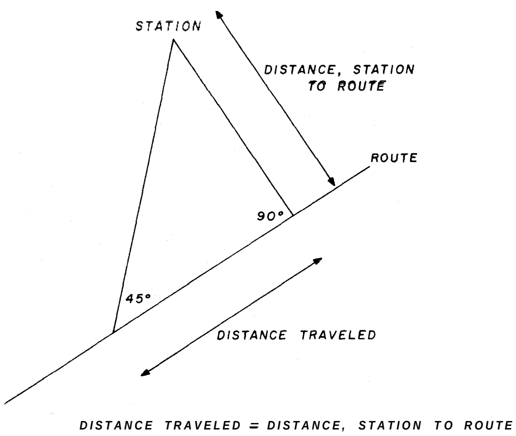


fig. 5. Method for pinpointing source location by using the navigational system of "doubling the angle on the bow."

an identifiable element of the signal that forms an identifiable characteristic. For example, the bias frequency in recorders varies. The frequency can be determined by comparing it to the 60-Hz hum frequency when present. Recorders, receivers, and transmitters may be identified by comparing the relative magnitude of hum components at 60 Hz, 120 Hz, etc. Sometimes special components are present, such as parasitics, peculiar background noise, or peculiarities of off-on characteristics.

Identification of individuals is also based on a form of signature analysis. For CW, this may be the ratio of dot to dash lengths, the way dash length changes from character-to-character, or individual tricks in spacing. For voice, the relative strength of components in the three major speech bands form identifier elements, as do word choice, syllable rates, and many other factors. It requires some detective work, but it can be done.

Signature analysis is certainly beyond the capability of most amateurs, and probably beyond the capability of most clubs. Those groups in areas with good laboratories can probably do the work or get it done: arrangements might be made to call on one of these groups if simpler steps are not sufficient. In any event, recordings should be made with the thought that analysis may be required as a part of the remedial action taken. Make at least ***some*** records with good-quality tape recorders; and, if there's reason to suspect an identified station or individual, make recordings of this station, giving identification.

action

In many cases interference of the type considered here disappears when the fact of an anti-jammer program becomes known. When this happens the plan should go into standby. It's a good idea to demonstrate that it works, say by a hidden transmitter hunt.

Unintentional interference will probably be corrected when the source is advised of its existence. If this type appears, a letter or telephone call discussing the problem appears to be in order.

Casual interference, say by kids, is usually not a great problem. If it persists, tracking down and notifying parents, the organization concerned, or even passing the information along the grapevine usually clears up the matter.

Interference caused by stolen rigs doesn't usually last long. If an attempt is to be made to recover the rig, however, fast action is needed. Several accounts of approaches to this problem have been published.

If interference continues and appears intentional or willful, there may be grounds for complaints to the FCC. Such factors as unidentified signals, or illegal forms of modulation, are direct grounds. However, interference per ***se*** is not necessarily adequate grounds. No amateur station has prior or exclusive rights to a frequency, and it's necessary to show that willful or malicious intent exists, which is difficult. In these continued interference problems, it's probably best to review the matter and the available records with the Commission field offices. But remember that these offices are understaffed and busy, so make sure you have a legitimate problem.

Some states have laws intended to apply to franchised communication services, which also apply to at least some amateur communications. Competent legal advice should be sought before any action is taken.

One thing should be watched. Individual action is not only unwise — it may even be dangerous. The news accounts of CBers' quarrels show this. Even group or club action should be carefully considered.

Fortunately, most interference problems can be handled without too much trouble. As noted, they tend to go away when there is a program to take care of them.

reference

1. Alf Wilson, W6NIF, "Happy Flyers — Helping People," Ham Radio Horizons, March, 1978, pages 12-22.

ham radio

high-performance rf-agc amplifier

Presenting a small step forward in front-end receiver design — an amplifier with high power gain and good agc characteristics

During construction of a 6-meter ssb transceiver, I needed an improved rf-agc amplifier. My requirements called for 15-dB minimum power gain, low noise figure, and good signal-handling capability that wouldn't deteriorate with the application of agc. Furthermore, a good input/output match to 50 ohms had to be maintained over the agc range. The well-known circuits I had used or seen in the past wouldn't meet all these requirements simultaneously. What follows is my approach to the problem of compatibility between a high-gain rf input amplifier and really good agc control. Also offered are some suggestions on further improving input/output linearity, and some equations for calculating power gain and impedance.

initial experiments

The dual-gate mosfet amplifier worked fine with constant bias, but signal-handling capability deteriorated with agc. Also, this circuit required a high-input impedance transformation for a match, a no-no in good front-end design. Next I considered the grounded-gate circuit, where an input match was easily obtained with a low-impedance transformation. Available power gain was disappointingly low, however, and input-impedance change with agc was totally unacceptable. From this initial investigation I reached two conclusions: A new amplifier configuration was necessary to meet gain, impedance match, and signal-handling requirements; and agc control

By Andy Borsa, WA1FRJ, 6 Colby Lane, Pelham, New Hampshire 03076

must not involve changing amplifier bias level in any way.[1]

final circuit design

To make a long story short, the circuit evolved into that shown in **fig. 1.** The basic arrangement is a cascode jfet amplifier. However the input fet, Q1, receives drive at both gate and source terminals. This provides two advantages, the first being an input match with a transformation only slightly higher than that of a simple common-gate stage. Second, this type input permits a choice of gain in the range between common source and common gate arrangements.

turn on, shunting more and more signal current away from Q2, decreasing stage gain accordingly. Maximum gain cut is determined by the saturated ON resistance of Q3 in series with the coupling-capacitor impedance, together with the g_m (forward transconductance) of Q2. With a good high-current switch at Q3 (such as a 2N2222), a very respectable agc range of at least 35-40 dB can be achieved.

There are other, less obvious, advantages with this agc circuit. First, control takes place at the point of smallest signal voltage, or lowest impedance. Second, the control element operates with increasing forward bias as gain decreases (input signal increases). Both points are extremely desirable from

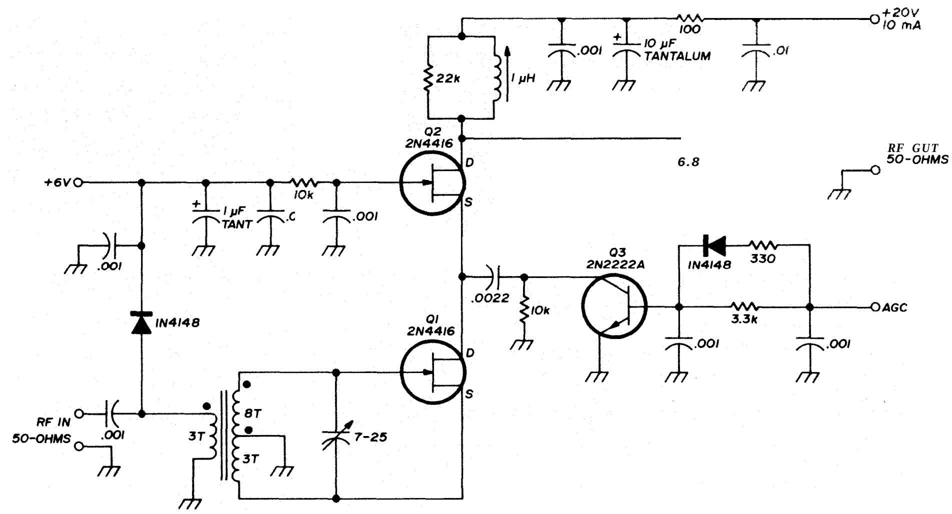


fig. 1. Schematic of the 50-54-MHz rf/agc amplifier. The input transformer is wound on a small toroid core; total secondary inductance is 0.5 µH. Coil data are sketchy because of the use of surplus cores and forms of unknown origin and characteristics.

In a related fashion, noise figure follows a similar relationship. The output fet, Q2, is conventional. Any reasonable output network may be used to meet gain and impedance requirements. The circuit isn't original with me, as a bipolar version is mentioned by Rheinfelder.[2] A similar arrangement has been described for use in vacuum-tube linear power amplifiers;[3] however, I've never seen its excellent capabilities put to use in modern ham equipment.

agc control

The method of agc control is unconventional (as far as I know) for an fet amplifier. It does bear some resemblance to a bipolar cascode circuit using a differential pair with current source, although operation does not involve changing amplifier bias level. In **fig. 1,** a transistor is ac-coupled to the interface between the fets. As long as this transistor remains biased off, it has little effect on the signal current passing from Q1 to Q2. With increasing agc voltage, Q3 begins to

a dynamic-range point of view. Finally, circuit complexity is only slightly greater than that of a standard cascode or mosfet stage.

circuit description

The circuit of **fig. 1** uses low-*Q* input and output networks to achieve an almost flat 16-dB power gain over the entire 6-meter band. A high-value shunt resistor across the output coil was sufficient to limit the gain to the desired value because of the low unloaded Q of the miniature coil I used. (A high-Q coil here would have resulted in a much lower shunt resistor to keep the gain down.) Input and output amplifier ports provide a very good match to 50 ohms, an important consideration for low-input vswr presented to the vhf antenna and cable and for the following multi-pole bandpass filter.

The agc transistor, a 2N2222A, receives its base drive through a nonlinear resistor-diode network. This arrangement linearizes the dB gain cut versus control voltage characteristic over the range I required. This or similar networks (using zeners, for example) can be used to tailor the agc levels and response curve to meet your specific requirements.

I found no need for parasitic suppression with this circuit; however, the layout is very compact with good bypassing and grounding, so other construction methods might require resistors or ferrite beads applied to Q2 gate and drain. A further point in this circuit is that Q2 was selected for a slightly higher I_{DSS} (zero-bias drain current) than Q1. This would not be necessary with sufficient source bias at Q1.

test results

Fig. 2 is a plot of the agc characteristics for the circuit. The linearity is acceptable but could be further improved by operating on the base drive to Q3. My test setup was limited to the 30-dB of gain cut shown, but at least 40 dB should be achievable.

Further tests of input-output linearity revealed a 1-dB gain compression level at +7 dBm output to 50 ohms, corresponding to a −9 dBm input level. Even at a gain cut of 30 dB, the input level for 1-dB compression didn't drop more than a couple of dB. Although this performance is good, further work is indicated, since this rf stage is still (on paper) the limiting strong-signal factor in my receiver.

Some possibilities for improvement: operate Q3 with some collector-to-emitter-current bias; use higher g_m fets; and maybe substitute PIN or rectifier diodes for Q3. Noise figure and IMD measurements are still awaiting completion of the receiver section, although I expect performance in this area to be good, at the very least.

calculating gain and impedance

When looking at published schematics, I've often wondered how other authors arrived at particular circuit values. So, at the risk of being blamed for too many technicalities, I'd like to present some equations and information for those wishing to experiment. For this, I'll use the simplified ac equivalent circuit of fig. 3:

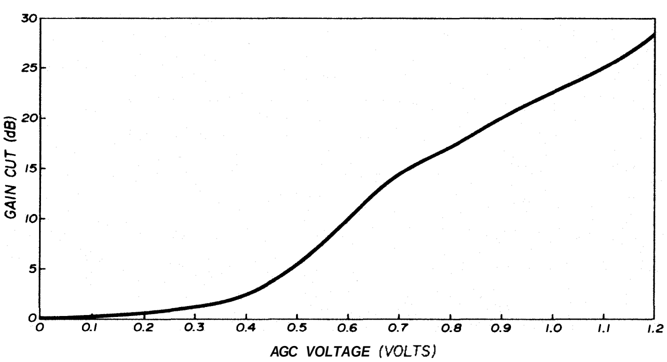


fig. 2. Agc characteristics for the circuit of fig. 1.

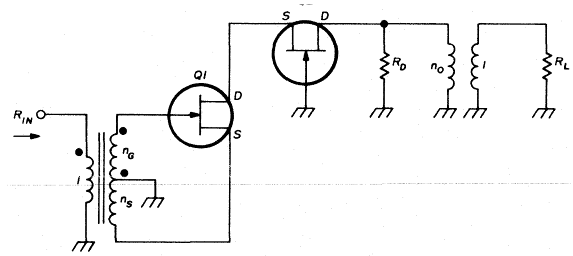


fig. 3. Simplified ac equivalent circuit for power gain and impedance calculations.

$$\text{power gain} = \frac{R_{in}R_L}{4}\,[g_m n_o(n_G n_s)]^2 \qquad (1)$$

where

n_o = output voltage step down ratio
n_G = gate winding step up ratio
n_s = source winding step up ratio
g_m = ac forward transconductance
R_D = combined resistive loading and coil losses
R_{in} = input impedance presented to input transformer primary
R_L = load resistance

Normally, $R_{in} = R_L = 50$ *ohms*. In **eq. 1** it's assumed that sufficient resistive and coil loading exists so that $R_D = (n_o)^2 R_L$, which makes the circuit output resistance equal to that of the load. Once R_{in} is set, the following equation should be used to define the input turns ratios:

$$n_s(n_G + n_s) = \frac{1}{g_m R_{in}} \qquad (2)$$

A few iterations between these equations may be required to achieve reasonable proportions for the input and output turns ratios. But reasonable numbers usually result by choosing $n_s = 1$ so that $n_G = \frac{1}{g_m R_{in}} - 1$. Then all that's left is to solve **eq. 1** for the output turns ratio corresponding to the desired power gain.

It is my hope that the data presented here will help produce some small advance in communications-receiver performance. Obviously, this amplifier circuit isn't limited to the 6-meter band; in fact, I use similar versions for my receiver's **9-MHz** i-f chain. I'd truly appreciate hearing from anyone who either uses or further evaluates this circuit.

references

1. Sabin, "The Solid-State Receiver," *QST*, July, **1970**, pages **35-43**.
2. Rheinfelder, *Design Of Low-Noise Transistor Input Circuits*, Hayden Book Co., **1964**, page **25**.
3. Pappenfus, Bruene, and Schoenike, *Single Sideband Principles and Circuits*, McGraw-Hill Book Co., **1964**, page **159**.

ham radio

modified quad antenna

If you haunt the 20-meter band for DX you've no doubt heard ZF1MA — here's a description of the antenna that puts out his big signal

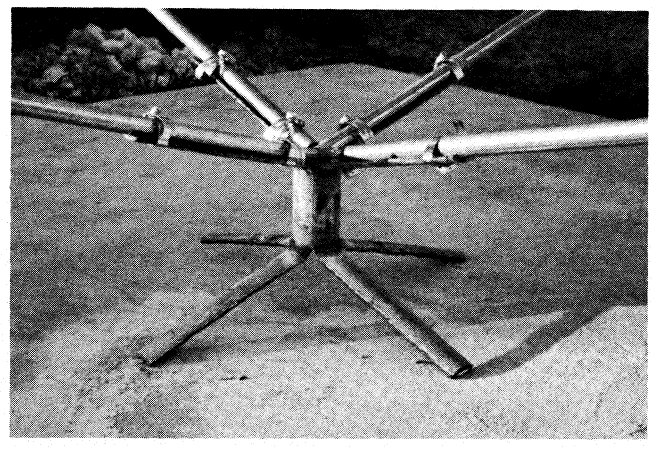
By Joseph B. W. Jackman, ZF1MA, P.O. Box 459, Grand Cayman Island, British West Indies

The quad has long been acclaimed a great DX antenna ever since Clarence C. Moore, W9LZX, made history with it in the mountains of Ecuador in 1939. Very few articles have appeared in the literature regarding its design changes. This article describes some novel changes to the quad, which makes it an efficient DX antenna.

construction

Fig. 1 illustrates the basic design of the quad. This antenna was made from locally available materials. Its design departs from conventional quads in several respects.

The spider. The first part to be constructed is the spider, which must be lightweight and strong. I made several spiders using different materials (aluminum and galvanized iron). These spider designs were abandoned as being either too heavy or too weak. Finally, a spider made from a 61-cm (24-inch) length of 51-mm (2-inch) diameter galvanized tubing had the necessary structural characteristics. This spider was cut as shown in **fig. 2.** The photo shows the spider in the half-finished stage.



Details of the spider used in the modified quad showing hose-clamp connections for the elements. All materials were obtained locally.

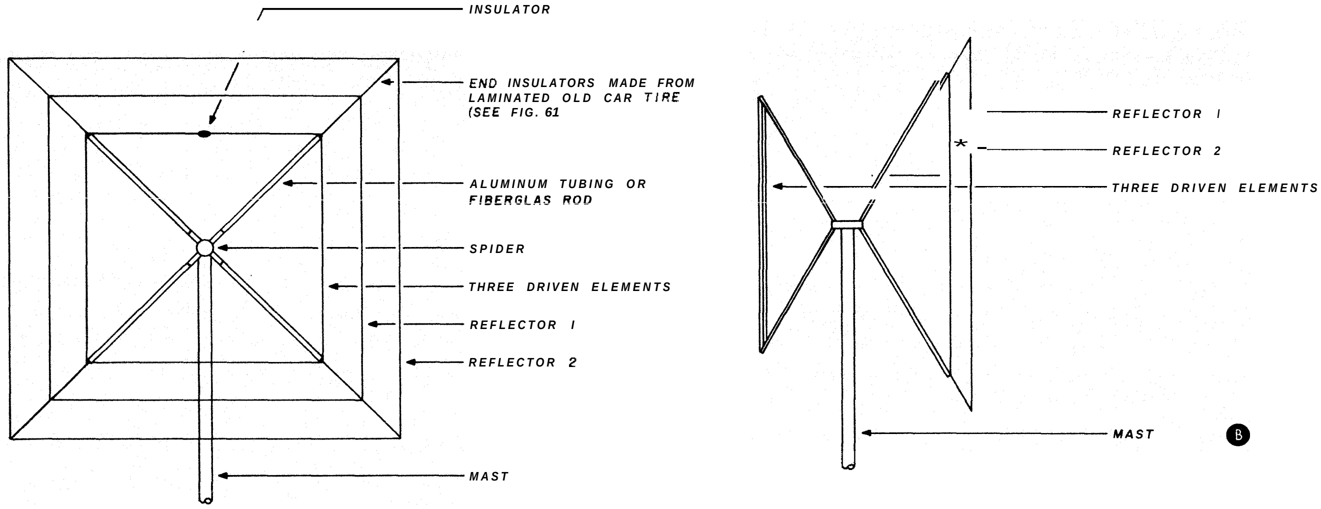


fig. 1. The modified quad antenna used at ZF1MA. Design uses three driven elements, which are fed with 450-ohm open-wire transmission line. Front view is shown in sketch A; side view in B.

Spider template. A template was used to bend the cut sections of tubing to the required angle consistent with the spacing used. The angle can be between 108 and 112 degrees. The template was made of stiff cardboard, which was cut as shown in **fig.** 3.

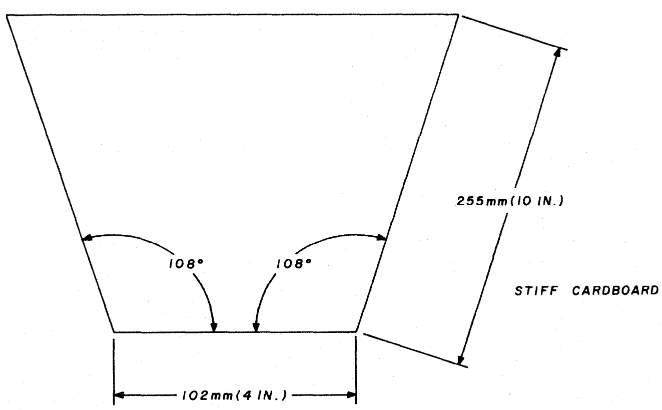


fig. 3. Template used for forming spreader-to-spider angles.

When tube bending has been completed, the spider should look as shown in the photo. The photo also shows how the spreaders are attached to the spider with stainless-steel 7.5-cm (3-inch) hose clamps.

To strengthen the spider, four pieces of 12.5-mm (1/2-inch) conduit was welded between each bent arm, so that the 108-degree angle, or whichever angle you choose, is maintained. See **fig. 4.** The drawing is simplified for clarity.

The spider can either be used as is, with only the addition of a mast-to-spider plate, or you can do as I did and weld a 0.9-meter (3-foot), 51-mm (2-inch) diameter pipe onto the spider, as shown in **fig. 5.** The mast is welded to the spider, depending on the shape quad desired.

Spreaders. Spreaders can be made from any of the conventional materials. In the tropics bamboo is the most available and is gratis; however, fiberglass, aluminum, or wood can also be used. This, of course, will depend upon where you live and what materials are most available. I had a few lengths of aluminum, so these were used. However, to prevent the spreaders from resonating, the horizontal spreaders on the driver and reflector elements were broken with insulators. For years I used wood, and at least one commercial company uses Bakelite. I decided to use rubber from old car tires cut into small strips and laminated together with contact cement. Incidentally, the same idea can be used to make standoff insulators or center insulators. Eight insulators were also

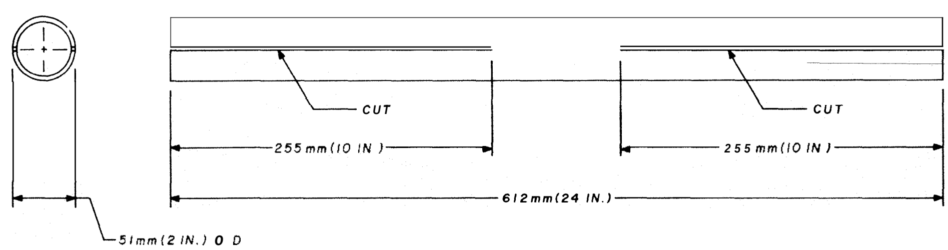


fig. 2. Method of cutting the galvanized tubing used for the spider.

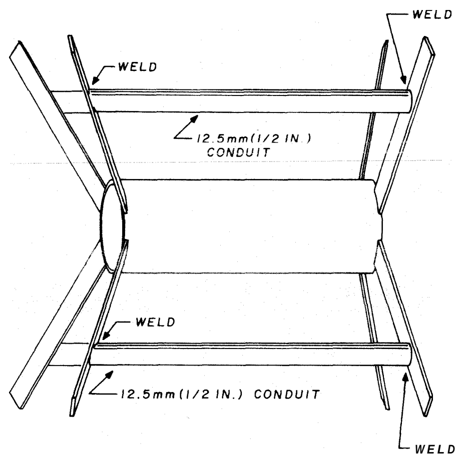


fig. 4. Galvanized conduit, 12.5 mm (1/2 inch) in diameter, is welded between spreader attachments for structural strength

made to put into the end of the aluminum tubing (fig. 6). This practice proved to be well worth the time spent in making them, as the flexible rubber moves with the wind and precludes broken elements on the quad. The three holes spaced 51 mm (2 inches) apart are for the driven-element spreaders only (fig. 6).

The per-side dimensions I used for the quad were:

three driven elements	5.3 meters (17.3 feet)
reflector 1	5.4 meters (17.8 feet)
reflector 2	5.6 meters (18.3 feet)

The three driven elements (fig. 7) were made up and passed through the holes in the end of the spreader insulator. The middle wire was open at the top and

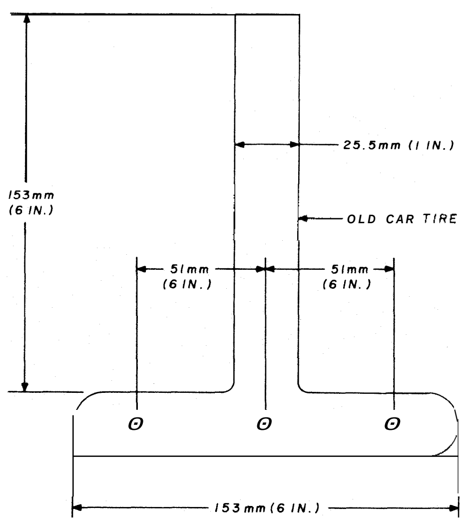


fig. 6. An old auto tire, which moves with the wind and eliminates broken elements, can be used for insulation material. Small strips of tire rubber were cut and laminated together with contact cement.

one of my car-tire insulators inserted. This was also done on the number-one reflector only. The feed-point was connected with 450-ohm open-wire transmission line to a universal transmatch, thence to a standing-wave-ratio meter, thence to the transmitter.

performance

The quad was mounted on a 12-meter-high (40-foot) mast and vswr checks were made. Having satisfied myself that all systems were go, I tuned up to 20 meters, pointed the antenna toward Europe and called CQ DX. I called only once, and all Europe seemed to call me! The first callsign I picked out was my longtime radio friend Paul, SM7CMC, who gave me a 5-9 + 20 report, and since it has been ever thus. That night I pointed the beam toward Australia and

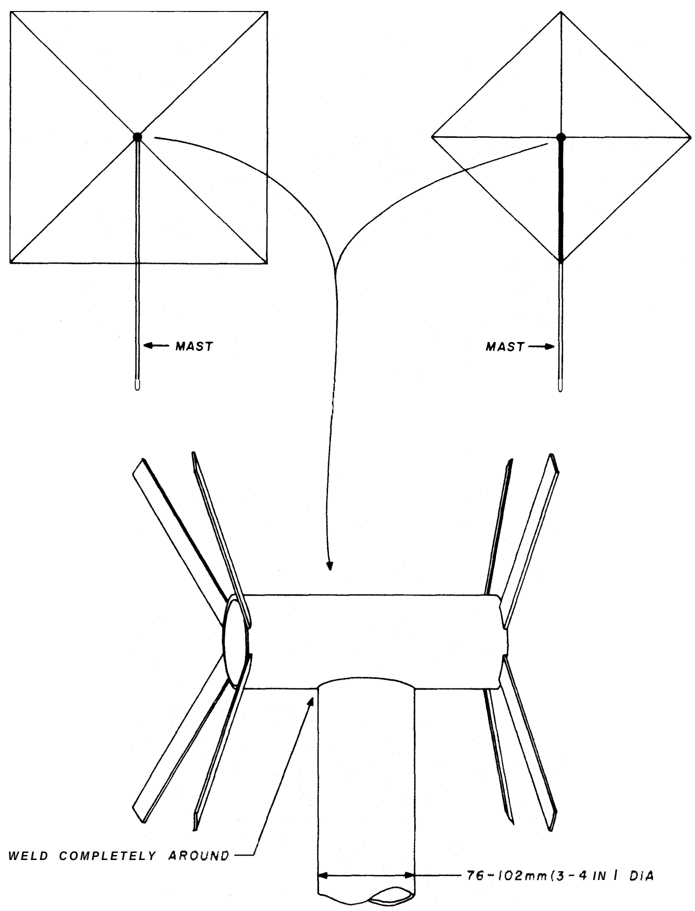


fig. 5. Method of welding spider-to-mast section to obtain a square or diamond-shaped quad.

the same thing happened. It seemed that all the Pacific stations were lining up to say what a great signal I had, including ZL and VK, but also P26, DU, F08, and others. At this time I ran tests with some of these stations. From their observations the following antenna performance figures were deducted:

	20 meters	15 meters	10 meters
Gain	8.5 - 9 dB	9.5 - 10 dB	10.5 - 12 dB
Front-to-side ratio	40-45dB	35 - 40 dB	40 dB
Front-to-back ratio	30 - 35 dB	25 - 30 dB	25 dB

The next morning at about 4:30 AM I crawled into bed, tired but contented, and next day I tried the beam first on 15 then on 10 meters. Needless to say, the results were even better than those received on 20 meters.

Since that time, whenever I get on the air on any of these bands, there's a pileup. So many amateurs have asked about my antenna system that I decided to write about it, and this is the story.

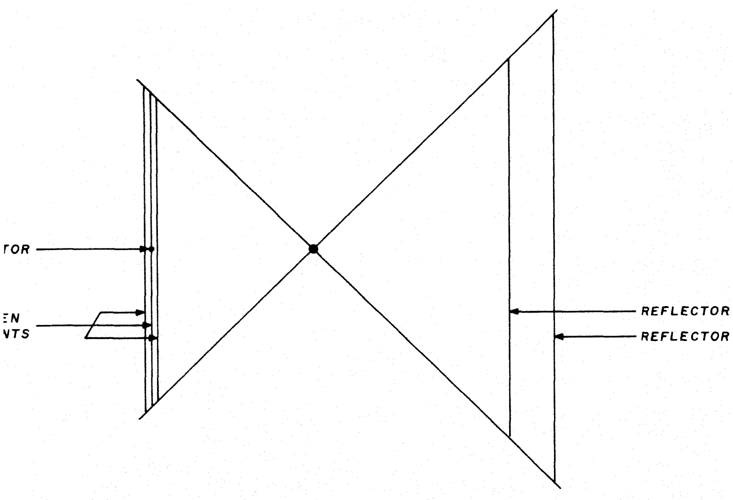


fig. 7. View looking down on the quad elements. Three driven elements are used; the center element is separated with an insulator to which a 450-ohm open-wire transmission line is connected. Reflector 1 is 3 per cent longer than the driven elements; reflector 2 is 6 per cent longer than the driven elements.

closing comments

I'd like to thank the many amateur radio operators around the world who have helped in testing this antenna. It's my hope that any amateur who decides to build this antenna will have as much pleasure with it as I've had.

bibliography

1. W. I. Orr, *All About Cubical Quad Antennas*, Radio Publications, Inc., Second Edition, 1976.
2. J. R. Fisk, W1HR, *Ham Notebook Volume 1*, Communications Technology, 1975.
3. *Radio Handbook*, 20th Edition, 1976, Editors and Engineers.

ham radio

fm demodulator using the phase-locked loop

Another application of the PLL for fm circuits, including design notes and theory of operation

The phase-locked loop has been around for some time and has many applications including frequency synthesis, clock synchronization, and a-m detection. IC technology has made the design of PLLs available at low cost and in small packages. This article provides some nonmathematical background on PLL theory and describes its application as an fm demodulator.

operation

A PLL is a loop circuit with feedback. It may be viewed as an electronic servo with the ability to lock onto and track a reference signal. The lock-and-track process is accomplished through phase-comparing circuits. However, because frequency is the rate of change of phase with respect to time ($\Delta\omega = d\theta/dt$), the PLL lock-and-track mechanism is meaningful in terms of frequency.

Basic PLL operation is as follows (**fig. 1**). The voltage-controlled oscillator (vco) output frequency depends on dc-voltage value from the phase detector and loop filter. As long as the phase difference between vco output and the reference signal is constant, the vco control voltage will be constant and its output frequency won't change. However, when the phase difference between vco output and reference signal begins to change (as would happen as the vco output frequency changes), the phase detector and loop filter provide a dc error voltage, which shifts the vco frequency to re-establish the original phase relationships. In this manner the vco output frequency is maintained at the value of the reference frequency. This is the lock condition of the PLL.

Even during lock a finite phase difference exists between output frequency and reference frequency because each PLL has a characteristic free-running frequency. That is, with no reference voltage applied to the PLL, some quiescent output frequency exists at which it rests. Hence, with a reference signal,

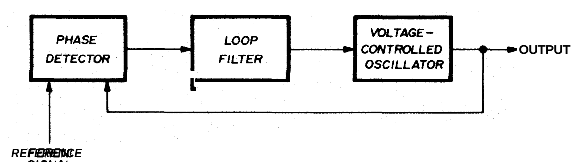


fig. 1. Block diagram illustrating PLL operation. A phase detector, filter, and voltage-controlled oscillator (vco) form an electronic servo in which vco output frequency and reference frequency are constant.

some phase difference between vco output and the reference signal must be present to provide an error voltage that will shift the vco output from is free-running value to the reference value. Often, terminals on the IC package are included for an external capacitor, which can be used to trim the free-running frequency

By Tim Daneliuk, KL7IPS, Room 0-408A, 820 North LaSalle Street, Chicago, Illinois 60610

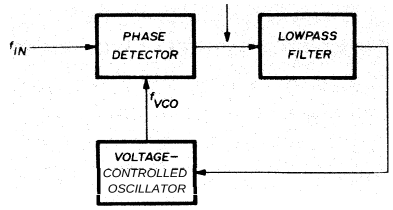


fig. 2. Example of a PLL with a phase detector used as a mixer so that $f_{in} \pm f_{vco}$ applied to a lowpass filter results in an output with a dc component that drives the vco so that loop lock is established.

as close as possible to the reference frequency. This action reduces pull-in time — i.e., the total time required for the PLL to establish the locked condition.

a different perspective

Fig. 2 shows the phase detector as a mixer whose output is $f_{in} \pm f_{vco}$. At lock, f_{vco} is such that the difference between the two frequencies is zero. The lowpass filter (LPF) attenuates the sum frequency and passes the difference frequency. Hence, at lock, the LPF output is a dc voltage (since the difference frequency is zero at lock). From this point of view it's seen that, if the PLL is not in lock but the difference frequency is within the LPF passband the filter output will not be dc but will be an ac beat frequency. This ac voltage is applied to the vco control terminal and, in effect, frequency modulates the vco output frequency. During the time f_{vco} is modulated so that it's closer to f_{in}, the error between the two signals decreases and the resulting beat frequency at the LPF output is lower. Thus the amplitude at the filter output is greater because, in a lowpass filter, less attenuation occurs for increasingly lower frequencies (fig. 3).

This entire process results in a filter output that is sinusoidal but decreases in frequency as a function

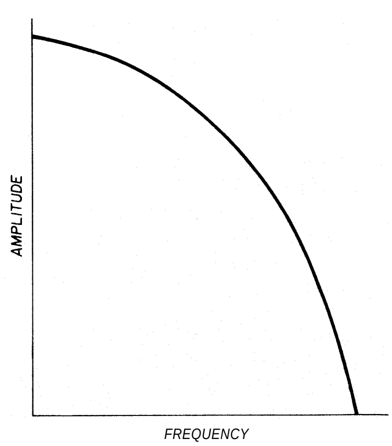


fig. 3. Amplitude as a function of frequency in a lowpass filter.

of time while increasing in amplitude (fig. 4). This waveform has a net dc component that drives the vco frequency to establish lock in the loop, which in turn results in a continuous dc output level from the filter. If the pull-in time is shorter than the period of the beat note appearing at the LPF output, the loop can lock without this oscillating error transient.

capture and lock range

It's important to understand the terms *capture* and *lock* and their mutual relationship. Consider the situation where the loop is not locked and the difference frequency is outside the LPF bandpass. The vco remains at its initial free-running frequency, and no real change occurs in the loop. However, as the difference frequency becomes smaller (as f_{in} approaches f_{vco}), a point is reached where an error voltage is passed bv the LPF to the vco control terminal, which in turn shifts f_{vco} so that the difference frequency is even lower. This is a *positive feedback* mechanism, which causes the loop to become locked.

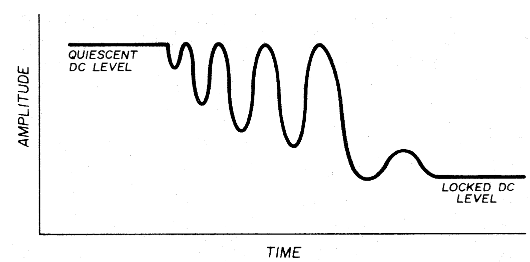


fig. 4. Lowpass-filter output response with time. The signal is sinusoidal but decreases in frequency while increasing in amplitude.

The capture range, then, is "the frequency range centered about the vco initial free-running frequency over which the loop can acquire lock with the input signal."[1] It is determined primarily by the LPF band edge and the system closed-loop gain. This action is contrasted with the loop lock range, which is "the frequency range usually centered about the vco initial free-running frequency over which the loop can track the input signal once lock has been achieved."[1]

The limiting factor of the lock range is the maximum amount of error voltage available and the corresponding maximum frequency swing the vco can produce. Because, in lock, the error voltage is a dc voltage, the lock range is not a function of the LPF bandpass characteristics. Note that the capture range is never greater than the lock range and is, in fact, often less. That is, a PLL may be able to track a signal and maintain lock over a greater frequency range than over which it was initially able to acquire lock.

table 1. Characteristics of several PLL ICs used for fm detection.

device	distortion (per cent)	upper frequency (MHz)
NE 560	0.3	30.0
NE 562	0.5	30.0
NE 565	0.2	0.5
NE 567	5.0	0.5

fm detection using the PLL

The basic circuit of a PLL fm detector is shown in **fig. 5.** Here, instead of a constant reference frequency, a frequency modulated signal is applied to the phase detector. The PLL is locked to the fm carrier frequency. With no modulation no frequency deviation occurs and the LPF output is a constant dc voltage. However, as soon as the wave is modulated and frequency deviation occurs, an error voltage appears at the LPF output which is proportional to the frequency deviation. The circuit demodulates the incoming fm signal. It's essential that the filter have a bandpass characteristic that doesn't attenuate the highest expected modulating frequency — *i.e.*, for a high-fidelity system, a bandpass all the way to 20 kHz would be desirable.

linearity

Another important consideration (particularly in high-fidelity systems) in the design of PLL fm detectors is linearity. The loop must detect the signal without introducing excessive distortion. Also, the PLL must operate at a frequency equal to the carrier frequency plus the *maximum positive frequency deviation* encountered. With modern devices this is usually not a great problem, since detection occurs after mixing and i-f amplification. For example, common i-fs are 455 kHz and 10.7 MHz. **Table 1** shows the distortion and upper frequency limits of several common PLLs used in fm detectors.

response in noise

One of the side benefits of this system of fm demodulation is its ability to detect signals in noise. Since the error voltage is filtered, the control voltage to the vco is "devoid of phase deviations produced by noise."[2] The main requirement is that the incoming signal be strong enough to activate the phase detector. This is not, perhaps, overly critical in the fm broadcast situation, where the receiver is in an area of fairly high signal strength (since fm broadcast is intended primarily for local coverage). However, incoming signal amplitude becomes exceedingly important, for example, in high-frequency RTTY FSK demodulation. Here, the incoming signal is often plagued by atmospheric noise and other disturbances. Unless the demodulated signal is an *exact* reproduction of the transmitted signal, the resulting printout will be gibberish. A similar application would be in the vhf fm telemetry systems.

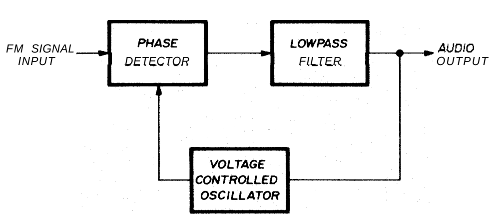


fig. 5. Fm detector using the PLL principle.

conclusion

Since the introduction of the initial IC PLL, technology has produced complete fm stereo demodulators available on one IC. **Fig. 6** shows one of these circuits. It's from National Semiconductor's *Linear Applications Handbook Vol. II*. This approach decreases parts count and improves reliability. It's safe to assume that this approach will be the industry trend in the future for such applications.

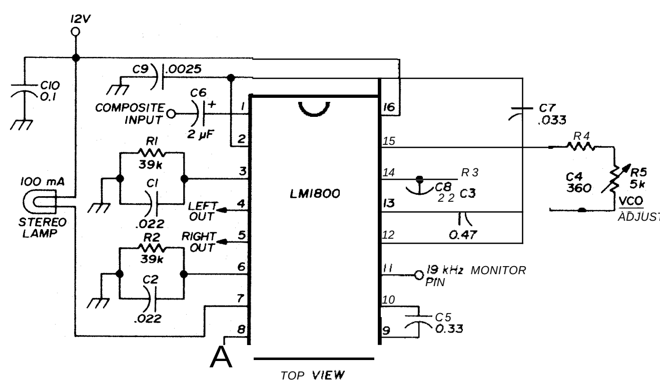


fig. 6. Typical fm detector using an IC chip from National Semiconductor. Additional data is available from their *Linear Applications Handbook*, Vol. II.

references

1. "Phase-Locked Loops," *Signetics* Application Notes, Signetics Corporation, Sunnyvale, California, page 6-5.

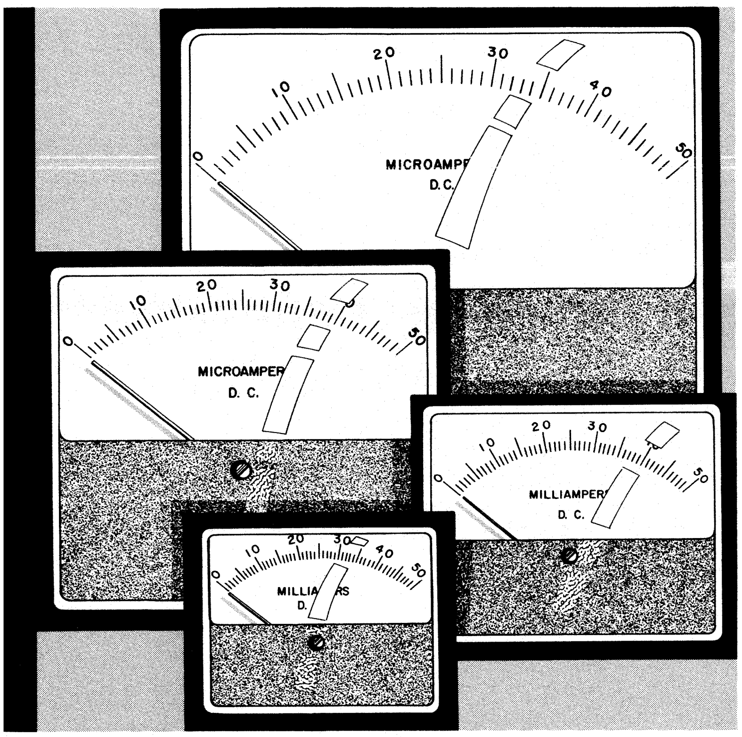
2. Nash and Garth, "Locking in on Phase-Locked Loops," Motorola Monitor, Vol. VIII, No. 2, page 23.

bibliography

"Phase-Locked Loop FM Stereo Demodulator," Application Note AN-81, Linear Applications, Vol. II, National Semiconductor, Santa Clara, California, 1976.

"Phase-Locked Loop Application Notes," Signetics Corporation, Sunnyvale, California, 1972.

ham radio



calibrating meter amplifiers

Meter amplifiers help provide increased sensitivity for high-current meters, but not without the need for a final accurate calibration

Meter-amplifier circuits have been around for several years, generally using conventional transistors, field-effect transistors, or operational amplifiers. If you want increased meter sensitivity in a wavemeter or grid dip meter configuration, then calibration may not be of much concern. In these cases you're more apt to want sensitivity (in the microampere range) to detect small radio-frequency currents.

When using meter amplifiers for absolute readings, however, several questions arise. How much amplification is expected, and in conjunction with what meter? What will be the final calibration of the new meter?

The data presented in tables **1,** 2, and 3 should provide an indication of what to expect from a meter amplifier, as it relates to current measurements.

The circuit used is shown in fig. **1.** It is a common-source configuration using an MPF102 field-effect transistor.[1] For testing purposes, I constructed the circuit in a mini-box, with six insulated binding-posts for the input, output, and 9-volt supply to the circuit. The amplifier was mounted on a small printed-circuit

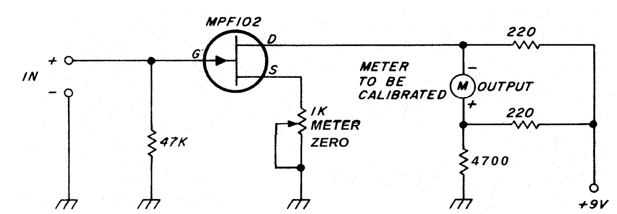


fig. 1. Schematic diagram of the simple meter amplifier. The pot connected to the fet's source is used to zero the meter to be calibrated.

board. The 1k pot was mounted in the center of the front panel. **Fig.** 2 shows a block diagram of the interconnections used to calibrate the meters.

As can be seen from table **1,** you should expect a seven- to eight-time increase in the readings of a

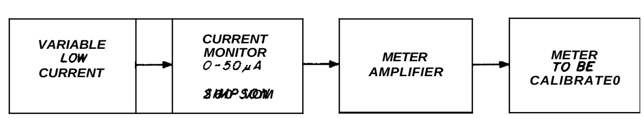


fig. 2. Test set-up for calibrating a meter movement. Using this system, the final calibration and accuracy will be determined by the current monitor.

By Howard J. Stark, W4OHT, 9231 Caribbean Boulevard, Miami, Florida 33189

table 1. Calibration for a COMCO 50μA meter.

Sirnpson 260	meter readings μA	reduced to unity μA
1 μA	7.0	7.0
2 μA	14.0	7.0
3 μA	24.0	8.0
4 μA	33.5	8.37
5 μA	44.0	8.8

50μA meter. The popular 1 mA meter is more likely to be used with an amplifier to raise its sensitivity. Not having a 1 mA meter available, I calibrated two identical 1.5 mA meters made by Marion Electric Company. **Table** 2 shows the results of the calibration.

table 2. Calibration readings for the Marion Electric 0-1.5mA meters.

meter 1

Simpson 260	meter reading	divisions on scale	μA per scale division
10 μA	48	8	1.25
20 μA	90	19	1.05
30 μA	128	28	1.07

meter 2

Sirnpson 260	meter reading	divisions on scale	μA per scale division
10 μA	43.0	8.5	1.17
20 μA	93.0	18.5	1.08
30 μA	135.5	27.0	1.11

The significant points to note are that 30 μA produced near full-scale deflection on the meters and that each division was slightly more than 1μA. As a final step, I checked Simpson 500 μA meter (see **table 3**).

table 3. Sirnpson 0 to 500 μA meters calibration.

Simpson 260	meter readings 0-500 μA	divisions on scale	μA per scale division
5 μA	127	17	.29
10 μA	325	35	.28
14 μA	500	50	.28

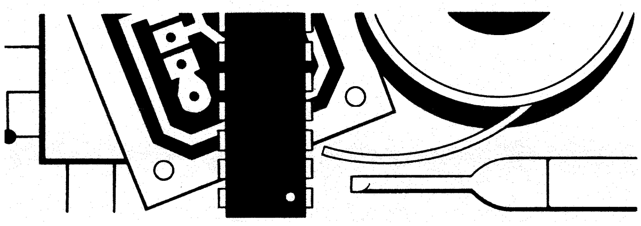
If the meter amplifier is to become a permanent part of the metering system, some relabeling of the face will be required to reflect the increased sensitivity. When using an amplifier with other meters, the basic resistance of the meter will effect the final sensitivity and calibration.

reference

1. Doug DeMaw, W1FB, editor, ARRL *Electronics Data Book,* American Radio Relay League, Newington, Connecticut, 1976, page *22.*

ham radio

the weekender



1-GHz prescaler for frequency counters

Amateurs **fall** into many categories — operators, experimenters, contesters, and so on. In addition, they may be full- or part-time mobile radio service or television technicians. These technicians, as well as the experimenters, often find themselves unable to measure frequencies in the low uhf region, especially between 500 and 1300 MHz.

While the 1-GHz prescaler described in this article will not quite reach 1300 MHz, it can be of considerable use in the range just above the limits of many of the inexpensive frequency prescalers available commercially. Although the Fairchild 11C05 used in this prescaler is guaranteed only to 1000 MHz, two that I have used have functioned well above 1200 MHz but couldn't make it to 1300.

In addition to possible application in the new 900-MHz mobile band, the prescaler may help you if you have to repair your TV set. The FCC's requirement for positive-detent tuning of the uhf TV channels has resulted in a proliferation of exotic front-end circuits. The use of digitally tuned oscillators and frequency synthesizers no longer permits us to check oscillator output by means of an electronic voltmeter. This will indicate only that the circuit is oscillating — but at what frequency? One of the solutions to this problem is this 1-GHz frequency prescaler, which can be used with virtually any counter and costs but a fraction of the price of a complete counter capable of operating at that frequency.

The prescaler will accept a signal between 25 and 1000 MHz, probably as high as 1200 MHz, and scale (divide) its frequency by either 10 or 100, depending on the position of a front-panel SCALE switch. Thus the frequency to be measured can be converted to one within the range of the counter which is connected to the prescaler. The accuracy of the measurement remains a function of the counter's accuracy multiplied by the scale factor.

If the prescaler is used with a 30-MHz counter, for example, frequencies between 30 and 300 MHz can be read on the counter by scaling by 10, which will result in read-outs between 3 and 30 MHz. Frequencies above 300 MHz can be scaled by 100, so that a 1-GHz signal can be read as 10 MHz on the counter. Obviously, it is necessary to multiply the counter reading by the scale factor to determine the input frequency, but multiplying by 10 or 100 should not tax anyone's mathematical prowess.

circuit operation

A block diagram of the prescaler is shown in **fig. 1**, and the overall schematic appears in **fig. 2**. U1 is an Amperex ATF417 broadband hybrid amplifier designed to operate from 40 to 860 MHz, and is usable from below 25 MHz to over 1200 MHz. Because the amplifier has a nominal input impedance of 75 ohms, a minimum-loss pad consisting of resistors R1 and R2 is incorporated to convert the prescaler input to 50 ohms. R2 also affords overload protection in conjunction with hot-carrier diodes CR1 and CR2. As shown in **fig. 3**, the input vswr (referenced to 50 ohms) is less than 1.5:1 below 1 GHz, and is less than 2:1 at 1.2 GHz. The sensitivity of the overall prescaler also appears in **fig. 3**.

The amplified signal is applied to U2, a Fairchild 11C05* divide-by-four circuit. The output of U2 feeds U3, a Fairchild 95H91 divide-by-five/six which is

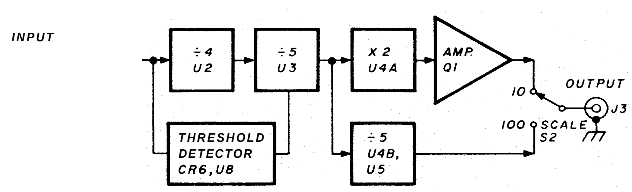


fig. 1. Block diagram of the 1-GHz prescaler.

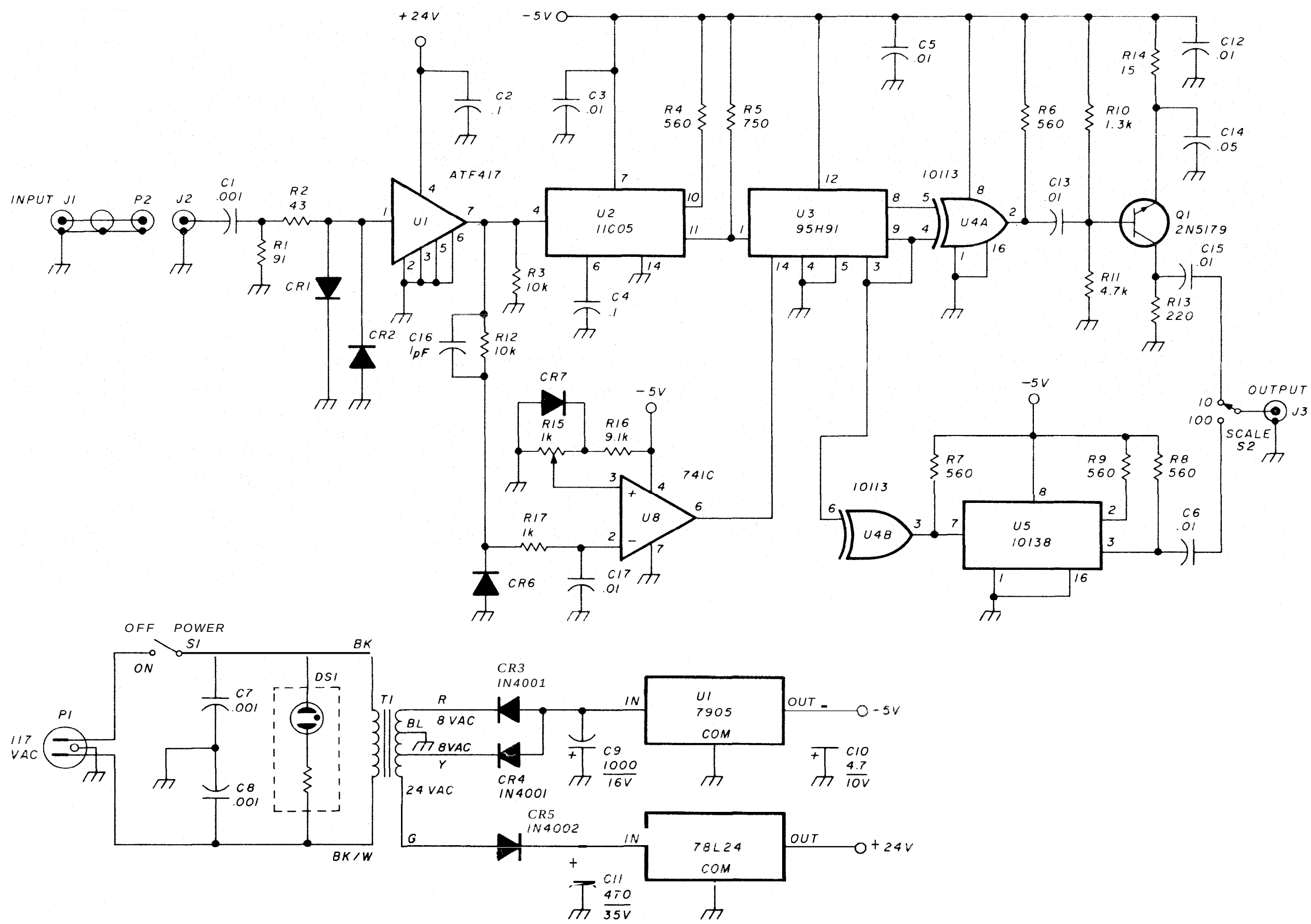
arranged to divide by five. Thus, the input frequency has been divided by twenty at the output of U3. To scale by 100, the frequency is again divided by five, this time in the quinary section of U5, a 10138 bi-quinary counter whose output is connected to the 100 position of SCALE switch S2. Gate U4B buffers the output of U3 from the loading effect of U5.

Since the input frequency has been divided by twenty in U2 and U3, it is necessary to multiply it by

By Robert S. Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022

*The Plessey SP8610 is a recently announced pin-for-pin replacement for the Fairchild 11C05.

two to obtain a scale factor of 10. This is accomplished in U4A, one section of a 10113 quad exclusive-OR gate. Frequency doubling is achieved by taking advantage of the finite rise and fall transition times of the complementary outputs of U3, as well as the slight difference in the propagation delays from input to each output. Because an exclusive-OR gate produces an output only when one, but not both, of



C1	0.001 μF, 50 V ceramic capacitor (Erie 8121-100-W5RO-102K, Centralab CY15C102M)	J2	Amphenol 27-800 or Sealectro 50-051-0000 miniature coax connector
C2, C4	0.1 μF, 50 V ceramic capacitor (Erie 8131-100-651-104M, Centralab CY20C104P)	J3	UG-1094/U BNC connector
C3	0.01 μF, 50 V ceramic capacitor (Erie 8121-050-651-103M, Centralab CY15C103P)	P2	UG-1465/U miniature coax connector (Amphenol 27-106, Sealectro 50-007-0000)
C5, C6, C12, C13, C15, C17	0.01 μF, 50 V disc ceramic capacitor	R15	Ik trimmer potentiometer (CTS X-201-R102B, Radio Shack 271-227)
C7, C8	0.001 μF, 1000 V disc ceramic capacitor	S1	SPST slide switch
C9	1000 μF, 16 V electrolytic capacitor	S2	SPDT slide switch
C10	4.7 μF, 10 V tantalum capacitor	T1	Triad F-91X transformer
C11	470 μF, 35 V electrolytic capacitor	U1	Amperex ATF417 amplifier
C14	0.05 μF, 50 V disc ceramic capacitor	U2	Fairchild 11C05 divide-by-4 (or Plessey SP8610)
C16	1 pF, 50 V disc or tubular ceramic capacitor	U3	Fairchild 95H91 divide-by-5/6
CR1, CR2, CR6, CR7	hot-carrier diode (Hewlett-Packard 5082-2835)	U4	10113 quad exclusive-OR gate
DS1	117 V pilot light assembly (Radio Shack 272-328)	U5	10138 bi-quinary counter
J1	Amphenol 86350 or Kings KC-19-54 BNC connector	U6	7905 negative 5-volt regulator, TO-220 case
		U7	78L24 or National LM340L-24 positive 24-volt regulator, TO-92 case
		U8	741C op amp, 8-pin mini-DIP case

fig. 2. Prescaler schematic diagram. All fixed resistors are 1/4 watt, 5% composition or carbon film.

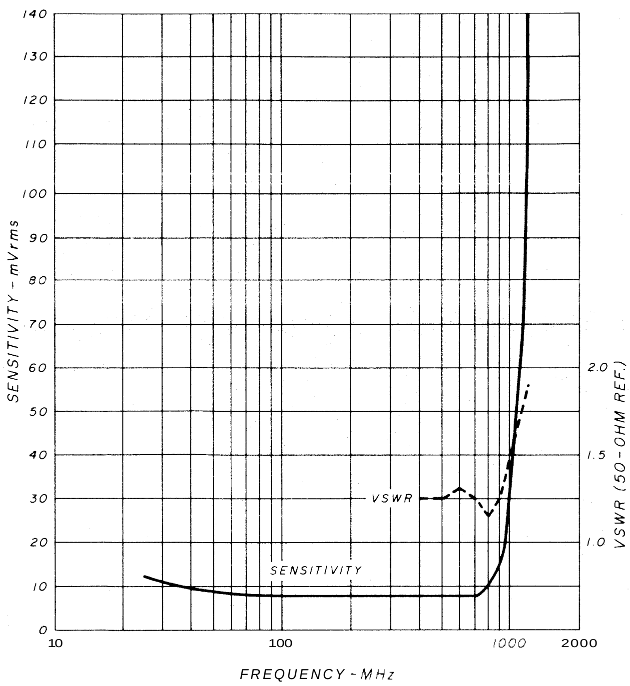


fig. 3. Prescaler sensitivity and vswr, plotted against frequency. Test equipment limitations precluded vswr measurements below 400 MHz.

its inputs is high, there is a very short period of time during the input transitions when *both* inputs are high or low, resulting in the output of the gate going low for this brief time. This is shown in **fig. 4,** as is the fact that two output pulses are generated for each cycle of the input frequency.

The extremely short time period during which the output pulses are developed prevents them from reaching full ECL-output level. Therefore, they are amplified by Q1 to a peak voltage level of at least 300 millivolts before being routed to the 10 position of switch S2. When S2 is set to the 10 position, the 25- to 1000-MHz input range is scaled to 2.5 to 100 MHz. When S2 is set to 100, the same input frequency range is scaled to 0.25 to 10 MHz.

As with all logic devices, U2 has a minimum input level below which it will not clock. However, at frequencies below 50 MHz, it may also divide by two or three just below this critical input voltage. To prevent this from appearing as a false reading on the counter, a threshold detector comprising CR6 and U8 is incorporated, which inhibits the output of U3 whenever the input voltage to the prescaler is below a predetermined level.

The output of amplifier U1 is sampled and rectified by diode CR6. The resultant dc is filtered and applied to the inverting input of U8, a 741C op amp configured as a dc comparator. The non-inverting input is connected to a voltage divider (R15 and R16) across the negative 5-volt supply; R15 establishes the reference voltage for the comparator. If the prescaler input level is insufficient for proper operation of U2, the voltage applied to the inverting input of U8 will be less than the reference voltage at the non-inverting input. This will cause the output of the op amp to go high, activating the master reset (pin 14) of U3 and inhibiting its output. As soon as the input signal reaches the required threshold, the op amp input conditions reverse, and the resultant low output enables U3.

construction

Except for power transformer T1, the primary power circuit, the INPUT and OUTPUT connectors, and the SCALE selection switch, the entire prescaler is built on a printed-circuit board measuring 8.9 x 7.9 cm (3-112 x 3-118 inches).* The board is two-sided; one side contains the printed-wiring traces, while the opposite side, on which the various components are mounted, provides a ground plane. I chose to mount the board vertically, parallel to the front of the cabinet, with the component side of the board facing the front. Two small right-angle brackets were used for mounting feet, as shown in the parts layout, **fig. 5.**

The first step in assembling the board is to mount connector J2. A complete bead of solder must be made to the ground plane around the entire periphery of the connector. Next, capacitor C1 must be soldered between the center pin of J2 and the adjacent trace, as shown in **fig. 6.** This arrangement is necessary to keep the lead inductance of C1 to an absolute minimum. The capacitor leads must be carefully bent and trimmed as shown. Then the shorter lead is soldered to the trace so that none of the wire is visible, and the longer lead is wrapped around the center pin of J2 and similarly covered with solder.

To keep the cost of the printed-circuit board

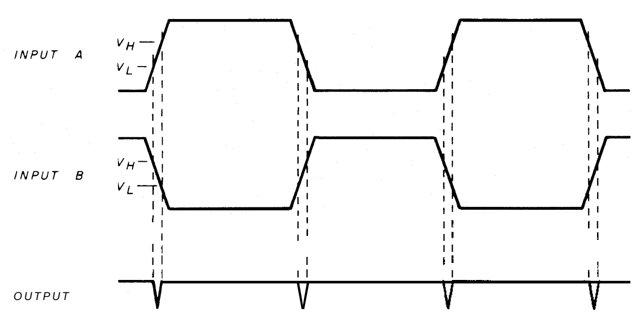
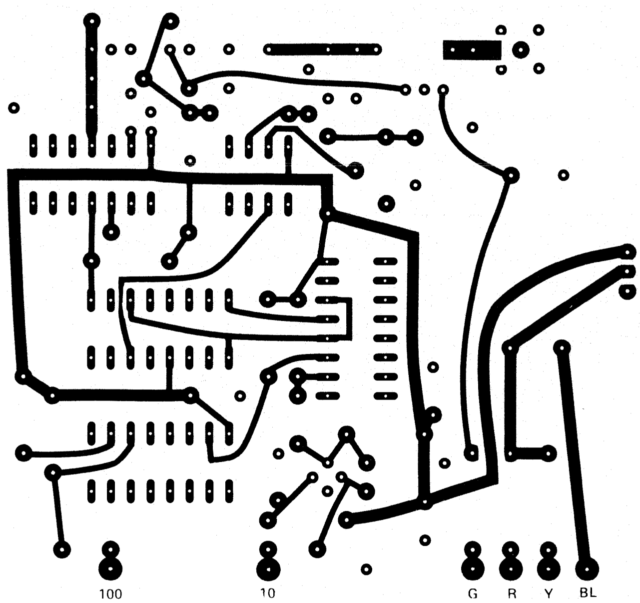


fig. 4. Waveforms showing operation of the exclusive-OR frequency doubler.

*If there is sufficient interest, the author will supply PC boards at nominal cost. Please enclose a SASE with all inquiries.

down, plated-through holes were not used. Instead, ground connectors are made by soldering the applicable component leads directly to the ground plane.

Before installing regulator U6 on the board, attach a TO-220 type heatsink to the tab of the regulator. Use a thin layer of silicone heat-transfer compound between the heatsink and the tab. The heatsink must not be grounded to the board, since the tab is internally connected to the input terminal of the device.



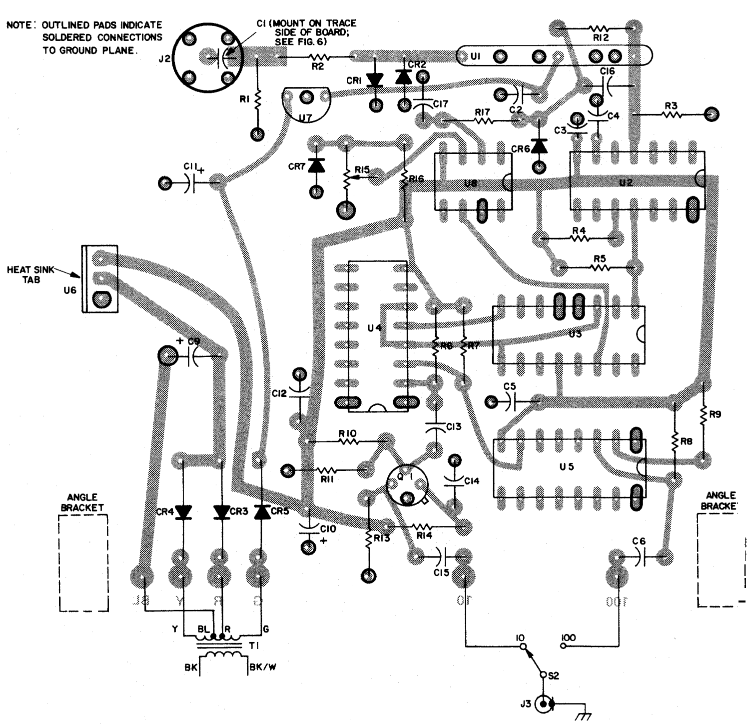


fig. 5. PC board component layout and wiring to transformer and output circuit.

Be sure that the polarities of the electrolytic capacitors and diodes are correct, and that none of the integrated circuits are reversed. After completing the board assembly, examine it carefully for solder bridges between traces or for any other foreign matter that might cause a short circuit.

Next, fabricate the input cable assembly which runs between the board and the front panel. Using a 7.6-cm (3-inch) length of RG-174/U coaxial cable, terminate one end with INPUT connector J1 and the other end with connector P2. Instructions for assembling the connectors to the cable appear in figs. 7 and **8.**

A recommended arrangement of the overall prescaler is shown in fig. **9.** OUTPUT connector J3, POWER switch S1, SCALE switch S2, pilot light DS1, and the input cable assembly (terminated by INPUT connector J1) are mounted on the front panel. Transformer T1 is mounted directly behind the printed-circuit board, and capacitors C7 and C8 are wired to a terminal strip located near the transformer.

Before mounting the circuit board in the cabinet, solder a short length of insulated wire to each of the output pads, designated 10 and 100, on the board. Then solder the secondary leads from transformer T1 to the pads marked G, R, Y, and BL on the board, matching the wire colors to the appropriate pad. Mount the board in the cabinet and connect the wires from the output pads to the corresponding terminals of SCALE switch S2. Finally, connect the input cable assembly between the front panel and J2 on the board.

operating considerations

It would appear obvious that after the prescaler has been wired and checked, all that remains is to connect it to the counter and plug it in. Maybe – and maybe not! Unfortunately no prescaler is all things to all counters. This one produces outputs of two types: the scaled-by-10 output is a train of very short, fast positive pulses having a peak-to-peak amplitude of 300 to 400 millivolts; the scaled-by-100 output is a square wave of approximately 800 millivolts peak-to-peak amplitude. Both are supplied from low-impedance sources.

Without going into a detailed explanation of transient response and reflections from improperly terminated transmission lines, it can be stated that what appears at the counter end of a coaxial cable connected to the prescaler output depends on the characteristic impedance of the cable, the length of the cable, the input impedance of the counter, and the frequency and waveform of the prescaler output signal. Since only the last of these factors is established (and varies over the prescaler's range), there may be considerable differences as to how the pre-

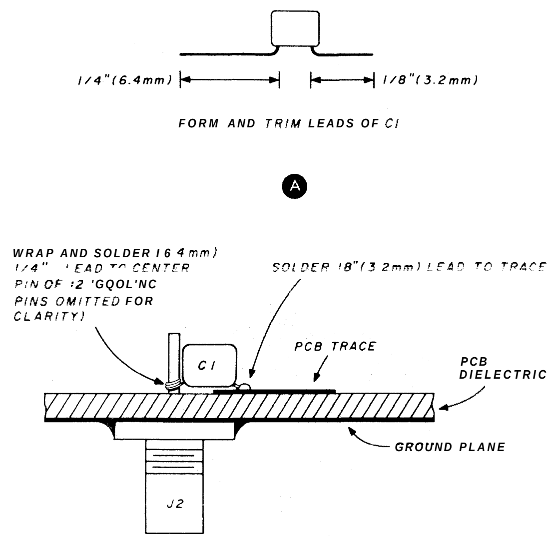


fig. 6. Method of installing capacitor C1 on PC board.

scaler will perform in conjunction with different counters and different cable lengths.

Now for the good news. Although you cannot tell how the prescaler will perform with your counter until you try it, there is a simple remedy if it does not appear to work properly. And that is merely to terminate the connecting cable in its characteristic impedance. Assuming that the connecting cable is 50-ohm coax (RG-58/U or equivalent), you need only to terminate it with a 50-ohm load at the ***counter*** end. If the counter already has a 50-ohm input impedance, the cable will automatically be properly terminated. However, most low- and medium-frequency counters have a high input impedance. Therefore an external termination may be required.

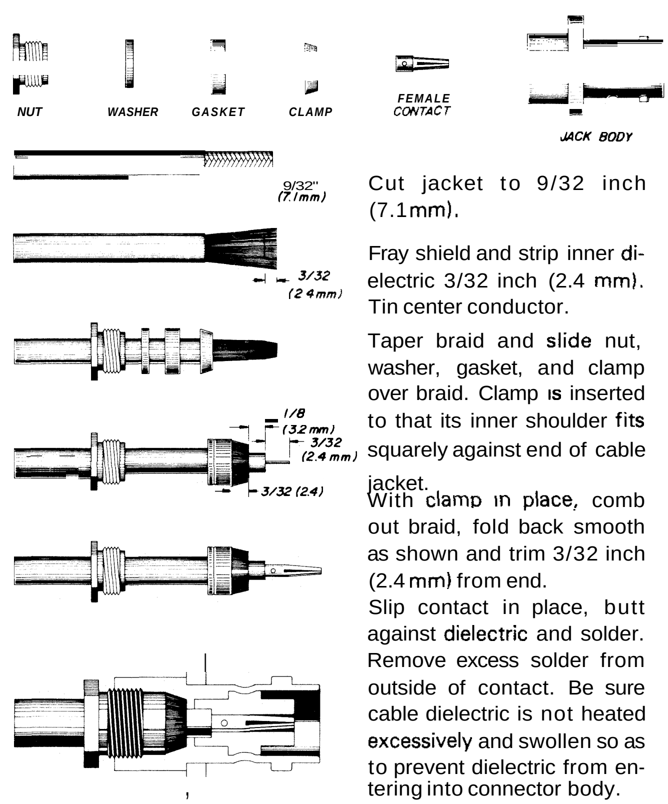
The simplest way of terminating the cable is to use a 50-ohm feedthrough termination, such as the Heath SU-511-50, Tektronix 011-0049-01, Hewlett-Packard 10100C, or Systron-Donner 454. An alternate method is to use a tee-connector at the counter input, with the connecting cable on one side of the tee and a 50-ohm termination on the other side. Or if you are ingenious, you can fabricate a cable with a 51-ohm composition resistor shunted across the counter end.

Another factor which enters into the picture is the characteristic behavior of the counter's input signal processor. All counters must convert the input signal to an appropriate logic level over the specified frequency range of the counter. Of the infinite variety of signals that can be applied to a counter, a symmetrical waveform is the easiest to process. Thus, your counter may exceed its specified maximum frequency when measuring the sine-wave output from a signal generator, but may not perform as well when counting an asymmetrical pulse train. Therefore, even if the counter's frequency range appears to be better than the published specifications, do not expect it to be so on the scaled-by-10 output from the prescaler.

As an example, on two different 80-MHz counters (one home-built, the other a Hewlett-Packard Model 5381A), the upper frequency limit when checked using a signal generator was well over 120 MHz for each. However, the maximum scaled-by-10 count from the prescaler which could be measured was between 80 and 90 MHz, corresponding to a prescaler input between 800 and 900 MHz. Of course, switching to the scaled-by-100 output permitted measurements to the limit of the prescaler. On the other hand, a counter ***specified*** to perform at 120 MHz should permit use of the scaled-by-10 output over the entire frequency range of the prescaler, as did a 500-MHz counter used to verify the prescaler output.

threshold-level adjustment

Before initially operating the prescaler, fully rotate the thumbwheel of potentiometer R15 toward the top of the board. This effectively disables the thresh-



Push assembly into body as far as it will go. Slide nut into body and screw in place with wrench until tight. For this operation, hold cable and shell rigid and rotate nut.

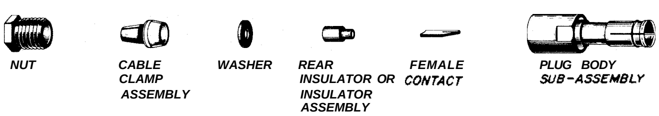
fig. 7. Assembling connector J1 to input cable (courtesy Amphenol).

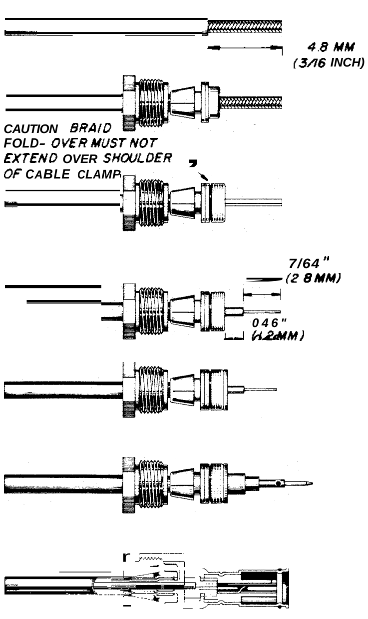
old detector and will permit a check of the prescaler in conjunction with the counter. Apply signals of known frequencies to the prescaler INPUT connector (noting the approximate minimum levels required, as shown in **fig. 3**) and check the counter readings in both the 10 and *100* positions of the SCALE switch. **Caution: Although U1 will withstand** a 10-volt peak-to-peak input without damage, it is recommended that inputs to the prescaler be limited to 1 volt peak-to-peak (approximately 350 millivolts rms, or 2.5 milliwatts across 50 ohms) to prevent clipping by diodes CR1 and CR2.

To adjust the threshold level, apply a 25- to 30-MHz input to the prescaler using a signal generator or any other signal source whose output level can be controlled. Feed enough signal to the prescaler to obtain a properly scaled reading on the counter, then slowly reduce the signal level until that reading is lost; as the signal level is reduced, readings of two, three, or four times the correct scaled reading will appear. Bring the input-signal level back up to the point where the last false reading is obtained, just before the correct reading appears, and adjust potentiometer R15 carefully to make the counter read zero. Then increase the signal level slightly to be certain that a true reading can be obtained. This is a critical adjustment, in that it establishes the sensitivity of the prescaler from 25 to 900 MHz. If the potentiometer is set past the position which just eliminates the false counts, the prescaler sensitivity will be reduced.

If you have no need to use the prescaler below 50 MHz or so, the overall sensitivity of the prescaler can be improved below 900 MHz by simply keeping the arm of R15 rotated fully toward the top of the board, thereby disabling the threshold detector. For those who desire maximum sensitivity above 900 MHz, a slight improvement will result from omitting C16 and R12 (in which case all other parts associated with U8 can also be eliminated).

ham radio





Cut end of cable even. Trim jacket to 3/16 inch (4.8 mm). Be careful not to nick braid.

Slide nut over jacket. Place cable clamp assembly over braid and push back against cable jacket as shown.

Comb out braid and fold evenly over cable clamp assembly and trim braid wires as shown, avoiding bunching.

Cut cable dielectric to .046 inch (1.2 mm). Cut must be sharp and square. Do not nick center conductor. Dimension of bared center conductor should be as shown; trim if necessary. Tin center conductor and remove excess solder.

Place washer over core.

Slide rear insulator or insulator washer over center conductor.

Make sure rear insulator or insulator washer butts flush against washer and cable core. Solder contact to center conductor — contact must butt against rear insulator as shown. Do not get any solder on outside surfaces of contact.

Insert prepared cable termination into connector body. Secure by wrench-tightening body while holding nut stationary.

fig. 8. Assembling connector P2 to input cable (courtesy **Amphenol**).

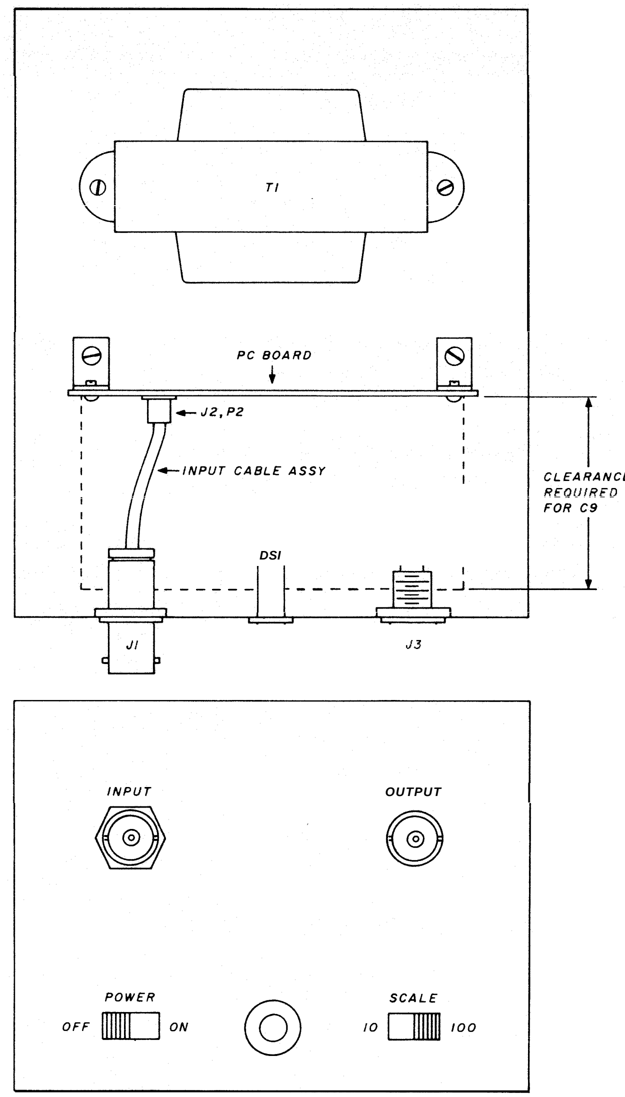


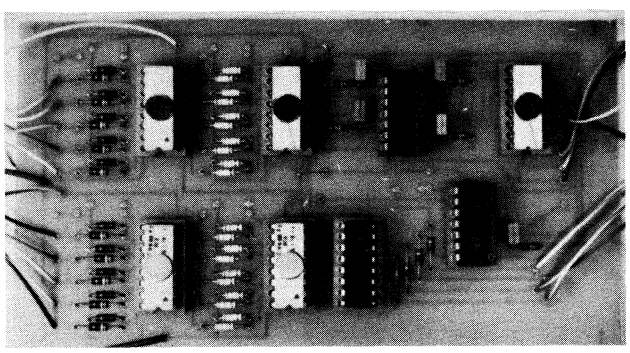
fig. 9. Suggested layout for complete prescaler. Wiring to and from the PC board is shown in fig. 5.

digital keyboard entry system

By sampling the signals from an inexpensive pocket calculator, you can build a complete data entry system

While working on a recent synthesizer project, it became apparent that the conventional methods used to enter the frequency control information, such as thumbwheel or rotary switches, are generally cumbersome. Keyboard entry has become popular with the widespread use of Touch-Tone* phones and hand-held calculators — so, it seemed to be a natural choice for a new design. This same choice has been made by a few manufacturers of digitally controlled signal generators and vhf/uhf scanning receivers. The advantage of a keyboard entry device is that, while providing a means of frequency control (as in our case), other digital information can be entered. If plans for the system change, new functions can be added by assigning a code to that particular function and inputting the code via the keyboard. In our keyboard-synthesizer system, we have incorporated the ability to step the frequency up or down by a fixed increment at the touch of a switch.

*Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.



The complete data entry system, with the exception of the 74175 latches.

After giving the keyboard-system problem some thought, the solution became a standard four-function calculator with a programmable automatic constant capability. At this point it should be stressed that our application was a frequency synthesizer, but any system that requires digital information in binary coded decimal (BCD) format could use the ideas presented in this article.

theory of operation

There are several problems when trying to build a keyboard encoder. First, the information from the keyboard is presented in 1-out-of-10 format; that is, you must encode the operation of 1-of-10 switches. Second, the keys will be presented in some sequence, and you must store the information as each new key is pressed. Finally, there is the problem of key bounce. As you actuate or de-actuate a key, the contacts of the switch bounce, giving a fairly noisy signal as illustrated in **fig. 1.**

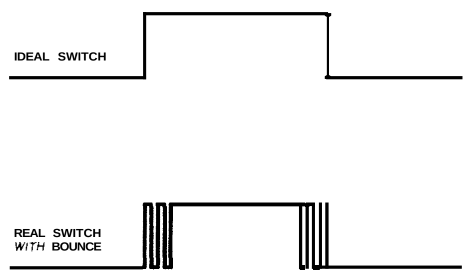


fig. 1. Waveforms generated by the ideal switch and a real switch which exhibits bounce.

A four-function calculator, of course, uses a key-

By Bruce **McNair, N2YK,** and Glenn Williman, **N2GW.** Mr. McNair's address is 12 Marion Avenue, Howell, New Jersey 07731; Mr. Williman's address is 145 Whalepond Road, Oakhurst, New Jersey 07755.

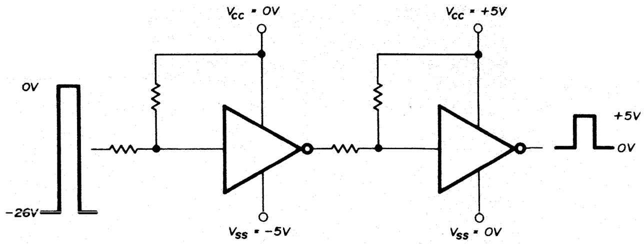


fig. 2. Level translators used to transform the signal voltages from the display to the standard TTL voltage levels.

board for information entry and has all the necessary circuitry to deccde, remember the inputs, and debounce the keys. An added benefit of the calculator is the fact that it also displays the entered information. You can take advantage of this fact by using the calculator display as a display for the synthesizer or system you are controlling.

However, there are certain problems associated with the use of a calculator. The information that you need is only present at the calculator display terminals. The information is not BCD encoded, but has been encoded to drive seven-segment displays. Further, to save pins on the calculator chip, the information is multiplexed. Seven lines carry the segment information to all the digits while (in the case of the calculator we used) eight lines enable the correct digit at the proper time.

To use the data from the calculator there are three steps to go through:

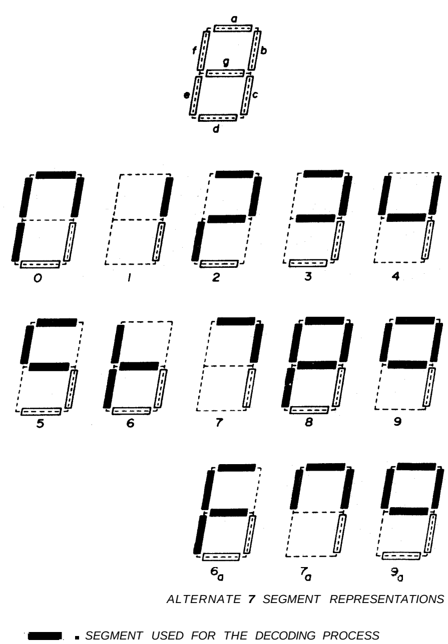


fig. 3. Segments used for the input to the 8223 PROM to convert between the 7-segment display information and the BCD code. As can be seen, segments c and d are not needed for code conversion. Each character is unique, regardless of whether c or d are present.

1. Shift the voltage levels of the display signals to that of the logic devices used in the decoder, and buffer these signals to prevent the calculator from being loaded down.

2. Decode the seven-segment information to BCD.

3. Demultiplex the BCD data and store the data.

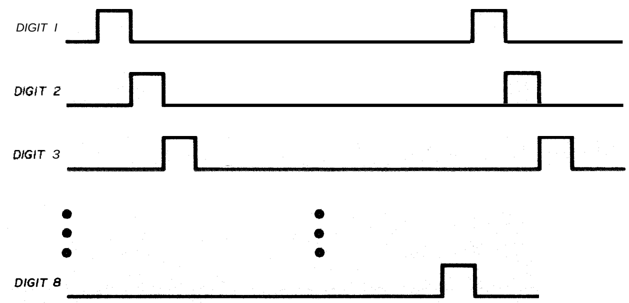


fig. 4. Timing diagram for the digit enable lines.

Each of these operations is accomplished as described in the following sections.

level shifting and buffering

The calculator we chose to use was a TI 1265. This is a simple, four-function calculator with automatic constant, non-blanking display, and low cost (in the $10 price range). Automatic constant is a convenient

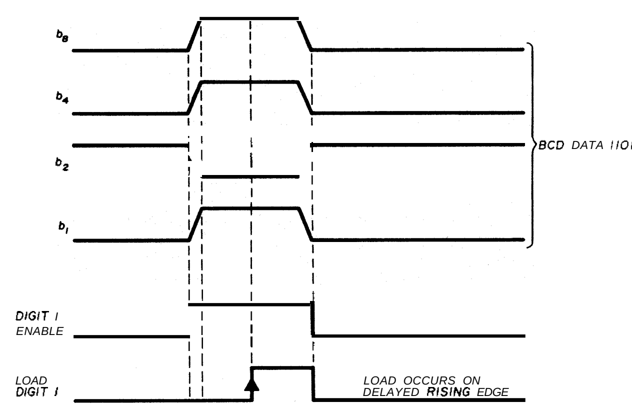


fig. 5. The latches will load the information on the leading edge of the respective digit's load pulse.

feature because by continually adding the fixed constant to the display, you can increment the synthesizer frequency by a constant amount. In 2-meter operation, for example, it would be best to store 0.015 as the fixed constant. Each time you press + the frequency goes up by 15 kHz, the channel spacing on 2 meters. To step down by a fixed amount, make the constant positive and the display negative (or vice versa), each time you press + you will be subtracting a fixed amount.

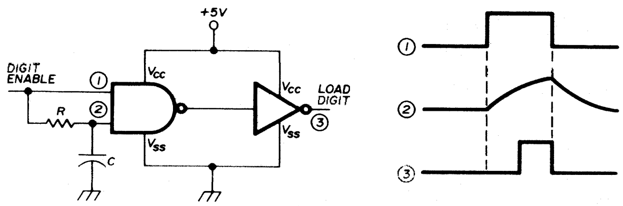


fig. 6. The time-constant formed by the R and C values provide a delay before the gate is enabled. This allows time for the data to settle.

A non-blanking display is an important feature to look for. Some of the calculators on the market will shut off the display, or even the entire calculator, after a period of inactivity to conserve battery power. Since the display conveys all the important information, this is unacceptable. If you already have a calculator with this feature, it may be possible to "fool" the calculator into thinking that there is always activity. Find a key that when pressed does not alter the memory of the display. If this key is regularly "pressed" by an added circuit the calculator will be under the impression that the operator is still there.

Finally, some calculators display the minimum number of digits to accurately represent the stored information. For example, 25. may represent the sum of *21.58* and *3.42*. To be usable for this system, the calculator must retain the same number of decimal places at all times. That is, $21.58 + 3.42 = 25.00$; the two zeros after the decimal place remain.

table 1. Truth table for the outputs of the 8223 PROM.

	prom address 7-segment code					prom outputs BCD output			
number	**a**	**b**	**e**	**f**	**g**	b_8	b_4	b_2	b_1
0	1	1	1	1	0	0	0	0	0
1	0	1	0	0	0	0	0	0	1
2	1	1	1	0	1	0	0	1	0
3	1	1	0	0	1	0	0	1	1
4	0	1	0	1	1	0	1	0	0
5	1	0	0	1	1	0	1	0	1
6	0	0	1	1	1	0	1	1	0
7	1	1	0	0	0	0	1	1	1
8	1	1	1	1	1	1	0	0	0
9	1	1	0	1	1	1	0	0	1
6_a	1	0	1	1	1	0	1	1	0
7_a	1	1	0	1	0	0	1	1	1
9_a	1	1	0	1	1	1	0	0	1

Now, consider the type of display the calculator uses; two of the more popular are LED and fluorescent. Most of the LED displays consume a great deal of battery power and blank the display after some period of time. The TI 1265 has a fluorescent display and is not blanked. Other TI calculators that are worth considering are the TI 1200 (very similar to the TI 1265) and the TI 1000 (a low cost calculator that uses an LED display but meets most of the other requirements). Since we did use a calculator with a fluorescent display, this is the type of interface we will describe, Interface with an LED-type display would actually be easier.

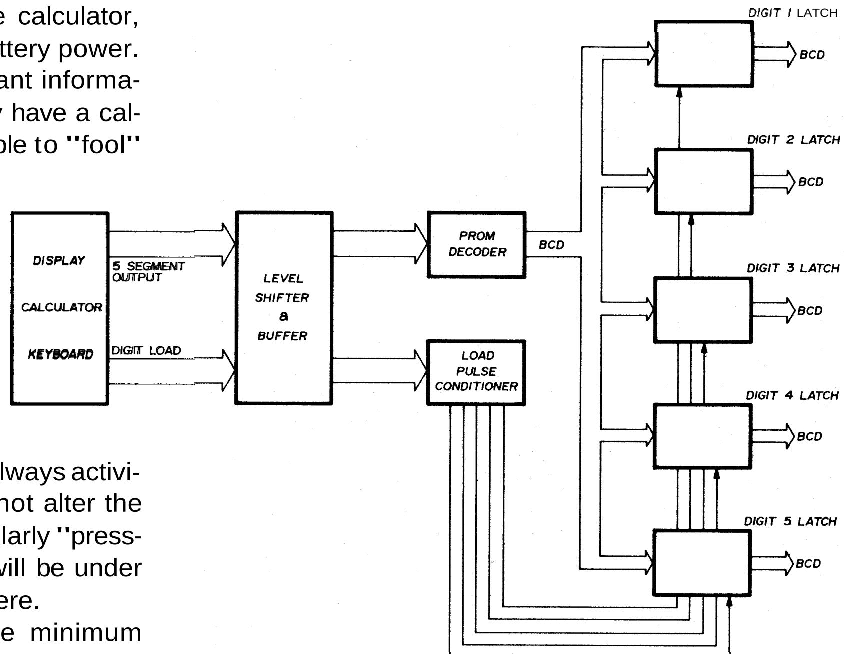
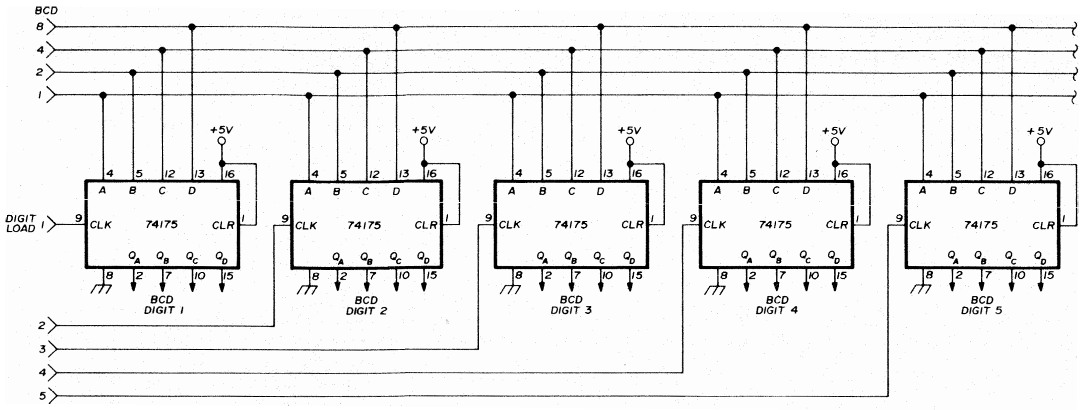


fig. 8. Block diagram of the complete system for using a calculator keyboard as a data entry system.

The display voltage levels in the TI 1265 are 0 and – 26 volts. This nonstandard signal can be translated to CMOS/TTL levels of 0 and +5 volts with the circuit shown in **fig. 2.** CMOS 4049 inverters are used to minimize loading on the calculator outputs. The same circuit is used in each data line for seven-segment as well as digit information. All resulting signals will drive CMOS or a single TTL load.

segment decoding

The information generated by the calculator chip is coded as a seven-segment signal to be directly used by the calculator display. To make use of this data, it is necessary to translate it to BCD. At first, it would appear that a complicated circuit with seven inputs and four outputs would be necessary. As it turns out, this will take four chips. Another possibility would be to use a special IC to perform the decoding (a 74C915). Unfortunately, these chips are not readily



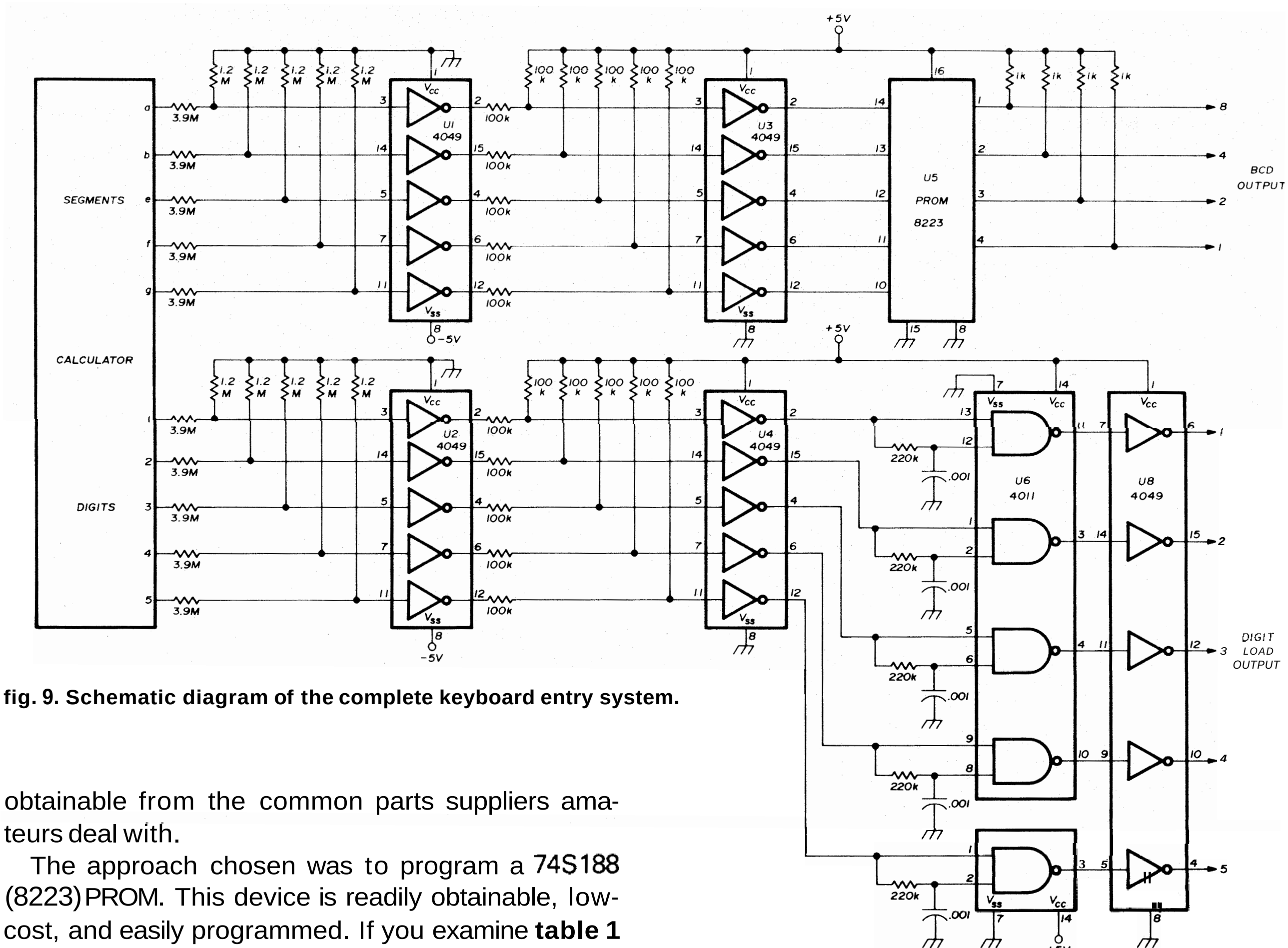


fig. 9. Schematic diagram of the complete keyboard entry system.

obtainable from the common parts suppliers amateurs deal with.

The approach chosen was to program a 74S188 (8223) PROM. This device is readily obtainable, low-cost, and easily programmed. If you examine **table 1** you'll see the code conversion table that must be burned into the PROM. It is also interesting to note that only five of the segments are needed to decode the information displayed. Segments *c* and *d* convey no new information. This is shown in **fig.** 3.

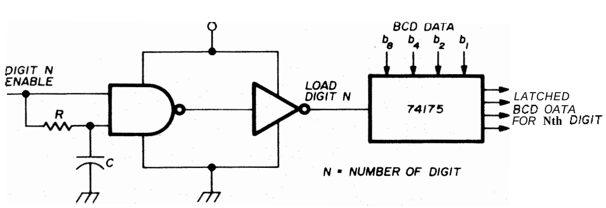


fig. 7. Overall diagram of the circuit necessary to demultiplex the BCD data for each digit.

demultiplexing

To minimize the number of outputs on the calculator chip, most manufacturers multiplex the display information. The segment lines go to the same seg-

digits in turn. For example, to light segment g of digit 1 (a minus sign in the first digit) this combination of digit and segment lines would be simultaneously ena-

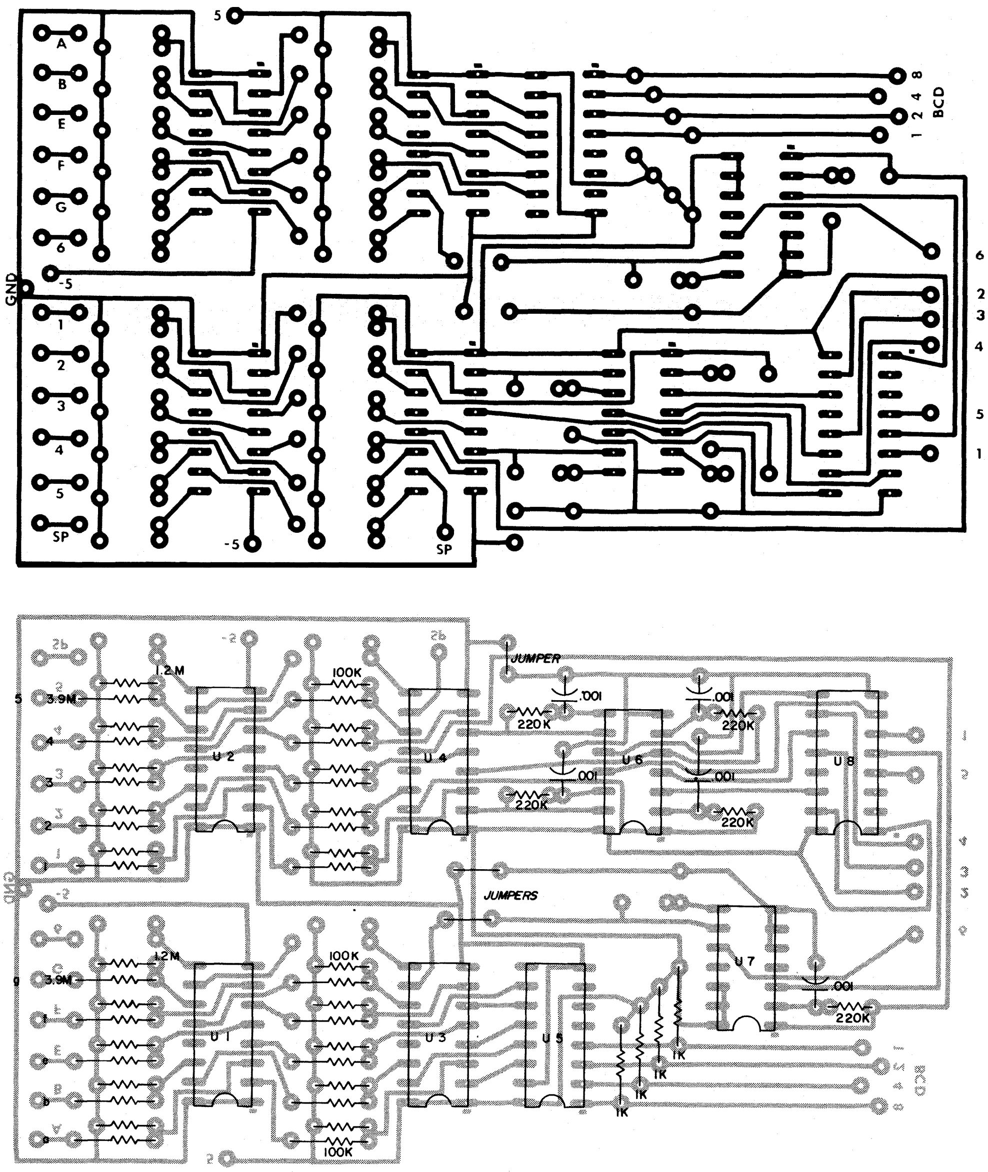


fig. 10. The foil pattern for the circuit board is shown in A. This does not include the 74175 latches. A parts placement diagram is also shown in B.

bled. A representative timing diagram of the digit enable lines is shown in **fig. 4.**

Demultiplexing the segment data can be done by generating a signal while the desired digit is turned on and using this signal to enable a latch to store the data. For this circuit 74175s are used. These ICs are four D-type flip-flops in one package, or one for each BCD bit. The timing diagram for this circuit is shown in **fig. 5.** The latch load signal is generated by the circuit shown in **fig. 6.** The RC time-constant delays the load pulse, so that data will be valid for any calculator that might be used. The overall demultiplexing circuit is shown in **fig. 7.**

If you are going to be transmitting the data from one board to another in your system (as we were), you should do the actual dernultiplexing at the data's destination. What we did was to locate the 74174s right next to the synthesizer's divide-by-N counters. In this way, interboard wiring was limited to 4 BCD bits plus one line for each digit used. Tying the whole system together results in the block diagram shown in **fig. 8,** with its schematic shown in **fig. 9.**

construction

As shown in the photographs, our original circuit was patched together on a number of breadboards. For our final system, and this article, a PC board was designed and is shown in **fig. 10A,** with the parts placement shown in **fig. 10B.** This board is designed for five segment lines, six digit lines, and a spare line.

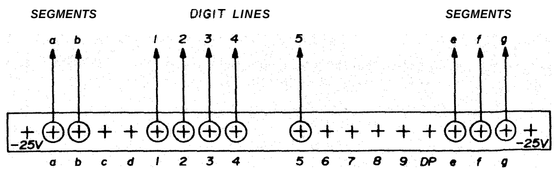
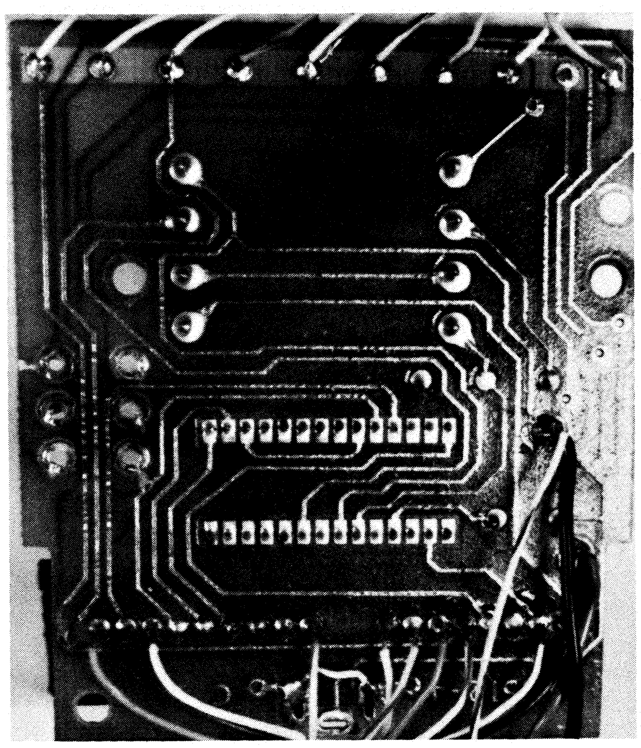


fig. 11. Connecting points on the calculator display board.

In general, construction is not critical. Just try to keep the connections to the calculator as neat as possible to prevent crosstalk and avoid loading down the calculator outputs. For those who choose to build this project with the TI 1265, the photographs also show the location for obtaining the necessary signals. A PC board and the 74S188 PROM are available from the authors; send a self-addressed, stamped envelope for details.

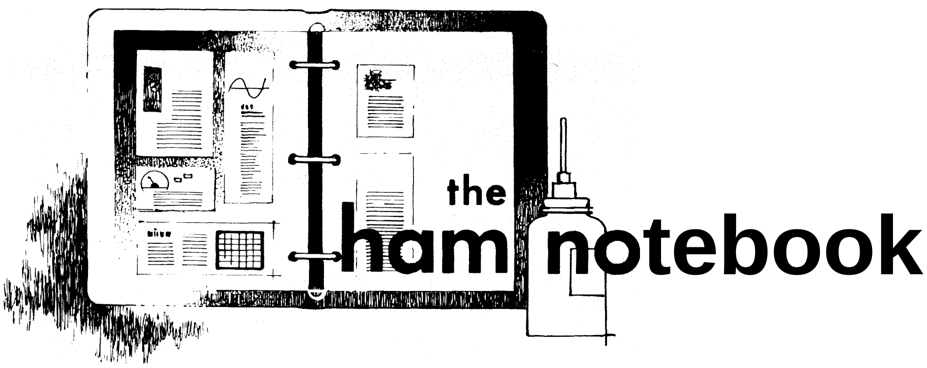


Connections on the calculator board.

What we have shown in this article is a simple, inexpensive system for using a calculator and its keyboard to control a digital system via BCD signals. Of course, if your system is not using BCD coding, you could devise your own programming for the code-converter PROM.

Once your system is designed, it would be a good idea to remount the calculator board, display, and perhaps use a different keyboard tailored to your individual needs, eliminating extra keys and labeling those with special functions.

ham radio



constant pitch monitor for cathode or grid-block keyed transmitters

After acquiring my old cathode-keyed Heath DX-40 transmitter and a new Novice ticket, WD4MNR found he needed a CW monitor. He didn't want to bother with either an rf-activated device that required connection to the coax or a separate signal pickup wire. Some published monitor circuits have obtained their operating voltage across a small dropping resistor connected in series with the key. This approach was tried, but we found that the final tube in his DX-40 keyed erratically if more than a few ohms were inserted in the cathode circuit. Even if this had not happened, the voltage drop across the resistor would have varied and changed the oscillator pitch every time the transmitter was loaded. A similar problem exists when rf pickups are used.

The circuit shown in **fig. 1** was developed to avoid problems. It keys reliably and at absolutely constant pitch, regardless of transmitter loading or band, and works equally well with other cathode keyed or grid-block keyed transmitters and transceivers. The open-circuit voltage across the key (about 75 volts in our case) turns on an optocoupler which gates a 555 oscillator circuit. Operation of the monitor is unchanged, even for manyfold variations in key-up voltage, as long as enough is there to turn on the MCT-2.

To use this monitor with rigs having different key-up voltages, temporarily substitute a 500k-ohm potentiometer for the 100k resistor connected to the key. With the monitor connected, the transmitter turned on, and the pot at maximum resistance, the monitor should produce a tone. Reduce the resistance of the pot until the oscillator just turns off; remove and measure the pot, and substitute a fixed resistor of the next lowest standard value. The monitor will now produce a tone when the key is closed.

Bill Clements. K4GMR

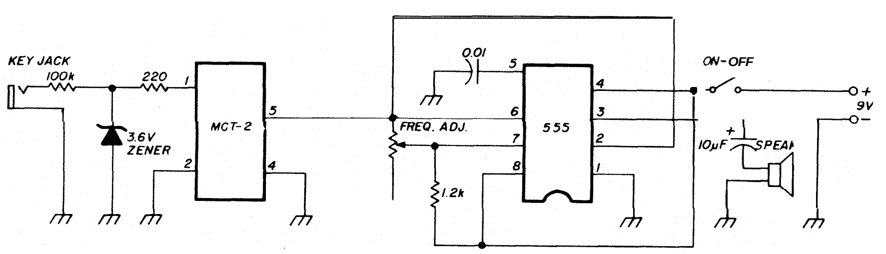


fig. 1. Schematic diagram of the constant-pitch CW monitor. The optical coupler is used to isolate the 555 from the actual transmitter circuitry.

direct-conversion receivers

I read with interest Dick Rollema's article on direct-conversion receivers (DCR) in your November, 1977, issue of *ham radio.* Since Dick also mentioned some previous work of mine, I feel compelled to give some additional information concerning DCRs.

It seems that the balanced transistor mixer which I used (as shown in **fig. 2** on page 46 of *ham radio)* has been surpassed through the work of a Soviet amateur, V. Polyakov (RA3AAE). So I must draw your attention to his basic article published in the December, 1976, issue of the Soviet magazine *Radio* (later articles appeared in the November, 1977, and December, 1977, issues). Using the first article, I built an experimental receiver and obtained excellent results, which I published in the September, 1977, issue of our Yugoslav radio amateurs' magazine, *Radio-amater.* I enclose partial copies of all those articles; however, I suppose you cannot read Russian or Yugoslav — so I shall give you a summary of the texts.

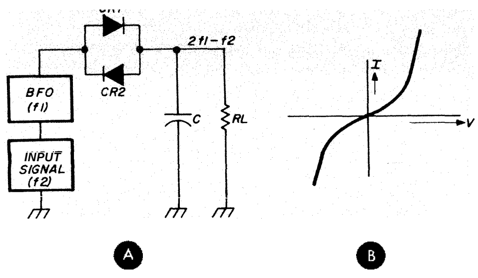


fig. 2. Mixer circuit for double-conversion receivers suggested by RA3AAE, and its characteristic curve.

On page 46 Dick Rollema rightly concludes that a good mixer for DCR should be a switching-type detector. Polyakov's solution is based on two matched silicon or germanium high-frequency diodes, connected in parallel and back-to-back. The local oscillator (in fact, bfo) must work on half of the received frequency, and the output from the detector is CW or ssb audio frequency. The integrated characteristic of the two diodes is represented with an inverted S-shaped curve (**fig. 2A**). Theoretically, the characteristic is similar to a cubical parabola ($I = AV + BV^3$), which means that this detector does not use square functions — thus making a-m detection impossible.

For popular explanation refer to the following. The lower cutoff points of the two diodes are used to obtain a switching device that can be opened and closed by the high-frequency voltage of the local oscillator (bfo). The so-called potential barrier for silicon diodes is about 0.5V and for germanium diodes about 0.1V. Therefore, in the middle part of the characteristic, neither of the two diodes conducts — which makes the switching function possible. The switching point occurs when the high frequency voltage of the bfo passes through zero and comes to one of the cutoff points. (Therefore the bfo's high frequency voltage, brought to the diodes through coil L-4, must exceed the potential barrier twice or more.) Since there are two zeroes in each cycle, it becomes obvious that the bfo must work on half of the received frequency. It can be equally said that the diodes CR1 and CR2 are closed sequentially when positive and negative half-waves of the bfo high frequency voltage reach their peaks. In those moments, the received signal — fed through coil L2 — will pass through one of the diodes and beat against the bfo voltage. At the same time the matched pair of diodes — connected back-to-back — cancel any a-m detection or a-m breakthrough, so this switching mixer is exceptionally insensitive to a-m detection.

In addition to simplicity, this mixer has several other qualities. Because the bfo works on half of the received frequency, it is easier to obtain good stability and to prevent unwanted radiation through the antenna. It is also very resistant to cross-modulation and overloading. Although Polyakov recommends silicon diodes — because of their higher potential barrier — I found germanium diodes to be more sensitive without losing other qualities of the mixer.

In the experimental receiver I built the rf resonant circuit C1-L2, for simplicity, was set on the middle of the 80-meter band once for ever (see **fig. 3**). The number of turns of the secondary coils L3 and L4 is one-fourth that of their primary coils, L2 and L5. Because there is no dc component after detection, the usual loading resistor was omitted and the high frequency components are shorted to ground through a 0.01 μF capacitor. I found it essential to use a low-noise audio preamplifier after detection — and one BF 173 did the job. The audio amplifier can be any integrated circuit or operational amplifier with possibility to include negative feedback through C_x and R, — thus peaking the resonant frequency of the amplifier at 1 kHz. Later I added also an rf stage ahead of the mixer — by using an fet — which made this little receiver a real marvel of clean CW reception, selectivity, stability, sensitivity, and resistance to cross-modulation and over-loading.

Because I am too busy for it at the moment, I hope that Dick Rollema and others will have some more time to experiment with this type of mixer.

Bozidar Pasric, YU2HL

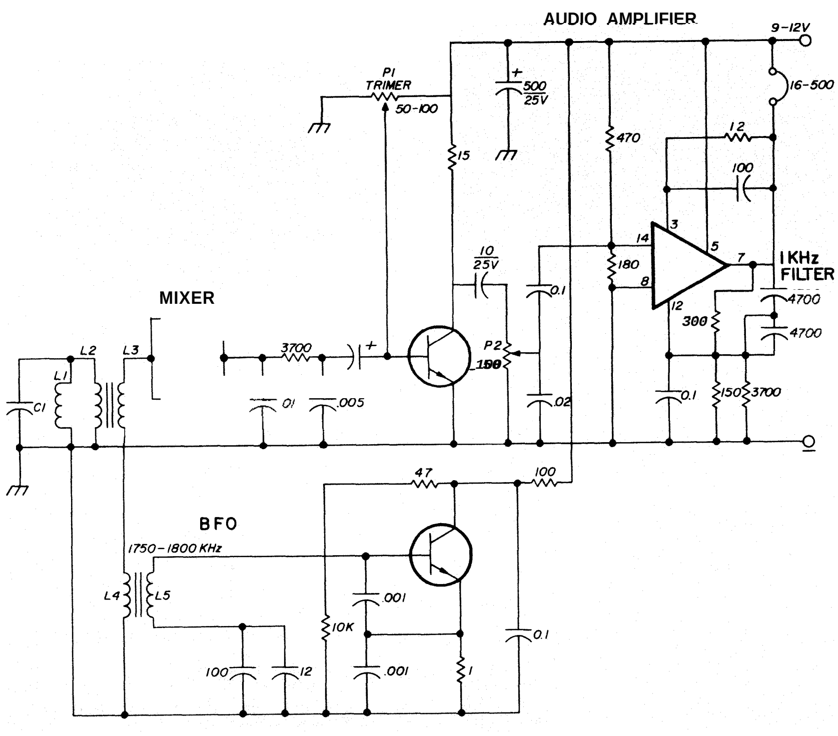


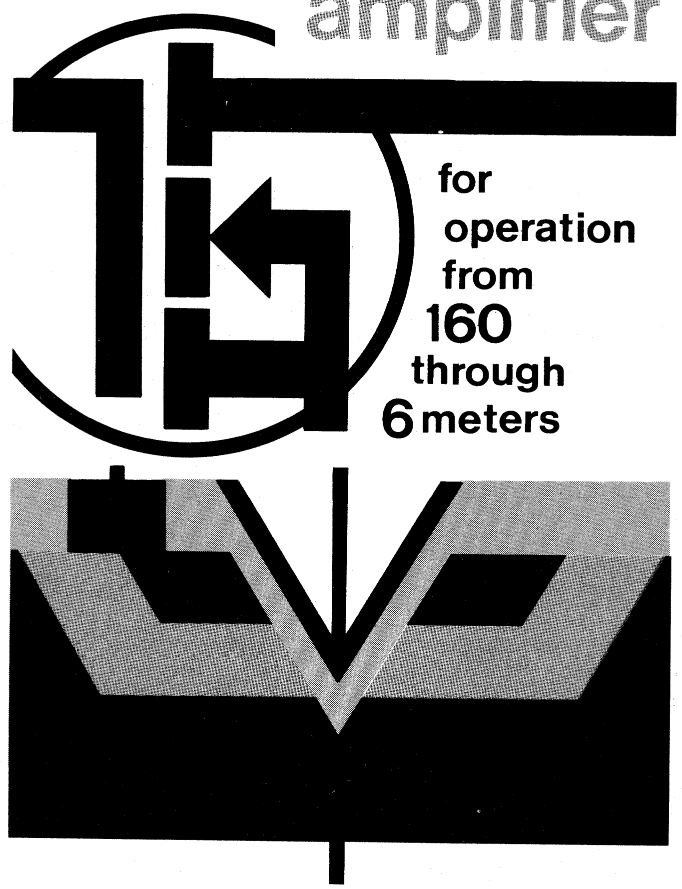
fig. 3. Experimental direct-conversion receiver built by YU2HL that uses the improved diode-mixer circuit.

$2.00

ham radio magazine

NOVEMBER 1978

mosfet power amplifier



for operation from 160 through 6 meters

ham radio magazine

NOVEMBER 1978

volume 11, number 11

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

United States: one year, $12.00
two years, $22.00; three years, $30.00

Canada: one year, $14.00
two years, $26.00; three years, $36.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

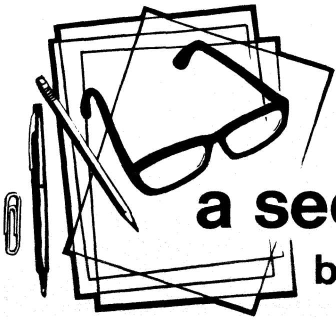
Foreign subscription agents are
listed on page 117

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

November 27th marks the 55th anniversary of one of Amateur Radio's most memorable events — the first ***two-way*** amateur communications across the Atlantic Ocean. It was a hard-won goal, its path marked with failure and frustration, but when the Atlantic, at last, had been spanned, it was conquered by short-wave Amateur Radio, on wavelengths that were considered by the "professionals" to be useless.

The first transatlantic tests, in December, 1920, were a dismal failure, as were another series of tests conducted in February, 1921. The 250 or so British stations which were listening for pre-arranged signals from the United States on a wavelength of 200 meters (1.5 MHz) jammed each other so badly with radiations from their own regenerative receivers that they couldn't hear any signals from across the pond!

A second transatlantic test was scheduled for December, 1921. In November, Paul Godley, 2XE, designer of the famous Paragon receiver, sailed from New York with two receivers under his arm — one a standard regenerative set with two stages of audio amplification, the other a 10-tube Armstrong superheterodyne built especially for the tests.

With this superhet and a wire antenna 390 meters long installed on the Androssan moor on the coast of Scotland, not far from Glasgow, Godley heard the first stateside signals coming through in the wee hours of the morning on December 8th. Over the next few days Godley copied the signals of 27 different American amateur stations; on December 12th he received the first complete message from the United States to Europe on the "short waves." British amateurs also participated in the test, and when all the reports were analyzed it was discovered that W.F. Burne, G2KW, had actually made the *first* positive identification of an American amateur signal.

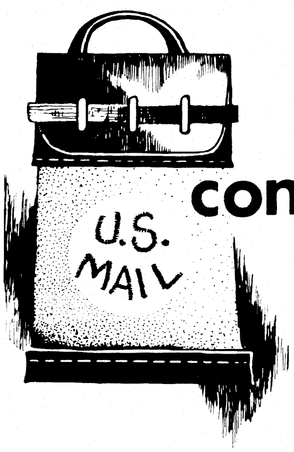
During similar tests a year later, two European stations, F8AB in Nice and G5WS in London, were heard by amateurs along the east coast of the United States. Many American signals were logged in Europe, but two-way communications were not established (but probably only because no one took the initiative to try).

A fourth series of transatlantic tests were scheduled for late 1923. However, these carefully laid plans were totally upset by the enterprise of one man, Leon Deloy, F8AB. Deloy came to the States during the summer of 1923 when he met with John Reinartz, 1XAM, and Fred Schnell, 1MO. Deloy picked up much valuable advice from his talks with Reinartz and Schnell, and before returning to France he acquired a new Grebe receiver and the details of a "trick" circuit which, he was told, would "go down to about 100 meters."

Deloy put his new 100-meter station on the air in early autumn, and, having satisfied himself that everything was in working order, cabled Schnell that he would transmit on 100 meters between 0200 and 0300 GMT on November 26, 1923. The signals from F8AB were heard by Schnell and Reinartz almost from the first dot he transmitted, but the Americans were not ready to transmit back. Unlike Deloy, who presumably did not think it was necessary to seek official permission to transmit on such a short wavelength, Schnell had to obtain the authority from the Radio Supervisor in Boston.

On November 27th, Schnell received special permits from Boston for himself and Reinartz; late that night they were both on the air. For an hour Deloy called the United States and then sent two messages. At 0330 GMT he signed off and asked for acknowledgment. Long calls followed from 1MO and 1XAM. Then came the eagerly awaited reply — Deloy had clearly heard both stations. Reinartz was asked to stand by as Deloy transmitted to Schnell, "R R QRK UR SIGS QSA VERY ONE FOOT FROM PHONES ON GREBE FB OM HEARTY CONGRATULATIONS THIS IS A FINE DAY — PSE QSL." It was, indeed, a fine day.

Jim Fisk, **W1HR**
editor-in-chief



comments

radiation hazards

Dear HR:

In the early 1960s I was civilian chief of a Radiation Hazards and Measurements Branch of European GEEIA — an arm of USAF. The branch had about 15 contract engineers and an equal number of service personnel checking RFI complaints, siting, and orientation accuracy of tropo and radar antennas, and noise measurements for prospective receiver sites at bases from Ireland to India. I seem to remember one or two radars rated at 3 megawatts!

Other than our concern with the possible (and to us tenuous) effects on personnel, there was the very real possibility that by some freakish concatenation of circumstances a radar beam might set off a squib on a bomb in an adjacent munitions dump, and touch off an international incident.

USAF used the 10 mW/cm^2 minimum, of course, but in deference to host governments, we had fences and warning signs installed at distances such that field intensity was perhaps one-half that minimum. Thus, we hoped that some peasant would not herd his sheep too close to a parabola and claim their milk had been soured.

Brodeur's article, which I read in the *New Yorker,* shook me somewhat. I was conditioned by an earlier series, also in the *New Yorker,* on the hazards posed by monster 200,000-ton supertankers — a hazard which has been brought to our attention rather frequently of late. Over a period of a month I pored over the composition of a letter to Mr. Brodeur; I didn't want to write anything challenging or in any way derogatory. I even had the local library try to ascertain whether Mr. Brodeur has academic credentials.

After reading W1HR's editorial in the August issue, I feel that Mr. Brodeur's articles and book must be relegated to the category of "The Bermuda Triangle." Both have been written more for shock effect than for any aim at veracity or objectivity.

Josef Darmento, W4SXK
Merritt Island, Florida

phasing networks

Dear HR:

In reference to the article on "SSB Phasing Techniques" in the January 1978, issue of *ham radio,* and WB9YEM's comments in the August issue, please note that I have published two articles on 90-degree phase-difference networks which may be of interest to your readers. The first appeared in *Electronic Design* in 1971* and described a simple, four-pole, 90-degree network consisting of four capacitors, six resistors, and one dual op amp; this is an improvement over WB9YEM's circuit by four resistors and seven years. It is also easier to calculate the element values from a given set of pole frequencies for this circuit.

In 1976, *Electronic Design* published a program in BASIC† which calculate the required pole frequencies for the A and B networks for any realizable set of design specifications. Note that until these pole frequencies are known, nothing can be designed. It is the only published program I know of which solves for these frequencies for any network. Computer hobbyists may wish to translate it to home computer BASIC.

I would be happy to send copies ot the original program listing to *ham radio* readers upon receipt of a self-addressed, stamped envelope.

Allan Lloyd, W2ESH
15 Greenwood Avenue
Hawthorne, New Jersey 07506

*Allan Lloyd, "Here's a better way to design a 90° phase-difference network," *Electronic Design 15,* July 22, 1971, page 78.

†Allan Lloyd, "90-degree phase-difference networks are simply designed with a program in Basic," *Electronic Design 19,* September 13, 1976.

windom antenna

Dear HR:

The article on the Windom antenna by K4KJ was of interest because I had written some articles on this antenna for *RADIO* magazine back in the 1930s. There was one variation I always used because I just don't like to cut wire into little pieces. I'd put my two meters in a jumper about 2 meters (6 feet) long, and jump this across the section of antenna to be measured; it can be moved along without cutting the wire. For a time I tied string from the feedline to the antenna to support the weight of the meters and feedline, so the jumper wire would stay reasonably close to the antenna.

I tended to use a wire about 85 meters (280 feet) long, and had no problems on 10, 20, 40, and 80 meters. With modern rigs, however, there might have to be some arrangement for coupling the single-wire feeder to the exciter or linear amplifier, especially for band switching without complications.

Bill Conklin, K6KA
La Canada, California

mosfet power amplifier
for operation from 160 — 6 meters

The potential of the recently developed vertical mosfet[1,2] to simplify and to improve rf power amplifiers has been shown in several recent papers.[3-6] These circuits usually employ class-A amplification, however, and so do not achieve the efficiency possible with class-B or class-D operation.[7,8] Where mosfets have been used in high-efficiency modes of operation, their power output has been relatively low.[9,10]

The circuit described here uses two Siliconix VMP-4 Mospower* fets in a push-pull configuration. The same circuit may be operated in either class B (for linear power amplification) or in class D (for highly efficient power amplification) by use of the proper output filter. The 160- to 6-meter frequency range and the 16-watt output make this circuit attractive for use in amateur equipment, particularly when multiband and multimode operation is desired.

The vertical mosfet has many advantages over its bipolar counterpart for both class-B and class-D rf power amplification. In particular, a complicated bias supply is unnecessary because of the high input impedance and the negative temperature coefficient of the fets; drive power is reduced and input matching is simplified by the high input impedance. Because of the negative temperature characteristic, secondary breakdown does not occur; hence the fets can withstand reactive loads in class-B operation.

In addition, the absence of storage time facilitates rapid switching in class-D operation. The absence of storage time eliminates the possibility of subharmonic oscillation in either class-D or saturated class-B operation. The ability of mosfets to pass current in both directions (*i.e.*, they are bilateral) allows them to operate in class D with reactive loads ***without*** diode protection (since suitable diodes for high-power high-frequency operation are not available, this is quite significant). The saturation resistance of an fet (in contrast to the saturation voltage of a bipolar transistor) allows the efficiency in class-D operation to remain high even at low power levels. This is also significant, since ssb voice signals[10] have relatively high peak-to-average power ratios.

The low gate-drain capacitance of the mosfet reduces feedback, making the amplifier easier to stabilize. The low gate-drain capacitance also reduces feedthrough, hence improving linearity (by supply voltage modulation) in class-D operation. And finally, the essentially constant gain of the vertical mosfet over the entire high-frequency band eliminates the need for gain-leveling circuitry.

theory

Class-A operation means that the fets are biased to remain in their active regions at all times.[7,8] While this mode of operation promotes high gain in bipolar transistor power amplifiers, it limits the circuit power efficiency to a maximum of 50 per cent. Efficiency is even less at less than full output (fig. 1), and attainment of full output is prevented in real circuits by the nonzero saturation resistance of real fets (or by nonzero saturation voltage in bipolar amplifiers).

A broadband class-B amplifier (fig. 2A) uses a pair of fets (or bipolar transistors or vacuum tubes) as current sources, controlled by the driving signal. For maximum linearity, the fets are biased so they have only a very small drain current when no driving signal is applied; therefore, application of drive alternately causes one transistor to be active while the other is cutoff (fig. 2B).

The parallel-tuned output circuit bypasses any harmonic components in the current, allowing only the fundamental (carrier) frequency component to reach the load. The sinusoidal voltage generated there becomes the drain voltage waveform with the addition of the supply voltage. The power output is then

$$P_o = \frac{V^2_{eff}}{2R} \qquad (1)$$

where R is the drain load resistance and $V_{eff} = V_{DD} R/(R + R_{on})$ accounts for the effects of fet on (saturation) resistance R_{on}.

At peak power output the efficiency (rf power out-

Mospower is the registered trademark of Siliconix Incorporated, Santa Clara, California.

This circuit was the runner-up in the 1977 EDN-Siliconix VMOS design contest, and is being published concurrently in EDN *Magazine*.

By **Frederick H. Raab, WA1WLW,** 240 Staniford Road, Burlington, Vermont 05401

put divided by dc power input) of a class B power-amplifier is

$$\eta \cong (\pi/4)\, R/(R + R_{on}) \leq 78.5\%$$

At lower output levels, class B efficiency is significantly higher than class A efficiency.

Saturation of the fets produces only a small increase in the output power and efficiency, because the sinusoidal drain voltage is maintained by the parallel-tuned circuit. During saturation, the fets are roughly equivalent to constant resistances, and the drain current is determined by application of the drain voltage to this resistance.

A class-D amplifier also uses a pair of fets, but they are controlled by the driving signal to act as switches. The voltage-switching configuration (fig. **3A)** has square-wave drain voltage (fig. **3B)** produced by the alternate switching of the two fets. The series-tuned output circuit passes current only at the fundamental (switching) frequency, and the sinusoidal output voltage is equal to the fundamental frequency component of the square-wave on the transformer secondary winding. Alternate half-cycles of the transformed output current flow through each fet. (If the load is reactive, the output current is out-of-phase with the switching, and at times must flow in the reverse direction through each fet.) The amplitude of the output is a function of the supply voltage, and is not affected by the amplitude of the driving signal if the drive signal is sufficient to cause switching. The power output of a class-D power amplifier is

$$P_o = \frac{8}{\pi^2} \cdot \frac{V^2_{eff}}{R} \qquad \textbf{(2)}$$

A class-D amplifier can ideally achieve an efficiency of 100 per cent, but in practice achieves $\eta \cong R/(R + R_{on})$ at most output levels. A gate bias voltage is not required, but is easily implemented with

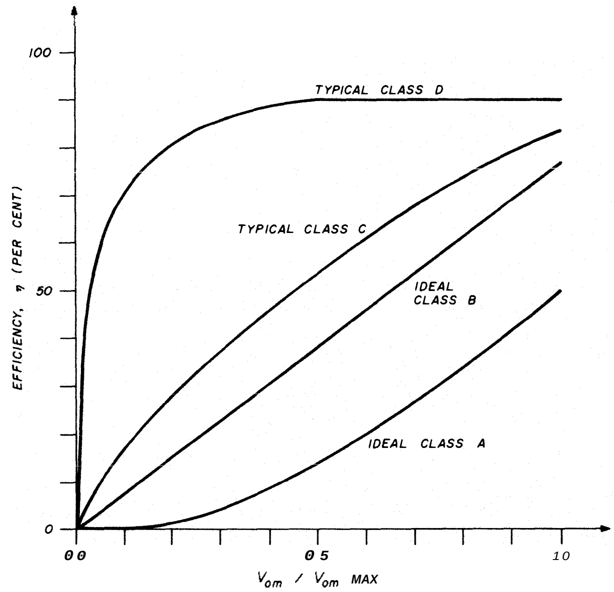
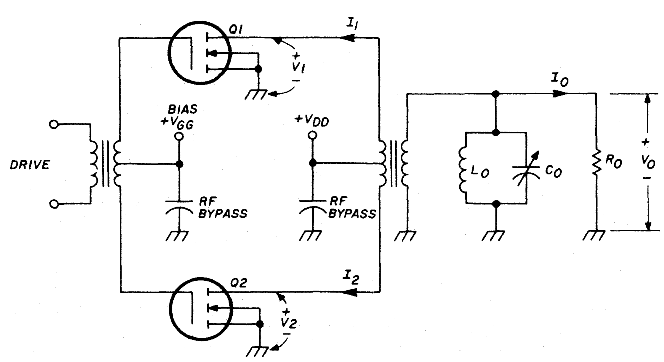


fig. 1. Efficiency of different classes of power amplifiers. Maximum theoretical efficiency of a class-A power amplifier is 50 per cent, while class B efficiency is 78.5 per cent. The efficiency of a class-D (switching) amplifier is much higher, particularly at lower output levels.

fets and can reduce the required amplitude of the driving signal.

Class-B operation of rf power amplifiers is typically used in the high-frequency bands for linear amplification of single-sideband suppressed-carrier signals. Class-B power amplifiers may be driven just into saturation to ensure maximum output and efficiency for CW and FSK transmissions. In real applications, a set of lowpass filters (with shunt capacitor inputs) is



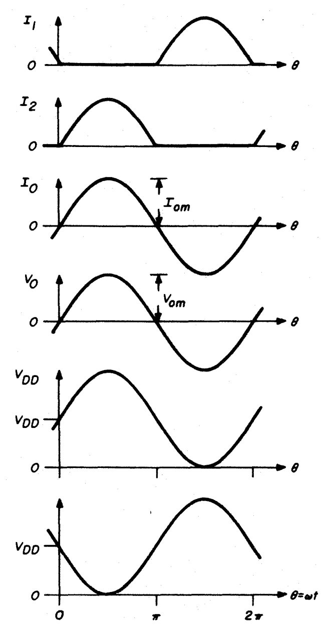
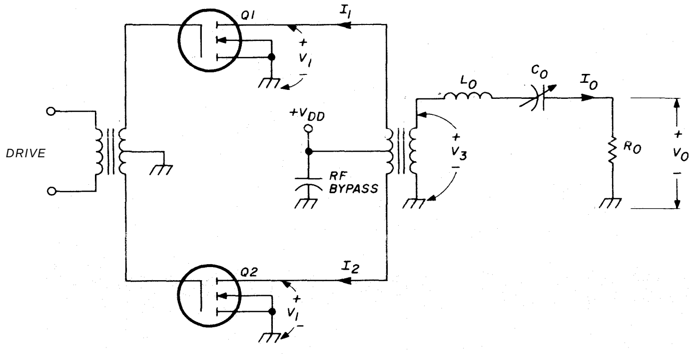


fig. 2. Operation of vertical mosfets (vmos fets) in class B. The fets act as current sources. Waveforms of the class-B amplifier are shown at right.



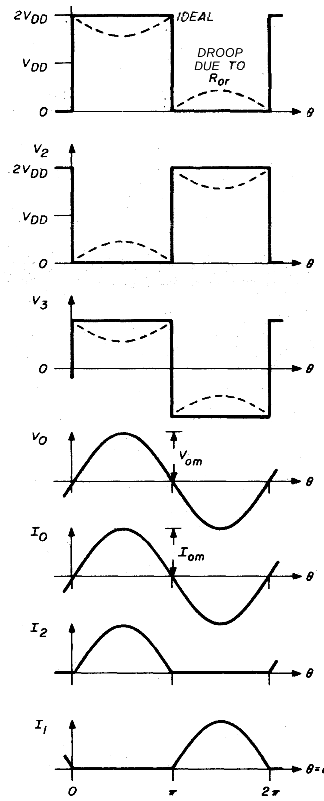


fig. 3. In class-D operation the fets operate as switches. Waveforms of the class-D amplifier are shown at right.

used instead of a simple parallel-tuned circuit. The cutoff frequencies of the filters are chosen to be slightly less than one octave from each other to provide adequate harmonic rejection for any frequency of operation.

Class-D amplifiers for high frequency use are just now becoming practical. Circuits such as this would be used directly for CW, FSK, and fm transmissions. Linear amplitude modulation is readily accomplished by variation of drain supply voltage. Single-sideband signals can be amplified through the envelope elimination and restoration technique,[7,8,10] which uses amplitude modulation to restore the envelope of the ssb signal and a limiter to ensure adequate drive for the power amplifier. A set of lowpass filters might similarly be used to span the range of operating frequencies; however, these would have series-inductor inputs, rather than the shunt capacitor inputs of the filters used with class-B amplifiers.

design

Since input requirements are based on the output current, design begins with the output portion of the circuit.

Output design. Inspection of the VMP-4 Mospower data sheet shows a maximum drain current of 1.6 amps and a maximum drain voltage of 60 volts; the latter allows a +30 Vac variation around a supply voltage of +30 volts. Were there no saturation voltage, outputs of 24 watts (class B) and 30.4 watts (class D) could be obtained. For these power outputs, load resistances of $30/1.6 = 18.75$ *ohms* (class B) and $(4/\pi)(30/1.6) = 23.87$ *ohms* would be required. The load resistance required to obtain the maximum output power when the saturation (on) resistance is non-zero is determined by subtracting the saturation resistance from the load resistance determined above for zero saturation resistance. Thus, for a typical value of $R_{on} = 2.6$ *ohms*, the maximum power loadlines are $18.75 - 2.6 = 16.15$ *ohms* and $23.87 - 2.6 = 21.27$ *ohms* for class-B and class-D operation, respectively.

Obtaining optimum broadband performance requires the use of transmission-line transformers;[8,11] this limits the choice of impedance transformations. A convenient transformation is 4:1; with a 50-ohm load, the load line is $R = 12.5$ *ohms*. This transformation is accomplished by a pair of transformers as shown in **fig. 4.** The hybrid (T2) requires a 25-ohm transmission line; it is built by winding two turns of transmission line through two parallel stacks of Ceramic Magnetics CN-20 ferrite* (each stack is 2.86 cm long). The 25-ohm transmission line may conveniently be made by connecting two 50-ohm miniature coaxial cables in parallel.

The balun (T3) requires a 50-ohm characteristic impedance line, and is fabricated by winding two turns of miniature coax through two parallel stacks of CN-20 material (also 2.86 cm long). The combination of T2 and T3 provides a mostly resistive input impedance over the entire 1.6 to 54 MHz band when connected to a 50-ohm load.

With a load line of 12.5 ohms, the output power is limited by the maximum drain current, thus

*If CN-20 material is not available, other rf ferrite material can be used. The reader can use the design procedures in references 8 or 11 or can borrow a design from another published PA circuit.

$$P_{omax} = \frac{I^2_{Dmax}R}{2} = 16\ watts \qquad (3)$$

(An output of 25 watts would be possible with the VMP-1, which has a 2 amps rating.)

The 1.6 amp peak drain current requires a dc supply current of $1.6(2/\pi) = 1.02\ amps$. In class-B operation, the drain voltage swing is simply $1.6 \bullet 12.5 = 20\ volts$ at peak output. To avoid saturation, the drain voltage must be greater than $1.6 \bullet 2.6 = 4.2\ volts$; therefore, the supply voltage must be 24.2 volts to allow maximum output power (higher supply voltages allow maximum output power, but reduce the efficiency).

For class-D operation, the 4.2 volt peak drop across the saturation resistance acts in opposition to the fundamental frequency component of the ideal square-wave, requiring a supply voltage of $24.2\ (\pi/4) = 19.0\ volts$ for maximum output. With $R_{on} = 2.6\ ohms$, you can expect efficiency of 82.8 per cent for class D, and 65 per cent for class B.

The values of the bypass and blocking capacitors and chokes are not especially critical. However, care should be taken to make sure there are no self-resonances within the band of operation. For testing purposes, simple tuned circuits with a loaded Q of 3 (parallel-tuned) or 5 (series-tuned) were used.

Input design. The input circuitry of a power amplifier using fets can differ considerably from the input circuitry of a bipolar amplifier. A relatively high voltage across a relatively high capacitive reactance is required. Both gate capacitances remain in the circuit at all times, in contrast to the reverse-biased base being effectively out of the circuit. Inspection of the VMP-1 data sheet suggests that a bias voltage of about 2 volts and a sinusoidal voltage of about 6 volts peak will be required to produce the 1.6-amp peak drain current. The VMP-1 data sheet also shows a gate-to-source capacitance of about 50 pF.

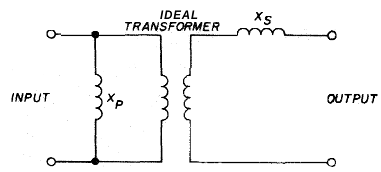


fig. 5. Equivalent circuit of the wirewound transformer used at the input of the amplifier (see text),

A 4:1 voltage reduction produces a more convenient output for the driver. Since the input impedance of a Mospower fet varies from about 2 kilohms at 1.6 MHz to 59 ohms at 54 MHz, true broadband matching is difficult. Since the drive power is relatively small in comparison to the output power, however, matching of the input to the power amplifier is not really necessary in most applications. This philosophy makes possible the use of a simple "wirewound" transformer. **Fig. 5** depicts the equivalent circuit of such a transformer; low-frequency performance is degraded by the shunt (magnetizing) inductance X_p, while high-frequency performance is degraded by the series inductance X_s. If the driver acts as a current source, only the shunt inductance reduces the current reaching the gates. If the driver acts as a voltage source, only the series reactance reduces the voltage applied to the gates.

The selected design for input transformer T1 is therefore a single-turn primary and a 4+4 turn secondary wound on a Ferroxcube 3B7 core (1 cm diameter). At midband, a driving voltage of 2.5 volts (peak) with a 70 mA (peak current) should be sufficient. The required driving current will be higher at the low end of the band, and both the driving current and the driving voltage increase at the high end of

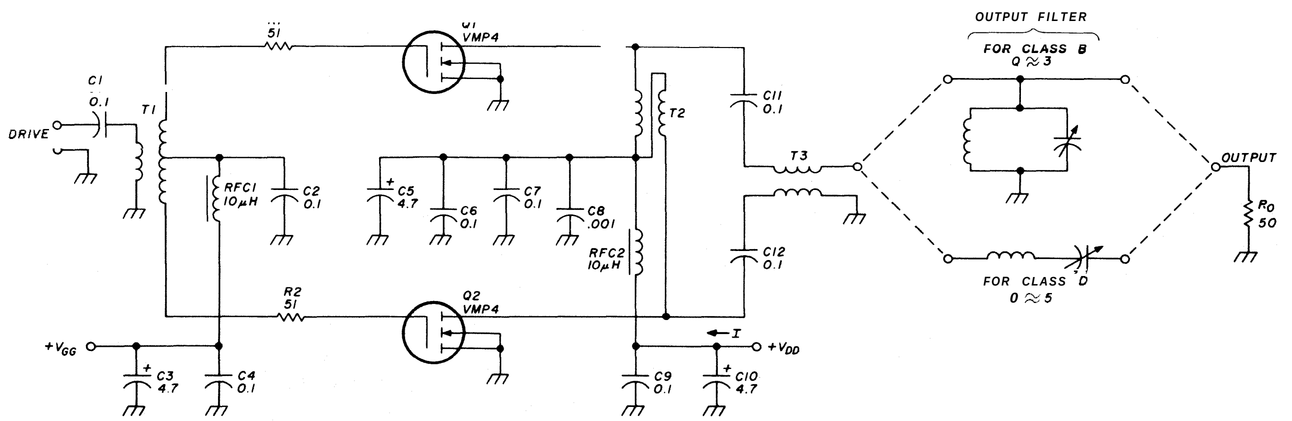


fig. 4. Schematic of a vmos fet power amplifier for 160 through 6 meters. The design of transformers T1, T2, and T3 is discussed in the text. Capacitors C5, C6, C7, and C8 provide an rf ground; combination may vary with different capacitors. RFC1 and RFC2 consist of 40 turns of no. 26 (0.4mm) wire on a Permacore 57-1677 form.

the band. Better overall performance might be achieved by using a core made from material with a lower permeability (3D3, for example). The 51 ohm resistors R1 and R2 dampen the tuned circuit formed by fet capacitance and the inductance of transformer T1, thus preventing high-frequency oscillations.

Construction is straightforward, not critical, and quite similar to that for other high-frequency power amplifiers. Fig. **6** shows the physical layout of the amplifier. The fets are in the middle, with input-related circuitry on their left and output-related circuitry on their right. Transformer T3 is on top of transformer T2, and both can be supported by pieces of PC board. The circuit was assembled on a piece of double-sided PC board to ensure a good ground plane; holes were cut in the board to allow the fets to be mounted directly on a small heatsink.

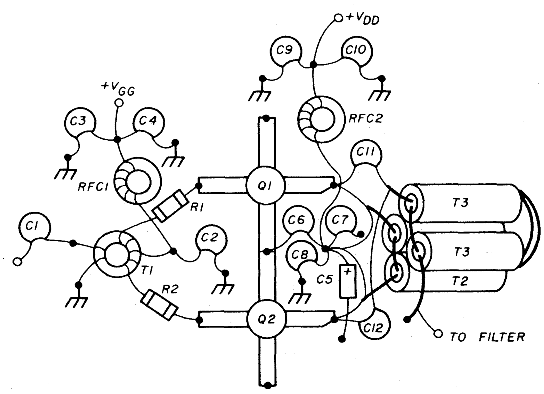


fig. 6. Parts layout of the mosfet power amplifier. Construction is straightforward, not critical, and similar to that used for other high-frequency amplifiers.

performance

Performance of this PA is measured in terms of its linearity and efficiency in both class B and class D operation, as well as by its maximum output and efficiency at various frequencies.

Class B linearity. Measurements of the linearity (output voltage) and efficiency as functions of the drive amplitude were made at 3.5 MHz and with $V_{DD} = 24$ *volts.* A gate bias voltage of 1.65 volts was used; this produced a total quiescent drain current of about 70 mA. Since a spectrum analyzer was not available, this value of V_{GG} was selected by increasing V_{GG} until cross-over distortion vanished from the combined current waveform at the output from T3.

Variation of the output voltage and efficiency with drive (peak voltages) is shown in fig. **7.** It is evident in both curves that saturation has an effect at a lower output than expected. Efficiency increases roughly in proportion to output amplitude, reaching the expected 65 per cent at peak output. A simple numerical analysis shows that this curve is roughly equivalent to a third-order IMD ratio of -20 dB with the peak output power at 16 watts. Decreasing the peak output power or increasing V_{DD} will keep the fets out of saturation and thus decrease the IMD. However, it is very possible that a different choice of V_{GG} will improve the IMD; this can easily be determined with the aid of a spectrum analyzer. Should this not produce a level of -30 dB or better, feedback might be added to the circuit.

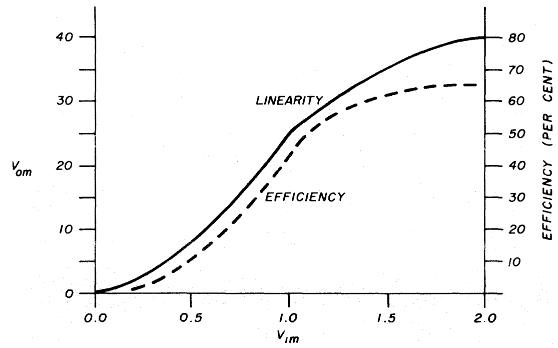


fig. 7. Class B performance of the mosfet power amplifier. Efficiency increases approximately with the output amplitude. reaching the expected 65 per cent efficiency at peak output.

Class D modulation linearity. Measurements of performance in class-D operation at various amplitudes were also made at 3.5 MHz. A driving voltage of 2.5 volts peak was sufficient to ensure saturation. Gate bias was reduced to 1.2 volts, since drive linearity was not required.

Variations of the output voltage and efficiency as functions of the supply voltage are shown in fig. **8.** The very linear relationship of the output voltage to the supply voltage indicates excellent amplitude

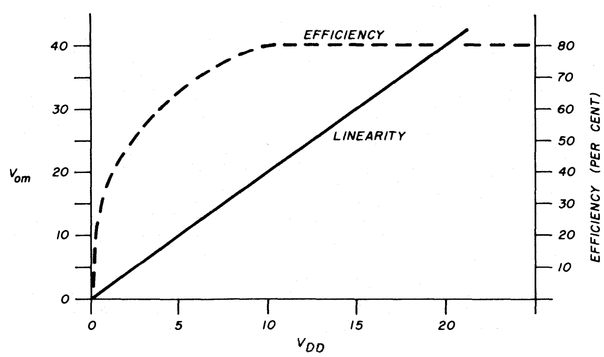


fig. 8. Class D performance of the high-frequency mosfet power amplifier. The very linear relationship of output voltage to supply voltage indicates excellent amplitude modulation characteristics. Efficiency remains close to 80 per cent except for low output levels.

modulation characteristics. The feedthrough (V_{om} with $V_{DD} = 0$) measured only 34 mV, which is –61.6 dB relative to peak output! Efficiency remains close to 80 per cent except for low output levels, where it decreases but is still several times greater than that for class-B operation.

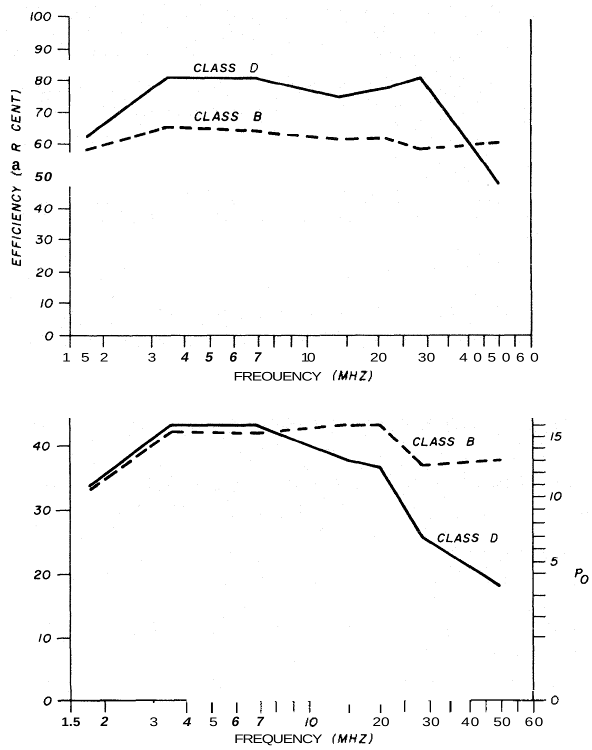


fig. 9. Performance of the mosfet power amplifier vs. frequency.

This amplifier can be tuned at full output power without fear of damage to the fets. This contrasts with the use of bipolar transistors,[10] with which tuning at other than very low power can destroy the transistors. Because of the exceptional stability of the *Mospower* fets, there is no oscillation when the drive is reduced to less than that required to maintain saturation; the output simply decreases gradually as the fets become current sources instead of switches. This may make possible a simplification of the class BD rf amplifier.[12]

frequency characteristics

Measurement of circuit performance in class-B operation was made with the supply voltage set at 24 volts; drive was then increased until maximum output was obtained. The resulting variations of efficiency and maximum output with frequency are shown by dashed lines in **fig. 9.** In the 80 to 15 meter bands, the output and efficiency are nearly constant at about 16 watts and 64 per cent, respectively. At both the high and low ends of the 160 to 6 meter range, both output and efficiency decrease as the performance of the output transformer degrades.

The performance of the circuit in class-D operation was measured with the supply voltage set at 20 volts and the drive increased until no further increases in output were observed. The resulting efficiency and maximum output curves are shown in **fig. 9** as solid lines. An efficiency of about 80 per cent is obtained in the 80 and 40 meter bands with an output of about 16 watts. Slight decreases in the output and efficiency can be seen in the 20- and 15-meter bands. More significant decreases in both output and efficiency occur near the high and low band edges.* It is not surprising that these are more pronounced than in class-B operation, since class D is a more exacting mode of operation.

This circuit has demonstrated the use of *Mospower* fets in both class-B and class-D operation in the high frequency range. Performance in class-B operation was generally similar to that obtained with comparable bipolar transistors; performance in class-D service was generally superior to that which can be obtained with bipolar transistors. In either type of operation, the VMOS fet offers many advantages stated previously that do not show up in the measurements made above, but do show up in the operation of real circuits. Similar design techniques can be used with other VMOS fets to achieve other power outputs.

*Because of the limitations of the author's test equipment, the 10-meter and 6-meter measurements should be regarded as approximate.

references

1. Ed Oxner, "Will VMOS Power Transistors Replace Bipolars in HF Systems?" EDN, June 20, 1977, page 71.
2. A. D. Evans, D. Hoffman, E. S. Oxner, W. Heinzar, and L. Shaeffer, "Higher Power Ratings Extend VMOS FET's Dominion," Electronics, June 22, 1978, page 105.
3. Larry Leighton, "Two-meter Transverter Using Power FETs," ham radio, September 1976, page 10.
4. Ed Oxner, "MOSPOWER FET as a Broadband Amplifier," ham radio, December 1976, page 32.
5. Gary D. Frey, "VMOS Power Amplifiers – This Broadband Circuit Outputs 8W with a 15 dB Gain," EDN, September 5, 1977, page 83.
6. Stephan Ludvik, "Vertical Geometry is Boosting FETs into Power Uses at Radio Frequencies," Electronics, March 2, 1978, page 105.
7. F. H. Raab, "High Efficiency Amplification Techniques," *IEEE* Circuits *and Systems* (Newsletter), December 1975, page 3.
8. H. L. Krauss, C. W. Bostian, and F. H. Raab, Solid-State Radio Engineering, John Wiley and Sons, New York, to be published.
9. F. H. Raab, "FET Power Amplifier Boosts Transmitter Efficiency," *Electronics*. June 10, 1976, page 122.
10. F. H. Raab, "High Efficiency RF Power Amplifiers," ham radio, October, 1974, page 8.
11. H. L. Krauss and C. W. Allen, "Designing Toroidal Transformers to Optimize Wideband Performance," Electronics, August 16, 1973, page 113.
12. F. H. Raab, "The Class-BD High-Efficiency RF Power Amplifier," IEEE Journal of Solid-State Circuits, June 1977, page 291.

ham radio

basics of the digital VFO — a tunable synthesizer

Basics of the digital synthesizer, from simple switch-controlled dividers to a full optical-encoder controlled system

It has been many years since something technically innovative has come to ham radio. Two meters has allowed development of repeaters, scanning, and Touch-Tone* phone patching, but these developments are generally extensions of already well-known and previously applied principles. The digital synthesizer is somewhat new to amateur radio, but has been in use in commercial and military equipment for many years. The digital synthesizer was a natural for two-meter operation because of the current practice of operating on fixed channels. By and large, amateurs have contributed very little to the synthesizer art.

On the high-frequency bands, due to nonchannel operation, the knob- or push-button-operated synthesizer has found little favor. Those of you who have tried to tune the hf ham bands using synthesized, military receiving equipment will readily understand why. Yet the advantages of the synthesizer over the VFO are tremendous. Apart from its inherent frequency stability and accurate readout, the synthesizer allows the transceiver designer to convert to an i-f frequency higher than the signal frequency, eliminating the need for multiple conversion and the attendant barrage of spurious and birdie signals that usually accompany multiple conversion techniques.

Additionally, each mixer placed at the front end of a receiver increases the potential for cross modulation, intermodulation, and image products.

The principal advantages of up-conversion design have been well documented, particularly in recent articles by Ulrich L. Rohde, DJ2LR,[1] and Wayne Ryder, W6URH.[2] So, no more will be said here except to point out that a high-frequency i-f amplifier requires a high-frequency local oscillator signal. Unlike the VFO, the synthesizer readily operates at the higher frequencies.

Recently, a new transceiver, the Astro 200, appeared on the ham market. To the best of my

*Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.

By Lester A. Earnshaw, Post Office Box 1584, Sedona, Arizona 86336

knowledge, the Astro 200 was an amateur first. Its design has probably set the pace for all new designs in ham radio, using a tunable synthesizer. To tune down in frequency a momentary switch is depressed; to tune up in frequency the switch is lifted. A digital readout keeps track of the frequency. The company's literature claims that this method of tuning will eventually be used in all ham transceivers and, with some reservations, I agree. However, I do not agree about the switch method of performing the tuning. In my opinion, the method lacks feel and does not control the speed with which the signal is tuned. But this is a matter of application, not principle. There are various variable-speed methods that overcome this objection. The point is, it is only a matter of time before the continuously tunable synthesizer does to the VFO what transistors have done to tubes.

Surprisingly, the manufacturers of the Astro 200 stayed with the conventional low-frequency i-f, reducing by perhaps 75 per cent the effectiveness of the system. To have "tamed" the multitude of harmonics and birdies that this low-frequency system must have generated is a task I am pleased was not allotted to me!

There has been little information in the amateur literature, or any literature for that matter, concerning tunable synthesizers, or, as I have chosen to call them here, digitally tuned VFOs (DTVFOs). Anyone experimenting with DTVFOs is virtually on his own. It is hoped that the information presented here, which has come mostly from breadboards and experiments, will be of some assistance. I have tried to present the information in a progressive way, much as I learned it myself. However, I should point out that the devices are not necessarily the latest or the best available. Most of the ICs are readily available junk box surplus.

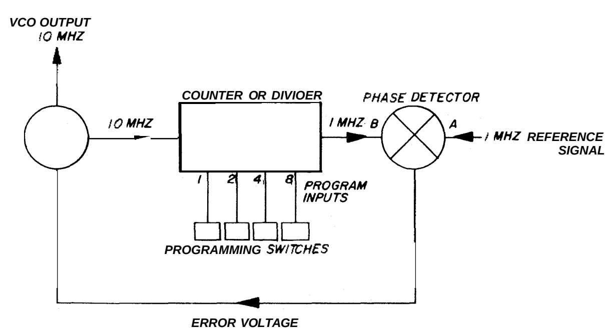


fig. 1. Block diagram of a basic digital synthesizer. The phase detector is used to determine the difference between the reference frequency and that supplied by the divider. The error voltage represents the difference and is used to change the frequency of the VCO until there is zero difference. The programming switches can alter the divider, and hence will change the output frequency of the VCO.

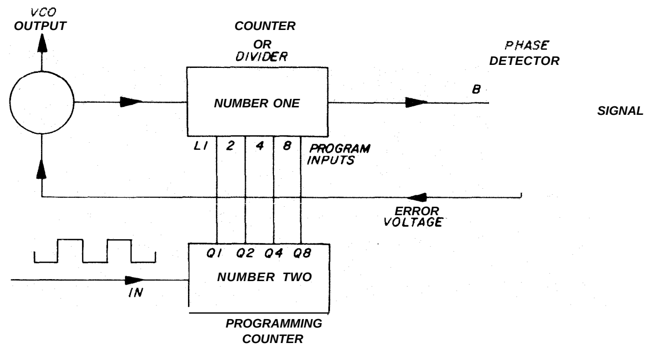


fig. 2. In this example, the programming switches have been replaced with another counter. Since the outputs of the second counter are in BCD, they are used to directly drive the load inputs of the first counter.

Amateurs with a technical background will have little trouble duplicating and improving the circuitry.

synthesizers

A DTVFO is actually a modification of a regular synthesizer which uses a number of panel-mounted BCD switches or pushbuttons to program counters to control the desired frequency. The DTVFO, instead of using switches, uses the Q outputs from a second parallel counter to do the programming. This is more clearly illustrated in the block diagrams of **figs. 1** and **2. Fig. 1** is a conventional, simplified digital counter using a phase-locked loop system and panel-mounted switches to select the frequency.

In this instance, let the reference signal applied to the phase detector be 1 MHz and the VCO output be 10 MHz. If the counter or divider has been set by the knobs to divide by 10, the second signal applied to the phase detector will also be 1 MHz. There will be no error voltage from the phase detector. Now, let the VCO be 11 MHz; the signal applied to the phase detector will now be 1.1 MHz. The phase detector will develop an error voltage and retune the VCO to 10 MHz so that the input to the phase detector is again 1 MHz. When no error voltage is developed the system is in lock.

If the divide ratio is set to 9, the VCO will retune to 9 MHz so that once again the phase detector input is 1 MHz. Thus, by changing the divide ratio, a number of VCO output frequencies, each locked to the crystal-derived reference frequency, may be obtained. Note that the counter cannot divide by fractions. Since the reference frequency in the example just discussed is 1 MHz, the VCO output can only be in 1-MHz increments. If smaller increments are required, the reference frequency must be lowered and the number of counters and synthesizer frequency-selector knobs increased. Obviously, it requires only one knob to tune a frequency synthesizer from 1 MHz to 9 MHz in 1-MHz steps, but it requires four knobs to cover the same frequency range in 1-kHz steps.

Fig. 3 shows a truth table for a 74192 programmable decade counter commonly used in frequency synthesizers. The program input pins (often called load inputs, program inputs, or jam inputs) are normally held low. By connecting specific inputs to +5 volts, a predetermined counter divide-ratio may be obtained. Highs (+5 volts) are shown as 1 on the truth table. To divide by 1, for example, input L1 (pin 15 on the 74192) is taken high. To divide by 2, input L2 (pin 1) is made high. And to divide by 3, both L1 and L2 are made high. These functions are normally carried out by the programming switches. It should be readily apparent that if a method of progressively raising the appropriate inputs can be devised, the VCO will sweep across a band of frequencies. If a smaller reference frequency is used, the VCO will sweep in smaller steps. When the steps are sufficiently small — less than 100 Hz — it will seem as if the VCO has tuned across a signal in a normal VFO manner.

To accomplish the tuning or sweeping action yet stop on command, a second counter may be substituted for the BCD switches. This is shown in **fig. 2** and, for the purpose of clarity, this counter is referred to as No. 2 counter. If we feed a square wave, one pulse at a time, into the input of counter No. 2, assuming the counter had been set to zero, the first pulse will cause Q1 output to go high. Since Q1 is connected to L1 of the No. 1 counter, the synthesizer will divide by 1. When the second pulse is entered into the counter, L1 will go to ground and L2 will go high. The counter will divide by 2. Similarly, a third pulse will cause both L1 and L2 to go high for a count of 3. And so on. If only one set of counters is used, the division will be from 1 to 9. If two sets of series-connected counters are used, the division will be 1 to 99; three counters, 1 to 999; and so forth.

DIVIDE BY	L 1	L 2	L 4	L 8
1	1			
2		1		
3	1	1		
4			1	
5	1		1	
6		1	1	
7	1	1	1	
8				1
9	1			1

fig. 3. Truth table for the programming inputs of a 74192 programmable divider.

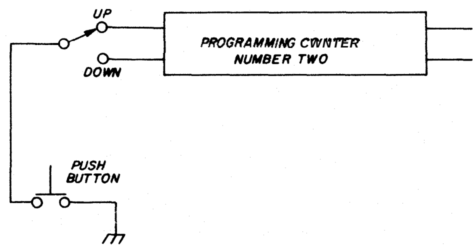


fig. 4. The single-pole, double-throw switch allows either an up or down count each time the button is pressed. Normally though, a low-frequency pulse train would be fed into the switches.

To tune or sweep a synthesizer, it is merely necessary to feed a number of pulses into the No. 2 counter. The pulses may be created by sequential action of a pushbutton-type switch. But, as this is a slow process, a square-wave oscillator may be turned on when the button is pressed and the output entered into the counter. This is the same method used to tune the Astro 200 transceiver.

One problem with feeding the oscillator into the counter is that it becomes necessary to tune across the whole band if the counter is at one end and the desired frequency at the other. This can be quite a chore on 10 meters, especially when the square-wave oscillator has a fixed speed. Fortunately, this problem may be overcome by presetting the load inputs. A ***Touch-Tone*** pad will allow the operator to preset the frequency, with the fine tuning accomplished by feeding the square-wave into the clock input.

With the counter connected as shown in **fig. 2,** it will sweep only in one direction. In the case of the single-decade unit, the counter will count to 9, reset to 0, and start over again. To move the frequency in either direction a special kind of counter is required. This is known as an up/down counter. The 74192 has this facility. The counter has two input terminals, one for counting up and one for counting down. **Fig. 4** shows how a double-pole, single-throw switch may be connected to route the tuning pulses to either counter. In practice, pulse conditioners, also known as debouncers, would have to follow the switches. (No matter how quickly and firmly the buttons or switches are activated, there will be bounce.) For those who wish to experiment with pushbuttons, a simple debouncing circuit is given in **fig. 5.3**

The pushbutton makes it relatively easy to observe the sequence of events, but you must realize that in an actual system with a 100-Hz reference, the VCO

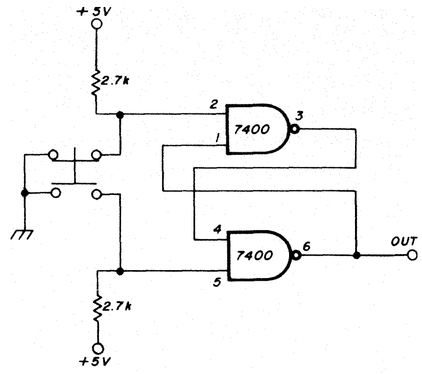


fig. 5. Example of a simple switch debouncing circuit.

will move only 100 Hz each time the button is pressed. Be prepared for an all-night session if you wish to tune across 10 meters!

As previously mentioned, a low-frequency, square-wave oscillator may be used to speed up the tuning: it may be switched between input terminals, or two separate oscillators can be used. Of course, only one oscillator may be activated at one time. It is important to note that when feeding a square wave into one input terminal, the other terminal must remain high.

Several switching methods for the square-wave oscillators have been used. The first used two separate pushbuttons, one for up and one for down. Better yet was a potentiometer with a center off position. When turned counterclockwise, the oscillator slowly started up, increasing frequency as the knob was turned. When the potentiometer was turned clockwise, past center, the other oscillator started.

frequency readouts

A major disadvantage of the DTVFO is that there is no simple, conventional, dial-type readout that can be used. Fortunately, the DTVFO also contains the essentials of a frequency counter that can be used instead of a conventional dial. Referring to **fig. 2,** the Q outputs of the No. 2 programming counter are exactly like those obtained from a frequency counter. **Fig. 6** illustrates the connections necessary to convert the BCD information to a digital readout. Only a few extra components are required. The system is simplified by the fact that no latching is required. Since the counter does not update, no flickering occurs and the figures change only when the VCO is tuned.

With the frequency counter measuring the local oscillator, an offset is required so that the display indicates the actual transmitted frequency. If the signal frequency is 14000.0 kHz and the VCO is 5000.0 kHz, a 9000.0 offset is needed to make the readouts indicate 14000.0 kHz. In an amateur-band transceiver, the MHz readout is often hardwired and activated by the bandswitch because the bands only cover a small part of the spectrum. In a complicated system, such as a general-coverage receiver, a PROM may well provide the programming link between the bandswitch and the readout.

A digital VFO is subject to all the pitfalls and problems normally encountered in a regular synthesizer. Of prime interest is error-voltage filtering and lock-up time, plus a couple of other hazards applicable to the tunable system. A tunable system requires a low-frequency reference, 100 Hz or less, preferably less. Even a 100-Hz system produces observable steps, especially noticeable on CW. Unfortunately, a low-reference signal makes it very difficult to adequately filter the reference spikes from the error-voltage line without unduly affecting the settling and lock-up time. The reference signal will show up as sidebands, removed from the carrier by the amount of the reference frequency. When a low-frequency reference is used, lock-up time takes considerably longer, and if the VCO frequency is high, many cycles of operation will take place before correction can occur. The VCO may then exhibit jitter or burble.

Sideband content is also highly dependent upon the type of phase detector used. Rohde has thoroughly discussed the advantages and the disadvantages of different types of phase detectors.' There is little I can add except to reiterate that the sample-and-hold-type detectors and the RCA CD4046 are preferred over most of their counterparts.

In many synthesizers, when the frequency is being changed the oscillator output is muted. With the oscillator muted, it will be impossible to recognize that a station is tuned in until the tuning stops. How-

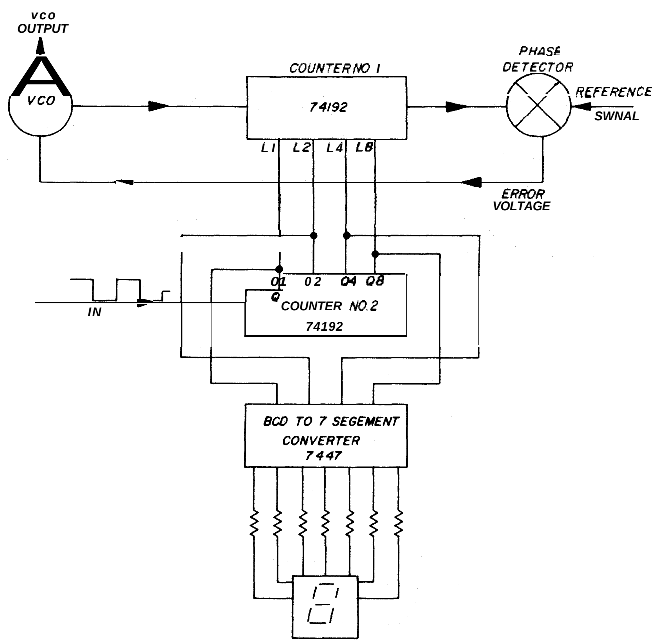


fig. 6. By connecting a BCD-to-7-segment converter to the outputs of the first counter, the frequency of the VCO can be directly read. This, in most cases, will not be the frequency to which the receiver is tuned.

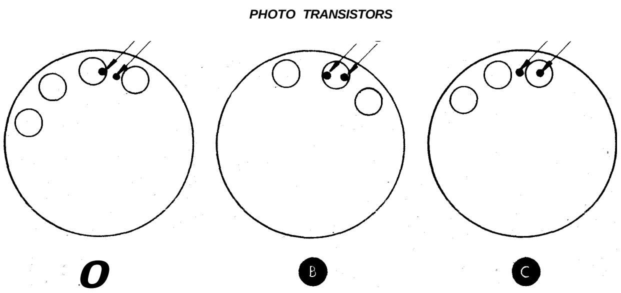


fig. 7. Diagram of the basic system for using a rotating disc to control the frequency of the VCO. Depending upon the direction of rotation, the voltages derived from the photo transistors will cause the counter to count either up or dawn, hence, changing the frequency of the VCO.

ever, if presetting is employed (a large VCO frequency change), muting may be necessary.

In a 100-Hz reference system, you will notice that the tuning action imparts a small fm effect to the signal during the tuning process. This is due to the fact the lock-up process is trying to follow the tuning, and unless lock-up is extremely fast, it must have a period of time in which there is an out-of-lock condition.

A problem encountered in tunable synthesizer systems that is not seen in the regular synthesizer occurs whenever the VCO approaches the end of the band. Assume a DTVFO is designed to cover the frequency range of 5 to 6 MHz, with the counter hardwired to 5 MHz. The counter counts up to 5997, 5998, 5999, and suddenly resets.back to 5000 kHz. This can be somewhat upsetting to the operator trying to tune in a signal near the band edge. Fortunately, the problem is readily overcome by using a few inexpensive gates. For every frequency, the outputs of the No. 2 counter will have a unique combination of highs and lows. For example, when the counter reaches 9, the Q1 and Q4 outputs will be high. Therefore, if the two highs are sampled by a two-input NAND gate, its output may be used to prevent any further pulses from reaching the counter.

optical encoder

After having built and operated a number of DTVFO systems, I soon realized that something better than spring-loaded switches or pushbuttons was required if the tuning is to have the right feel. Using the button- or switch-activated square-wave oscillator, you either run right over the station or you stop short of it. Then it becomes necessary to "milk" the switch in the manner of an aircraft flap system, to creep up on the station. Then, in the middle of the milking process, the other person stops speaking! The potentiometer-controlled oscillator was better, but still lacked feel.

*Rohde & Schwartz, Munich, applied for a patent on the shaft-encoder, controlled-frequency synthesizer in May of 1961. Patent 977,780 was issued on July 13, 1962.

The next logical step called for the incorporation of the rotating disc tuning system.* In this system, a series of holes was punched around the perimeter of a 8.9-cm (3-112-inch) diameter disc. The disc was mounted on a flywheel, which in turn was connected to a 2.8:1 Jackson Brothers drive. The entire mechanism was then mounted on the receiver front panel and turned by an ordinary, large-diameter tuning knob. The flywheel not only gives the tuning a nice heavy fee!, but also allows the knob to be quickly spun.

Using the same principles as the commercial versions (light sources and photo transistors), this tuning system creates a series of pulses as the knob is rotated. The problem was then one of sensing direction of travel, or tuning the DTVFO up or down in frequency as the knob is turned one direction or the other. The answer turned out to be quite simple, and is shown in **fig. 7.** Initial conditions start with photo transistor A receiving light and delivering an output (see **fig. 7A**). As the disc is turned clockwise, **fig.** 7B, both transistors receive light and deliver output voltages. In **fig.** 7C, transistor A has lost its light, accompanied by the resultant decrease in output voltage. Thus if the disc is turned clockwise, transistor A always receives and loses its light ahead of B. However, if the disc is turned in the opposite direc-

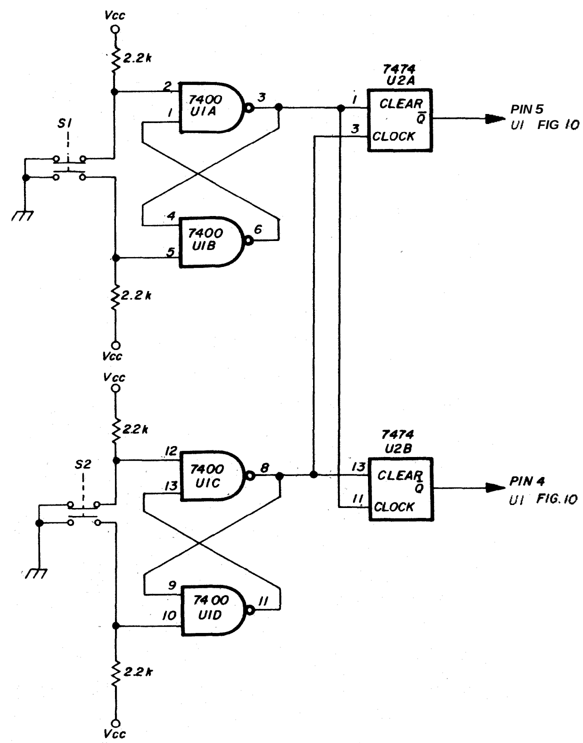


fig. 8. Pushbuttons have been used to simulate the pulses received from the photo transistors in the rotating disc, or optical encoder, system. The NAND gates are used to debounce the switches. The two 7474s determine in which direction the disc is rotating, thereby feeding the pulses to the appropriate input of the up/down counter.

tion, B will receive and lose its light ahead of A. Therefore, the voltage pulses from the photo transistors, in conjunction with the appropriate logic, can be used to control an UP/DOWN counter. By detecting which pulse occurs first, A or B, the second pulse can clock the counter, either up or down.

To prove the theory, I constructed the actual circuit, with pushbuttons used to simulate the pulses from the photo transistors (see **fig. 8**). As previously mentioned, it is necessary to use debouncing circuits to ensure that only one pulse is generated when the button is pressed; hence, the 7400 NAND gates with normally low outputs. U2A drives the count-up input of the 74192 counter, with U2B driving the count-

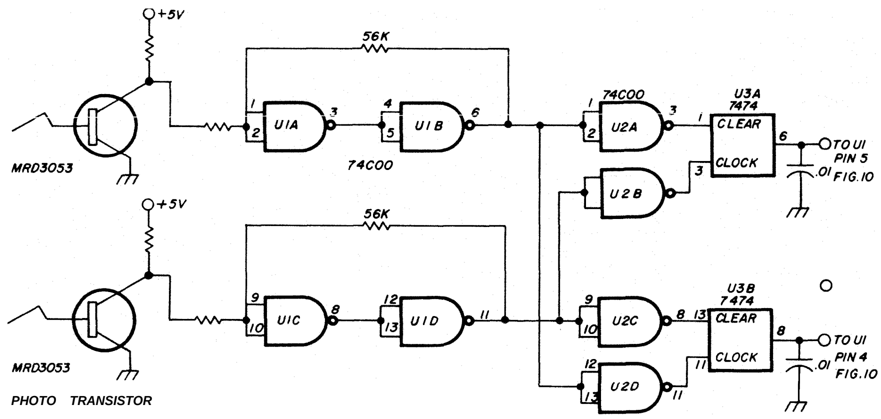


fig. 9. Final circuit for the homemade shaft encoder. Each time the light path is interrupted, a pulse is generated. This pulse causes the counter to increment or decrement by one count, depending upon the direction of disc rotation.

down terminal. Using the switches, I was able to simulate the sequence of events as the disc is turned.

1. S1 is pressed and held down. U2A is preset by the high level on pin 1.

2. S2 is pressed and held down. U2A is clocked, with pin 3 going high and pin *6* low.

3. S1 is released. Pin 1 of U2A goes low. Pin 6 of U2A goes high causing the 74192 counter to advance by one count. (The counter operates only on positive-going pulses.)

4. S2 is released. Pin 3 of U2A goes low.

If the buttons are pressed in the opposite order (S2 pressed and held down, S1 pressed and held down, S2 released and then S1 released), U2B will be preset and clocked. The output pulse applied to the 74192 will cause the counter to decrement one count.

After the pushbutton system was operating correctly, the circuit was modified to that shown in **fig. 9.** A small light bulb was positioned so as to shine through only one disc hole into the two photo transistors. The hole diameter and the spacing between the holes should be at least twice the diameter of one of the photo transistors. The transistors should receive an equal amount of light when the hole is exactly above the two. There must be no stray light, which causes the photo transistors to bias up into an intermediate state. When the light is shaded, the collector voltage should be close to +5 volts. When the photo transistor receives full light, its collector voltage should drop to less than 1 volt.

Just as the pushbuttons were followed by debouncing circuits, so must the photo transistors be followed by similar circuits. In this instance, the debouncing circuits are schmitt triggers fashioned from CMOS gates. CMOS was used to prevent loading of the photo transistors. Further gates were necessary to prevent the 7474 from loading the schmitt triggers and causing interaction between the inputs. Before the gates were added, the 7474 flip-flops were delivering a stepped-output waveform which was causing the counter to misfire. Since the 74192 counters are edge triggered, a stepped waveform edge can cause havoc. If the CMOS version of the 7474 is used, the buffering may be unnecessary.

practical DTVFO

A complete, tunable synthesizer with 100-Hz resolution is shown in **figs. 10** and **11;** although I earlier mentioned that a DTVFO readily allowed up-conversion operation, the unit to be described operates in the low-frequency range, 5 to 6 MHz. This frequency was chosen so that its output could be used to operate an ALDA 103 transceiver. The frequency and output level are both compatible with this particular transceiver. **Fig. 10** contains the two counter chains and readouts, with **fig. 11** showing the reference oscillator, dividers, phase detector, VCO, and buffers.

Referring to **fig. 10,** the synthesizer divider chain

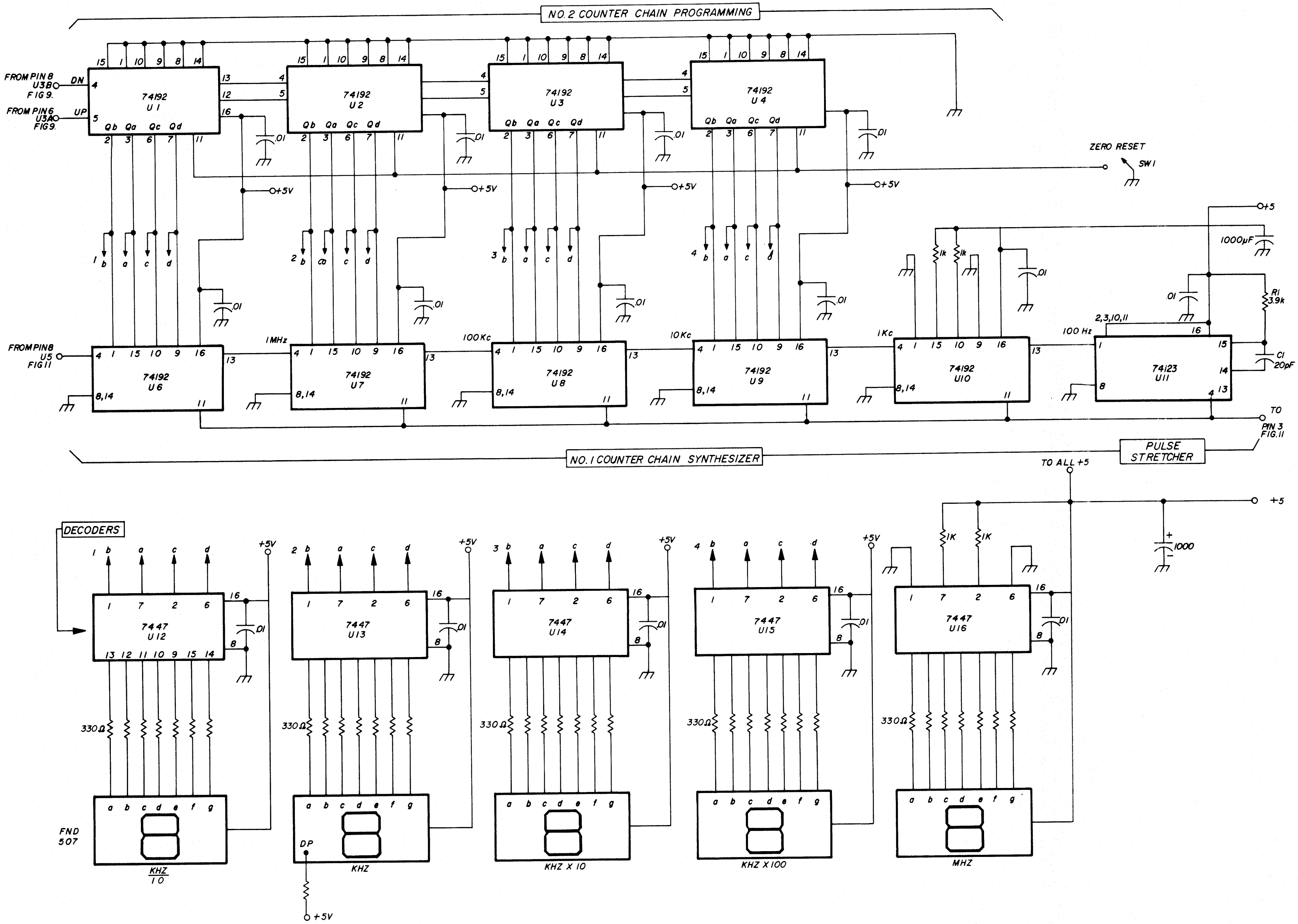


fig. 10. Schematic diagram of the counter chains used in the 5-MHz digitally tuned VFO. The up/down inputs for U1 come from the optical shaft encoder system described in the text. In addition to driving the load inputs of the first counter chain, the Q outputs also drive the BCD-to-7-segment converters. The load inputs for U9 have been hardwired since this counter always divides by five.

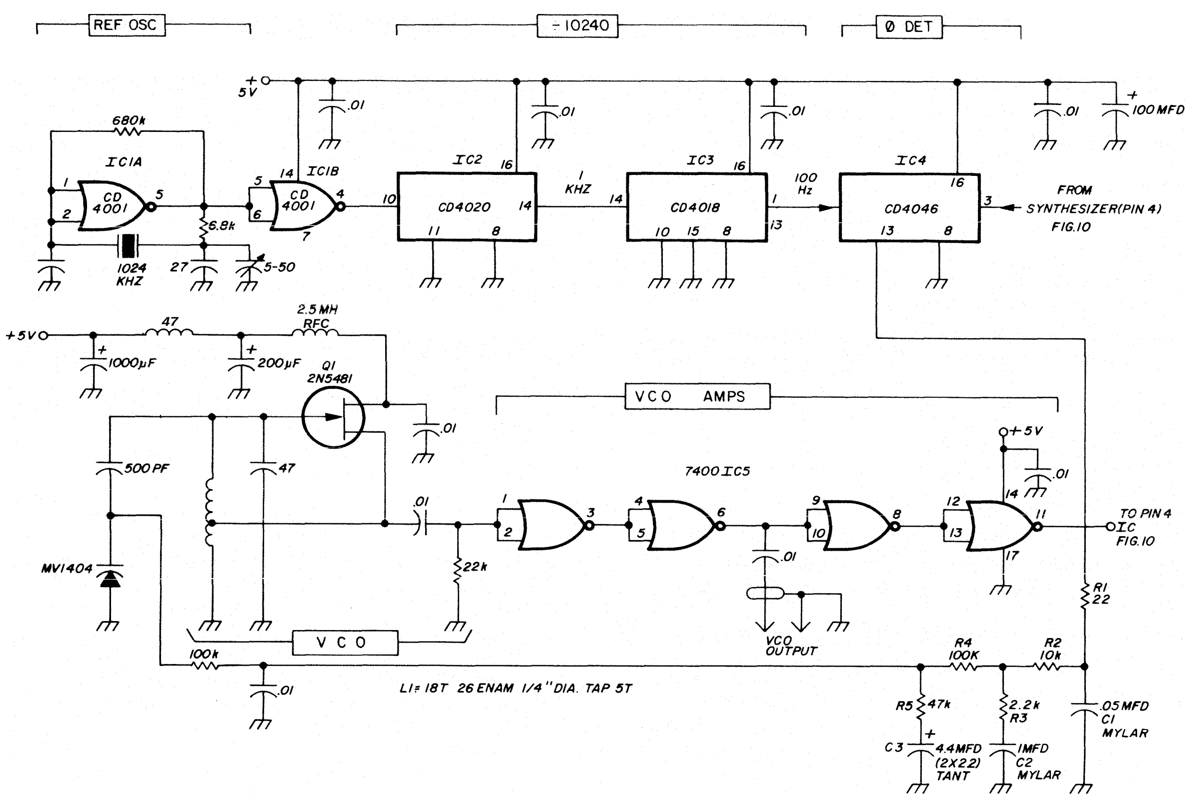


fig. 11. Diagram of the VCO and reference oscillator portion of the digitally tuned VFO. Using standard CMOS ICs, the oscillator is divided down to 100 Hz prior to being applied to the phase detector. Transistor Q1 is a Hartley oscillator which covers the 5 to 6 MHz range under control of the phase detector. L1 is 18 turns of no. 26 AWG (0.4 mm) wire wound on a 6.5-mm (114-inch) diameter form. The tap is located five turns from the bottom of the coil.

consists of U6 through U10. Pulse stretcher U11, in addition to supplying the reset pulse for the counters, also drives the phase comparator in **fig. 11.** U1 through U4 constitute the programming chain, with pulses from the optical encoder used to provide the count-up and count-down inputs. The Q outputs from the four respective counters are used to program the load inputs of the synthesizer divider chain and also to drive the BCD-to-7-segment converter.

U10 has been hardwired to divide by five so that the synthesizer output covers from 5 to 5.9999 MHz. Provided the VCO-tuned circuit is adjusted accordingly, U10 may be hardwired to cover other 1-MHz increments.

U16 may be dispensed with and the appropriate segment resistors grounded to obtain the digit 5, eliminating the 7447 BCD-to-7 segment decoder.

When first turned on, the counter outputs cause odd readout displays; momentarily grounding the load line (S1) will set all counters and displays to zero. Of course, this function could be made automatic or gated to the bandswitch so that each band is started at the low-frequency end.

In **fig. 11,** U1A is connected as a 1024-kHz crystal oscillator, followed by buffer U1B. U2 and U3 subsequently divide the reference oscillator output down to 100 Hz prior to application to pin 14 of the phase detector.

The resultant error voltage from the phase detector is filtered by the following RC network. While this filter does not represent the ultimate in filtering, it is satisfactory for normal listening. A Singer spectrum analyzer shows the sidebands well down into the noise. However, when the VCO is beat against a crystal oscillator to create a low-frequency audio beat note, a slight "burble" can be detected. As the DTVFO was an experimental project rather than a construction project, no further effort was made to improve the filtering.

I found in the breadboarded circuit that considerable sidebands were created not by poor lock-line filtering, but by lock-line pick up. The line was changed to coax and the sideband content dropped considerably.

The VCO tuning slug should be adjusted for the middle of the range to be covered. This is more readily accomplished if the lock line is disconnected from the phase detector and connected to a variable source of voltage. The voltage should be set to 2.5 volts and the slug set to the center of the frequency range to be covered, or in this case 5.5 MHz. When the lock line is reconnected and the circuit tuned to

5.5 MHz, the lock-line dc voltage should read close to 2.5 volts.

The VCO must be capable of swinging at least 20 per cent beyond the frequency range to be covered when the lock line is varied from 1 to 4 volts. Except for the photo transistor part of the encoder, the circuit is pretty well foolproof and, provided the correct connections are made, should work right off. Because my junk box yielded two different photo transistors, some difficulty was encountered obtaining a match from the two devices and one of them was somewhat temperature conscious. However, it was felt that proper devices would solve this problem so it was shelved until better devices were obtained.

When tuning ssb signals, the DTVFO will feel exactly like a conventional VFO. After using the DTVFO for a few days I was amazed at how sloppy my previous listening habits had been. I'm quite sure that I'd been listening to stations with the receiver 100 Hz or more off frequency. Consequently, the fact that the DTVFO comes in 100 Hz steps made no difference. On CW, however, as the signal is tuned, the steps are quite noticeable, but make little difference to the copy.

references

1. Ulrich L. Rohde, DJ2LR, "Optimum Design for High-Frequency Communication Receivers," *ham radio,* October, 1976, page 10.

2. Wayne Ryder, W6URH, "High-Performance General Coverage Communications Receiver," *ham radio,* November, 1977, page 10.

3. Don Lancaster, *TTL Cookbook,* Howard W. Sarns, Indianapolis, 1974.

bibliography

Allen, George R., W1HCI, "Synthesize Yourself," 73, October, 1977, page 182.

Gardner, Floyd M., *Phaselock Techniques,* John Wiley and Sons, Inc., *New* York, 1965.

Rohde, Ulrich L., DJ2LR, "Modern Design of Frequency Synthesizers," *ham radio,* July, 1976, page 10.

Phase-Locked Loop Data Book, Motorola Semiconductor Products, Phoenix, Arizona, 1973.

ham radio

printed-circuit layout using the longhand method

Electronic circuits require some method of connecting the parts together. This can be done by the printed-circuit board, which is made using photographic methods, or by point-to-point wiring. Obtaining the photo negative for the PC board is usually beyond most home workshop capabilities. I'd like to describe what I call the "longhand circuit layout" method. This method allows you to make PC boards without using a photo negative. With it, you can quickly lay out and build a PC board. My method differs from the photo approach in that *you* draw the circuit pattern on the copper-clad board.

materials

Fig. 1 shows the materials required for laying out a circuit. The graph paper should have 10 squares per 25.4 mm (10 squares per inch). This allows for IC pin spacing of 2.54 mm (0.1 inch). The copper-clad board can be cut to size by scoring a line on both sides and snapping it in two. Of the three types of etchant resist shown, I prefer the felt-tipped pen because it's easiest to use. The pen and etchant resist can be obtained from Radio Shack and other sources. (You might have to thin the resist to draw fine lines with it.)

By John A. Burton, WB9QZE, 2282 McKinley Ave., **Columbus,** Indiana 47201

procedure

The first step is to draw your parts onto the graph paper – the "longhand method." Do this by placing a dot for each end or pin of the part. For resistors I use 13-mm (0.5-inch) spacing between the dots. Then I draw the schematic symbol for the part between the dots. This helps keep track of the parts. If I

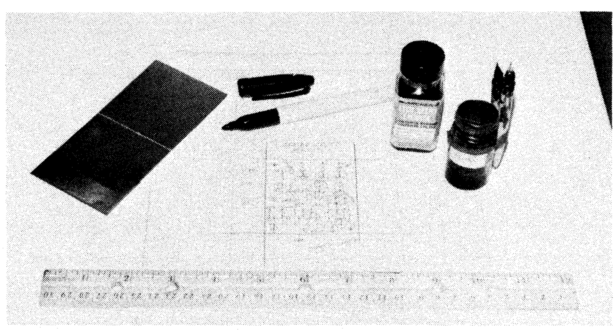


fig. 1. Materials needed for the "longhand layout method" for making PC boards. Shown are a scored copper-clad board, etchant, etchant resist, etchant-resist pen (preferred), and the circuit laid out on graph paper.

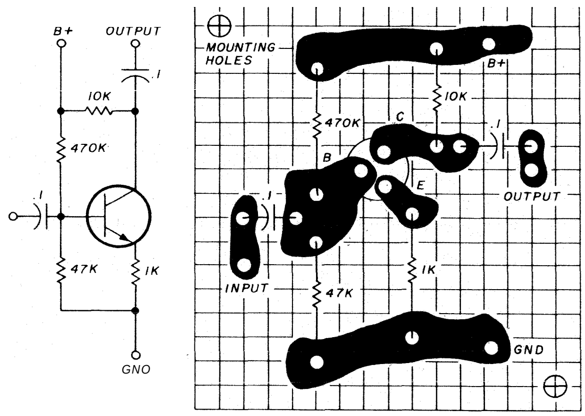
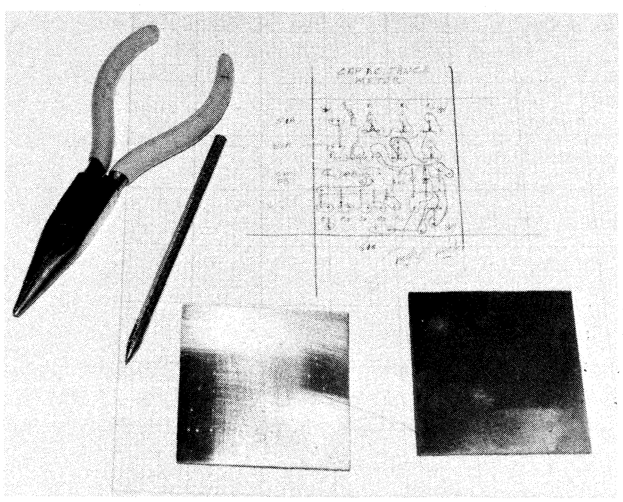


fig. 2. Typical circuit layout on graph paper, which has 10 squares per 25.4 mm (10 squares per inch). The copper-clad board should be taped to the back of the circuit layout so that the dots in thepattern can be punched onto the board.

have the room, I put the value of the part next to it, which helps me later when I insert the parts into the board.

Once the pattern is drawn onto the graph paper, the next step is to transfer it to the copper-clad board. With the copper-clad board beneath the graph paper, take a sharp-pointed tool and tap the dots on the graph paper. Important — the board must not slip while you're doing this. I prevent slippage by taping the board to the back of the graph paper.

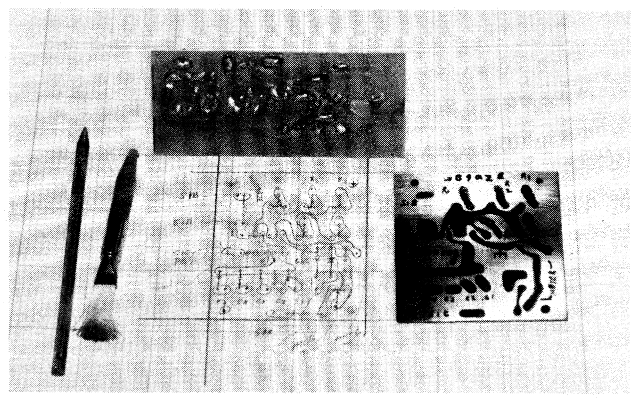


Tools and materials used in the "longhand layout method." The long nose pliers are used to tap the center punch. which transfers component-mounting holes from the graph paper to the copper-clad board. Note that one of the copper-clad boards (left)hasn't been thoroughly cleaned.

When all the dots have been punched, remove the board from the graph paper. With the etchant resist, draw the same pattern on the copper-clad board by connecting the correct dots together. Don't worry should you connect the wrong dots. If this happens, let the resist dry. It can be removed with a scribe or typewriter eraser.

The board can now be etched. When it has been etched, remove the resist by rubbing the board with steel wool or a household cleanser.

Now drill the holes, which will be easy since you



The typewriter eraser is used to remove unwanted etchant resist. At top is an example of a pattern on a copper-clad board using the "longhand layout method." A completed built-up PC board is shown at right.

have already centertapped each hole when you transferred the dots from the graph paper to the copper-clad board. I use a no. 60 drill, which is 1 mm, or about 0.04 inch.

You are now ready to insert your parts. Refer to your graph-paper pattern and put the parts into the correct holes on the board. That's all there is to it. Once you've used this method I think you'll agree it's a fast, simple way to make circuit boards.

A couple more things need mention: one is to clean the copper-clad board before applying the resist. If the board isn't free of dirt and oil, the resist will not stick to the copper. Also, the more information you can print on your pattern, the easier it will be to insert parts and wire to switches or any other part that's not on the board. So it's a good idea to plan your layout with this in mind.

I hope the photos and drawing will answer questions I may have overlooked in the text. I've been using this method for ten years and haven't found its equal.

ham radio

monolithic crystal filters

Introducing the MCF, a new and improved development in crystal-filter technology — a report on its applications in communications electronics

When high selectivity in electronic circuits is a must, mechanical filters, ceramic filters, or crystal filters are used. A new type of crystal filter is becoming popular: the monolithic crystal filter (MCF). It replaces the conventional crystal filter and leads to new applications for this new product, because the MCF shows comparable electrical features but is smaller and can be made more economically in large-scale production. What is a monolithic crystal filter, and what is the difference between it and conventional crystal filters?

conventional crystal filter

Conventional crystal filters consist of one to ten single quartz crystals coupled together by inductors, capacitors, and resistors in a distinct way to yield the filter characteristics required by the design specifications. As a typical example, fig. 1, shows the internal circuit of the well-known XF-9B bandpass filter of KVG* which is a standard filter for single-sideband applications. It's an eight-pole filter, which means it includes eight crystals. The bandwidth is ±1.2 kHz at −6 dB attenuation (related to the passband) at a center frequency of 9 MHz.

Each stage consists of tuned half-lattice bridges with one crystal in each branch. The stages are coupled through $C4$ and $C5$. For matching the filter impedance to 500 ohms, the input and output circuits have stacked windings. To compensate for strays, both resonate circuits are tuned to a higher frequency. With external trimming capacitors the filter can be aligned to the exact center frequency, which is also the point of minimum passband ripple.

monolithic two-pole (dual) filter

Instead of single crystals, the monolithic crystal filters use multiple resonators, which consist of several vibrating systems plated onto one common crystal blank. Typically, these vibrators are of the thickness shear type (AT cuts) with a frequency range from 5-30 MHz in the fundamental mode (piano-parallel crystal blanks). Most types of multiple resonators work in the fundamental mode, but the same principle can be applied to overtone resonators as well.

The simplest arrangement of multiple resonators is the monolithic two-pole filter, referred to as "dual." What is its difference compared with single crystals?

The equivalent electrical circuit of a single crystal is shown in fig. 2. It involves a high-Q series-resonant circuit consisting of the motional parameters $L1$, $C1$, $R1$ and the static capacitance, C_0, in parallel. The Q is greater than 50,000.[1,2]

The series-resonant frequency is

$$f_s = \frac{1}{2\pi \sqrt{L_1 C_1}} \qquad (1)$$

and parallel-resonant frequency above f_s is

$$f_p = \frac{1}{2\pi \sqrt{L_1 \left(\frac{C_0 C_1}{C_0 + C_1} \right)}} \qquad (2)$$

By Bernd Neubig, DK1AG, Westliche Ringstrasse 42, D-6921, Epfenbach |FRG|, Germany

*Kristallverarbeitung Neckarbischofsheirn, West Germany; available in the United States and Canada from Spectrum International, Post Office Box 1084, Concord, Massachusetts 01742.

The monolithic two-pole filter has two pairs of electrodes, which are plated onto one common crystal so that both resonators are mechanically coupled to each other by the crystal blank in a well-defined magnitude. This leads to the equivalent electrical circuit shown in **fig.** 3.

Both resonators consist of the motional quantities $L1$, $C1$, $R1$ and $L1'$, $C1'$, $R1'$. The static capacitances across each pair of electrodes are C_0 and C_0'. The four-pole in **fig.** 3 (dashed lines) involving C_k and the negative capacitances $-C_k$ — a so-called impedance inverter configurations — represents the mechanical coupling between both systems.4

The coupling factor, k is

$$k = \frac{C_1}{C_k} \tag{3}$$

and has an order of magnitude of 10^{-4} - 10^{-3} in fundamental-mode duals. Overtone duals have smaller coupling factors. The coupling factor of a particular dual is determined by the configuration of the electrodes, the gap between them, the crystallographic direction of coupling, the mass of the electrode plating, and the thickness of the quartz blank.[5]

The resonant frequencies of both resonator pairs are normally equal or very close to each other:

$$f_1' \approx f_1 = \frac{1}{2\pi\sqrt{L_1C_1}} \tag{4}$$

Because of the coupling effect these frequencies can't be measured directly. For example, between pin A and pin B, with pin C open **(fig. 3)**, you can measure a frequency f_1^* that's lower than that obtained with **eq. 4.***

Two new resonant frequencies occur, which are the characteristic vibration modes of monolithic dual resonators. These frequencies are of great significance for the application of duals as filter components:

1. When both resonator systems are connected in parallel, as shown in **fig. 4A,** series resonance appears at the so-called "symmetric frequency":

$$f_{sym} = f_1\sqrt{1-k} \tag{5}$$

In this circuit both systems vibrate in equal phase. This means that the mechanical displacement (of the thickness-shear motion of the crystal) takes place in the same direction in both systems.

2. Connecting both resonators as shown **in fig. 4B,** series resonance appears at the "antisymmetric frequency," f_{asym}, which is above f_{sym}:

$$f_{asym} = f_1\sqrt{1+k} \approx f_{sym}(1+k) \tag{6}$$

In this configuration both systems vibrate in phase opposition.

Additionally, both frequencies, f_{sym} and f_{asym}, can be measured between pins A and B with short circuited output pins (B and C). In this case, symmetric and antisymmetric frequencies are the frequencies of maximum input admittance of the four pole.[6]

The frequency difference between both characteristic frequencies — often called "mode spacing" — increases with higher coupling factor, k, as you can see in **eq. *6.***

monolithic multipole resonators

The principle of the monolithic dual resonator can be expanded and leads to monolithic multipole resonators with up to eight or ten resonator systems on the same crystal disc.

The mathematical synthesis of such vibrators is complex. Also measuring and production techniques are difficult. Each type of filter needs a certain configuration of the resonators. This restricts the feasibility of economically producing a large number of filter specifications in smaller quantities. Furthermore, with a larger number of resonators, the problem of spurious responses increases. This is why the multipole monolithic crystal filter hasn't become popular except for special applications such as channel filters.7 The tendency is to obtain multipole crystal filters by stacking several dual resonators, as explained below.

*More precisely, two frequencies exist:

$$f_{1,2}^* = f_1\sqrt{1+\frac{C_1}{2C_0} \pm \sqrt{k^2+\left(\frac{C_1}{2C_0}\right)^2}}$$

The lower frequency is **between** f_{sym} and f_{asym}, where f_{sym} and f_{asym} are the "symmetric frequency" and "antisymmetric frequency" as described in the following text.

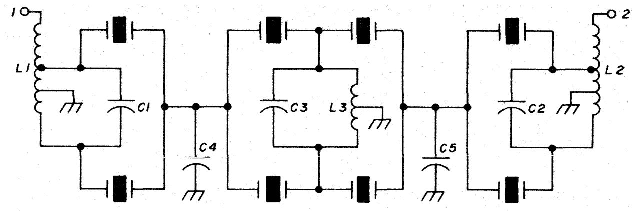


fig. 1. Internal circuit of KVG filter XF-9B.

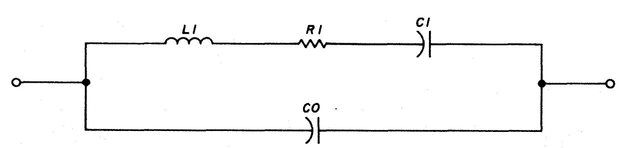


fig. 2. Equivalent electrical circuit of a crystal.

monolithic crystal filters

We now consider the MCF in its role as a practical addition to electronic circuits. Examples are given to show how the dual can be used in noise blankers, noise filters, and fm discriminators. We then examine the MCF in another practical application in which more than two poles can be synthesized by connecting several duals in series. First, some background on two-pole filters.

Two-pole filter characteristics. As shown by the theory of network synthesis and by the theorem of Bartlett,[8] the equivalent electrical circuit of a dual (fig. 3) can be transformed into an equivalent half-lattice bridge with a differential transformer, which has the same response of amplitude and phase *versus* frequency.[9]

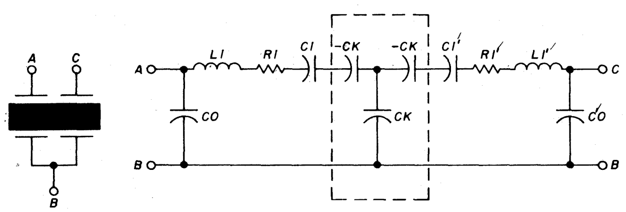


fig. 3. Configuration and equivalent electrical circuit of a dual resonator.

When a dual is terminated at its input and output with an impedance value, $Z_{in} = Z_{out}$, the elementary form of a two-pole crystal filter results. The identical selectivity curve is shown as the equivalent half-lattice, two-pole filter with single crystals terminated with the same impedances (fig. *5).*

Comparing the number of components in both filters shows clearly that a monolithic filter is much easier to construct: there is no need for a differential transformer and, instead of two crystals, only one crystal component is required.

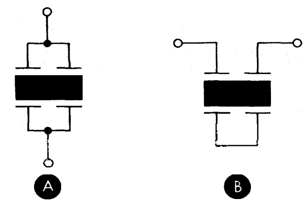
The bandpass response of such a filter depends on the magnitude of the termination resistors. The characteristic impedance is

$$R_0 = 2\pi L_1 \Delta f \qquad (7)$$

where L_1 is the motional inductance of one resonator (see fig. 3), and Δf is one-half the filter bandwidth. Typical duals have characteristic impedances of several kilohms at, for example, 10.7 MHz.

*The classical wave-parameter theory yields the following example: For $\frac{R}{R_0} = 0.8$, the passband ripple is about 0.1 dB, while the bandwidth at attenuation is at the three-fold bandwidth a -3 dB. With $\frac{R}{R_n} = 0.5$, the ripple increases to 0.22 dB, while the -20-dB bandwidth is only 2.8 times the -3-dB bandwidth.

fig. 4. Pin configurations for measuring symmetric (A) and antisymmetric (B) frequencies.



Reasonable bandpass filter responses (with curves having a near-rectangular passband characteristic) can be achieved with terminating impedances of $R = Z_{in} = Z_{out}$ smaller than R_0. The smaller the R value, the higher the passband ripple, but the skirts of the filter characteristic will be steeper.*

By proper selection of the termination impedances and characteristic frequencies of the dual, every filter response known from the filter theory of effective parameters (*i.e.*, Chebyshev, Gaussian, Bessel, Legendre)[9] can be synthesized.

The skirts of this filter can be made steeper by introducing a coupling capacitor, C_A between both resonators. This capacitance produces an attenuation peak at both sides of the passband. But at the same time the stopband attenuation decreases, as shown in fig. 6, for several values of C_A (as multiples of C_0).[10]

To fulfill the total selectivity demands of a special device (*e.g.*, a receiver), such a two-pole filter will surely not be sufficient. Despite this disadvantage, there are some interesting application examples for this simple filter component.

Dual as an i-f preselector in noise blankers. Noise blankers are designed to blank out short-duration noise pulses with high amplitude, especially in shortwave or mobile receivers. The best point at which to insert a noise blanker in a receiver is ahead of the i-f stages before the crystal filter, which provides the main selectivity. This is because narrow-bandwidth filters produce ringing, which distorts the signal.[11]

The block diagram of a typical noise blanker is shown in fig. *7.* It includes two alternatives to obtain the noise information. Version A derives it from the i-f signal; version B obtains it from a separate noise receiver tuned to a "silent" frequency.[12]

Following the mixer a broadband i-f filter must be

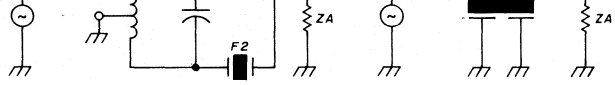


fig. 5. Two-pole crystal filter with differential transformer and equivalent monolithic crystal filter.

inserted, one narrow enough to cut off strong out-of-band signals but broad enough to avoid ringing. A monolithic dual is a unique device for this i-f preselector, as it can be inserted directly with pure ohmic termination and without any alignment effort – if the dual spurious responses are negligible.

For the well-known 9-MHz crystal filter line, KVG offers its dual type XF-912 for a bandwidth of ±7.5 kHz (at –3 dB). It is housed in a 3-pin HC-18 case and needs terminating resistances of 4.0 kilohms for a Chebyshev response. Different types with other bandwidths are available on request.

Dual as a noise filter. Normally, the main selectivity of commercial receiver i-f strips is produced by a high-performance (*e.g.*, eight-pole-type crystal) filter ahead of the i-f amplifier stages. But usually the following broadband high-gain amplifier stages generate broadband noise, which reaches the second detector in full magnitude. This noise, which can be very inconvenient, especially at extremely narrow bandwidths (as with the CW-filter type XF-9 NB from KVG), can be reduced drastically by a simple filter directly ahead of the demodulator stage. This is one more typical application for a monolithic dual, which can be easily inserted without any adjustments. For example, the type XF-912 can be used again. Similar duals exist also for other i-fs such as 10.7 MHz or 21.4 MHz.

Dual as an fm discriminator. IC quadrature detectors are frequently used to demodulate fm signals. The principle on which these demodulators work is as follows.

The frequency-modulated i-f signal is applied to one port of an AND gate; the other port is connected to a phase-delayed portion of the same signal. The phase delay depends on the frequency. The gate output yields the demodulated audio-frequency signal if followed by an integrating RC lowpass filter.

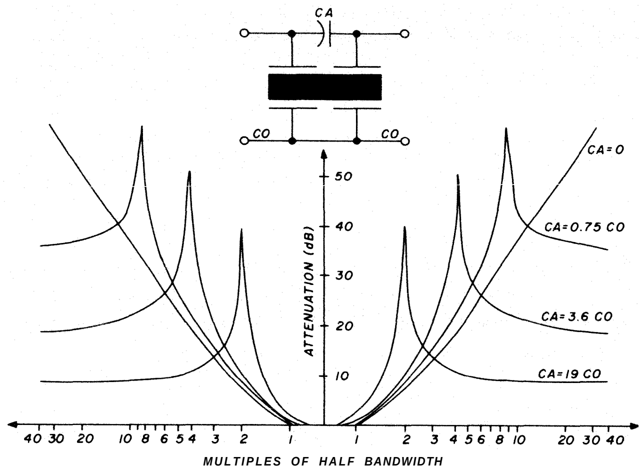


fig. 6. Frequency response of a dual with attenuation poles achieved by an additional capacitor, C_A.

The phase-delay circuit is a parallel-resonant circuit, which is coupled through a choke or a small capacitor, as shown in **fig. 8A.** This circuit provides a phase delay that increases or decreases linearly with frequency changes in the vicinity of the resonant frequency.[13] Higher slopes of the phase vs frequency curves, and thereby higher recovered audio, can be realized by using a bandpass filter with a coupling coefficient of about $Q \cdot k = 0.7$. This is shown in **fig. 8B.**[14]

The bandpass-filter can be substituted with a conventional dual resonator, which is designed as a Bessel-function filter with linear phase characteristics. The result is excellent demodulator linearity as

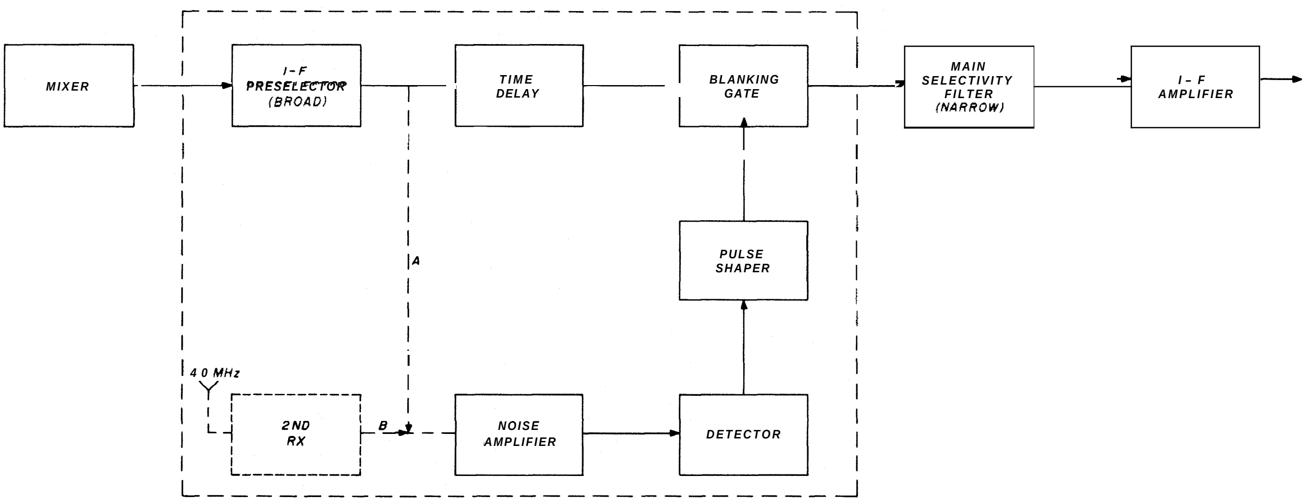


fig. 7. Block diagram for a typical noise blanker. Version A: noise information derived from i-f. Version B: noise information derived from separate noise receiver.

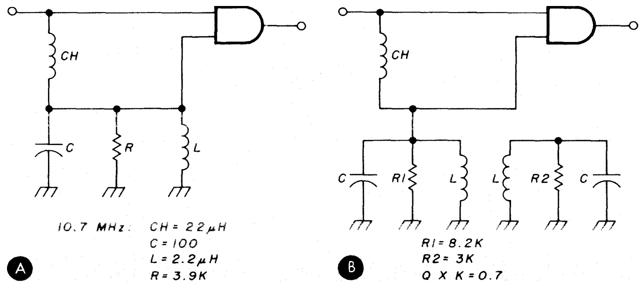


fig. 8. Schematics of f-m quadrature detectors – (A) with single resonant circuit; (B) with bandpass filter.

well as high audio yield. The demodulator response is the typical S-shaped characteristic. The two peaks are approximately at the symmetric and the antisymmetric frequencies.

Fig. 9 shows a circuit working on this principle. It consists of the RCA integrated circuit CA3089E. Both terminating resistors, $R1$ and $R2$, are chosen for best phase linearity (*i.e.*, constant group delay) in the passband between the two peaks. Their values depend on the motional parameters, L_1C_1, and the mode spacing of the dual.

This principle can be generalized for other quadrature detector ICs such as ULN2113A (Sprague), TAA661 (Signetics), or TBA120S (Siemens).

multipole filters with n>2

Monolithic crystal filters with more than two poles can be synthesized by connecting several dual resonators in series whereby they are coupled to each other by capacitors to ground (*i.e.*, the common electrode). As an example, **fig.** 10 shows the internal circuit of the KVG monolithic filter XFM-107B (bandwidth ±7.5 kHz at 10.7 MHz). Input and output are terminated with tuned circuits, which transform the filter impedance to a standard value of 910 ohms (with $C_{ext} = 25pF$ in parallel).

As with single duals, such composite filter structures can be terminated directly with a pure ohmic resistance given by the filter synthesis. In this case both tuned circuits can be omitted.

Furthermore, with increasing bandwidth up to about 1 per cent (*i.e.*, 1 part in 100) of the center frequency, the coupling capacitors become so small that they are realized by the static input and output capacitances of the coupled duals plus stray capacitances alone. Then the simplest structure of a monolithic crystal filter can be achieved. It consists only of

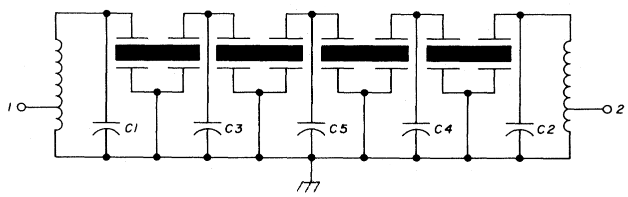


fig. 10. Internal circuit of KVG-Monolithic filter XFM-107B.

a chain of directly coupled duals, as you can see in **fig.** 11 for the KVG monolithic filter XFM-107S03 (10.7 MHz bandwidth ±10.6 kHz).

Comparing this circuit with that of a discrete crystal filter as in **fig.** 1, the simplification is evident. Further, the half-lattice filters with more than five crystals usually need a third tuned circuit (see L_3, C_3 in **fig. 1**), which gives additional insertion losses that can't be compensated for by the termination. This isn't necessary with monolithic filters because they present much smaller amounts of insertion loss than conventional crystal filters.

The electrical properties of monolithic crystal filters are equivalent to those of crystal filters with discrete components. That's why you can realize all known filter responses in monolithic structures, but with an upper limit for the bandwidth. Generally, filters with monolithic crystals vibrating in the nth overtone mode can be obtained with relative bandwidths smaller by the factor $\frac{1}{n^2}$.

The theoretical filter curves for ideal (lossless) filters – that means, the responses of attenuation and phase vs. frequency – are cataloged in normalized representation in the literature on the subject.[9,16]

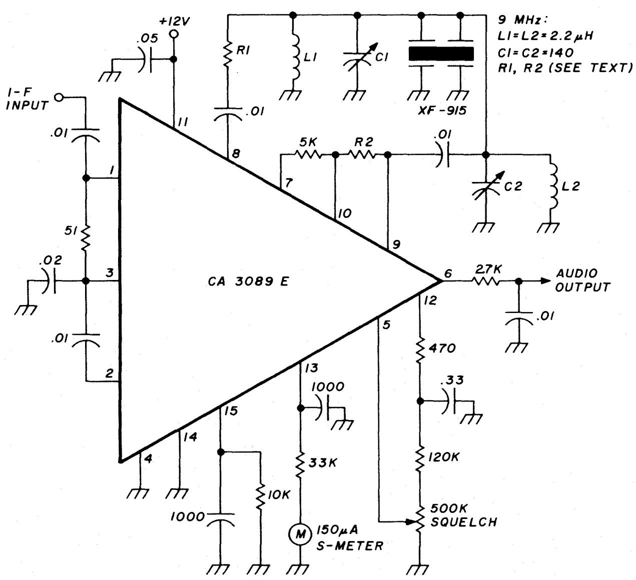


fig. 9. F-m demodulator with dual as discriminator.

summary

Monolithic crystal filters (MCFs) stand for the sim-

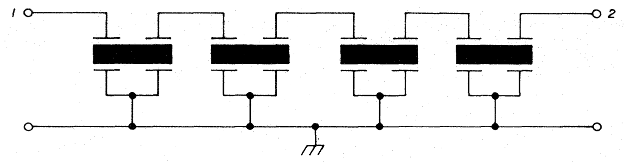


fig. 11. Internal circuit of KVG-Monolithic filter XFM-107S03.

plification of the structure of multipole crystal filters. Instead of discrete crystals, MCFs contain multiple crystal systems on a common quartz blank, which are mechanically coupled between each other by the quartz blank.

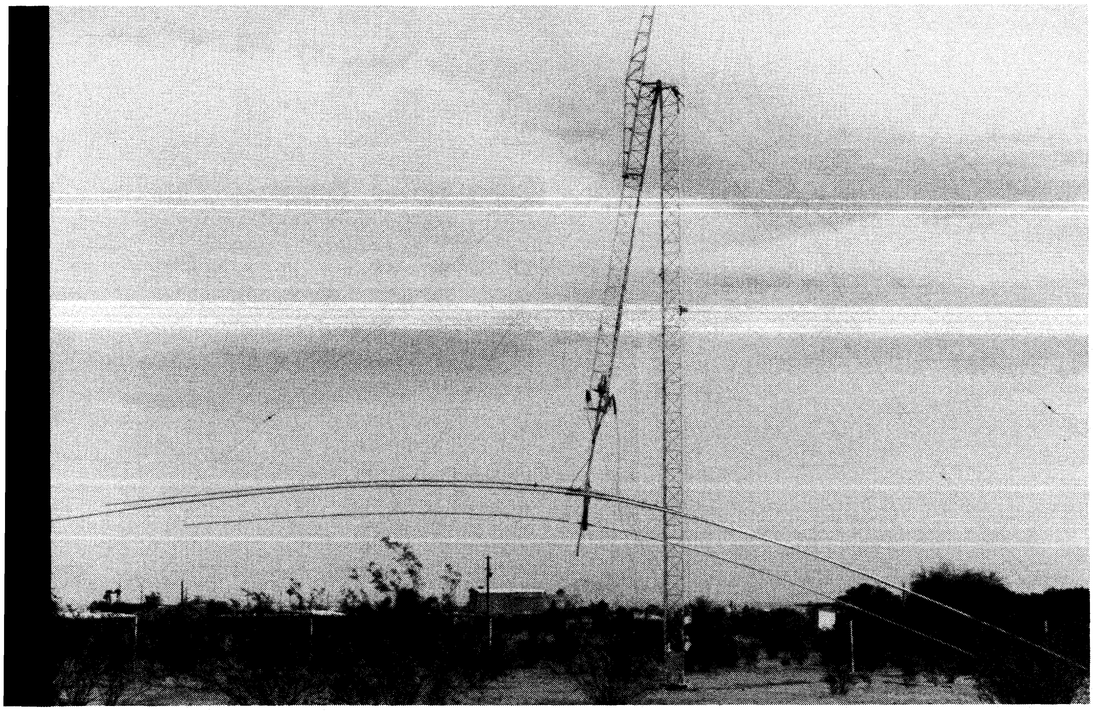
The typical application is in the frequency range of AT-cut crystals, most in the fundamental mode, but with increasing importance also as third or higher overtone monoliths. However, bandwidth is somewhat reduced in overtone applications.

The most usual forms of MCFs are dual resonators and series configurations of duals-to-high-pole filters. Beyond this, there are new interesting applications for duals as i-f preselectors, simple noise filters, or f-m demodulators.

references

1. Raymond A. Heising, Quartz Crystals for Electrical Circuits, D. Van Nostrand & Company, Inc., New York, 1946.
2. Virgil E. Bottom, The Theory and Design of Quartz Crystal Units, V. E. Bottom, Abilene, Texas, 1974.
3. S. B. Cohn, "Direct-coupled Resonator Filters," Proc. IRE, Vol. 45, February, 1957, page 189.
4. W. L. Smith, "The Application of Piezo-electric Coupled Resonator Devices to Communication Systems," Proc. of the *22nd* Annual Symposium on Frequency Control (1968), page 206.
5. W. D. Beaver, "Theory and Design of the Monolithic Crystal Filter," Proc. of the *21st* Annual Symposium on Frequency Control (1967), page 179.
6. G. Kohlbacher, "Vorschlag zur Messung der Ersatzgroessen von zweikreisigen monolithischen mechanischen Filtern fuer 5-100 MHz. Bericht Telefunken Fl, Ulm, Germany.
7. S. H. Olster et *al*, "A 6 Monolithic Crystal Filter Design for Manufacture and Device Quality," Proc. 29th Annual Symposium on Frequency Control (19751, page 105.
8. W. Herzog, "Siebschaltungen mit Schwingkristallen," Wiesbaden, 1949.
9. A. I. Zverev, Handbook of Filter Synthesis, John Wiley & Sons, New York, 1967.
10. R. A. Sykes and W. D. Beaver, "High Frequency Monolithic Crystal Filters with possible Application to Single Frequency and Single Sideband Use," Proc. 20th Annual Symposium on Frequency Control (19661, page 288.
11. Wes Stewart, K7CVT, "Noise Blanker Design," ham radio, November, 1977, page 26.
12. G. Laufs, DL6HA, Amateur SSB Technik, Franzis-Verlag, Stuttgart. Germany, 1965.
13. Otto Limann, "Der 'Ratio' bekommt Konkurrenz," third part, Funkschau 1411970. page 467.
14. L. S. Baar, "Application of the CA3089E FM-IF Subsystem," RCA *Application* Note ICAN-6257.
15. H. Olson, W6GXN, "FM-Detectors," ham radio, June, 1977, page 22.
16. R. Saal, Der *Entwurf* von Filtern *mit* Hilfe des Katalogs *normierter* Tiefpaesse, Frequenz, Vol. 15 (1961), page 111.

ham radio



rotary beam antenna
for 40 meters

Recipe for a really big signal on 7 MHz: a 3-element Yagi beam on a foldover tower

A full-sized, three-element 40-meter rotary beam has been a dream of mine for many years. They are scarce in the Midwest, where I operated as W9ERU for many years. I personally, at one time, got as close as the Hy-Gain DB24, a shortened two elements on 40, and three elements on 20. When I moved to the Phoenix area, I initially tried a conical monopole. However, not being satisfied with the results, I finally put up the 40 meter only version of the DB24. This antenna had two folded elements, each about 13.5 meters (44 feet) long, on a 4.9 meter (16 feet) boom.

tower design considerations

I mounted the two-element short beam, and later the full-sized antenna, on a 22-meter (72-feet) Rohn model 45 foldover tower. Legs on this tower are about 46 cm (18 inches) apart. I decided to mount the hinge between the fourth and fifth sections above the ground, giving three sections above the hinge. When the tower was folded over, this would allow the antenna to hang about 4 meters (13.3 feet) above the ground. In addition, it appeared that there would be enough clearance to allow the ends of the elements to pass the ground in such an operation. And, the portion above the hinge, would only be about 9 meters (30 feet) long, reducing the strain on the hoisting mechanism. Guy wires, two sets of four, were attached to anchors about 15 meters (50 feet) out from the base, and about 21 meters (70 feet) apart in a square configuration. See fig. 1. These dimensions were chosen to allow the use of a 12-meter (40-foot) boom at the top and still get it to pass through two of the lower guy wires.

construction

There are many parts to a beam antenna installation such as mine. I've provided photos showing

By Gene Hubbell, W7DI, 6633 East Palo Verde Lane, Scottsdale, Arizona 85253

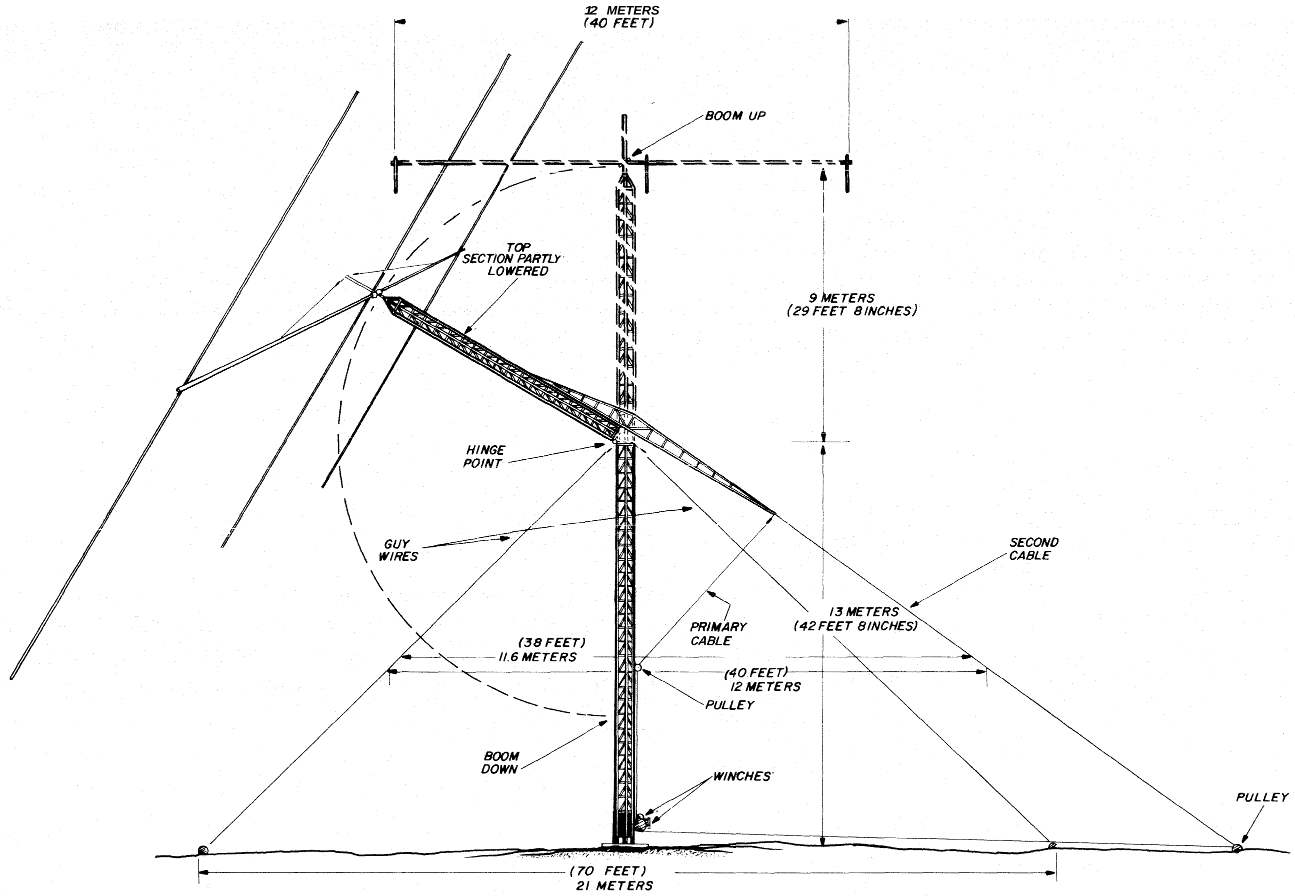


fig. 1. Overall dimensions of the 40-meter, three-element beam. Antenna is mounted on a Rohn model 45 foldover tower 22 meters (72 feet) high. Two sets of guys are used and two winches assure safe operation.

essential hardware and some views showing the antenna during and after installation.

Tower and rotator hardware. The tower lowering and raising mechanism is extremely important. A runaway when raising or lowering the antenna would result in a disaster. So I used two winches and cables. One, supplied by Rohn, was hand operated and located at the tower base. Its cable is passed through a pulley part way up the tower and to the tailpiece or tower boom, which went up as the top end of the tower came down. A second cable from a motor-driven winch at the tower base went out to a pulley some 18 meters (60 feet) from the base and up to the same tailpiece. Thus, pull could be exerted from two directions (see **fig. 1**).

This article is sprinkled with many metric conversions, which tend to disrupt continuity and ease of reading. But there's a good reason: we must face the fact that the outmoded and cumbersome English system of measurements is rapidly becoming extinct in the technical literature. So to help in the transition, author Hubbell's article has been edited to show first the metric dimensions, followed by English equivalents in parentheses. We apologize for this slight inconvenience. Sometime in the not-too-distant future we'll discard all references in our articles to the English system of measurements. We've graduated from tubes to solid-state devices without too much trouble. Let's progress further with the metric system, a totally logical and convenient method of defining measurements.
Editor.

The prop-pitch rotator and selsyn indicator chain-drive system were conventional. I welded a bell housing to the prop-pitch motor top gear and connected the housing to a piece of 41-mm (1.61-inch) pipe*, which passes through the top of the tower and is held to the boom saddle on top. This pipe is actually in two pieces, joined by a larger diameter sleeve and bolted through the drive pipe and the sleeve. It is not a one-piece shaft. A bicycle chain-and-sprocket system drives a 115-volt ac selsyn, which was offset from the main rotator pipe drive. It also drives another similar selsyn in the tower base. A limit-switch system allows the rotating pipe to turn about 450 degrees.

*The pipe used was called "Inch and a half" water pipe. In actuality, the pipe has a standard inside diameter of 4 cm (1.61 inches) and an outside diameter of 4.8 cm (1.9 inches).

Boom construction. The boom was made of two 6-meter (20-foot) lengths of 102-mm (4-inch) diameter irrigation tubing. To couple these tubes required a sleeve. I found only 0.4 mm (26 gauge) galvanized iron readily available, so two pieces 0.6 meter (2 feet) long and about 0.3 meter (12.5 inches) wide were rolled into tubes, one inside the other, and filed down until this double-walled sleeve just fit inside the 102-mm (4-inch) tubing, 0.3 meter (12 inches) each way. All three pieces were drilled for several rows of blind rivets, which made a strong joint. Now I had a boom 12 meters (40 feet) long (see **fig. 1**).

At the junction of the two boom halves, a hole about 33.5 mm (1-5116 inches) in diameter was drilled through the boom to pass a piece of "one inch" pipe.

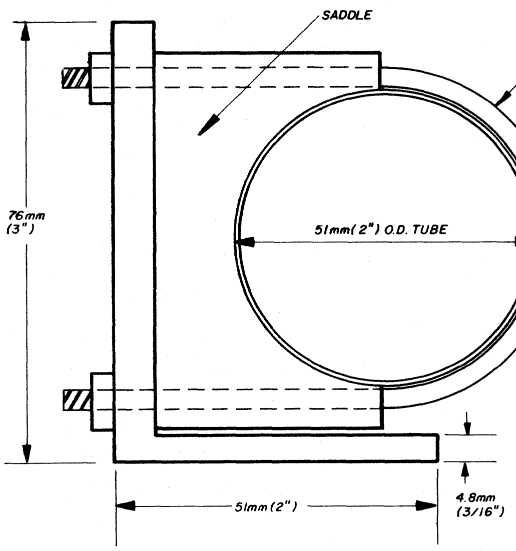
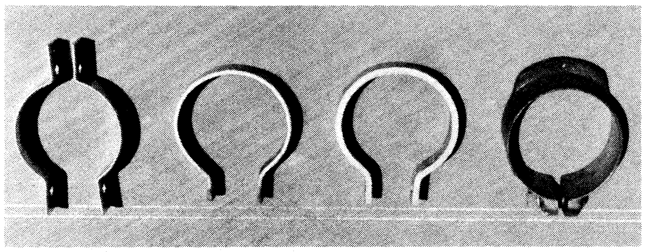


fig. 2. Details of boom-to-element clamps. The clamps are known as "muffler clamps" and were obtained from an automotive supply store.

This pipe is used as the vertical support for a truss system to take up the vertical strain on the boom.

Boom mounting. I made a U-shaped saddle of 6-mm (0.25-inch) thick steel with two sides 76 by 508 mm (3 by 20 inches) and a bottom 102 by 356 mm (4 by 14 inches). A 41-mm (1.61-inch) pin, 152 mm (6 inches) long, was centered in this piece (see photo). All items were welded together and holes were drilled through the sides of this saddle 25.5 mm (1 inch) from each end and 51 mm (2 inches) above the inside bottom piece. This allowed for 12.5-mm (0.5-inch) bolts to be passed through the saddle and boom. The bolts were 45.7 cm (18 inches) apart. Enough clearance was made to allow the boom to be tilted on a single bolt if desired.

A 16-mm (0.625-inch) hole was drilled axially into the 41-mm (1.61-inch) pin. The hole was 76 mm (3 inches) deep. This hole was for another pin large



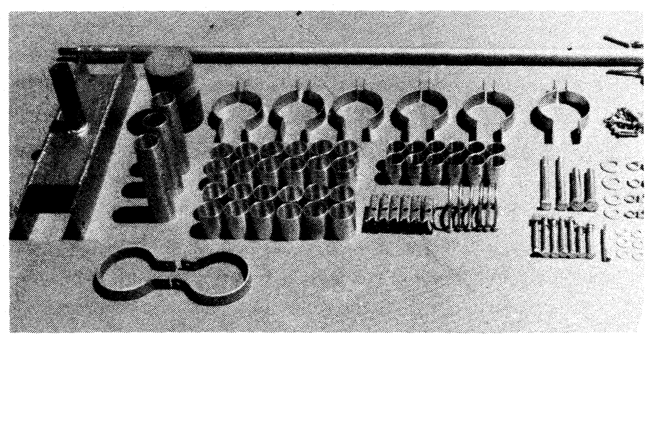
The boom clamps. The two-piece steel clamps, left, were used in the final version.

enough to fit inside a 27-mm (1.049-inch) diameter aluminum pipe turned down to 16 mm (0.625 inch), for 76 mm (3 inches) projecting from the pipe, which was about 2.5 meters (8 feet) overall. When saddle, boom, and this guying pipe were assembled, a guying system took the strain 3 meters (10 feet) out from the center of the boom on each side, up to the top of the 2.5-meter (8-foot) vertical guy support. Turnbuckles allowed adjustment.

Boom-to-element mounts. A horizontal aluminum angle holds the center of the element. This angle is 81 cm (32 inches) long, made of 51 by 76 by 4.8 mm (2 by 3 by 0.187 inch) material. Two bolts near the center hold the element to the 51-mm (2-inch) face, and two muffler clamps near the outer ends of the angle hold the element to the 76-mm (3-inch) face **(fig. *2*).** These muffler clamps were marked "1-718 inch" but fit 51-mm (2-inch) tubing perfectly. They were bought at an automotive parts supply store.

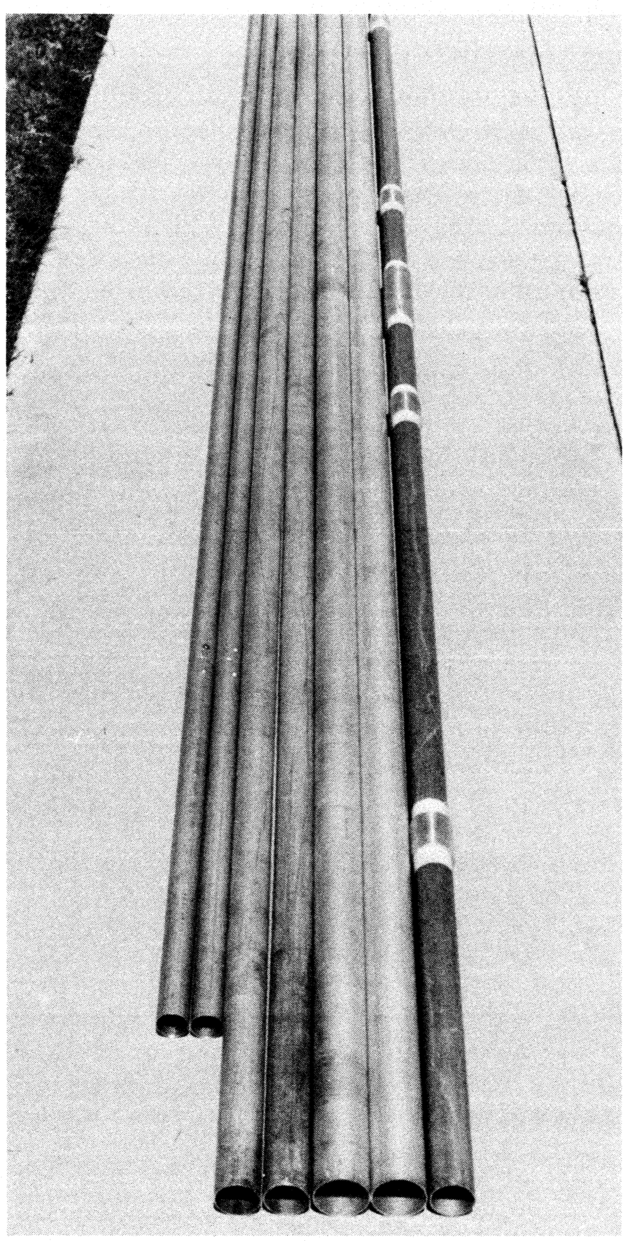
These angles are mounted crosswise with respect to the boom. They are bolted to two 30-cm (12-inch) lengths of 38 by 38 by 4.8 mm (1.5 by 1.5 by 0.188 inch) aluminum angle, positioned 41 mm (1.63 inches) apart (see photo). These 30-cm (12-inch) lengths were in turn bolted to two clamps, which passed around the boom. Two bolts hold the 38-mm (1.5-inch) angles to the clamps. All three angles are secured by a triangular gusset made of aluminum 3-mm (0.125-inch) thick, 45 by 45 by 63 cm (18 by 18 by 25 inches) with its corners cut off about 25 mm (1 inch) from the ends on the long dimension.

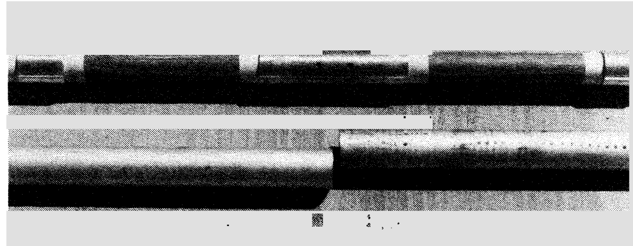
Hardware including boom saddle, boom guy, sleeves, clamps, and other items described in the text.



The t clamps changed over th years. In Bill 's *eam Antenna Handbook* (1955 , clamps were shown made of cast aluminur d to ld irr tio tubing together. When some of e broke I replaced them with clamps ına from 25.5 by 6.5 m m (1 y 0.25 inch) uminu bar bought at a hardware stor When these broke I had others I at a blacksmith shop fr 25.5 × 3 mm (1 by 0.125 inch) steel. All had a flaw. The degree of tightening needed on the clamp made a difference in the spacing of the aluminum angles mentioned above. The two-piece clamp shown at the left-hand end of the four in the photo solved the problem by allowing the adjustments to be made independently of each other.

All parts for the driven element (seven pieces of tubing). Light-colored bands are tape for positioning the sleeves.





Element center showing reinforcing tubing.

Antenna elements. Now I could mount elements to the boom, but I needed the elements. On the first try I used two 9-meter (30-foot) lengths of 51-mm (2-inch) irrigation tubing joined at the center by a 152-mm (6-inch) sleeve of aluminum pipe turned to fit inside the irrigation tubing. Short extensions on the ends brought the element lengths to the design figures of 19.8 meters, 20.3 meters, and 21.3 meters (64.8, 66.6, and 69.7 feet) for the director, driven element, and reflector respectively. I arrived at these numbers for the design frequency by the formula:

140.2/7.1 meters or	*460/7.1 feet*	*(director length)*
144.2/7.1 meters	*473/7.1 feet*	*(driven-element length)*
150 9/7.1 meters	*495/7.1 feet*	*(reflector length)*

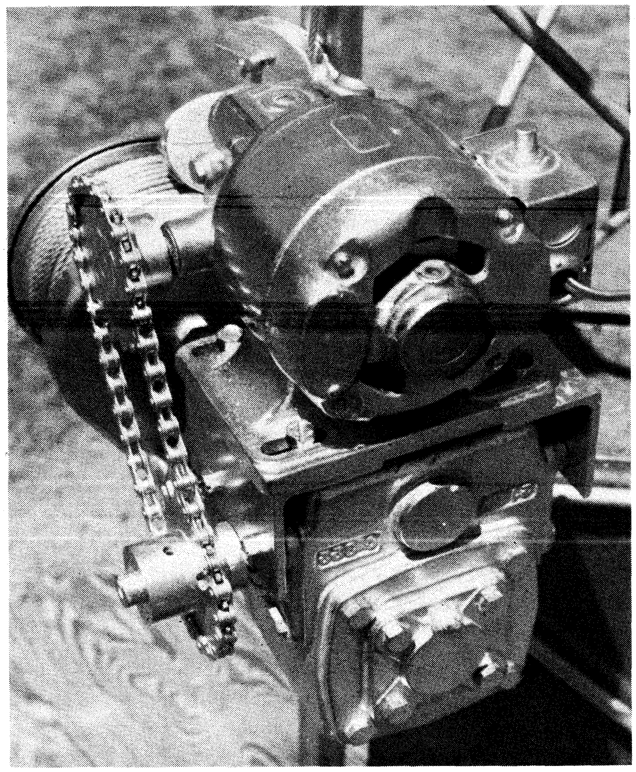
Results were unsatisfactory. Besides being plagued by an intermittent short circuit, which I eventually found in my gamma match, a windstorm soon bent the soft irrigation tubing. Back to the drawing board.

Through a local office of a tubing-supply firm, I bought six lengths of 6061T6 duraluminum 3.7 meters (12 feet) long, 51 mm (2 inch) diameter, with 1.2 mm (0.049 inch) wall. I also got six lengths of 3.7 meters by 41.3 mm by 1.2 mm (12 feet by 1.63 inches by 0.049 inch) tubing, and, from my stock of used tubing, six 3.7-meter (12-foot) lengths of 32-mm (1.25-inch) material with 1.7-mm (0.065-inch) wall — thicker than needed, but I had it on hand.

By overlapping the tubing about 30 cm (1 foot) at the joints between the 51-mm, 41.3-mm, and 32-mm (2, 1.63, and 1.25-inch) tubes, I had 10.4-meter (34-foot) lengths, and two of these gave me elements 21 meters (68 feet) long. The butt joint in the center (two 51-mm, or 2-inch, tubes) and the overlapping joints of different size tubing were all joined by sleeves turned from short lengths of aluminum pipe nominally 48 mm (1.9 inches) OD and from aluminum pipe couplings that fit inside the 41.3-mm (1.63-inch) tubing over the 32-mm (1.25-inch) tubing.

Additional sleeves were inserted inside the 51-mm (2-inch) tubes for support where the muffler clamps were placed. All overlapping joints and sleeves were held in place by steel 10-32 (M5) bolts held with lock-washers and nuts.

I did some more reading on element lengths for



Closeup of the homebrew slow-speed winch.

close-spaced, three-element yagis, and changed element lengths according to:

138 7/7 080 meters or	*455/7 080 feet*	*(dzrector)*
144 8/7 080 meters	*475/7 080 feet*	*(driven-element)*
152.4/7.080 meters	*500/7.080 feet*	*(reflector)*

This gave me new element lengths of 19.6, 20.5, and 21.5 meters (64.3, 67.1, and 70.6 feet).

I had written Bill Orr, W6SAI, for advice on the element lengths and he suggested that tapered elements might give a higher resonant frequency than the design figures indicated. The change was slight: from 7080 kHz to about 7120 kHz.

Matching section. Bill also strongly recommended that I use a small-diameter gamma rod, "as large diameter gamma rods detune the driven element." So I used triple-stranded 2-mm (12-AWG) copperweld guywire. The gamma rod terminated in a 127- by 152- by 229-mm (5- by 6- by 9-inch) aluminum box mounted on the gusset plate of the driven element-to-boom mounting. Inside was an air-variable capacitor for the gamma match, about 35 to 200 pF, driven by a reversible motor with a 4-rpm gearing. An added fixed capacitance of 50, 100, or 150 pF could also be switched in parallel with the variable capacitor.

The other end of the gamma rod was grounded to the driven element by an aluminum arm clamped to the gamma rod and the driven element. To get a good contact to the three strands of 2-mm (12-AWG) wire, the strands were soldered together, and a 19-mm (0.75-inch) OD aluminum cylinder, with a hole to fit snugly over the three strands, was placed over the soldered area and fastened with three set screws.

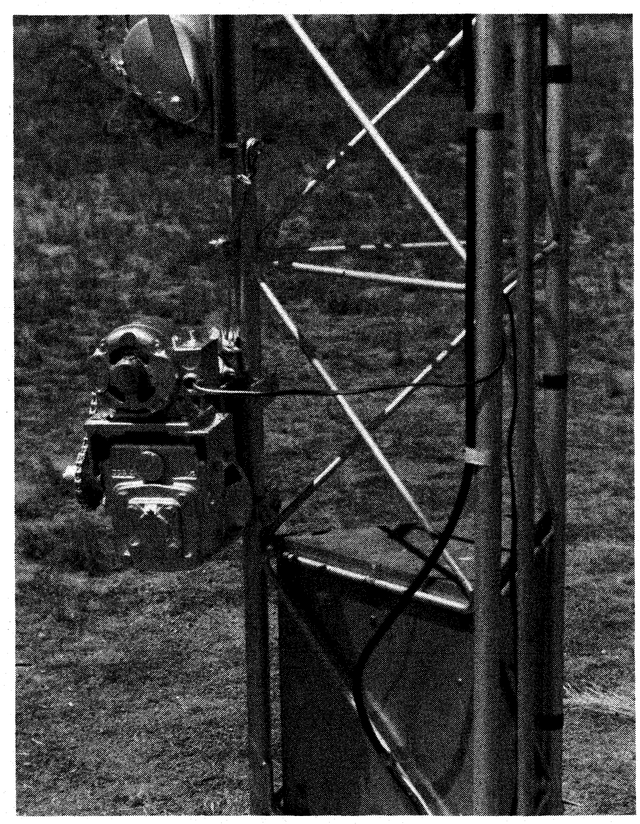
The aluminum grounding arm from gamma rod to driven element was hinged in the middle to allow for motion of the driven element, and a piece of heavy shield braid paralleled this joint. The ends were formed to fit around the aluminum cylinder and the driven element. Mating clamps held this assembly. The remaining outer end of the gamma rod wires was fastened to a long, coiled spring. The other end of this spring was clamped to the driven element with a stainless-steel hose clamp. The length of the gamma rod is about 188 cm (70 inches).

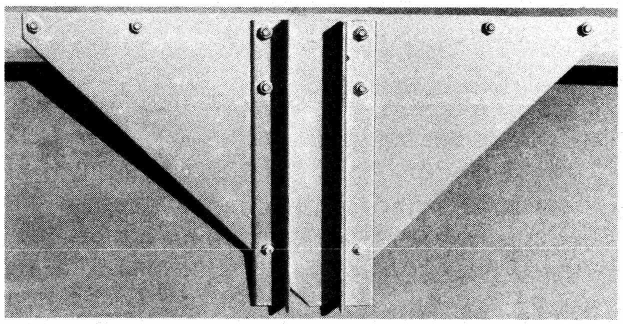
The gamma-capacitor box was arranged so that all items were supported on the cover, which could be removed by removing six thumb nuts. The gamma rod and ground connections were made by mating banana plugs and jacks.

adjustment and testing

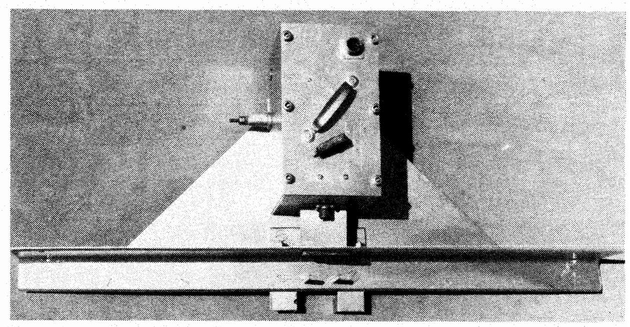
These procedures were performed with the beam in the down position, hanging about 4 meters (13 feet) from the ground. The motor-driven capacitor gamma match was operated from the ground at the tower base. The only adjustment made from a ladder was the length of the gamma rod, and the added tap-switched capacitors were reached from the tower

Tower base showing the two winches and metal box housing the selsyn and prop-pitch motor supply.

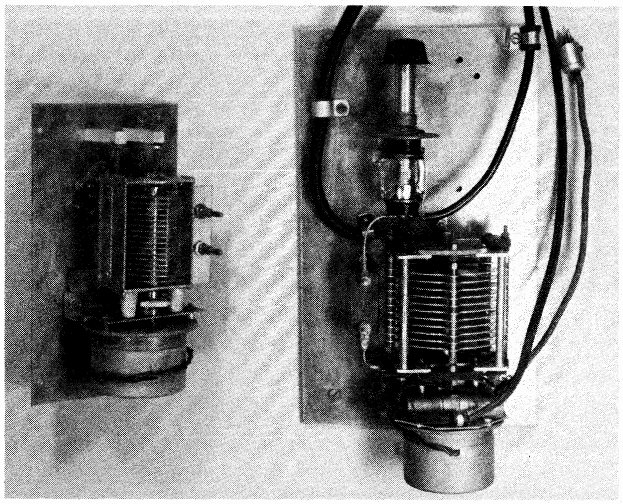




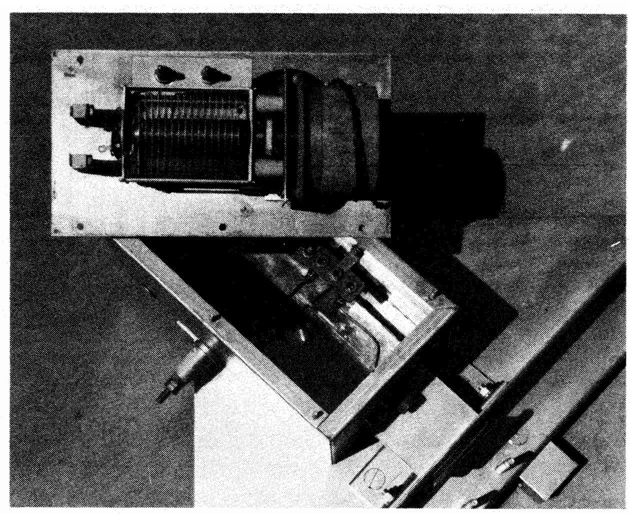
One side of an element-to-boom mount.



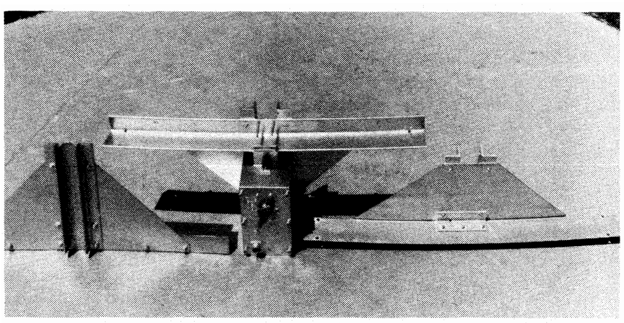
Driven-element-to-boom mount with gamma match box.



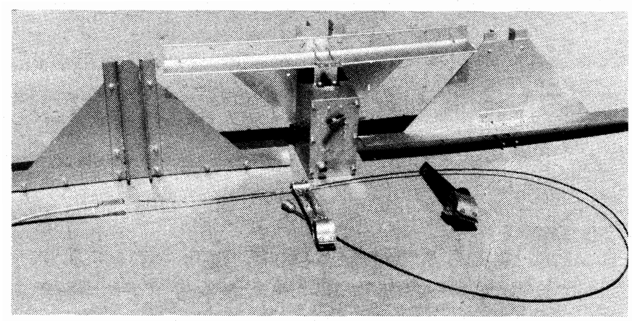
Gamma capacitor and motor-drive assembly, left. At right is a matching network for using the tower on 80 meters.



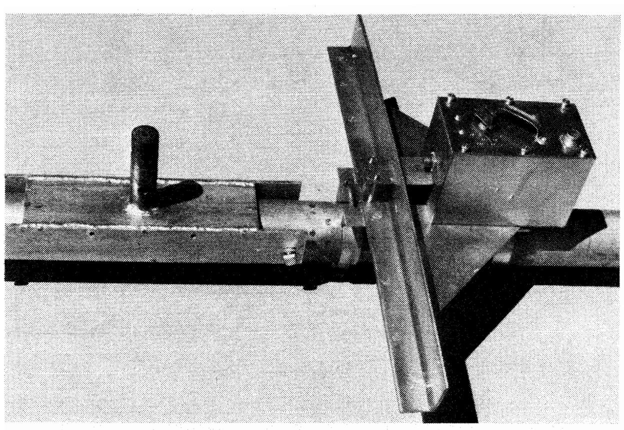
Driven-element mount with gamma capacitor and motor drive.



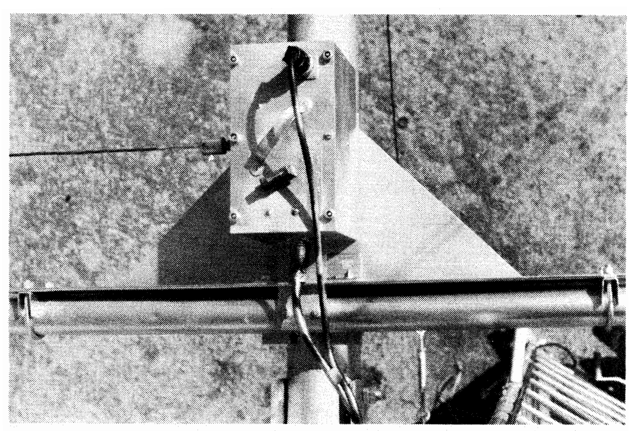
Element-to-boom mounts. Center assembly is for driven element with gamma box.



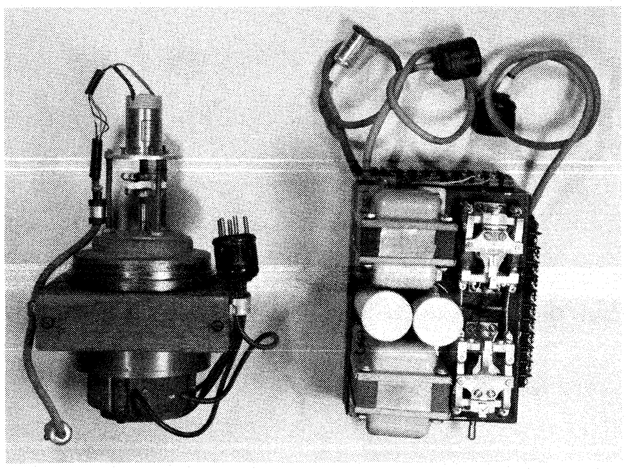
Element-to-boom mounts; gamma rod and clamp; and insulating bar. which clamps to driven element and reduces gamma-rod vibration.



Boom center with boom saddle and driven element-to-boom support in place.



View looking down from the driven-element mount.



Selsyn repeater system, left, and dc supply for the prop-pitch motor.

itself. A small vfo-controlled transmitter on the ground, with an swr bridge, provided the readings for adjustment. Unfortunately, the test readings in the down position didn't hold precisely for the beam up position. And the readings made with a 10-watt scale on the Bendix Micro-Match didn't directly compare with those taken with the output of a linear. Also, 60 plus meters (200 feet) of coaxial cable gave better readings in the shack than were obtained at the tower.

However, my 4-1000A linear[1] was quite tolerant, as was my Johnson "desk kilowatt." So I managed quite well with an swr of 1.3 at 7000 kHz, 1.0 at 7050 kHz, and 1.25 at 7100 kHz, all with the same gamma capacitor setting. I also had 1.5 at 7150 kHz, 2.0 at 7200 kHz, 2.25 at 7250 kHz, and 1.9 at 7300 kHz, each with the gamma capacitor retuned for lowest swr. Many commercially built transmitters wouldn't stand up with the higher swr readings, though a simple matching system would take care of the problem.

precautions

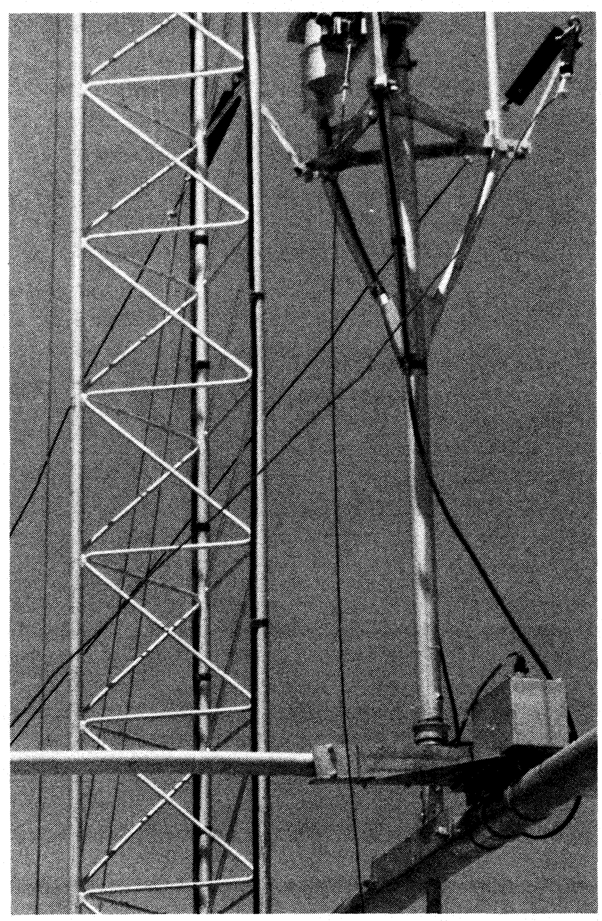
Many details haven't been covered so far; one is important. The weak point in the element-to-boom support was where the muffler clamps hold the elements. When the elements failed at this point in a 1974 wind storm, I decided that added strength was needed here and the stress should be distributed over a considerable section of the center portion of each element. Heavier wall in the 51-mm (2-inch) tubing or the addition of an internal tube to strengthen the center would do it. I used the latter method and put 3.7-meter (12-foot) lengths of 41-mm (1.63-inch) diameter tubing with 1.5-mm (0.058-inch) wall inside the elements, overlapping the 51-mm (2-inch) tubing 1.8 meters (6 feet) on each side. I used sleeves between the reinforcing tube and the inside of the element tube at the center, with the muffler clamps at the outer ends. They were all held in place by masking tape.

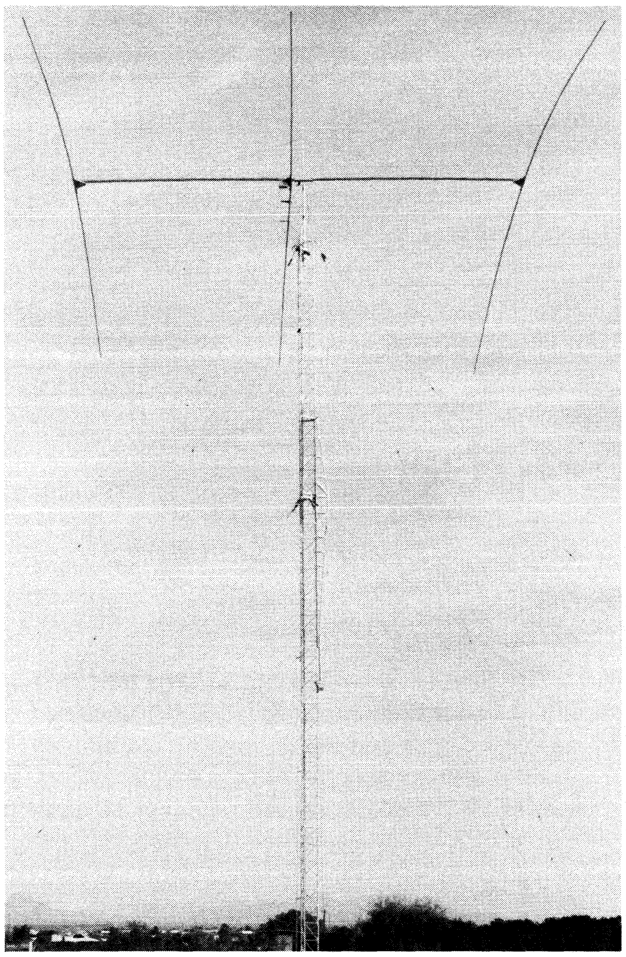
Internal support must be provided for the boom where the clamps are applied, as the tube wall is only 1.3 mm (0.05 inch) thick. Disks of 51-mm (2-inch) wood were turned to fit tightly inside the boom and driven into place where the clamp bands are installed. Friction at these points was all that kept the elements level and in place – except that the elements do hang under the boom rather than being placed on top, a difference of about 152 mm (6 inches).

Vibration dampers made of 13-mm (0.5-inch) rope were installed inside the elements, just a bit short of the full length, and fastened at the center with a bit of wire, making a loop where bolts would pass through during assembly to keep the rope from falling out when the beam was raised.

I built this antenna by myself and assembled it to the tower, which I also assembled. All the help I had was from W7EH, who helped me pour the guy wire anchors, and from the crane and operator who set the tower upright. I raise and lower the beam alone, not because I couldn't use help, but I like to go slowly and check and recheck for safety.

Tower inverted showing rotator, driven element and mount, and the gamma matching system.





The 40-meter beam on fully erect tower ready for action.

A word of warning; the Rohn tower is an excellent piece of equipment but is *not* rated to carry the top load I use when raising and lowering the tower. I have no idea what the margin of safety is. Mine has stood up verywellfor nearly ten years. But be warned – the foldover system is overloaded! Once up and in place, the two sets of four guy wires, well out from the base, make the tower safe even if the antenna is torn to pieces. The guy wires are insulated from the tower and ground. They're broken up with insulators so that the tower can be used for a vertical radiator on 80 and 160 meters.

Does it work? It sure does. It's very nice to enter a pileup of East-Coast hams working Europe or the Near East and make contacts right along with them. I'm getting too old for 48-hour contest stretches, but I thoroughly enjoy having a ***big sig*** on 40 meters.

reference

1. Gene Hubbell, W7DI, "Ecology Linear," *ham **radio,*** March, 1972, pages 6-15.

ham radio

the Micoder:
some improvements

Modifications to the Micoder tone encoder, eliminating the 9-volt battery and adding a crystal-controlled encoder

Amateur Radio magazines have published articles on improving the early production models of the Heath Micoder.[1] The faults found were in the tone outputs. Imbalance between high and low tones and poor frequency stability were prime complaints. My Micoder seemed to work well in these respects, but the ***Touch-Tone**** feature is not used that much and almost never at temperature extremes. My main complaint is that the 9-volt battery poops out about twice a year and disables the microphone as well as the tone encoder.

The solution is to power the device from the 12-volt supply in the rig. The Micoder cable contains an extra wire that's used as a ground in parallel with the shield. With the encoder components available today, this modification makes the improvement simple and inexpensive.

circuit description

The circuit is shown in **fig. 1.** Note that the microphone- and tone-circuit voltages are obtained from separate 9-volt and **5-volt** zeners. The filtering provided in the rig (an **HW-2036** in my case) and the **9**-volt zener provide adequate rejection of hum (from the ac supply) and hash (from the mobile supply).

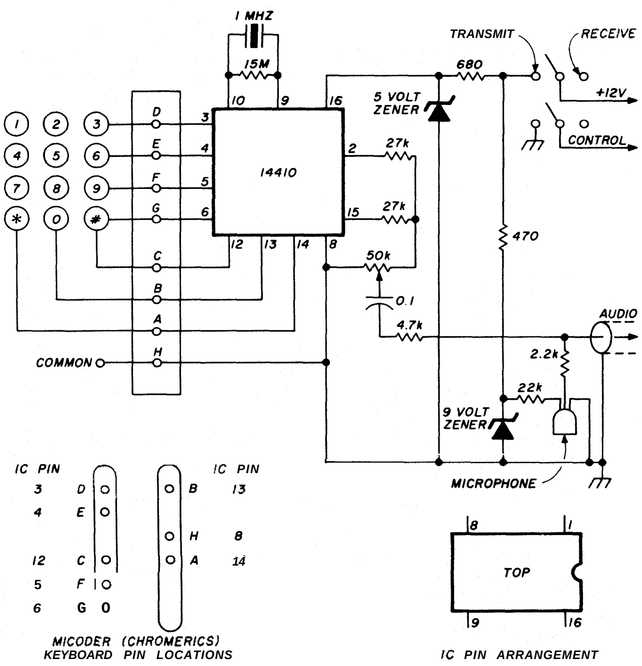


fig. 1. Modifications for improving the Heath Micoder. Circuit eliminates the 9-volt battery and adds a crystal-controlled encoder. Parts are available from Data Signal, Inc., and Radio Shack.

By George A. Wilson, W1OLP, 318 Fisher Street, Walpole, Massachusetts 02081

**Touch-Tone* is a trademark of American Telephone and Telegraph.

construction

The original circuit board was stripped of parts and reused to mount the keyboard and the output-level control. Make sure no solder bridges occur and that solder splashes don't cause shorts between circuits. This makes it possible to use the PC-board paths outboard of the keyboard pin receptacles as connection points for the wires to the encoder circuit. The battery and its connector and cable are removed and discarded.

Lift the black wire in the cable at each end. Solder a small tie point to the switch on the end opposite the existing tie points. The black wire (+12 volts) goes to the switch lug where the +9 volts from the battery was connected. The encoder circuit and 9-volt zener circuit are fed from the switch lug that previously accepted the red lead from the original circuit board.

The encoder is easily built using a PC board such as that sold by Data Signal, Inc.* They can also supply the chip and the crystal. A minor board modification is required to accommodate an extra connection for the output level control — a simple no. 60 (1 mm) hole drilled in a blank spot does the trick. When wrapped with plastic foam, the completed encoder circuit occupies the space that was previously occupied by the battery.

The level-adjustment pot can be a miniature 6.4-mm (0.25-inch) shaft-type with a screwdriver slot. If this type is available, mount it where the LED was previously. Its shaft will protrude into the hole in the outer case previously occupied by the LED. A PC-type control can be cemented to the circuit board if the miniature unit previously described isn't available.

Connect the HW-2036-end of the black wire in the cable to the +12-volt leads of the LEDs on the front panel. Lengthen the lead by about 25 mm (1 inch). The joint should be insulated with a short piece of sleeving.

This mod costs less than $20, requires no new circuit board, and solves both the battery and encoder problems.

"Data Signal, Inc., 2403 Commerce Lane, Albany, Georgia 31707

reference

1. Fallenbeck, "Micoder," *QST,* April 1978.

bibliography

1. DeLaune, "Digital *Touch-Tone* Encoder," *ham radio,* April 1975.
2. Lowenstein, "Hand-Held *Touch-Tone,*" *hamradio,* September 1975.

ham radio

vhf/uhf preamplifier burnout

Soon after arriving in New England, I started to experience random burnout of my vhf/uhf preamps, both bipolar transistors and fets. Initially I blamed the burnout on electrical storms, but the problem increased drastically in the winter, a time when electrical storms are usually at a minimum. Occasionally I even lost second-stage preamps and multipliers in my local oscillator chain.

How could this be? All normal methods of burnout protection failed to reduce the failure rate. The plot thickened when I left one of the preamps terminated in a 50-ohm load and it still blew out!

Then I connected a high-impedance, battery-operated digital VTVM to the B+ supply to the preamps. Everything looked fine until I keyed my kilowatt linear on 80-meter CW (where I spend most of my operating time during the winter) while using my east/west dipole; the dc voltage on the preamp went wild. When the west sloper was connected to the linear, the problem disappeared.

It developed that the problem was twofold: rf pickup on the power-supply lines to the preamps, and rectification in the reverse-voltage protection diode (also known as the *idiot* diode). Eliminating the idiot diode or shortening the power-supply leads are not good solutions; placing a 0.01-μF mylar or ceramic disc capacitor just ahead of the idiot diode as shown in **fig.** 1, however, prevents rf from reaching the diode and, hence, from being rectified.

No burnouts have been experienced with zener diode biasing as described in *ham radio.*" In the zener bias circuit the zener diode clamps and prevents the transistor voltages from soaring.

The circuit in **fig.** 1 won't solve all your preamplifier burnout problems, but it should give longer life to those expensive low-noise semiconductors where large rf fields and high-frequency operation are prevalent.

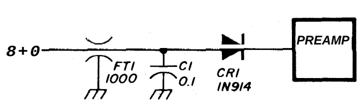


fig. 1. Circuit for preventing burnout of sensitive vhf/uhf preamplifiers in strong rf fields. C1 is a 0.1-μF mylar or ceramic disc capacitor installed between the feedthrough capacitor FT1 and the idiot diode CR1.

Joe Reisert, W1JR

*J. Reisert, W1JAA, "Ultra Low-Noise UHF Preamplifier," *ham radio,* March 1975, page 8.

multiple quarter-wave matching transformers

There are many ways of matching the impedance of an antenna to the output impedance of a transmitter. These methods often have limited bandwidths over which they provide a match. For example, the popular quarter-wave transformer shown in **fig. 1** has a bandwidth of approximately ± 10 per cent for a vswr of less than 1.5:1. Fortunately, the bandwidth of this method can be easily increased by cascading several transformers (see **fig. 2**).

To speed up the design of multiple quarter-wave matching transformers,[1] several nomographs have been developed. To make these nomographs more useful to amateurs, only the characteristic impedance of standard coaxial transmission lines is shown. **Fig.** 3 shows a nomograph for a one-section, or standard quarter-wave transformer. The impedance values shown are values of characteristic impedance for coaxial type transmission lines. **Fig. 4** shows the design nomograph for a two-section transformer and **fig. 5** shows the nomograph for a three-section transformer. To decide whether to use the one-, two-, or three-section transformer, another nomograph is provided in **fig. 6.**

The nomographs will give the impedances of the individual transformer sections. The lengths of the sections can be found by the following formula:

$$L(meters) = \frac{75}{f_{MHz}} \times vf \text{ or } L(feet) = \frac{246}{f_{MHz}} \times vf \qquad \textbf{(1)}$$

where L = length of the section
f = center frequency
vf = velocity factor

The velocity of propagation can be obtained from table 1.

design examples

The use of these nomographs can best be explained by the use of several examples.

Example **1.** Assume you are trying to match an antenna impedance of 100 ohms to a 50-ohm transmitter. One solution to this problem is to use a one-section transformer. Use a straight edge to connect 100 ohms on the R_L line and 50 ohms on the R_O line. Read the characteristic impedance of the transformer as slightly under 72 ohms. A **73-** or 75-ohm coax from table 1 may be used with little difference.

Example 2. You want to match a 250-ohm antenna to your 50-ohm transmitter, with the vswr not to exceed **1.5:1**. The antenna is a log periodic that covers the 80- and 40-meter bands.

The first step is to compute the bandwidth of the antenna. Assume the frequency range of the antenna is **3.5** to **7.5** MHz. The bandwidth is 72.7 per cent $\left(\frac{7.5-3.5}{5.5} \times 100\right)$, or about ± **36** per cent. Since this exceeds the bandwidth of a one-section transformer, **fig. 6** is used to determine the correct number of sections.

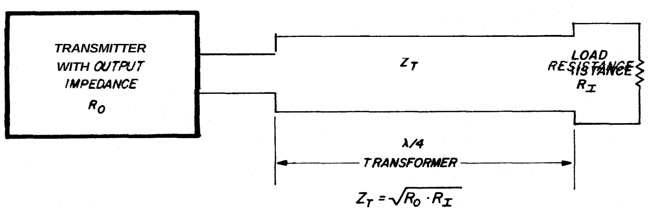


fig. 1. Schematic diagram of a standard one-section transformer. When terminated in a resistance R_L, the impedance seen at the transmitter end will be Z_T^2/R_L. By varying the impedance of the quarter-wave transformer, the load presented to the transmitter can be made to more closely match the transmitter's output impedance.

Lay a straight edge on **fig. 6** so that it lines up with 1.5 on the S scale on the left and with 5 on the R_L/R_O scale on the right. Draw a line between these two points and indicate where this line crosses the unmarked vertical line between the **S** and N scales. Now, move the straight edge so that it lines up with the dot on the unmarked line and the BW scale. Draw a line between these two points.

By **Samuel Guccione, K3BY, 110** Chalet Court, Camden, Delaware 19934

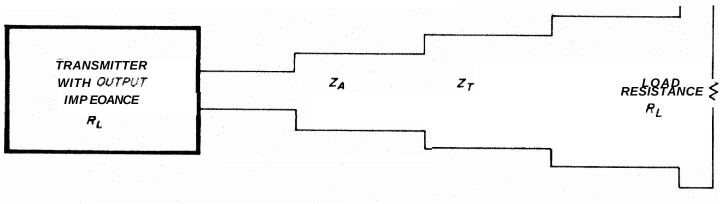


fig. 2. A multiple quarter-wave matching transformer can be used to match two impedances with a large difference in value, or to lower the vswr over the frequency range on which the transformer is to be used.

Now, note where this second line crosses the N scale. It should cross at 2, meaning that a two-section transformer will just satisfy the initial requirements. If desired, a three-section transformer could be used to achieve a lower vswr. Using **fig. 4,** for the two-section transformers, the impedances needed are 74 ohms and 170 ohms. The closest standard values that wili work are 73- and 185-ohm cabies. Using these values will give a small error, which will increase the vswr above 1.5:1. A three-section transformer (see **fig. 5**) may be more appropriate for lower vswr.

In this example, I've assumed that the antenna presented a constant impedance with frequency, and that the antenna could accept an unbalanced feed without disturbing its radiation-pattern characteristics.

table **1.** Characteristic impedances of coaxial cable **iden-**tified by **RG** type.

nominal characteristic impedance in ohms	**RG** type	velocity factor
50	RG-9B	0.659
	RG-58A, 58C	0.659
	RG-142, 142A, 1428	0.695
	RG-174	0.659
	RG-178B	0.695
	RG-196A	0.695
	RG-213	0.659
51	RG-9, 9A	0.659
52	RG-8, 8A	0.659
	RG-14A	0.659
	RG-17, 17A	0.659
	RG-18A	0.659
53.5	RG-55B	0.659
	RG-58	0.659
58	RG-54A	0.659
73	RG-59	0.659
75	RG-11, 11A	0.659
	RG-59B	0.659
	RG-140	0.695
	RG-179B	0.695
	RG-187A	0.695
	RG-216	0.659
83	RG-212	0.659
93	RG-62, 62A, 62B	0.84
	RG-71A, 71B	0.84
95	RG-180B	0.695
	RG-195A	0.695
125	RG-63, 63B	0.84
	RG-79B	0.84
185	RG-114, 114A	0.88

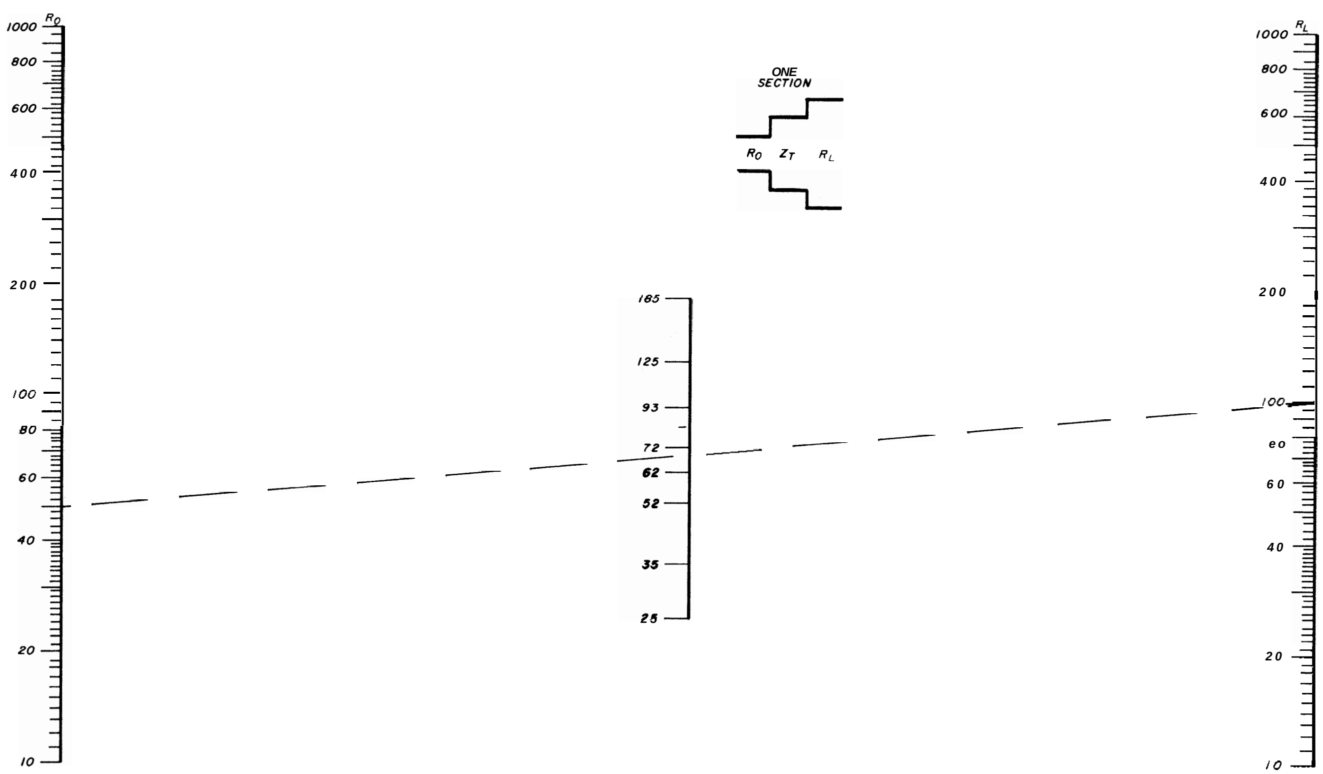


fig. 3. Design nomograph for a one-section transformer. The dashed line illustrates example one in the text. By connecting the points representing the two known values ($R_O = 50$ ohms and $R_L = 100$ ohms) the impedance of the matching section can be determined.

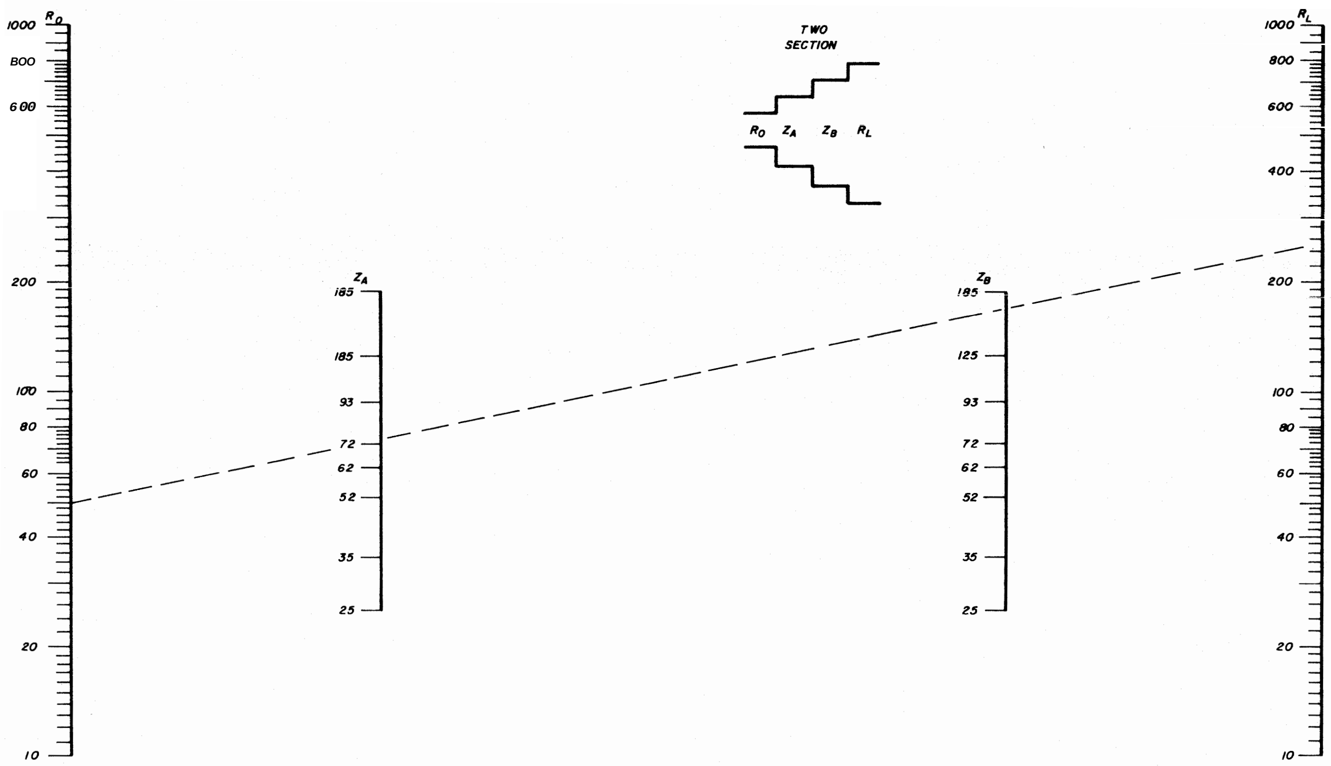


fig. 4. Design nomograph for a two-section transformer. In this illustration, the dashed line represents example two in the text. As in fig. 3, a line is connected between the two values to be matched. The impedance of the two matching sections is read from the nomograph.

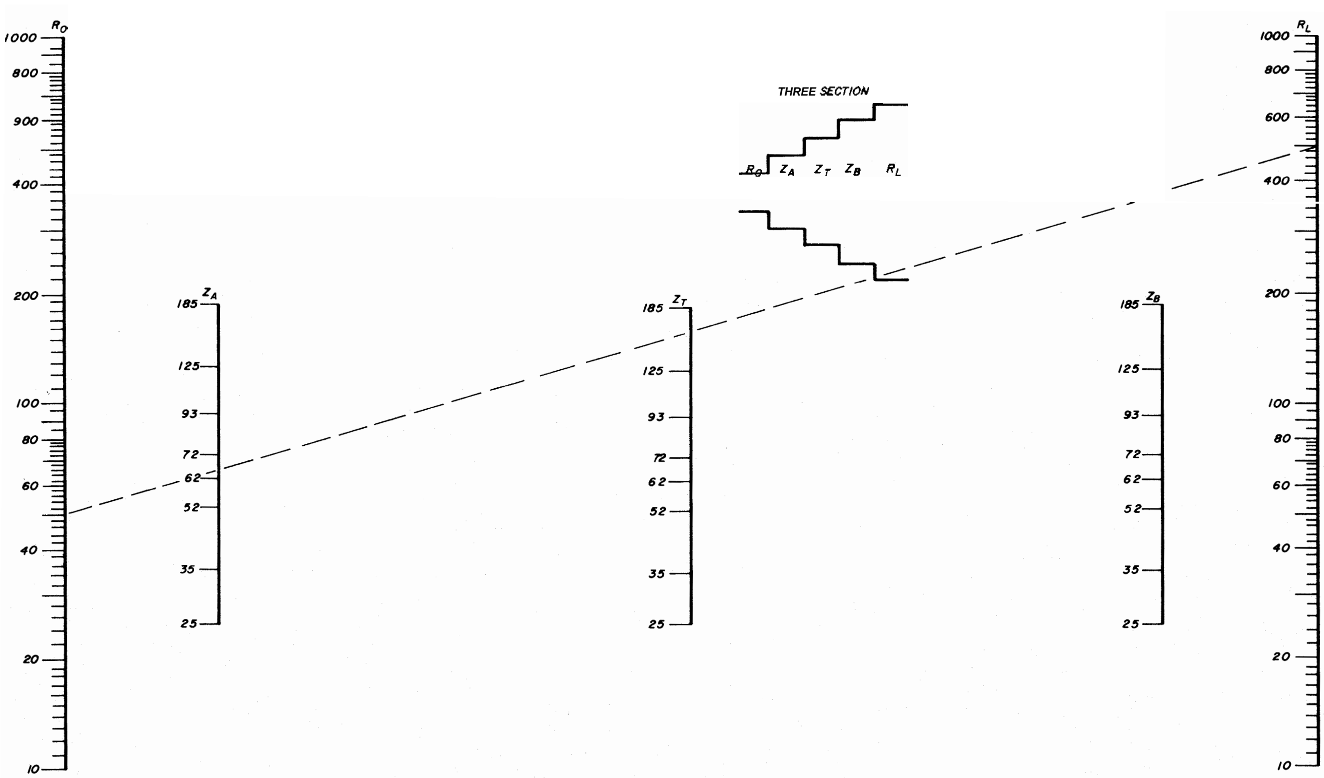


fig. 5. Nomograph for a three-section matching transformer. The impedances necessary to fulfill the bandwidth requirement in example three are illustrated by the dotted line.

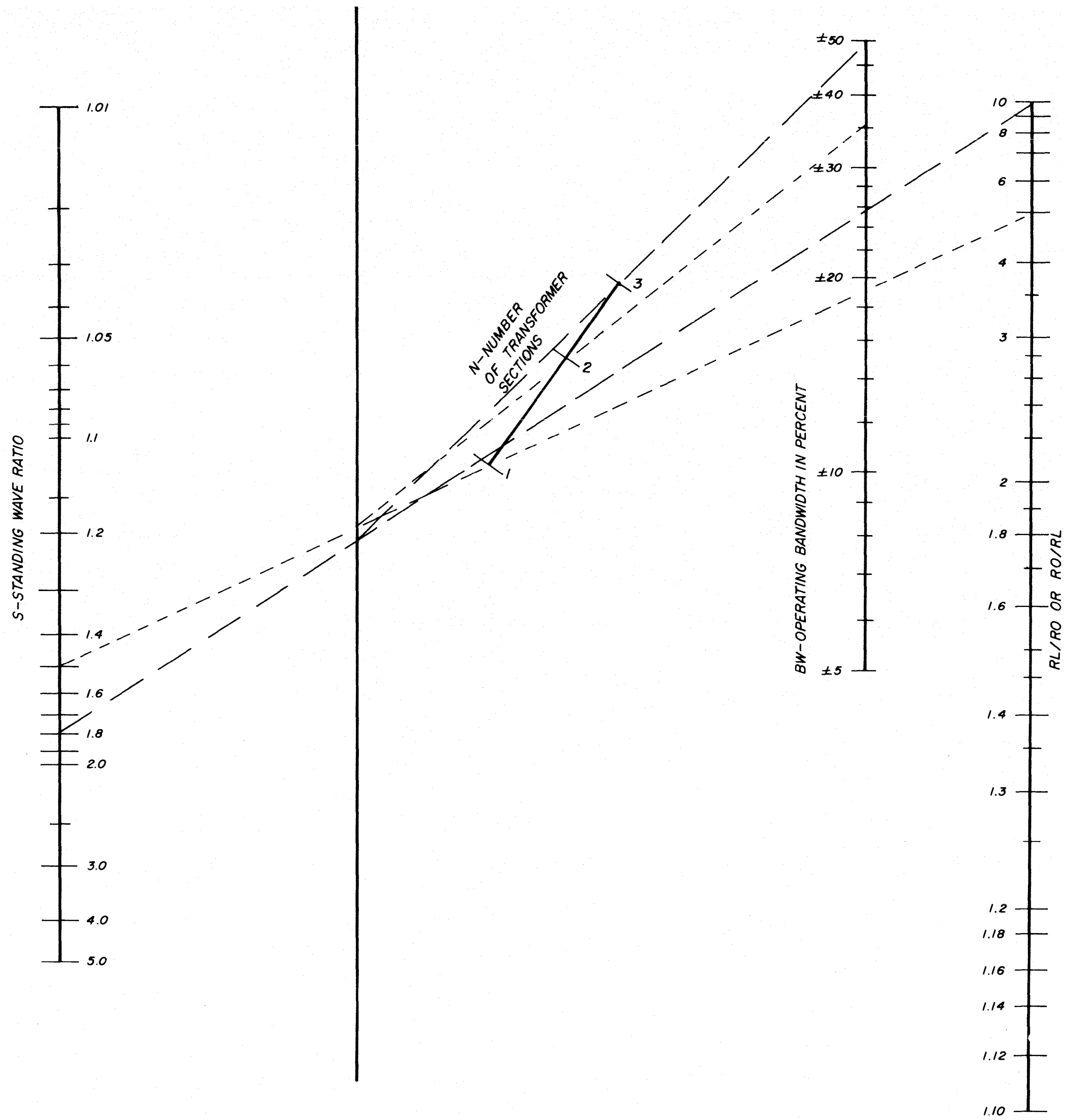


fig. 6. Nomograph for transformer selection. In example two, as represented by the small dashed lines, this figure has been used to determine the number of sections needed for matching two impedances within specific vswr limitations. The large, dashed lines represent example three.

Example 3. What will be the maximum vswr of a three-section transformer used to match a 500-ohm load to 50 ohms over a ±50 per cent bandwidth?

Using **fig. 6,** lay a straight edge so that the 50 on the *B*W scale and 3 on the *N* scale line up. Draw a line through these two points and extend this line to the unmarked scale. Place a dot where the line crosses the unmarked scale. Now lay the straight edge so that it lines up with the 10 on the R_L/R_O scale and the dot on the unmarked scale. Read off the vswr of 1.8:1 on the S scale. Thus, a three-section transformer has a vswr of 1.8:1 over a ±50 per cent bandwidth with a load resistance ten times the input impedance.

If you have a problem where a quarter-wave transformer can be used, these nomographs will greatly speed up the process of determining the optimum number of transformer sections and impedances of the sections.

reference

1. Samuel Guccione, "Nomograms Speed Design of λ/4 Transformers," Microwaves, August, 1975, page 48.

ham radio

phase-locked 9-MHz BFO

Construction of a 9-MHz BFO system which can be phase-locked to a 1-MHz reference standard

The frequency synthesizer system for an ultra-stable receiver or transceiver must include a BFO which is phase locked to the external frequency standard. **Fig. 1** shows the diagram for a crystal beat-frequency oscillator which can be phase locked to a 1-MHz reference. The BFO delivers a signal exactly 1 kHz below 9 MHz. When driving the detector of a receiver using a narrowband 9-MHz i-f filter, the receiver will deliver an audio pitch of exactly 1000 Hz for an input carrier producing an i-f signal of exactly 9 MHz. Variations on this basic design permit operation at any desired BFO frequency near 9 MHz.

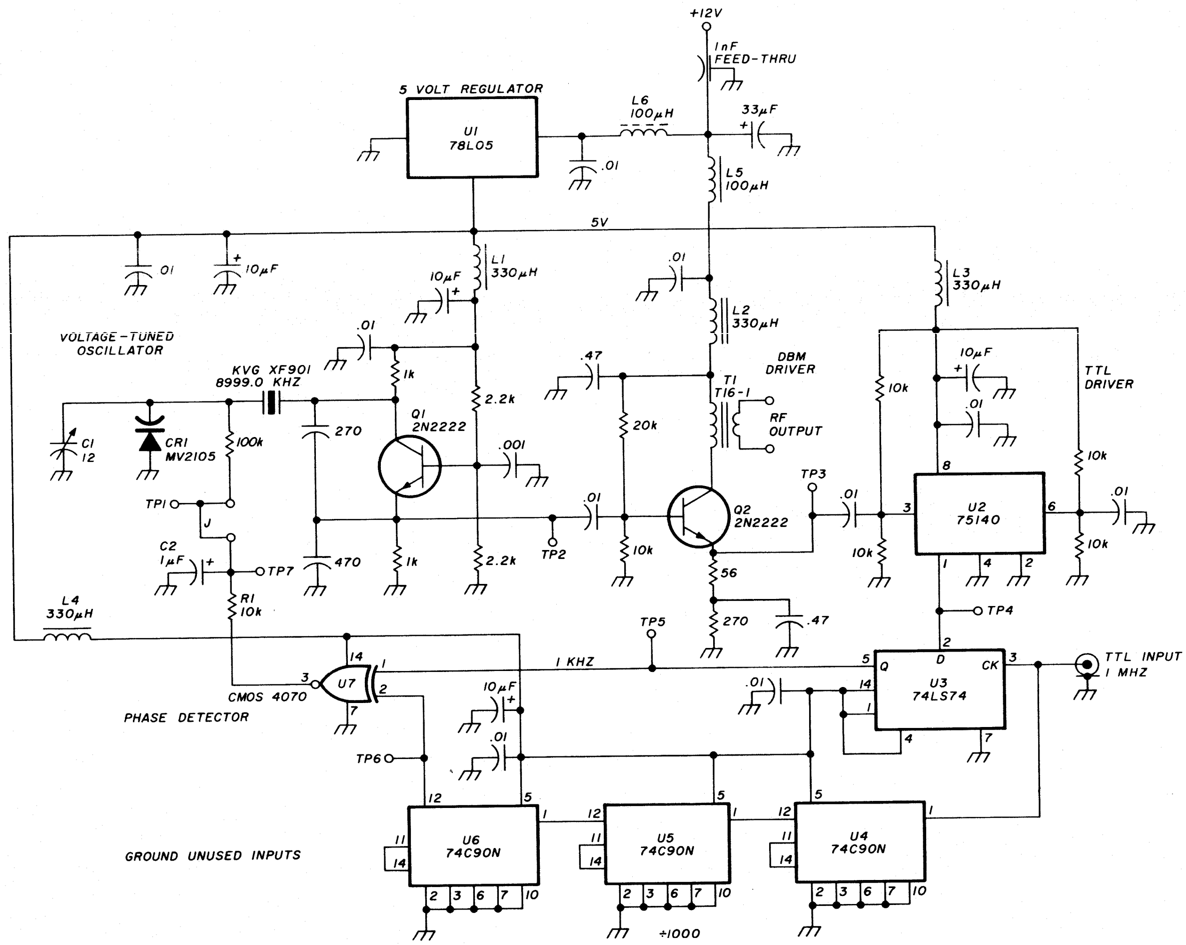
In this circuit Q1 is the voltage-tunable crystal oscillator; Q2 is a power-amplifier/driver stage which produces output suitable for the local-oscillator port of a double-balanced mixer. Output from the Q2 emitter feeds a 75140 TTL line receiver which supplies the appropriate levels for the D input of the 74LS74 "stripper" circuit. The type-D flip-flop is clocked by the 1-MHz reference signal; its output is a TTL-level squarewave at a frequency equal to the crystal oscillator frequency minus the nearest harmonic of the reference input. Thus, this circuit "strips" off the 9-MHz portion of the oscillator frequency, leaving only the difference frequency at approximately 1 kHz.

A chain of three 74C90 decade dividers delivers a l-kHz reference to pin 2 of the CMOS 4070 phase detector; the stripper output goes to the other input, pin 1. The lowpass filter R1-C2 reduces the ripple output from the phase detector and delivers the control voltage to the tuning diode CR1.

design approach for low noise

Because this system uses a 1-MHz reference signal, radiation from the reference oscillator must be prevented from reaching the receiver's i-f system. I have found that, with reasonable care, the 9-MHz reference harmonic can be eliminated. I used power-supply filters consisting of L1 through L6 with low value ceramic and 10-μF tantalum electrolytic capacitors to clean up the power supply system. By specifying CMOS and LS-TTL devices where possible, I keep the noise sources themselves as quiet as possible. All power-supply lines are bypassed to a massive circuit-board ground plane with capacitors having

By Raymond C. Petit, W7GHM, Post Office Box 51, Oak Harbor, Washington 98277



C1 3-12 pF miniature trimmer capacitors
CR1 MV2105 varactor diode (Motorola)
R1 10k nominal; may require adjustment
T1 16:1 broadband rf transformer (Mini-Circuits T16-1)

fig. 1. Circuit for the phase-locked 8.999 MHz BFO with output suitable for driving a double-balanced mixer. Output is +7 dBm into 50 ohms (5 mW). L1 through L4 are rf chokes.

the shortest possible leads. The entire circuit is enclosed in an rf-tight aluminum enclosure with connectors for the reference input, BFO output, and power. The shielded case itself does not carry any rf currents.

assembly and checkout

First build the Q1 and Q2 circuits, including the 5-volt regulator. Apply power and check the regulator output. Then check for about 2 volts at TP2 and about 4 volts at TP3. Place a 50-ohm load on the rf output and check for about 1.5 volts peak-to-peak. Connect a source of 2.5 volts to TP1 and adjust C1 to obtain exactly 8999.000 kHz at the rf output. If you cannot obtain this frequency, check the crystal and increase or decrease the net value of capacitor C1.

Wire the 75140 TTL driver and check TP4 for the desired TTL-level output. Wire the 74LS74 and connect the 1-MHz TTL-level reference signal. TP5 should show about a 1-kHz TTL squarewave which varies in frequency when the voltage at TP1 is varied. Wire the decade divider chain and look for a 1-kHz squarewave at TP6. With the jumper J removed from between TP1 and TP7, observe the waveform at TP7 while varying the voltage at TP1. With the TP1 voltage at 2.5 volts, TP7 should show a very low frequency triangle wave going from zero to 5 volts. As the TP1 voltage is raised or lowered, this waveform should pass through zero beat, rise in frequency, and diminish in amplitude.

Wire in the jumper J and remove the voltage source from TP1. The junction TP1-TP7 should now show a 2.5-volt steady dc level. When C1 is varied slightly, this dc level should rise or fall slightly. If C1 is moved too far, TP1 will suddenly show a low-frequency triangle wave. By moving C1 back, this should suddenly "capture" and again show a steady level. Adjust C1 until this level is again at 2.5 volts.

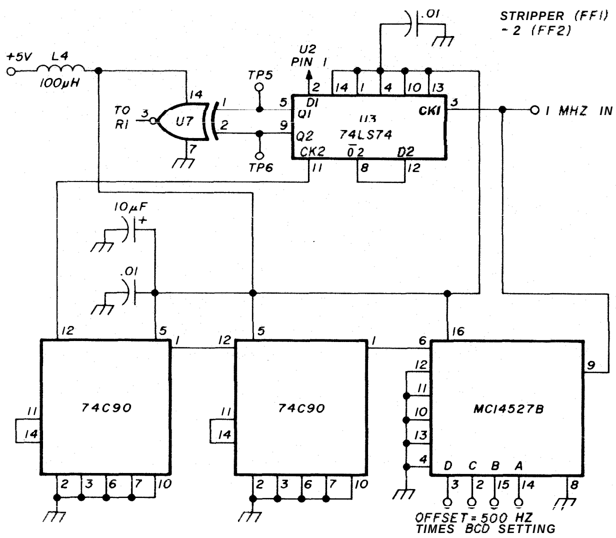


fig. 2. Schematic of a rate multiplier offset synthesizer for the phase-locked BFO.

modifications for other frequencies

If the desired BFO frequency is 9001.000 kHz, substitute a crystal of this frequency. No other modifications are required. For single-sideband work, where the BFO frequency will be plus or minus 1500 Hz from 9 MHz, use the appropriate crystal and substitute the rate-multiplier circuit of **fig. 2.** With this circuit you can select any frequency offset from 500 Hz to 4500 Hz. The MC14527B rate multiplier accepts a 1-MHz input from the reference oscillator and generates unevenly spaced output pulses, the average rate of which can be programmed with the four rate input settings D, C, B, and A. The rate input is in BCD, positive logic. If the setting is 3 (BCD 0011), for every 10 input pulses from the reference there will be three output pulses. The two 74C90 decade counters and the second half of the 74LS74 serve as a divide-by-200. With this arrangement a rate setting of three produces an average frequency output of the rate multiplier of 300 kHz, and the divider brings this to 1500 Hz.

The next article will describe the phase-locked up-converter which translates the 1100 to 1600 kHz output of the first loop[1] to the required local-oscillator frequencies for each of the amateur bands from 160 meters through 30 MHz.

reference

1. Raymond C. Petit, W7GHM, "Synthesized High-Frequency Local-Oscillator System," *ham radio,* October 1978, page 60.

ham radio

magnetic mount for mobile antennas

A novel antenna mount made from a cast-off surgical device called a *Bio-Pump*

It's not often you run across the perfect piece of junk. In this case, it's a magnet with all the qualities you'd want in a magnetic mount for an antenna:

1. Holding power
2. Shape
3. Plastic case
4. Coax-fitting-sized hole

The device that houses this magnet isn't piled up at your local surplus store; in fact it won't be there at all. Fortunately, however, it's available to those wishing to seek one out. More on this later.

My son Brian had a strange looking contraption among his collection of junk, and I paid little attention to it for months. The thing finally aroused my curiosity one day when it started collecting nuts and bolts on its own, so I attempted to figure out its use and purpose. The marks on its side said *Bio-Pump,* whatever that meant.

It looked expensively made and had a magnet buried deep inside. Since it was headed for the scrap pile anyway, I struck it with a hammer and ended up with a magnet encased in plastic. And that's the heart of this story.

development of an idea

I didn't immediately visualize the gadget as an antenna mount. It took a while since I didn't need one at the time. Holes for an antenna mount had been recently put into my International Scout, justified by the fact that it was the only mount I had without buying one. I also needed strong support for the colinear antenna that would be used.

When I finally got around to the Bio-Pump's mounting potential I checked out its holding power. Using a metal file cabinet, I wasn't prepared for how it was grabbed out of my hands by the magnetic attraction.

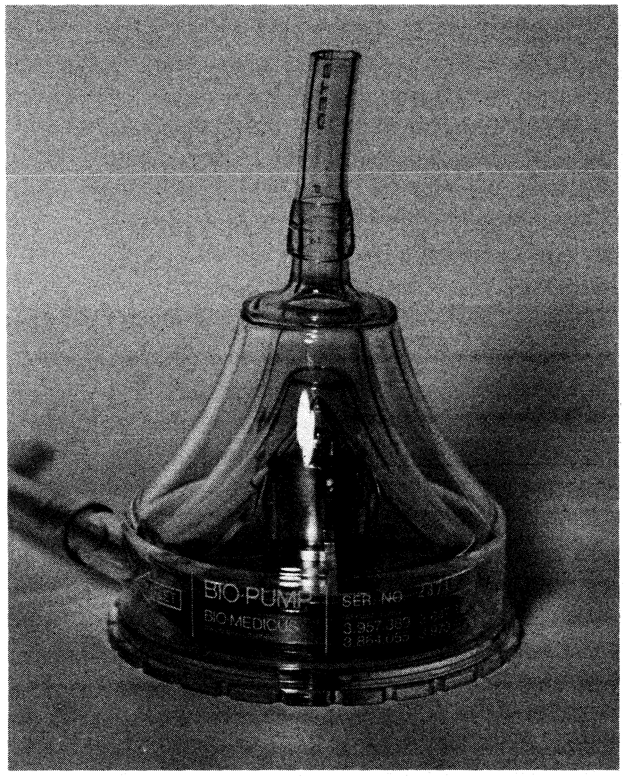
Since my location is near metropolitan Minneapolis-St. Paul, and normal driving takes me near the repeater I use often, I decided to build a quarter-wave whip for the mobile rig. Result: I put the colinear into permanent storage. Now I don't whack tree limbs and can drive into my garage.

how to get one

It turns out that this pump is a disposable item after being used during open-heart surgery. In the Twin Cities alone, 1500 to 2000 open-heart operations are performed annually. Not all of them use this particular pump; however, many do, and its use is increasing.

I can't supply direct contacts for you to obtain

By W. H. Kelley, WØHK, Route 1, Box 295A, Maple Plain, Minnesota 55359



The Bio-Pump, which contains a healthy magnet that can be used to mount a mobile antenna. These devices are expendable after heart-surgery operations and can be obtained free from many hospitals.

one, but knowing someone who works in a hospital is how I've obtained three of them for this article. Since they're disposable, I believe a few phone calls will put you in touch with the right people. If they're reluctant to give you one and prefer instead to throw it awav or destroy it, offer to destrov it for them in their presence just for the encased magnet. In any case, be open about who you are and what you want it for.

I'm not usually in favor of articles in which an exotic hard-to-obtain component is required, but obtaining this pump is the heart (pun intended) of this design. Finding one will be worth your effort.

description and use

The plastic covering is perfect for not marring the car surface. Construction is simple, as you will see.

> Amateurs have a proclivity for making do. The name of the game is innovation; if you don't have it, you make it. This article shows what one amateur operator did with a cast-off piece of junk. The result is a mag mount for a mobile antenna that sticks to metal like a bum to a ham sandwich. Does your mobile whip hit the dirt when a big semi tractor-trailer rig goes by? Try this mount and forget your worries. Editor.

It's shaped for minimum wind resistance and the holding power is unbelievable.

To accent that last point, I went on a fishing trip where four-wheel drive was periodically required. You'd probably be impressed if I told you that the antenna stayed on top without moving. Better than that, I moved it inside the vehicle on a vertical panel and it did not move in the slightest for several days.

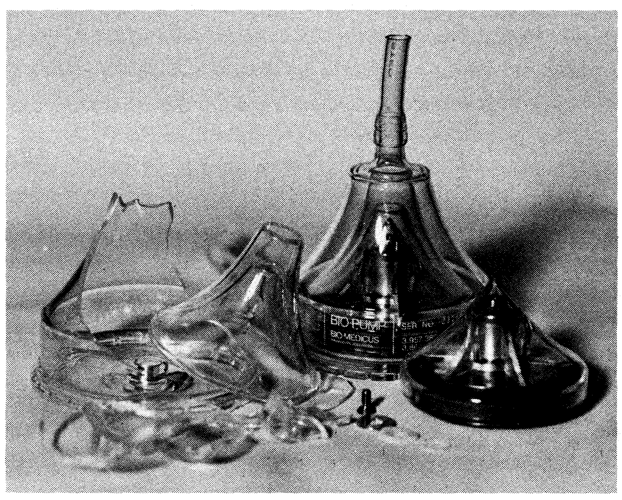
When you get one, don't use a hammer as indiscriminately as I did, especially if you only have one plastic pump. When breaking the outer plastic case, start at the opposite end of the magnet. The plastic that houses the magnet is the fourth wall in as you look inside. The second and third walls are actually one piece and do a good job of protecting the magnet as you break through the outer wall.

The plastic chips very easily, so protect your eyes and don't go barefoot until you clean up. Once you remove enough plastic to expose what you're after, use an allen wrench to remove the screw, washer, and seal from where it's held on the bottom. The photo shows how.

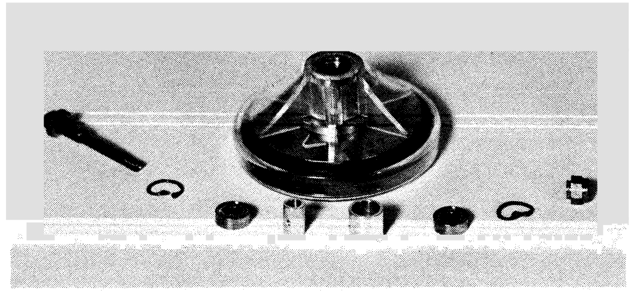
You'll see a couple of strips of something wandering around inside and the inclination is to figure out a way to remove them. ***Don't!*** I tried and it's impossible without doing damage.

A hacksaw can be used to cut the pointed plastic end of the inner container. Cut just above the inside metal cylinder top surface. You'll find that the plastic is molded around this corner, leaving about 0.8 mm (1/32 inch) lipped over the top tends to seal at that point.

Both ends of the rotating works in the center will now be exposed and can be disassembled. Don't



The Bio-Pump after applying a hammer to break the plastic outer shell, left. A complete instrument before destruction is shown at center. At right is the inner plastic container that houses the magnet.



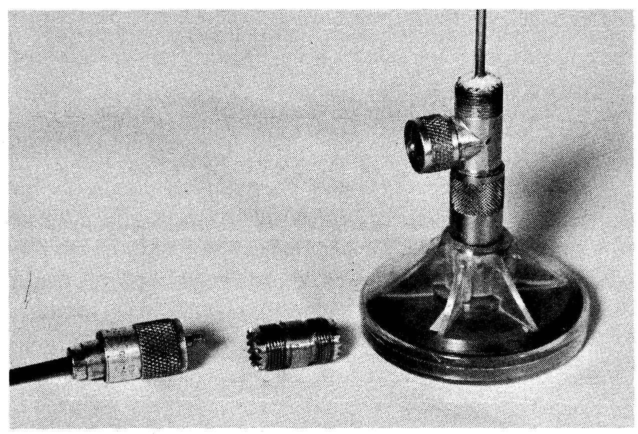
Inner part of the Bio-Pump with pointed end sawed off.

drop the pump on a hard surface, as it will be damaged.

Disassemble the parts in the same order as shown in the photo. The two snap rings can be bent and pulled out with a sharp-pointed object. Now is the time to set the magnet onto a metal surface and be impressed.

The following description shows how I used this magnet. It was chosen more because of parts on hand than anything else. (See photo.) The rear of a PL-259 coax fitting will be a loose fit. Some solder on the fitting, filed down to a press fit, will do the job.

If you have some no. 277 ***Loctite*** as I did, it will also work. A tee coax fitting allows a point to feed the antenna.



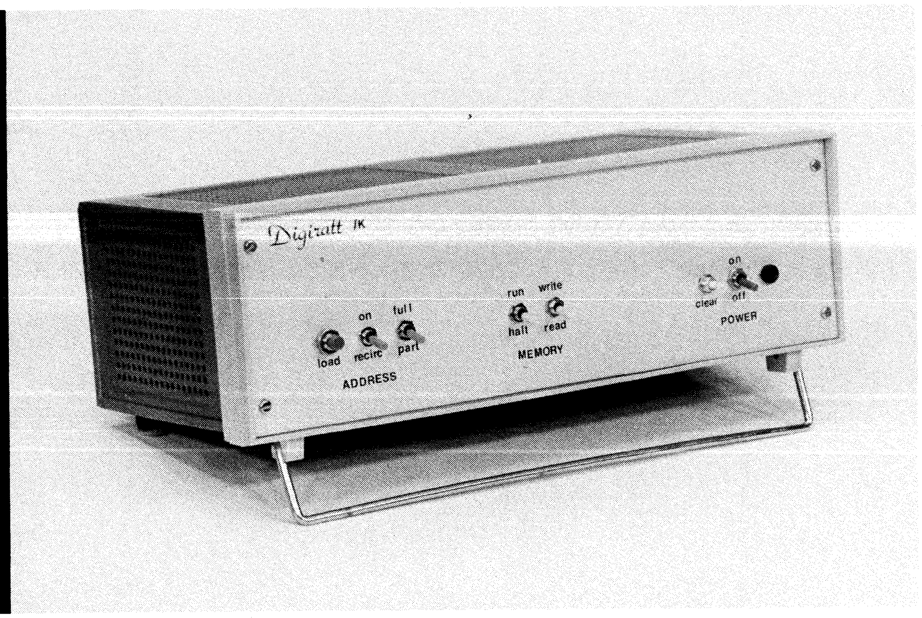
Complete mag mount showing coax Tee fitting and coax-cable accessories. Antenna is a quarter-wavelength whip made from 3-mm (118-inch) diameter brass welding rod.

For a whip, I used 3 mm (1/8 inch) diameter brass welding rod. Put some sealant at the whip base for a weather seal after soldering the rod into the fitting.

acknowledgment

Special thanks to Neil Gravatt for taking the pictures.

ham radio



Digiratt 1K — a digital reperf/TD

Continuing in the Digiratt series, KB9AT has developed a 1024-character digital RTTY reperf/TD

Most ideas become reality because someone defines a need and develops an approach to meet that need. The original Digiratt RTTY AFSK Generator and PLL Demodulator1 was designed so I could simply and rapidly place a vhf RTTY station on the air with few frills. The Digiratt RY Test Generator2 came into being because of a badly misadjusted model 15 and the resulting need for a stable test signal. When a model 28 KSR arrived, I found that a model 28 ASR would be more desirable than the standard KSR. Therefore, of necessity, a digital replacement for the usual mechanical device was designed.

design concepts

I initially decided to design a unit which was the equivalent of a Reperf/TD, which is to say it could copy RTTY off the air as well as from the local keyboard. Also, since all my RTTY equipment was TTL based, it followed that the new unit would also be TTL. It had to have the ability to edit mistakes without re-entering the entire message and it also had to have a memory capability of a minimum of 16 lines of 64 characters. I decided to include a recirculator so that all, or a portion, of the memory could automatically be repeated over and over for calling CQ, printing R/Y's, and doing "Quick Brown Fox" type testing.

By John Loughmiller, KB9AT, RR1, Box 480C, Borden, Indiana 47106

The resulting Digiratt 1K allows the user to write (enter) up to 1024 RTTY characters into the memory, either from a local keyboard or off the air (see fig. **1).** Entry speed can be as slow as desired so long as the Baud rate is 45.45 Hz. Full 60 wpm entry speed is allowed, with other speeds accommodated by changing the system clock frequency.

In the read mode, the data is clocked out at a constant rate approaching *60* wpm. Either 128 characters or all 1024 characters can be recirculated. (This could be 2 lines of 64 characters or 16 lines of 64 characters.) To change or delete a character, the memory is allowed to read (print) up to and including the character immediately preceding the one to be changed or deleted. The unit is then switched to the write mode and the correction is made. Read is re-established and the next correction made in a like manner.

theory of operation

Clock generator. The clock generator board* is composed of a 1-MHz oscillator followed by a divide-by-1375 circuit which furnishes an output of 727.2 Hz (fig. **2).** This frequency is 16 times the 45.45 **Hz** Baud rate used for *60* wpm RTTY and is also the frequency required by the UARIT for input data processing.

Data entry. The 727.2-Hz clock signal, in addition to being applied to the UARIT, is also divided by 16 (U1 and U2), becoming the shift register clock. It is further divided by 8 (U3) and used as the shift register load/read pulse (see fig. 3).

Serial RTTY data enters the board and is applied to the UARIT. When the character has been processed and is ready to be transferred to memory, a 5-bit parallel data word appears on pins 9 through 12 of the UARIT, and the received data available flag changes

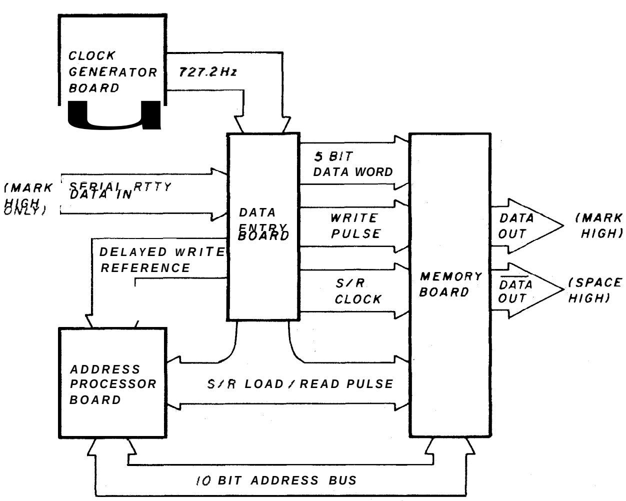


fig. 1. Functional block diagram of the Digiratt 1K reperf/TD.

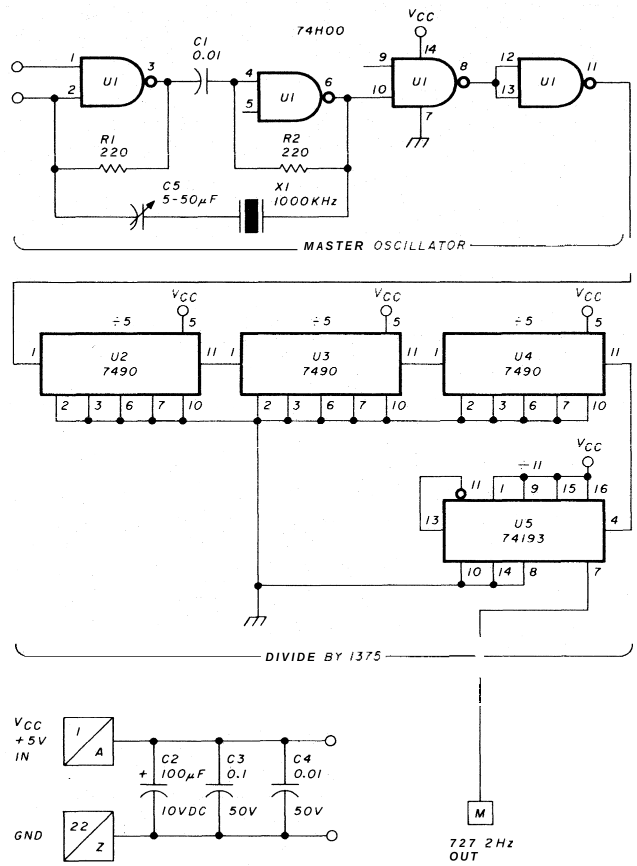


fig. 2. Schematic diagram of the clock and divider board. A basic 1-MHz oscillator is divided by 1375 to produce the 727.2-Hz waveform.

state. This flag is then used to fire U4. The output from U4 triggers U5, and also becomes the write pulse to load data into memory. U5's output is called delayed write reference and is used to decrement address processor one number and simultaneously clear the UARIT for reception of the next character.

Address Processor. The three binary counters (U1 through U3 in fig. **4**) receive either the delayed write reference or the shift register load/read pulses as a clock input, depending on whether the write or read mode is selected. When the load button is pressed, a predetermined address is loaded into these counters, either 1024_2 or 1282 depending upon the position of the full/partial switch. This allows the user to instantly return to a predetermined point. If the *Recirc* switch (recirculator) is on, the predetermined

*A copy of the printed circuit board layout is available by sending a self-addressed, stamped envelope to ham *radio,* Greenville, New Hampshire 03048. In addition, a complete set of etched, drilled, and plated circuit boards is available for $17.50 from Circuit Board Specialists, P.O. Box 969, Pueblo, Colorado 81002. A complete parts kit is also available for $55.00.

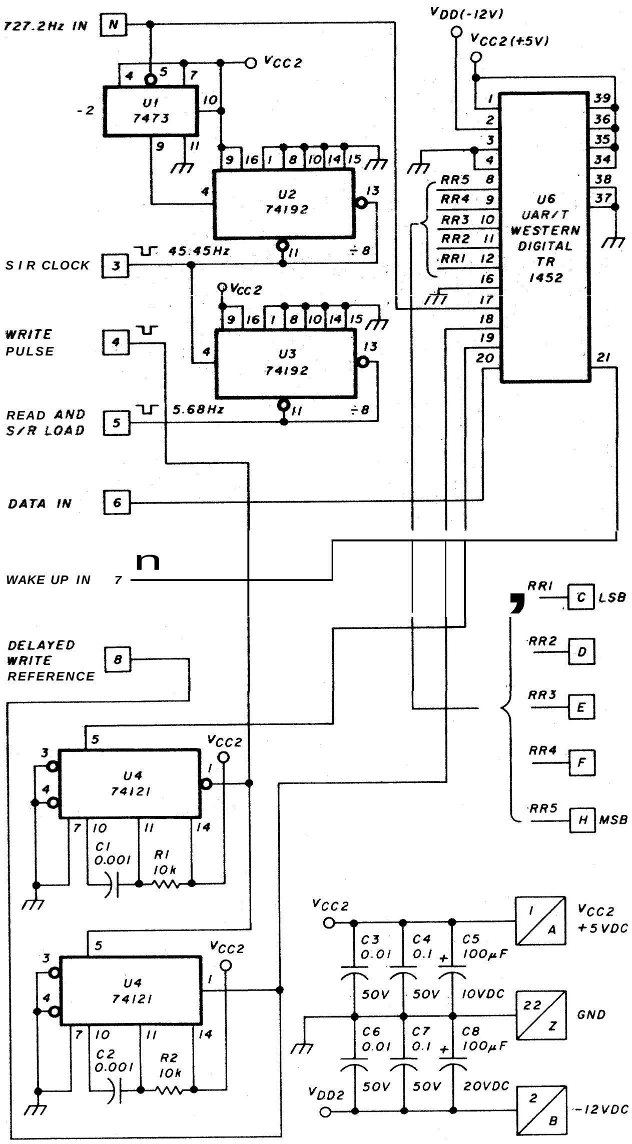


fig. 3. The UAR/T on the data entry board is used to convert the serial RTTY data to a 5-bit parallel form. U1 through U3 provide additional division of the 727.2-Hz signal to provide the correct frequency for the output shift register.

address will constantly be reloaded at the end of the countdown cycle, resulting in recirculation of either 128 or 1024 characters, indefinitely. The outputs from the three 74193s is the binary address used to control the 2102 RAMs during both read and write modes.

Memory. In all modes, the binary address is applied to all five RAMs, one bit per RAM (fig. 5). The write pulse enters the data into memory. The delayed write reference pulse, which follows, increments the address processor to the next lower address. This procedure then repeats for subsequent characters.

In read mode, the shift register-load/read pulse both decrements the address processor and loads the data from the RAMs into the 74165 shift register (U6). Finally, the clock pulse into pin 2 of U6 shifts out the serial data at a 45.45 Hz Baud rate.

Power supply. The power supply, as shown in fig. 6, is quite straightforward, using the popular LM309K regulators. I do suggest that the reader not deviate from the schematic with respect to the use of separate regulators. Also, bypassing on the individual boards should not be omitted.

With the power supply and on-board decoupling shown, I've placed the unit on top of an operating linear power amplifier, with no problems resulting from the rf field. In extreme cases, however, you may have to bypass the RAMs with 0.01-μF capacitors at their V_{CC} pins. For that matter, any of the ICs may need bypassing if erratic operation is experienced in a high-intensity rf field.

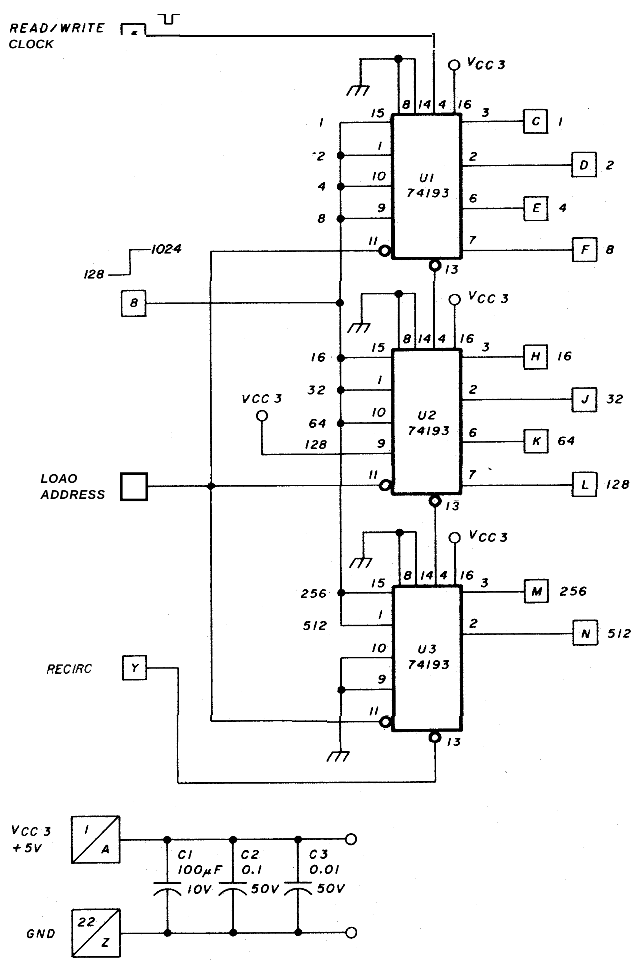


fig. 4. The address processor generates the 10-bit binary address for the RAMs. The 74193s count down from the number loaded into their pre-set inputs.

initial start-up

Apply power and press the clear button to reset the UAR/T registers. This must be done each time the system is turned on, or you can build in a power-up system using a resistor, capacitor, and inverter. In any event, the UAR/T registers must be reset at turn on.

Select full address, write, run, no recirc, and press the load button. Enter 1024 Blanks and select run and read. Press load address once and observe the playback. If you get anything other than blanks on playback you probably have a defective RAM. Consult a baudot code chart and determine which RAM would have to be defective to cause the resultant printed letter; the RAMs correspond directly to the baudot chart. Next, load ***Ltrs*** into all 1024 spaces and repeat the procedure outlined above. If you can store 1024

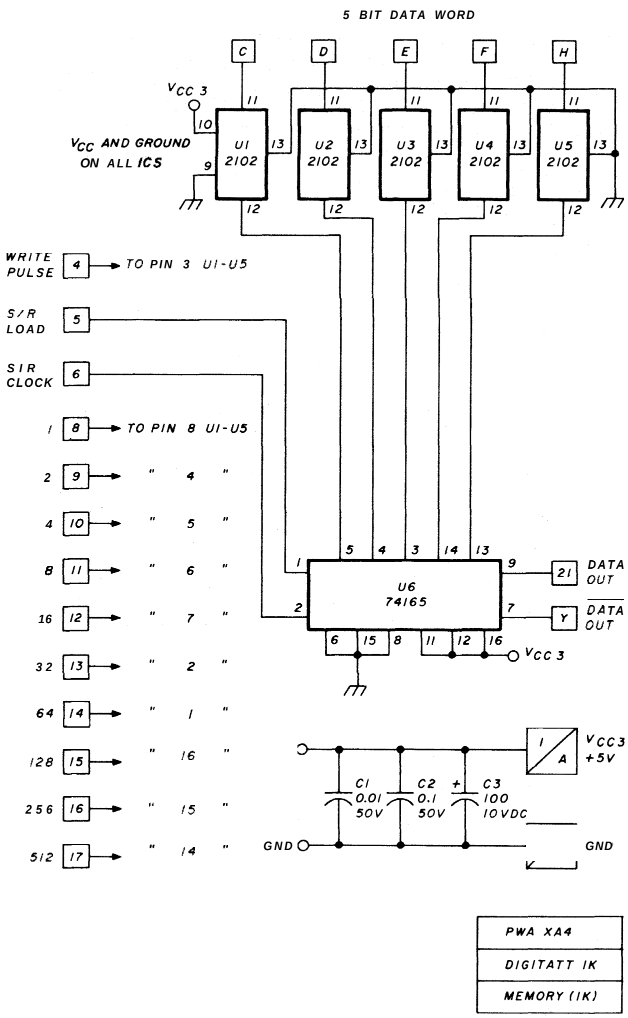


fig. 5. The memory board contains five 2102 RAMs and the output shift register. The 10-bit address lines are simultaneously applied to all 2102s. The hardwired inputs of the shift register are used to generate the stop and start pulses.

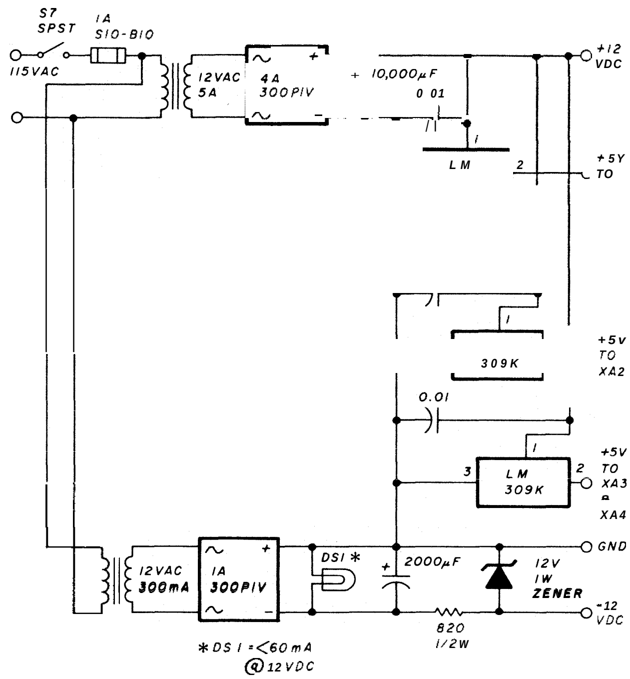


fig. 6. Schematic of the power supply for the Digiratt 1K. The small indicator lamp should draw less than 60 mA at 12 volts. The three LM309K regulators are mounted on the back of the case.

blanks and 1024 ***Ltrs*** and recover them without error, the RAMs are probably acceptable.*

Obviously, you'll need a baudot code chart (found in most RTTY books) to trouble-shoot RAM problems. I bought eight "prime" 2102 ICs, and three were bad at certain addresses. It was a blessing in disguise, however, since this trouble-shooting procedure was the result of that problem.

system interface

The UAR/T requires the input mark signal to be a TTL high. Therefore, any system devised to enter data will have to furnish a high logic level when the loop is in the mark condition. With respect to outputs from the Digiratt 1K, both mark and space signals are available with TTL high levels as their true conditions.

One suggested input circuit is the system devised by John Alford, WA4VOS,[2] using a diode bridge and optical isolator. Another method would be to key a reed relay from the local loop. Connect the relay contacts in such a manner that the center arm is connected to 5 Vdc when the loop is in the mark state and ground when the loop is in space condition. You may find it necessary to debounce the contacts of the reed relay.

*It should be noted, however, that special memory testing routines are required to disclose some failures. This procedure is not a complete test but should suffice in most cases.

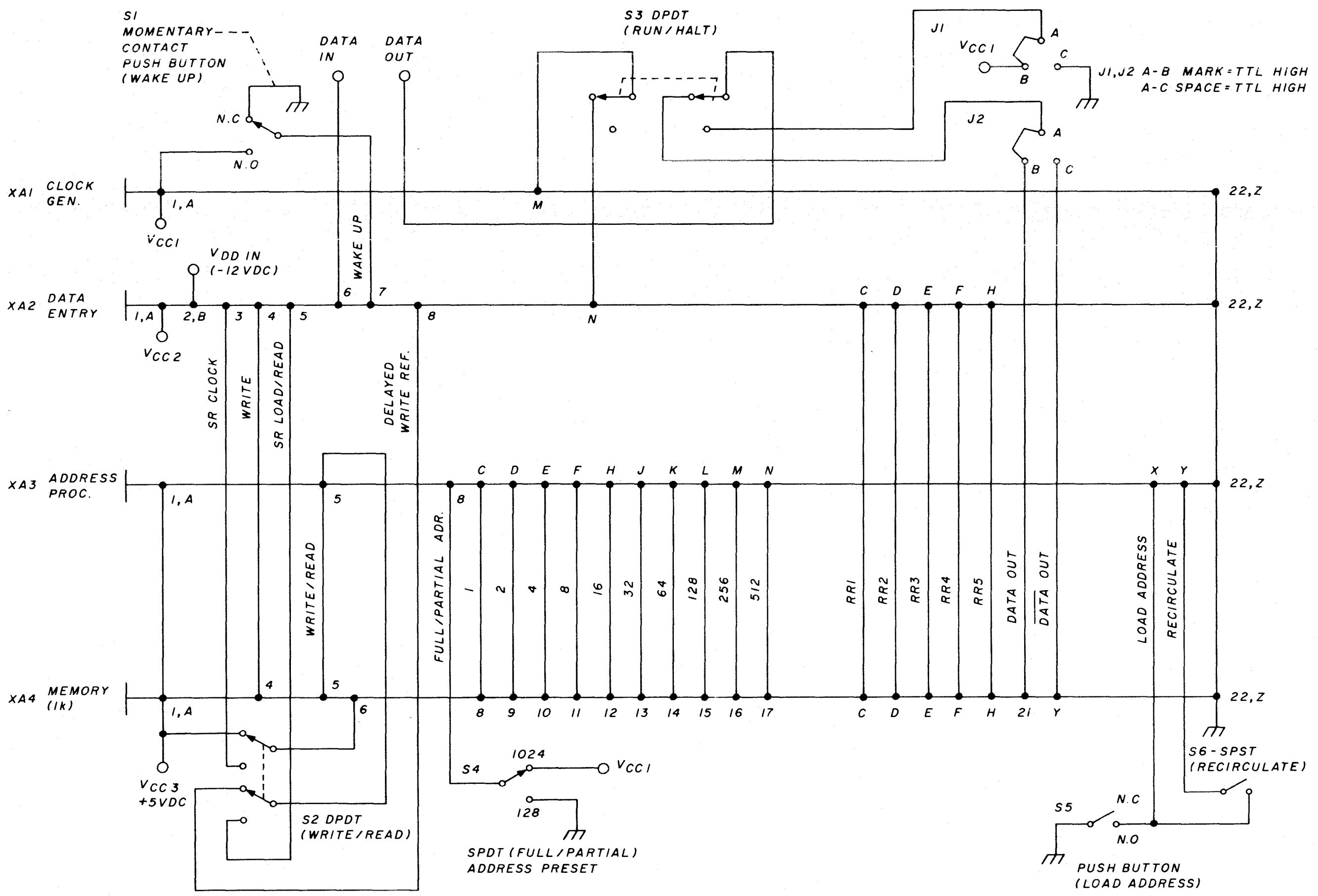


fig. 7. Interconnection wiring diagram for the Digiratt 1K. With jumpers J1 and J2 connected to A-B, the mark output will be a TTL high level; connected to A-C the mark level will be a low level.

operation

Write. 1. Select either full or partial memory.

2. Press load address.

3. Turn the recirculator off (even if it will be used during read).

4. Set the write/read switch to write.

5. Set run/halt to run.

6. Enter the data.

If the recirculator is to be used; fill in all unused memory slots with blanks, or random data will be printed until reset occurs.

Read. 1. Set full/part switch as required.

2. Press load address.

3. Set write/read to read.

4. Set recirc as desired.

5. Set runlhalt to run when ready to read out the data.

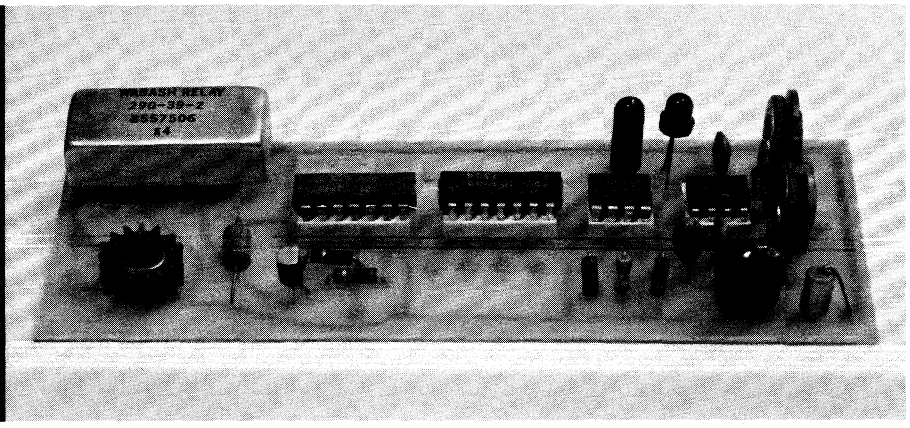
future expansion

I've designed a couple of options which are not on the prototype unit. The first is a one-step forward/backward text editor which allows correction of mistakes. Second, an address preset using BCD thumbwheels and 7-segment displays, making the memory completely random access. You merely dial up the place where a particular message is stored, press the load address button, and read out the contents. Finally, a 2K version and an ASCII version are also designed but not yet built.

references

1. John Loughmiller, WB9ATW, "Digiratt – AFSK Generator and Phase Locked Loop Terminal Unit," *ham radio,* September, 1977, page 26.
2. John Loughmiller, "RTTY Text Generator," *ham radio,* January, 1978, page 64.
3. John Alford, WA4VOS, "Improved Digital AFSK," *ham radio,* March, 1977, page *22.*
4. Johnathan A. Titus, "The UAR/T and How It Works," *ham radio,* February, 1976, page 58.

ham radio



tone-alert decoder

Construction of a simple tone-alert decoder for emergency call-ups of RACES groups and the Amateur Radio Emergency Service

How do you monitor for a possible call-up of the local ARES or RACES group on the local repeater while you're at the office – without subjecting other office employees to the normal traffic on the repeater? How do you monitor for emergency calls throughout the night without being jolted out of bed by an amateur who works the sign-off shift at the TV station and puts out a call on his way home from work? One solution is a telephone call-up using the telephone tree system. The problem is that if one or two members are not home, it can really slow down or even stop the fan-out process.

After considering and experimenting with several alerting systems, the Wayne County (Michigan) Amateur Radio Public Service Corps (ARPSC) developed an alerting system over the local repeater that is based upon a tone-alert. The basic concept of the system is that any ARPSC member equipped with the tone-alert decoder who wishes to be notified of an alert – but does not want to listen to the repeater audio – can switch the tone-alert decoder to monitor mode. This allows the audio from the receiver to pass to the decoder but does not allow it to pass on to the speaker.

To activate the tone-alert the amateur calling the alert transmits the appropriate audio tone for the proper duration; the decoder latches up and the relay contacts switch the audio onto the speaker.

So far the tone-alert system has been used for several tornado watch "sky warn" net call-ups in Wayne county. Washtenaw, Monroe, and Genisee County groups have also started the development of tone-alert systems using the circuit described in this article.

There were four primary objectives of the decoder design:

1. A decoder which would not give false alerts by detecting voice peaks or alternator whine.

2. A decoder which would cost less than $15, because the higher the cost, the fewer that would be placed into service by volunteers.

3. A decoder which was built from readily available parts; this would increase the number which would be built by volunteers.

4. A decoder circuit which could be reset by the appropriate signal.

After the criteria were established, a search of the literature was made and no simple circuit was found that met our objectives. (After our system was developed a decoder meeting some of the design requirements was published in *QST.*)

A prototype was developed and evaluated for several months on a receiver monitoring the local repeater. Tests were made of various call-up systems and tone encoders. When the design was considered satisfactory a batch of circuit boards was produced and kits of parts made available to ARPSC members who wanted to participate in the program. The record of performance is good for the units now in service.

By Harold C. Nowland, W8ZXH, and **Stan** Briggs, **W8MPD,** Hal-Tronix, Post Office Box 1101, Southgate, Michigan 48195.

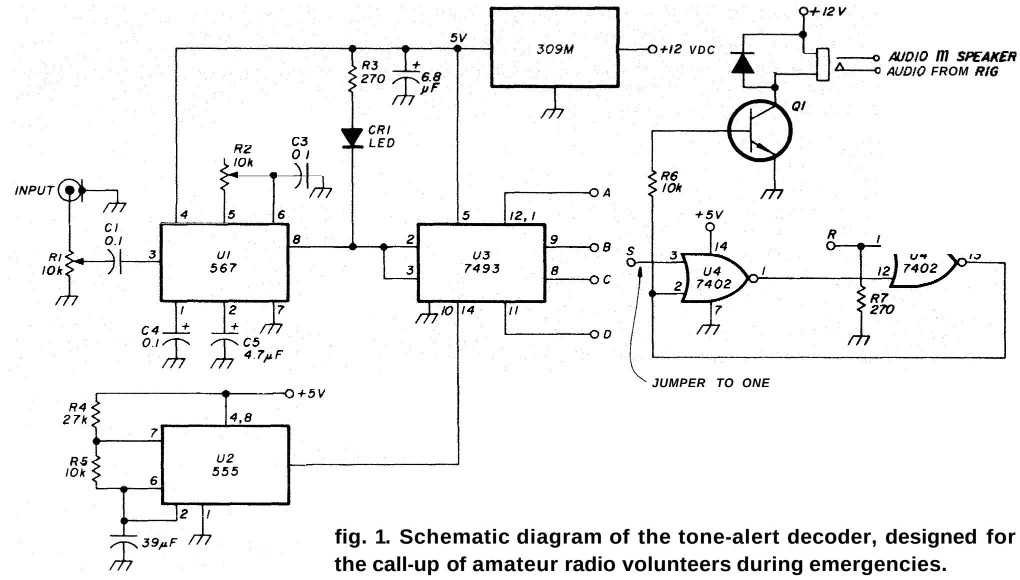


fig. 1. Schematic diagram of the tone-alert decoder, designed for the call-up of amateur radio volunteers during emergencies.

circuit description

Integrated circuit U1 is a NE567 PLL tone-decoder timed to the system alert tone frequency (see fig. 1). The audio from the receiver passes through the level control R1 to the 567 input. R2 and C2 control the 567's operating frequency. When the proper tone is not being received the 567 output is about +4 volts, which is applied to the reset input of U3. U2 is an oscillator operating at about 1 Hz. The output is fed to the clock input of the 4-bit counter, U3; as long as the reset is held high by the output of U1, however, the counter stops at zero.

When the proper tone frequency is received at the input of U1 the output voltage goes to zero. This turns the LED CR1 on, indicating that the PLL has locked up. This also allows the reset input to the counter to go low, allowing the counter to start to count. The outputs of the counter will go high in turn as the counter counts up: (1) high after the first pulse, (2) high after the second pulse, (3) high after four pulses, (4) high after eight pulses.

The desired delay is obtained by selecting the output to be jumpered to the SET input to the latch, U4. If a delay of four clock pulses (about four seconds) is desired, for example, a jumper would be placed between C and S. Once the count reaches four, the output of the latch will turn on the relay driver transistor Q1.

If the tone is not on for the whole four clock pulses, the counter will reset when the PLL output goes high. This protects from lock-up on voice peaks, alternator whine, or other spurious signals.

The relay contacts may be placed in series with the external speaker circuit so that audio can be heard when the relay is operated.

To reset the latch a positive voltage must be applied briefly to the R input. The circuit can also be reset remotely by using a diode from the D output to the R input. This allows the tone of eight clock pulse duration to reset the decoders remotely if desired. The format is thus: ON, four clock pulses; OFF, eight clock pulses.

construction

The layout of the circuit is not very critical. The only critical component is the 0.1-μF capacitor (C3). This should be a good quality Mylar capacitor to limit frequency drift.*

The tune-up procedure is as follows:

1. Set R1 and R2 at the center of their ranges.

2. Apply the desired tone frequency to the decoder input. R2 should be adjusted until the LED lights, indicating the lock-up of the PLL.

3. The signal level should be reduced and R2 should be adjusted to locate its "center of lock."

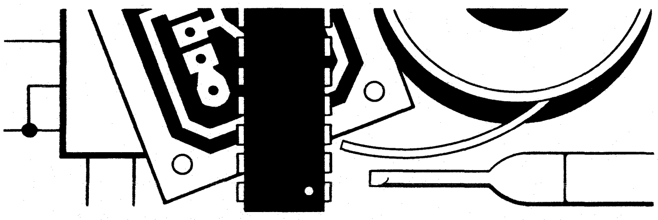
4. On-the-air checks should be made to determine the appropriate level setting of R1.

Various types of tone generators may be used. The Wayne County ARPSC group uses a quartz crystal oscillator which is counted down to the proper audio frequency by a 14-stage ripple-carry binary counter/divider integrated circuit such as the CD4020. Other groups have found that simple oscillators using 555 timers with good components and voltage regulation are adequate.

ham radio

*A parts kit, less relay, and an etched and drilled circuit board are available for $10.95 plus $1.00 postage from Hal-Tronix, Box 1101, Southgate, Michigan 48195.

the weekender



an antenna swr meter

In keeping with the Weekender theme, here is a useful station accessory that can be constructed in the course of a weekend and should provide many years of service.

The antenna meter (or vswr bridge, as this type of device is usually called) provides an indication of match or degree of mismatch between the transmitter and antenna system. I don't like to refer to the unit as a bridge, because that implies a device of extreme accuracy and brings to mind instruments built by General Radio and the like. Rather, the antenna meter is a simple instrument that samples the forward (incident) and reflected voltages in the coaxial feedline between the transmitter and antenna system. This easily built device is useful for tuning your transmitter for maximum output, adjusting an antenna tuner for minimum reflected voltage, and tuning your antenna system to a favorite spot on the band.

This antenna meter is for use with 50-ohm coaxial lines, such as RG-8/U and RG-58/U. It covers the high-frequency bands and will operate with any type rig, other than QRP. The unit contains no exotic or hard-to-find parts, with construction greatly simplified through the use of a printed-circuit board. If you do not have the facilities for etching the PC board, an etched and drilled board is available as shown in the parts list. This form of construction eliminates the mechanical difficulties often associated with instruments of this type. The sensitivity of the meter varies with frequency, about ***60*** watts on 80 meters and approximately 3 watts on 10 meters for full-scale deflection.

circuit description

The circuit of the antenna meter shown in **fig. 1** has been around a long time. It has been constructed in many forms, ranging from wires snaked inside coaxial cable to copper tubing spaced inside a minibox. Although construction techniques vary, the basic theory and function remain the same. In this configuration, the signal from the transmitter is applied to the input jack, J1, passes along the center conductor of the etched board, and exits via J2.

Two sense lines are etched parallel to the center conductor. The lower line along with its associated diode detector, is called the incident, or forward, line; the upper line is called the reflected, or reverse, line. The sense lines are identical, but are connected to detect voltage in opposing directions.

A forward/reflected switch connects a metering circuit which reads the voltage developed in the sense lines as the rf signal passes along the surface of the center conductor. By using the calibration control, R3, the reading from the forward sense line can be adjusted to deflect the meter to full scale. Then, through the use of the forward/reflected switch, the reflected voltage can be read to develop forward-to-reverse voltage ratio.

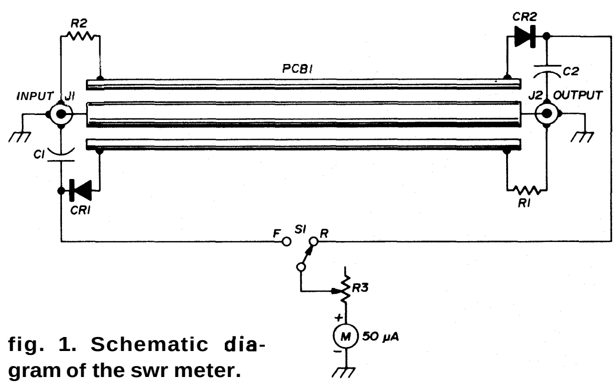


fig. 1. Schematic diagram of the swr meter.

S1	rocker switch, dpdt	RS 275-030
J1, J2	single-hole SO-239 connector	RS 278-195
CR1, CR2	germanium diodes 1N34A	RS 276-1123
C1, C2	.005/500 V dc disc caps	RS 272-130
R1, R2	51-ohm 5 per cent resistors	RS 271-1308
R3	100k-ohm pot	RS 271-092
M1	50 μA meter	RS 22-051

The reflected reading represents the portion of the voltage that has been applied to the feedline but has been reflected back from the antenna system due to mismatch. The primary objective is to maintain the reflected reading as low as possible, with a reading near zero indicating a purely resistive 50-ohm termination at the far end of the coaxial feedline; that is, a good match between the transmitter and the antenna system.

construction

The heart of the antenna meter is the printed-circuit board shown in **fig. 2.** The board is double-sided glass epoxy with a thickness of 2.5 mm (3132 inch).

By **Ken Powell, WB6AFT, 6949** Lenwood Way, San Jose, California 95120

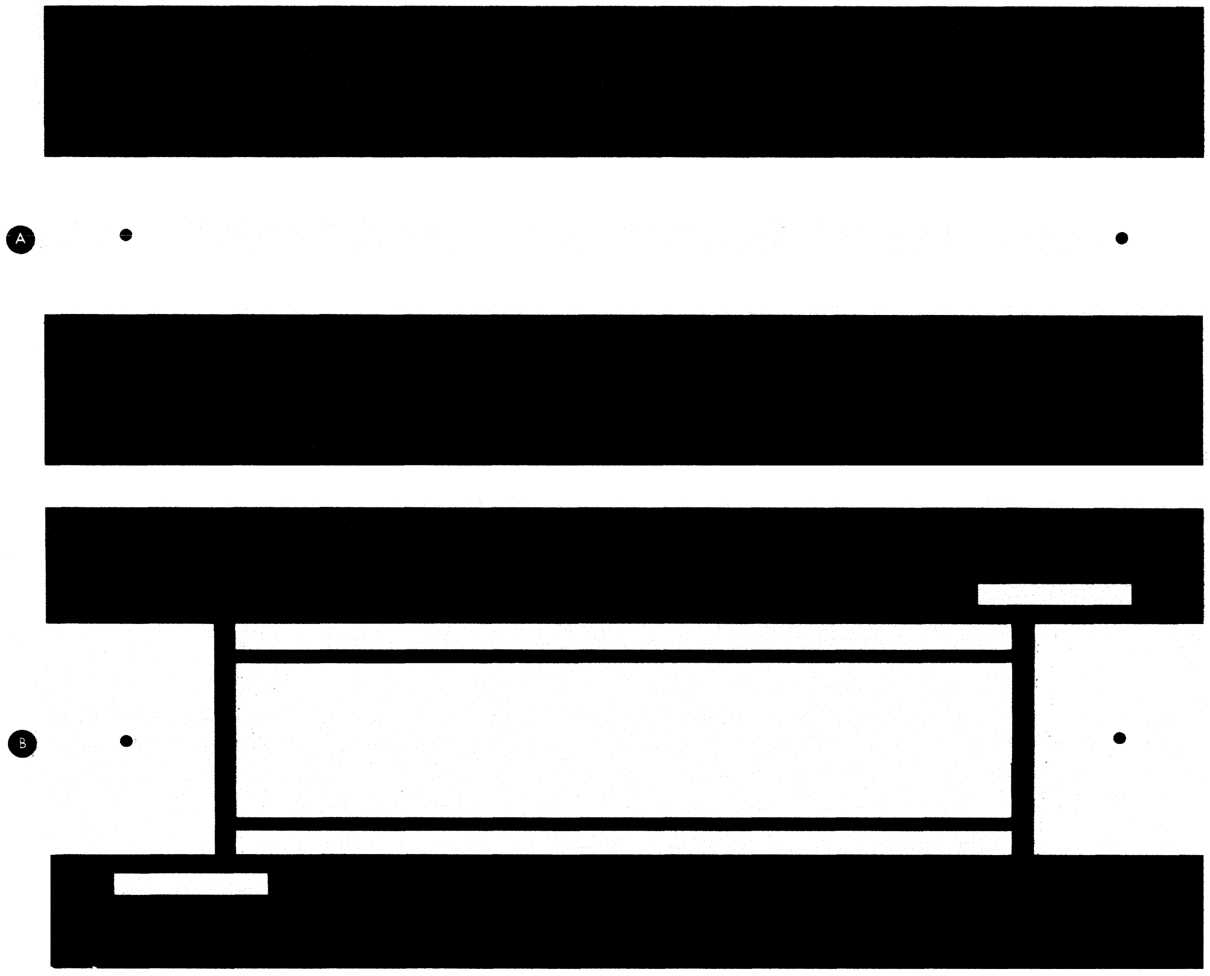


fig. 2. Circuit board layouts for the board used in the swr meter. The top side is shown at (A); the back of the board at (B). Contrary to normal *ham radio* style, the foil is represented by the white areas.

The only holes in the board are the 16-mm (518-inch) holes for the coax connectors. After etching and drilling are completed, place a piece of masking tape on the reverse side of the board (fig. 3). Lay the PC board down, component side up, and tin the areas shown in fig. 3 with a small iron and rosin-core solder. This pre-tin operation will aid in component mounting.

Form the resistors, diodes, and capacitors as depicted in fig. **4,** and, using needle nose pliers to grasp the lead being soldered, mount the diodes, resistors, and capacitors as shown in fig. **5.** Solder the output leads from both sense lines and leave them about 30 cm (12 inches) long. Next, mount the single-hole SO-239 UHF connectors to the circuit board, but *do* not over-tighten the hardware. Using no. 14 (1.6 mm) AWG solid copper wire, form two pieces of wire, as shown in fig. **4,** to connect the center leads of the UHF connectors to the printed circuit board. Solder the wires to the UHF connectors first, then align the wires with the Center of the board and solder in place, being sure the wires do not touch the grounded frame of the connectors. This completes the circuit board wiring.*

Drill the required holes in the case, being careful not to damage the paint in the process. After making sure that all components fit correctly, wash the case assembly with a mild detergent to remove any wax or oils which may be present. Prior to mounting the components, apply the lettering to the case using rub-on letters (I used the Datak K61 Letter Set). Then apply a couple of light coats of Krylon Clear or Datak Matte Finish to protect the lettering and finish. This will give a professional look to your project.

Remove the lock nuts and washers from the UHF connectors and mount the circuit board and other components in the case. Wire the unit as shown in the schematic diagram, using the meter's negative terminal as the common ground point for the ground wires from both sense lines. Parts placement is not

*An etched printed circuit board is available from J Oswald, 1436 Gerhardt Avenue San Jose, California 95125 for $4 00 postpaid Order board number 1004J.

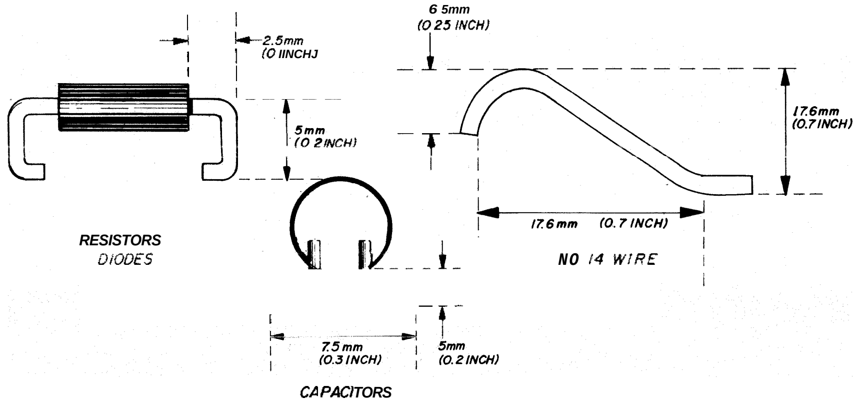


fig. 4. Component preparation prior to mounting on the circuit board.

critical, but try to keep the wires from the sense lines perpendicular to the board for a distance of a few centimeters. After assembling the cabinet, remove the plastic snap-on cover from the meter and remove the meter face. Using a common eraser, remove all the numerals from the meter face, exercising caution not to damage the scale graduations. Now, apply the new digits (see **fig. 6**) using the lettering set and

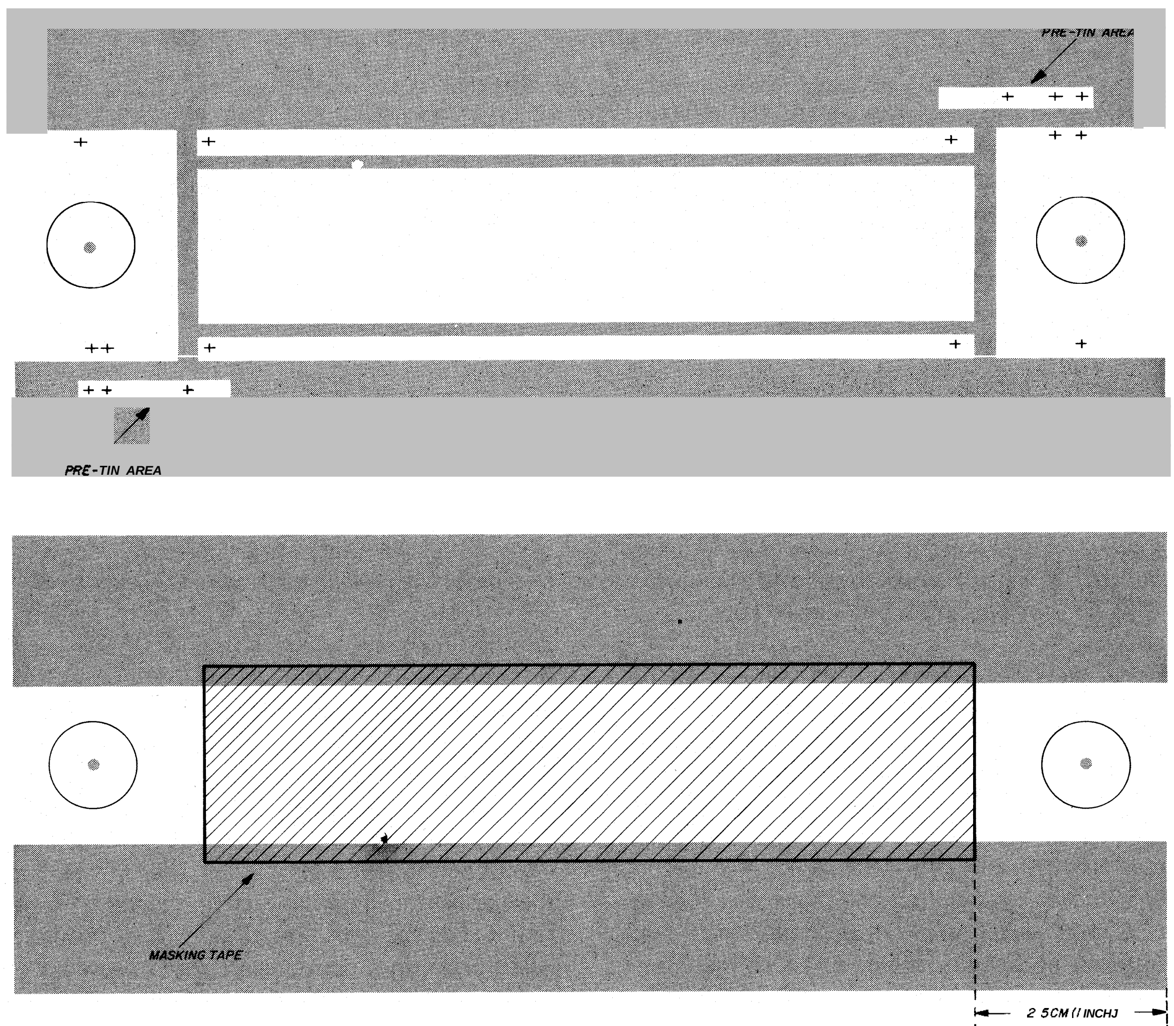


fig. 3. The top of the circuit board should be pre-tinned in the areas marked. The back plane is covered with masking tape as indicated.

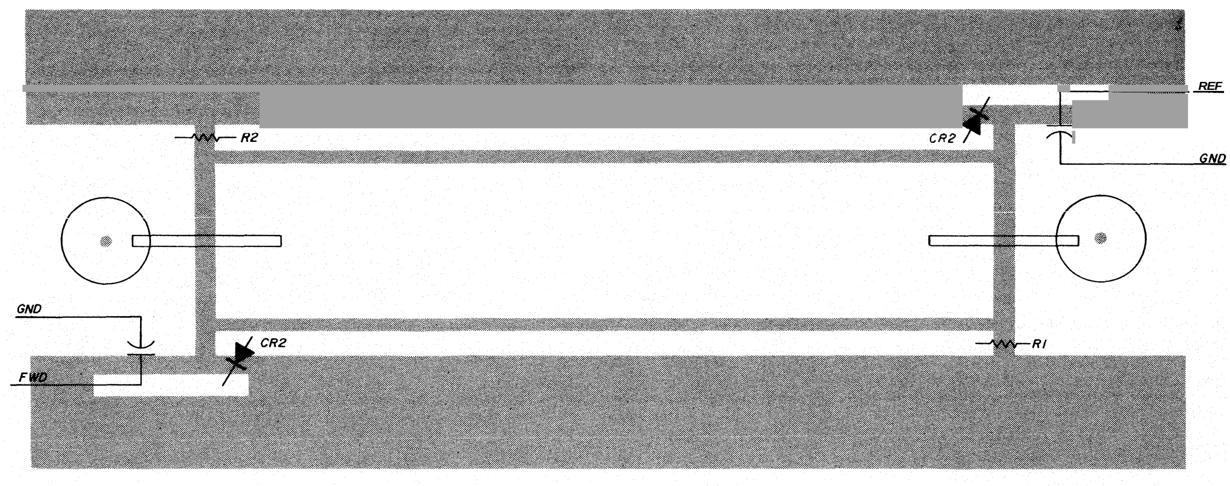


fig. 5. Parts placement diagram for the top of the printed circuit board.

finish the meter face with a light coat of clear finish. This will protect the face and hide the erasure marks. After the finish has dried, assemble the meter and mount the calibration knob.

testing and use

To test the completed unit, connect the transmitter to J1 and the antenna feedline to J2. Set the forward/reflected switch to the forward position, key the transmitter, and adjust the calibration control for a full-scale meter reading. Now, flip the switch to the reflected position and note the reading on the meter. Next, transpose the antenna and transmitter leads, key the rig, and adjust the calibration control to full scale again. Flip the switch to the forward position, and a reading close to that noted in the first test should be observed. This is due to the fact that the sense lines are the same. Transpose the antenna and transmitter leads again and the unit will be restored to normal operation.

The first antenna I put the meter to work on was my 40180-meter dipole. I took measurements on both bands, recording the readings as shown in **fig. 7**. The readings indicated the 80- and 40-meter elements had the lowest reflected readings at the bottom of the bands. Shortening the 40-meter elements

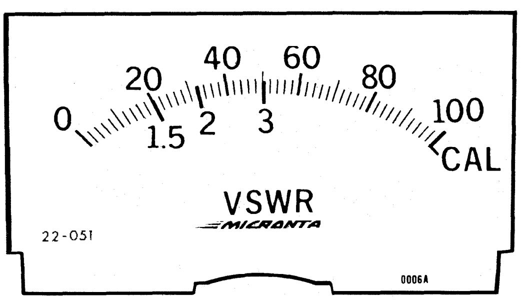


fig. 6. New calibration markings for the meter face.

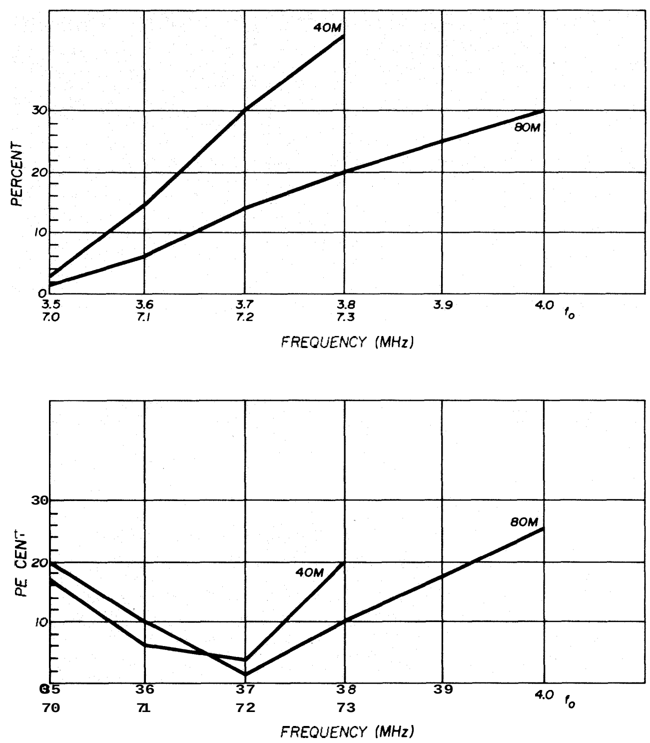


fig. 7. Readings obtained with the swr meter on a dipole for 40 and 80 meters.

first, and then the 80-meter elements, eventually yielded the reading shown in **fig. 7B.** This was more in keeping with the section of the band I use most. Now the antenna meter is left in the line and I use it to squeeze the last drop of output out of the little transceiver. I honestly cannot say that there has been a dramatic improvement in the overall performance of the antenna, but I am happy knowing that the vswr is more in keeping with the specifications for the transceiver.

ham radio

IC tester using the KIM-1

One useful tool that's sure to be appreciated by any electronic or computer experimenter is a reliable, fast, and easy method for testing digital integrated circuits. The concept presented in this article uses the KIM-1 microprocessor as one approach to testing 7400 series ICs. While only 7400-type numbers are listed, the program is equally applicable to those devices in the 74H, 74L, 74S, 74LS, and 74C series, as well as to others whose pinout arrangement is identical to that of the 7400-type shown. Also included is an optional "search and identify" procedure to help in the identification of unmarked devices.

program basics

The program itself is basically simple. A flow chart showing the concepts involved is seen in **fig. 1.** Operation of the program is broken into three main parts: selection of the proper set of input combinations from a table of possible combinations; application of inputs to the device under test; and comparison of the results with the set of correct results contained in another table.

A third table of "pointers," actually the lower address byte of locations in program page 1 and 3 where the input and results tables are located, is provided in program page 0. The third table is used to identify the starting address for each variable in the program for each type of chip listed. The correct set of pointers is selected by inserting into location 0000 an appropriate code number for the chip to be tested. The ICs that can be tested are shown in table **1.**

When an IC has successfully completed the testing, the device number appears in the address locations of the KIM-1 display. For type numbers above 7499, a hexadecimal digit is used to reduce the display to four digits. For example, a 74107 would be shown as 74A7 and a 74157 as 74F7. Failure of any tested chip to match expected results will result in FFFF being displayed in the address locations.

*A complete program listing, IC lists, and sequence tables are available by sending a self-addressed, stamped envelope to ham *radio,* Greenville, New Hampshire 03048

By Robert E. Babcock, W3GUL, 1706 Fawcett Avenue, McKeesport, Pennsylvania 15132

table 1. Listing of 14- and 16-pin ICs which can be tested by the KIM-1 microprocessor using the program described in this article.

14-pin ICs			16-pin ICs
7400	7415	7439	7442
7401	7416	7440	7445
7402	7417	7451	7446
7403	7420	7453	7447
7404	7421	7454	7448
7405	7422	7474	7449
7406	7426	7486	7485
7407	7427	74107	7490
7408	7428	74125	7492
7409	7430	74126	7493
7410	7432	74128	74145
7411	7433	74132	74148
7412	7437	74136	74151
7413	7438	74140	74153
7414			74155
			74157
			74158

To use the program, connections must be made between the KIM-1 and a suitable test socket. When testing a 14-pin IC, all pins are connected to the KIM-1 for maximum flexibility. However, when a 16-pin device is tested, all pins ***except*** ground and V_{CC} are connected to the KIM-1. The 14-pin 7400 series devices shown in table **1,** with three exceptions, all use pin 14 for V_{CC} and pin 7 for ground. The three exceptions are the 7490, 7492, 7493. The 16-pin devices all use pin 16 for V_{CC} and pin 8 for ground. Note that the KIM-1 ports cannot supply enough current as a V_{CC} source, therefore, a separate connection must be made.

The following procedure is recommended before inserting the first IC into the test socket.

table 2. Connections between the test sockets and the KIM-1 applications connector.

14-pin socket	16-pin socket	KIM-1 applications connector	KIM-1 port designation
1	1	8	PA7
2	2	7	PA6
3	3	6	PA5
4	4	5	PA4
5	5	2	PA3
6	6	3	PA2
7	7	4	PA1
8	9	14	PA0
9	10	16	PB5
10	11	13	PB4
11	12	12	PB3
12	13	11	PB2
13	14	10	PB1
14	15	9	PB0

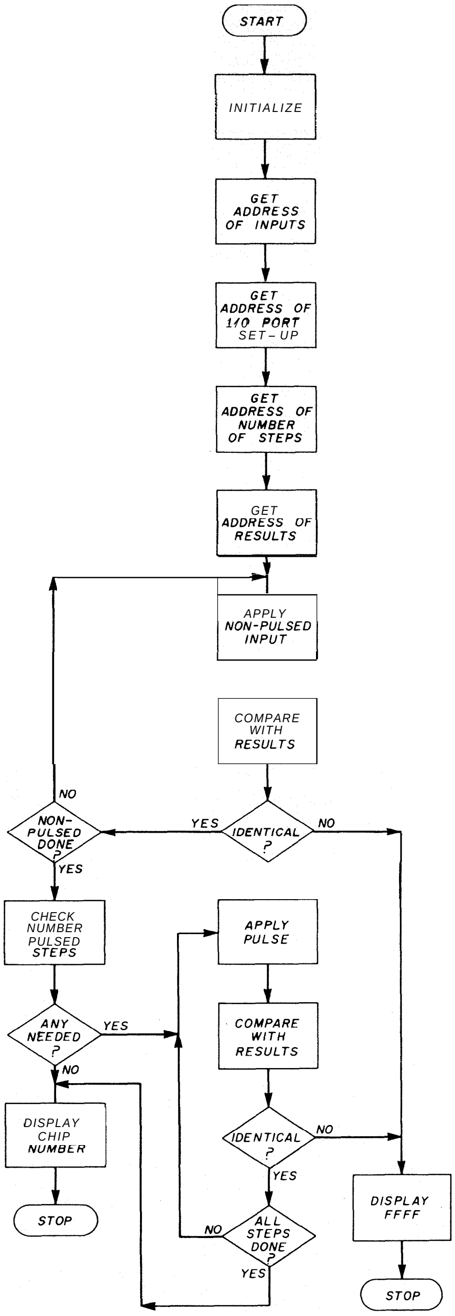


fig. 1. Flow chart for the program used with the KIM-1 microprocessor to evaluate and identify ICs.

1. Put the code for the selected IC into location 0000.
2. Put the starting address for the program in the ST vector (00 in 17FA, 02 in 17FB), then set address to 0200.
3. Press GO button — display should read FFFF IC.
4. Attach V_{CC} and ground connections to the correct pins of the IC. If adequate, the KIM-1 power supply may be used. If an external supply is used, the grounds must be connected together.
5. Insert the IC and press ST button — display should read type number if good, FFFF if not.
6. Repeat step 5 for each additional IC of the same type.

This procedure ensures that the input-output ports are correctly configured for the IC to be tested, which will reduce the possibility of damage to the 1/0 port due to shorting of outputs and "fighting for the bus." To change the code in location 0000, leave the test socket empty and press ST. The display will again read FFFF. Pressing the + button will put the address in location 0000.

A delay subroutine is included as part of the program so that the changing values may be observed by attaching indicators (with drivers) to the various ports. To increase the delay for this purpose, put FF in 02D6 and 07 in 0208. If observation is not desired, the delay may be removed by inserting NOP's (Op Code EA) in 0263, 0264, and 0265, and also, 0290, 0291, and 0292.

additional test capabilities

If you desire to include additional IC types, caution must be exercised. One factor that must be considered is the load imposed on the driving source by any input of a TTL chip. Some devices, such as the 7475, impose as much as four loads on certain inputs, far beyond the unbuffered capacity of the KIM-1. A second, less critical problem is the number of memory locations that are needed to provide the input and output sequences for some of the more complex ICs. Finally, in addition to the table of pointers in page 0, three other locations in that page are reserved for use by the program; these are OOEC, OOED, and OOEE. Any tables to permit testing devices not included in **table 1** should avoid these locations.

IC identification

Through a small program modification, it is possible to use the KIM-1 to determine the characteristics

table 3. Example of binary values necessary to check a 7400 quad, dual NAND gate.

KIM-1 port designation	IC pin number	pin function	binary entry to DDR	hex equivalent
PA7	1	in	1	D
PA6	2	in	1	
PA5	3	out	0	
PA4	4	in	1	
PA3	5	in	1	8
PA2	6	out	0	
PA1	7	ground	0	
PA0	8	out	0	
PB7	—	—	0	3
PB6	—	—	0	
PB5	9	in	1	
PB4	10	in	1	
PB3	11	out	0	6
PB2	12	in	1	
PB1	13	in	1	
PBO	14	V_{CC}	0	

of unmarked chips. To operate in this automatic search and identify mode, enter code 00 in location 0000, enter the program starting address (0200) into the ST vector, insert the unknown IC in the test socket, and press ST. The program change will cause the code in 0000 to be incremented by one if the IC fails the test for the device coded 00. If the IC coded 01 also fails, then the code will be incremented to 02 and so on until all codes for which input and output sequences have been provided have been checked. If no match is found, the display will show FFFF as before.

The *first* code which is satisfactorily matched will cause that device type to be indicated by the display. This can lead to errors in identification. As an example, suppose the IC being tested is 7420, a dual 4-input **NAND** gate. Investigation of the characteristics of the other devices in **table 1** will show that a 7413 is also a dual 4-input **NAND** Schmitt trigger. Thus, a 7420 would actually show as a 7413 in this mode. However, the fact that the device is identified as a dual 4-input **NAND** is helpful, and further tests can be made to completely identify the device. After the IC has been identified, remove the chip and press ST. This will make certain that the code is initially at 00 for the next chip tested.

The sequences of inputs and results, as well as the configuration for the input-output ports, can be determined by analysis of the chip and its connection to the KIM. Consideration must be given to the fact that any pin designated as an IC output must have the corresponding KIM-1 port set as an input, and ***vice versa.*** Unused IC pins, V_{CC}, and ground connections are made inputs to the KIM. Since a port location is designated as an output terminal by writing a **1** into the data direction register for that position, and similarly an input by a 0, the known pattern of a particular IC establishes the values which must be used to set up the ports for that device.

This procedure is probably best explained by an example using the 7400 two-input **NAND** gate. The pin arrangement for the 7400 and the truth table for each of the four identical gates contained in the chip are shown in **fig. 2.** By referring to the pinout, a table of the KIM-I port designations, the IC pins attached

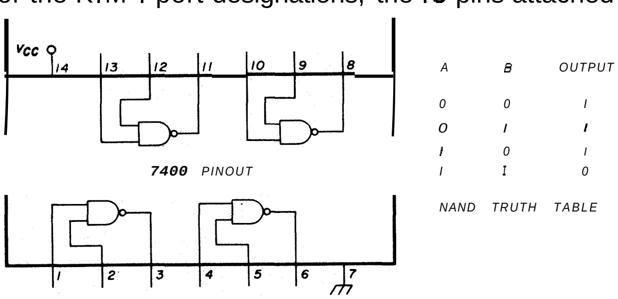


fig. 2. Pinout diagram and truth table for the 7400, a quad, dual-input NAND gate.

table 4. Binary inputs for the 00, 01, and 10 testing combinations.

KIM-I port designation	IC pin number	input 00	input 01	input 10	results 00	results 01	results 10
PA7	1	0	0	1	0	0	1
PA6	2	0	1	0	0	1	0
PA5	3	0	0	0	1	1	1
PA4	4	0	0	1	0	0	1
PA3	5	0	1	0	0	1	0
PA2	6	0	0	0	1	1	1
PA1	7	0	0	0	0	0	0
PA0	8	0	0	0	1	1	1
PB7	—	0	0	0	0	0	0
PB6	—	0	0	0	0	0	0
PB5	9	0	0	1	0	0	1
PB4	10	0	1	0	0	1	0
PB3	11	0	0	0	1	1	1
PB2	12	0	0	1	0	0	1
PB1	13	0	1	0	0	1	0
PBO	14	0	0	0	1	1	1

to each, and the binary value which must be placed in each location can be made (see **table** 3).

As can be seen from the hexadecimal equivalents for the binary values needed at the data direction registers, D8 must be written into PADD (1701) and 36 into PBDD (1703). The truth table also indicates that the same hex digits can be used to apply the 1-1 input combination to all four gates simultaneously. Thus, these values fulfill two functions and thereby save space in the input sequence table.

The result digits can also be determined by reference to the truth table; each 1-1 input should result in an output of 0. In fact, when the ports are sampled for correct response, the only change from the written value will occur at PBO where V_{CC} will cause a 1 input at that port. The result digits would then be D8 at Port A and 37 at Port B. The values to be written into PAD (1700) and PBD (1702) (to apply the 1-1 combination) and the values which should read at PAD and PBD if the chip responds correctly have now been determined. The combinations for the remaining 00, 01, and 10 inputs are shown in binary form in **table 4.**

The portion of the program that applies the pulses for the 7490, 92, and 93 or other similar counters requires only two input entries for all of the pulses applied. The pulse is obtained by starting with the input at a 1 level, applying a 0 level, then returning the pin to a 1 level. After each pulse, the index that determines the input values is decremented two places; the next pulse is applied by incrementing twice and so on. This also saves program space and permits the testing of more chips with a given set of tables. The process of determining the needed tabulations is not difficult and it should be possible to find combinations that can test virtually any logic chip.

ham radio

simplified capacitance meter

A revised version of the capacitance meter uses a second 555 timer to replace the original programmable unijunction

In a previous article I described what I thought to be a very easy-to-build, direct-reading capacitance meter.[1] It would linearly read on a meter any capacitance value from 1.0 μF down to as low as 1.0 pF, just the ticket for measuring those unmarked variables and micas you can get so cheaply at the surplus and flea markets.

Well, the article proved to be quite popular, at least I got a lot of letters and phone calls about it. Only trouble was most of them asked, "Where can I get that programmable unijunction transistor (PUT), or what can I substitute for it?" Information on substitute devices was published,[2] but I'm still getting phone calls nearly three years later. The circuit presented in this article is an alternative; no PUTs, just two of the popular 555 timer ICs.

theory

As it turns out, the output pulse width of a 555 timer is a linear function of the timing capacitance connected to it. If the 555 is triggered with a constant frequency, then the average dc value of the resulting pulse train is a linear function of pulse width. Putting these two facts together means that a dc meter connected to the 555's output will provide a linear indication of the timing capacitor's value. Decade capacitance ranges are implemented by switching the value of the 555 timing resistor.

Since the output waveform of the 555 doesn't quite get down to zero volts between pulses, and since there is some stray capacitance in any circuit, zero adjustments are needed to buck out that small voltage and stray capacitance. In the original circuit, I used a PUT and a transistor inverter to generate the constant-frequency trigger for the 555; here I have replaced the original circuit with another 555 operating in the free-running mode. A more detailed discussion of theory may be found in **reference 1.**

construction

The schematic of the capacitance meter is shown in **fig. 1.** It is convenient to mount most of the parts, including the four trimpots, on a piece of perforated board which is mounted to the meter terminals. The two 555 ICs will plug into a 16-pin IC socket.

Layout is not critical, but the wiring associated with the range switch and test terminals should be as short as possible to minimize stray capacitance. Use a metal box to house the instrument, and connect circuit ground to the box. Make sure the 2-pole, 5-position switch is wired so that trimpot R11 is selected when the range switch is in the 100 pF position.

All resistors may be 1/2 watt, 5 per cent, but greater overall accuracy will result if the five resistors connected to the range switch are of one per cent tolerance. C1 should have a tolerance no greater than 10 per cent.

power supply

The circuit is designed for a six-volt power supply, to facilitate battery operation. However, the battery drain is heavy, up to about 50 mA, and varies with the range-switch setting. Since calibration stability is directly related to supply-voltage stability, batteries really shouldn't be used unless portable field-use is required. If batteries are used, they should be alkaline types.

My solution to the problem of infrequent use *vs.* cost of a built-in power supply is as follows. A simple

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248

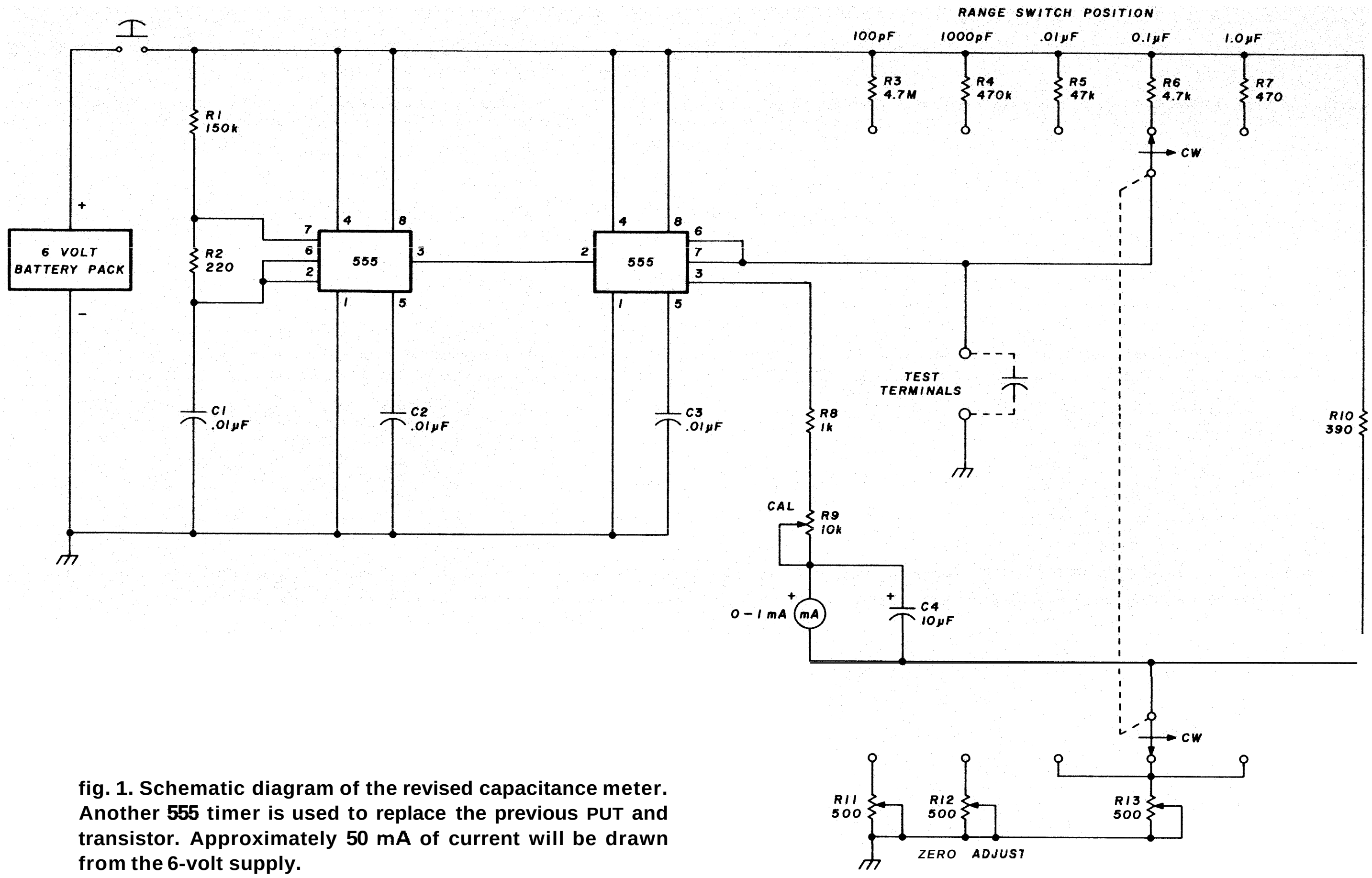


fig. 1. Schematic diagram of the revised capacitance meter. Another 555 timer is used to replace the previous PUT and transistor. Approximately 50 mA of current will be drawn from the 6-volt supply.

six-volt zener regulator, such as shown in **fig. 2,** is mounted in the capacitance meter box with two wires brought out. These wires are connected to a 12-volt bench power supply when I need to use the meter. One of the three-terminal, six-volt IC regulators should give even more stable performance than a zener.

calibration

The three zeroing trimpots, R11, R12, and R13, must be adjusted with nothing connected to the test terminals. Zeroing is accomplished by adjusting these trimpots for a zero meter reading when the push-button is depressed. Adjust R11 with the range switch in the 100-pF position. R12 is for the 1000-pF range, and R13 is for the three highest ranges. If the meter cannot be zeroed, the trimpot is probably defective, having too much "end resistance." Smaller value trimpots may be used, down to 200 ohms, or possibly smaller, depending on the stray capacitance in a particular version of construction.

After the zeroing trimpots are adjusted, connect a 100-pF mica capacitor, with a tolerance of five per cent or better, to the test terminals. Set the range switch to 100 pF. Depress the push-button and adjust trimpot R9 for a full-scale reading on the meter. Calibration is then complete.

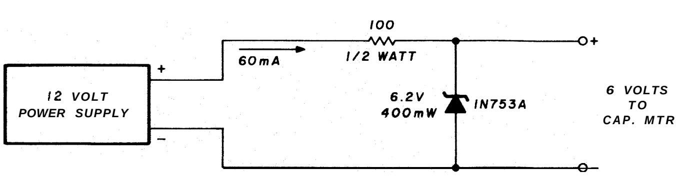


fig. 2. Zener regulator for use with the capacitance meter.

operating tips

When the capacitor being measured is too large for a particular range-switch setting, the circuit may be driven out of its linear range of operation. Under these conditions, the meter may read less than full-scale even though the actual capacitor value is more than the full-scale setting. To avoid such erroneous readings, test an unknown capacitor on the 1.0-µF range first, then move the range switch to lower settings until a usable reading is obtained. Keep the original calibration capacitor handy for calibration checks.

references

1. Courtney Hall, WA5SNZ, "Direct-Reading Capacitance Meter," ***ham radio,*** April, 1975, page 32.
2. Courtney Hall, WA5SNZ, Comments, ***ham radio,*** October, 1975, page 31.

ham radio

improvements to the Measurements Corporation 59 grid-dip oscillator

Used in labs for years, the model 59 grid dipper is still a great instrument. Presented here are some ideas for updating this old workhorse

This article describes modifications to a Measurements Corporation model 59 grid-dip meter. However, it pertains to most other tube-type grid-dip meters as well. The B&W, Heath, and Millen grid-dip meters are basically similar (in the rf head) and should be adaptable to the circuitry described.

I found an rf head from a model 59 in a local surplus emporium without the meterlpower-supply module and with only four of its normal complement of seven hf-vhf coils. Since the model 59 (in my opinion) is the best grid-dip meter ever built, I felt that a reconstruction project was worthwhile, so a new meterlpower-supply unit was built to replace the original. Here's the story.

circuit design

Rather than copy the Measurements Corporation circuit, I designed a new supply using all solid-state components. The 5Y3 rectifier was replaced with silicon rectifiers, and the OD3/VR150 voltage regulator tube was replaced with a 10-watt zener. As neither an O.E.M. transformer or choke was available, I used standard Triad components. I added a B+ switch; this allows you to first switch on the power, which heats the 955 heater in the rf head, and *then* switch on the B+ after a few moments wait –

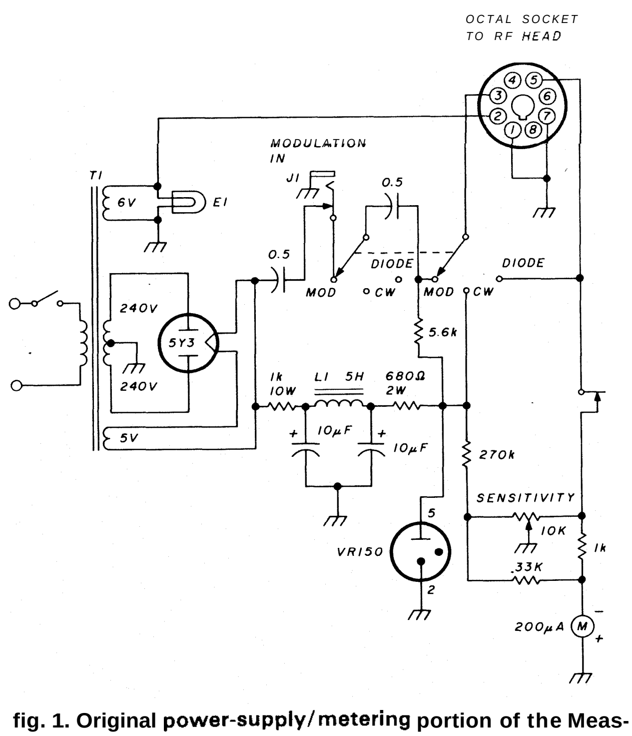


fig. 1. Original power-supply/metering portion of the Measurements Corporation model 59 grid-dip oscillator.

By Hank Olson, W6GXN, P.O. Box 339, Menlo Park, California 94025

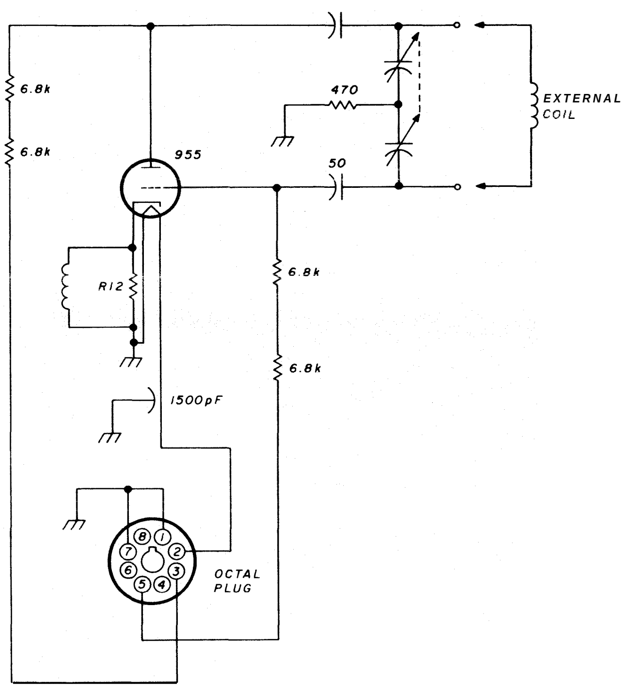


fig. 3. Schematic of the Measurements model 59 rf head. The 955 could perhaps be replaced with a modern high-performance fet.

a conservative addition because replacement 955s are now about $10 on the surplus market.

Looking at **fig. 1,** which shows the original schematic diagram, we see tubes used for the rectifier and voltage regulator. **Fig.** 2 shows the replacement supply using solid-state devices. The rf-head is shown in **fig.** 3, and is the same for original and modified GDOs.

internal modulation

Note that in **fig. 1,** when the function switch is placed in the MOD position and no plug is inserted into closed-circuit jack J1, 120 Hz is capacitively coupled to the B+ going to the rf head. The equivalent circuit is shown in **fig. 4.**

This simple method of plate-modulating the GDO leaves a lot to be desired, mostly because of the prevalence of 120 Hz as an incidental modulation on almost every oscillator. So this is the question: Is the 120-Hz modulated carrier you're hearing really from the GDO or from some other source?

To make the modulation frequency more distinctive, I added a simple, solid-state, 1-kHz modulator. This modulator uses a four-layer diode relaxation oscillator followed by a solid-state amplifier to increase level **(fig. 5)**. Although this modulation isn't sinusoidal, it has a more definitive pitch (400 Hz would be another good frequency and could easily be obtained by changing the RC constants in the relaxation oscillator).

I made no attempt to modify the rf head, although it's tempting to make the GDO entirely solid state by replacing the 955 tube with an fet.

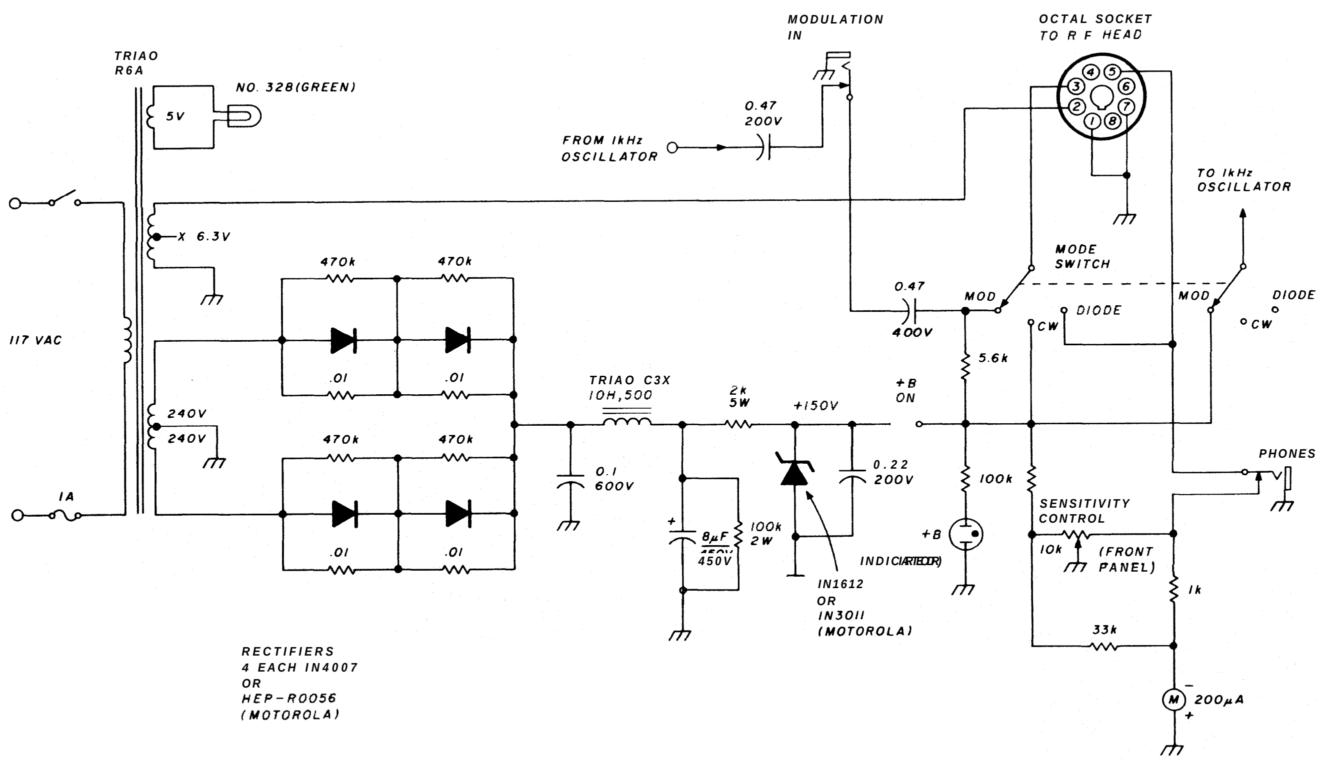


fig. 2. Improved power supply for the Measurements Corporation model 59 using modern devices. Design results in higher efficiency and longer life for this essential part of the circuit.

I built the meter/power supply into an LMB-W1A box, and screwed the wooden coil block onto the back. I then drilled a small slot into one upper edge of the W1A cabinet so that the hook on the rf head could engage it.

Measurements Corporation's successor, as well as their "standard of the industry" grid-dip meter, are both very much alive. The company is now a subsidiary of the McGraw Edison Company, Manchester, New Hampshire. The new GDO is called the model 159, but it remains pretty much the same old unit as its predecessor, the model 59. Incidental-

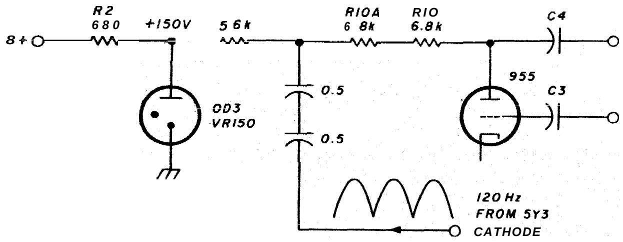


fig. 4. Equivalent circuit in the model 59 when using the unit for internal modulation. The 120-Hz modulation from the 5Y3 cathode, which appears on the 955 oscillator plate, causes ambiguous results. A better circuit is shown in *fig.* 5.

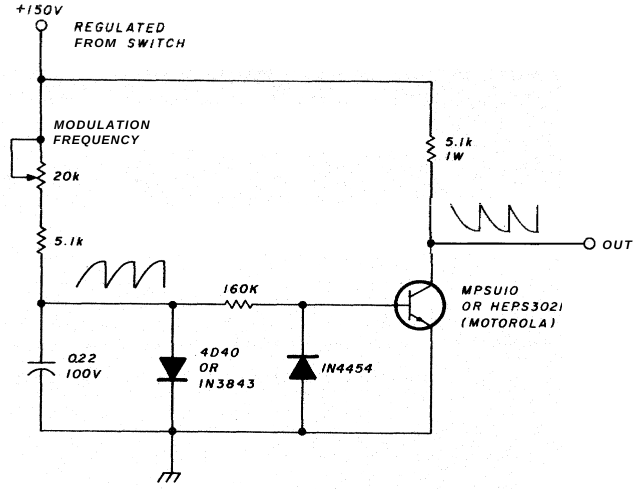


fig. 5. Simple modification to the Measurements Corporation model 59 GDO to previde l-kHz modulation.

ly, for those (like myself) who have a couple of coils missing, it's still possible to get single replacements from McGraw Edison; they are located at Grenier Field Municipal Airport, Manchester, NH 03103.

ham radio

the ham notebook

positive lead keying for the Heath HD—10 keyer

This is a simple and fast modification for the Heath HD-10 keyer for use with the newer solid-state transmitters and transceivers that require positive keying to ground. The HD-10 is designed for negative keying. Examination of the circuit in the manual shows a PNP transistor (2N398A) with the collector to the keyed line. On page 15 of the assembly manual, the lead arrangement for the keying transistor is shown with instructions to bend the base lead closer to the emitter. For solid-state units, remove Q8 from its socket, bend the base lead closer to the collector, and replace the transistor in its socket.

In my case, I used a 2N1305. Since the keying current is small (5 milliamperes), most any PNP transistor will do if one wishes to keep the original 2N398A for resale purposes. This modification takes about ten minutes because of the time required to remove the printed circuit board and replace it. No external buffer is required.

Richard Jasper, W4VAF

cleaner audio for the R-4C

The Drake R-4C offers many attractive features for the weak-signal enthusiast (excellent AGC, noise blanker, and selectable i-f filters). However, a major annoyance is the large amount of hiss in the audio. This hiss can be traced to the 50-kHz BFO feeding into the audio circuit. An effective cure is to roll off the audio response and put a filter between the product detector and the subsequent audio stages.

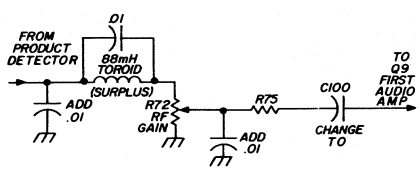


fig. 1. Schematic diagram of the changes made to the Drake R4C to eliminate the 50-kHz BFO feedthrough into the audio system.

The following simple circuit changes (see **fig. 1**) are an effective cure for the problem.

1. Change C99 to 0.1 μF

2. Connect a 0.002-μF capacitor in parallel with R83

3. Change C100 to 0.47 μF, improving the low-frequency audio response

4. Connect a 0.1-μF capacitor from the wiper of R72 to ground

5. Connect an 88-mH toroid, with a 0.01-μF capacitor in parallel, between the audio gain control and the product detector output

6. Connect a 0.01-μF capacitor across the output of the product detector

7. Replace C103 with a 0.005-μF capacitor

8. Change C175 to 100 μF

These changes will eliminate the hiss and also clear up the low-frequency distortion.

Steve Powlishen, K1FO

crowbar circuit for the HWA-2036-3

Recently I had a problem with my HWA-2036-3 power supply which caused the driver transistor to fail. The cause of this failure was that the collector of the 2N3055 pass transistor and the mounting screw (on rear chassis) failed to make contact, thereby placing the entire load on the driver transistor. An inspection showed that I had forgotten the small 3-mm (1/8-inch) spacers in the plastic cover over the pass transistor. These spacers are very important because they force contact between the mounting screws and the pass-transistor case.

The resultant failure caused approximately 24 Vdc to be applied to the HW-2036. This increase in voltage caused the receiver's audio chip (TBA820) to burn out. I was lucky that nothing else went up in smoke. If the HW-2036 had been in the transmit mode, many hard to replace components would have been destroyed.

To solve this problem, I decided to build a simple "crowbar" circuit, within the power supply, which would blow the dc fuse if the output voltage exceeded 15 volts. The crowbar circuit is quite simple and easy to construct; it can be placed on a small printed circuit board or mounted on terminal strips. But, modification of the HWA-2036-3 requires some mechanical work. The dc fuse has to be relocated to a point ahead of the regulator instead of at the power supply output. This was done to protect the transistors and regulator chip. When the fuse is blown, it removes voltage from the entire regulator.

The following steps are required for this modification:

1. Referring to the circuit board X-ray picture on page 30 of the manual, cut the trace at a point opposite the diode D2 (it's pretty wide here). This will separate the filter section from the regulator section.

2. Drill two small holes (no. 55 drill [1.3 mm]) in the trace just below the one cut in step 1. This trace connects to Q1's collector and one side of R2. (These holes will be used for wire connections.)

3. Disconnect the red wire from point C on the circuit board and resolder it in one of the holes just drilled.

4. Disconnect all wires from the dc fuse socket.

5. Disconnect the wires from the DC Regulated post on the printed circuit board.

6. Disconnect the white wire from A on the board.

7. Solder a jumper wire from point **A** to DC Regulated on the board.

8. Solder the red wire which goes through the grommet to the plug on the rear panel to the DC Regulated post on the printed circuit board.

9. Solder a red wire to the remaining drilled hole and solder the other end of this wire to one side of the dc fuse holder.

10. From the other side of the dc fuse holder, solder a red wire to point C on the board. (Just above diodes D1 and D2.)

This in effect relocates the dc fuse between the filters and the regulated section of the power supply (referring to the circuit diagram, at a point between C1 and R2/Q1's collector).

Now that the power supply has been modified, the crowbar circuit can be constructed and installed. Referring to **fig. 1**, you'll note that there are only 5 components, an **SCR**, 2 resistors, a Zener diode, and a bypass capacitor. Also note that there are only 3 connections, marked **PLUS, FUSE** and **GND.**

PLUS is connected to the printed circuit board at the DC Regulated terminal post.

FUSE is connected to the dc fuse holder on the same terminal as the red wire from R2/Q1.

GND is connected to chassis ground.

Once the **SCR** has fired and blown the dc fuse, the transceiver is protected from damage due to overvoltage. But, don't just replace the dc fuse and expect it to take off and work properly. The reason the fuse blew in the first place indicated trouble in the power supply. First, disconnect the transceiver, take the case off the power supply, and discharge the filter capacitor, C1. The **SCR** will blow the fuse and save your transceiver, but it will not discharge the filter capacitor. After the capacitor has been discharged, normal troubleshooting procedures can be used to locate faulty components. As a final note, use a fast blow fuse in the dc fuse holder.

James T. Conner, W3HCE

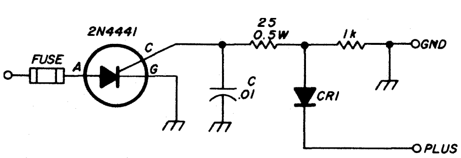


fig. 1. Schematic diagram of the simple crowbar circuit used to protect the output of the HWA-2036-3 power supply. The Zener diode is rated at 15 volts, 1 watt. The SCR is rated at 100 volts, 8 amps, but other SCRs can be used.

external speaker and tone pad for the HR-2B

The Regency HR-2B transceiver provides screw terminals at the rear for connecting an external speaker. I found this method to be rather awkward since I'm frequently removing the rig from the car. By carefully prying the ears up on the terminal strip, the individual screw terminals may be removed. The remaining holes are large enough to permit the installation of miniature phone jacks. I used one closed-circuit jack for normal-through internal speaker operation and an open-circuit jack for plugging in the ***Touch-Tone**** **pad.** When the mating plug is inserted into the speaker jack, the internal speaker is muted and the external speaker is operative. Thus, connect/disconnect for mobile/base use is rapid and easy.

Paul Pagel, N1FB

*Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.

setting 2-meter receivers using hf harmonics

In the absence of a counter or other suitable piece of test equipment, a surprisingly good job of setting a 2-meter fm receiver on frequency can be done using the harmonics from a 15- or 10-meter transmitter. This assumes that the transmitter or its companion receiver has reliable dial calibration. In my case, I depend on my Collins 75A4 as the frequency standard, since its dial can be read to 100 Hz increments with little difficulty, and it tracks to within 100 Hz between 100 kHz calibration points.

To set the 2-meter receiver on frequency, divide the desired channel frequency by five, tune the 10-meter receiver to that frequency, zero beat the transmitter, and zero the 2-meter receiver on the transmitter harmonic. If your transmitter doesn't cover all of 10 meters (147 MHz/5 = 29.4 MHz), the same trick can be done on 15 meters using the seventh harmonic for frequencies above 147 MHz. (147 MHz/7 = 21.0 MHz). It may be necessary to temporarily remove your low-pass filter to hear these harmonics. Make sure you are not tuning up on top of someone on 15 or 10 when you are doing this.

Using this technique, it should be possible to set a 2-meter fm receiver to within 1 kHz or less. Once the receiver frequency is set, most radios have provision for zeroing the transmitter to the receiver.

John Becker, K9MM

$2.00

radio

DECEMBER 1978

ham radio magazine

DECEMBER 1978

volume 11, number 12

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

United States: one year, $12.00
two years, $22.00; three years, $30.00

Canada: one year, $14.00
two years, $26.00; three years, $36.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

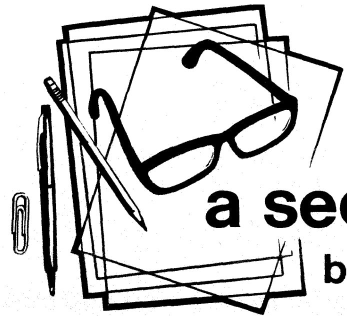
Foreign subscription agents are
listed on page 107

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

contents



a second look

by Jim Fisk

Small-scale integration, large-scale integration, counters on a chip, 64K memories — are the possibiiities iimirea only by the designer's imagination? I can think of no other segment of the electronics industry which has come so far, so fast, as integrated circuits. When vacuum tubes were 20 years old they were mired down by bitter patent litigation which had slowed development work to a standstill. Marconi, the Fleming patent holder, even placed a full-page advertisement in QST warning that amateurs who even ***used*** other than Marconi-licensed tubes for radio purposes were "liable to a suit for injunction . . . and they will be prosecuted." A scare tactic, perhaps, but it was effective and held back the progress of the entire radio industry.

No such patent litigation ever developed in the transistor industry because the inventors placed the basic idea in the public domain, so by the time transistors reached their 20th birthday there were more than 2000 registered types, and the industry was thriving. Circuit development, however, was stymied by engineers schooled in vacuum-tube techniques who struggled to make solid-state devices work as well as tubes. If you used any of the early solid-state ham gear you know that they were not always successful. In fact, there are some amateurs who still believe that, in a communications receiver, a solid-state front end is inherently inferior to a vacuum tube. In truth, solid-state front ends are actually several hundred times ***better*** than their vacuum-tube counterparts — both in strong-signal handling and sensitivity — but only when properly designed.

The rapid growth of integrated-circuit technology has been hampered neither by patent squabbles nor inept circuit design — to engineers who cut their teeth on transistors. ICs were the next logical step. So in this, their 20th year, ICs have already affected every industry, every level of society, and the end is nowhere in sight.

It was during the summer and fall of 1958 that Jack Kilby built the first integrated circuit at Texas Instruments. Other semiconductor manufacturers had been working on ways to miniaturize electronic circuits, but most of them relied on miniature discrete components. Kilby was the first to use semiconductor material for both the active and passive elements (resistors and capacitors) to build a complete circuit on a single piece of germanium. (Germanium was used because germanium manufacturing techniques were well established, and those for silicon were not.) Kilby's first circuits, a phase-shift oscillator and multivibrator, demonstrated the feasibility of this approach. On top of the germanium substrate were the contacts of the diffused transistors, junction capacitors, and resistors. A gold-plated metal frame protruded from the lower surface of the wafer and thermally-bonded gold wires were used for connections between those elements not linked by the substrate itself.

The first circuits were crude by today's standards — large and irregular; the photo masks and resists necessary for precision IC manufacturing were yet to be developed, so the patterns were hand painted on the semiconductor chip with black wax. About the same time Kilby was working on the first ICs in Texas, Fairchild Semiconductor developed the *Planar* process; this process made semiconductors more reliable and less expensive to produce, and greatly accelerated IC progress and acceptance.

In the 20 years since those early discoveries, prices have decreased and the number of circuits per unit area has increased dramatically. In 1962, for example, a typical IC flip-flop was about 2 mm square; a similar circuit in 1968 was ten times smaller, and today an entire 8-digit frequency counter with all control functions is available on a single chip. In 1962 a decade counter required several counter chips and logic gates for a total cost of twenty or more uninflated dollars; by 1968 the cost of a one-chip decade counter had dropped to about seven dollars; today you can buy a TTL decade for fifty cents! And you must remember that 1962 counter did well to count reliably at 1 MHz; the 1978 counter is guaranteed to30 MHz.

Progress in the field of linear ICs has been just as impressive, and the many functions which are built into modern amateur transceivers are a direct result. Some of the newer transceivers have 25 or more controls on the front panel; can you imagine how many 6-foot racks of vacuum-tube circuits it would take to do the same job? And even if you had room for the racks, you wouldn't want to pay for the electric power to run them (and cool them). If progress in the next ten years is as rapid as in the past ten, the commercial equipment that will be available to amateurs in 1988 really boggles the mind.

James Fisk, W1HR
editor-in-chief

comments

transient suppression

Dear HR:

I have read the article in your June 1978, issue, "Protecting Solid State Devices From Voltage Transients," with great interest. Mr. Prudhomme has written a very informative article that apparently was well researched, but I question his statement that varistors are excellent in suppressing transients on the line side of equipment power supplies.

The surge energy capacity of the MOV is proportional to the volume of its active material, and the breakdown voltage is proportional to the thickness of its material. Page 57 of the GE transient voltage suppression manual, referenced in Mr. Prudhomme's article, addresses itself to (1) Overstress Near Ratings, and (2) Extreme Overstress. The varistor has its place in the proper environment, but its application must be carefully chosen. As pointed out in Prudhomme's article, transients cover a wide range of frequencies and amplitudes. How do you select the proper component to protect against an unknown?

As a manufacturer of transient voltage suppressors, we discarded the varistor a number of years ago because it will age, and aging increases the leakage current. The higher the current, the faster the aging — a runaway condition. Also the higher the ambient temperature, the sooner this can happen. We have marketed a 110-volt plug-in unit that is extremely effective, using a *Tranzorb** as the heart of a circuit; it is cost comparable to the circuits you describe. A comparison report of *Tranzorbs* versus various metal oxide varistors is available from General Semiconductor Industries.

Nanosecond ranges are becoming antiquated; picosecond ranges are the ones you must be concerned with.

Stephen J. Sorger,
Vice President
W. N. Phillips, Inc.
Lake City, Michigan

English, *si!* Metric, no!

Dear HR:

I have subscribed to ***ham radio*** almost since its inception, and have enjoyed it more than any of the other amateur magazines. When you added the metric units after the English measurements it was annoying, but I just cussed the confusion under my breath; when you made metric primary and English secondary, however, it was more than I could put up with. I refuse to have the metric system shoved down my throat.

S. D. Brokhausen, W5VMN
Georgetown, Texas

metrics made easy

Dear HR:

By choosing metrics for your primary system of measurement, you have again set the pace for the other amateur radio magazines. It's only through everyday usage that people will become conversant in the metric system; once they can speak ***metric*** as well as they do English, opposition to metric conversions will disappear.

David L. Campbell, W1CES
Boston, Massachusetts

* *Tranzorb* is a registered trademark of General Semiconductor Industries, Inc., Post Office Box 3078, Tempe, Arizona 85281.

radiation hazard

Dear HR:

W1HR's editorial in the August issue should serve as a caution to the ham community to exercise great care in the operation of vhf and uhf installations. This is particularly timely in view of the accelerating trend toward these higher frequencies.

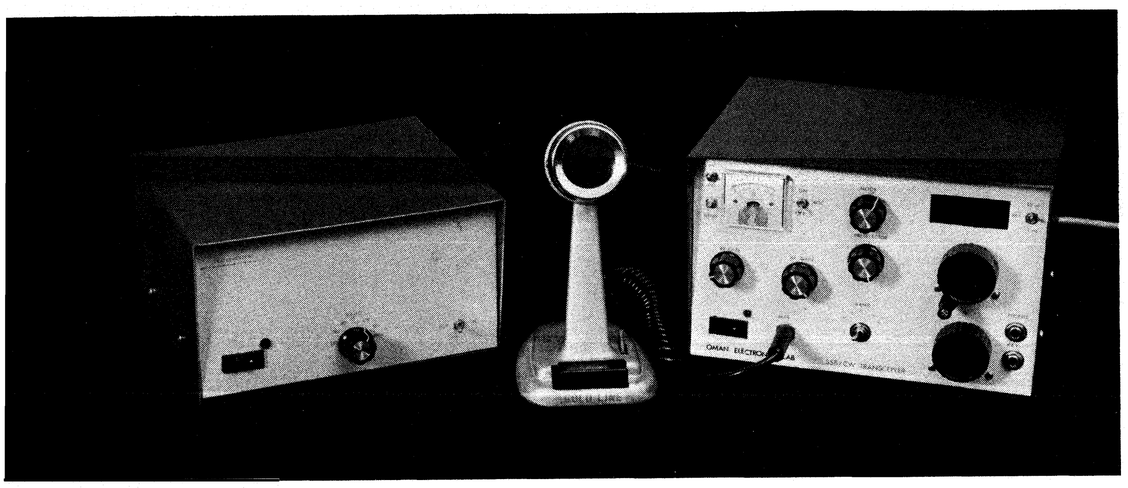
There is, however, another possible hazard area which seems to have been ignored by amateurs and commercials as well. This is the use of uhf/vhf handy-talkies. Here, too, their use is multiplying and our improving technology is compounding the problem. In the past, 1 watt was a high power for these units; now 5-watt units are common, and higher power is probably on the way.

Now it seems that this should be low enough to be negligible, but the inverse square law does apply, so that power densities in close proximity to the antenna can reach significant levels. In operation, the HT antenna is situated but a few inches from the eyes of the operator.

A quick calculation, based on an isotropic source radiating 1 watt, shows that the 5 mW/cm^2 occurs about 4 centimeters from the antenna. A 4-watt power would thus give this power level at 8 centimeters, which is about the distance that an HT would be located from its operator's eyes if he close-talks the unit — as most people do.

I recognize that the 5 milliwatt level is recognized as being safe by most authorities. Furthermore, we are protected by the fact that most HTs will not function for an extended period of time at the 5-watt level. But I do not believe that it would be prudent to continue to ignore this possible source of eye damage, and we certainly should limit HT power levels to no more than 5 watts.

Harley C. Gabrielson, K6DS
La Mesa, California

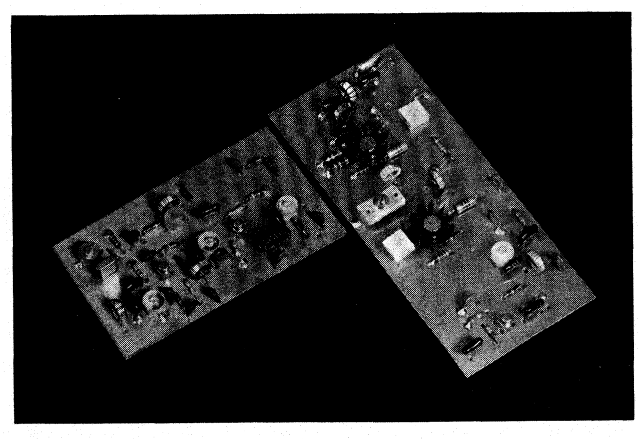


low-power transverter

for the high-frequency amateur bands

Is your QRP transceiver band limited? Try this simple transverter to obtain multiband operation

Low-power transceivers are becoming more popular each day on the amateur bands. Many of these rigs are single or dual band, usually for 40 and 20 meters. My home-brew 20-meter ssb transceiver fits this category. The need for 80-meter operation resulted in the transverter described here. A savings of both money and construction time can be realized if a transverter is used — rather than constructing an entire assembly for another band. The transverter eliminates the need for another ssb filter, VFO and associated components, and dial drive and indicator. The transverter can be used with any 20-meter transceiver or transmitter/receiver combination. Com-



Printed-circuit boards for the low-band transverter. Receiver board is to the left, transmitter board to the right.

By Mark Oman. WAØRBR, 427 East 11 St., Loveland, Colorado 80537

mercial QRP transceivers, such as the HW-7 and Ten-Tec PM series, can also be used with this transverter. Both ssb and CW modes can be used.

I built the transverter for 80 meters, but other bands can also be covered. With appropriate component changes for inductors and capacitors, the same PC-board layout can be used. **Table 1** lists values for these components for 160, 40, 15, and 10 meters. The transverter provides CW or ssb output of approximately 1-2 watts and needs a minimal amount of rf drive.

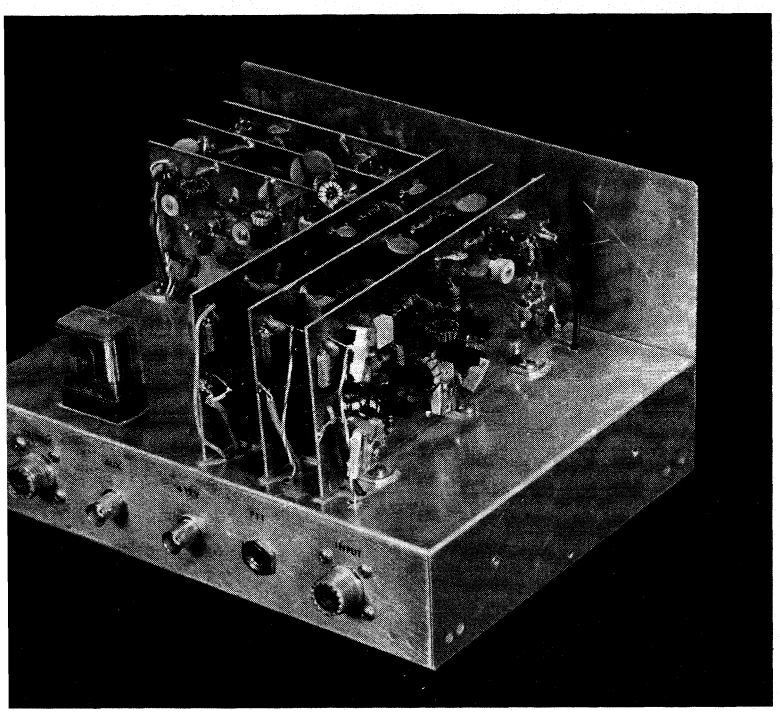
operation

The basic principle of this transverter is to receive a 3.5-4 MHz signal, convert it to 14 MHz, and transmit a 3.5-4 MHz signal by converting 14 MHz to 3.5 MHz. A crystal oscillator of 18 MHz provides the signal to convert these frequencies. The oscillator operates on both transmit and receive modes.

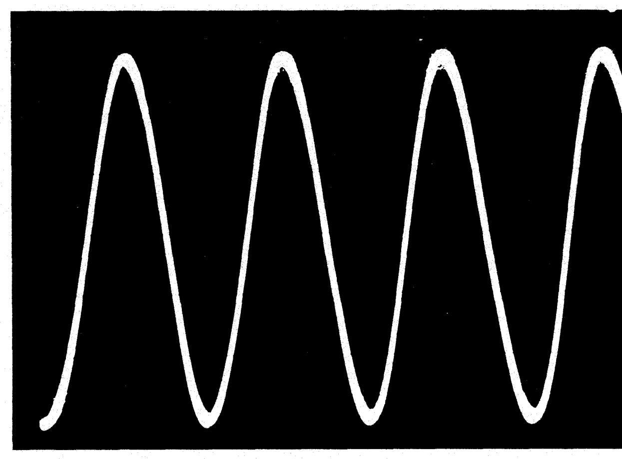
receive mode

Referring to **fig. 1**, Q3 operates as an 18-MHz, third-overtone crystal oscillator with output tuned by L5 and C14. Q1 is a dual-gate mosfet rf amplifier tuned to the 80-meter band. The 80-meter signals are mixed with the 18-MHz oscillator output in Q2, resulting in a 14-MHz output across L4. A 50-ohm

Rear view of the transverter showing the 10, 40, and 80 meter circuit board assemblies. The receiver converters are in the background, while the transmitter converters are in the foreground.



Transverter output on 80 meters showing srb (upper) and CW (lower) modes as viewed on a Hewlett-Packard model 182 oscilloscope.



output impedance is obtained by using a capacitive divider across L4.

transmit mode

Q1 operates as a mixer, this time mixing 18 MHz with the 14-MHz input signal to produce a 4-MHz signal at Q2 base. Q2 is a class-A amplifier with about 100 mW output to drive Q3. A "T" network matches Q2 collector impedance to Q3 base impedance. Q3 is a linear amplifier. Its bias is adjusted by R13. L4, L5, C15, C16, and C17 are a pi network to match the Q3 collector impedance to 50 ohms. (This matching network also provides good harmonic attenuation.)

The transverter matching network design was based on the excellent articles in reference 1. Design was a lot easier than anticipated and circuit debugging was fairly easy.

Attenuator **Z1** is a critical part of the circuit for transmitting clean signals. Most QRP rigs will need about 20 dB attenuation, as only 10-20 mW is needed to excite Q1. Too much drive will cause distortion on

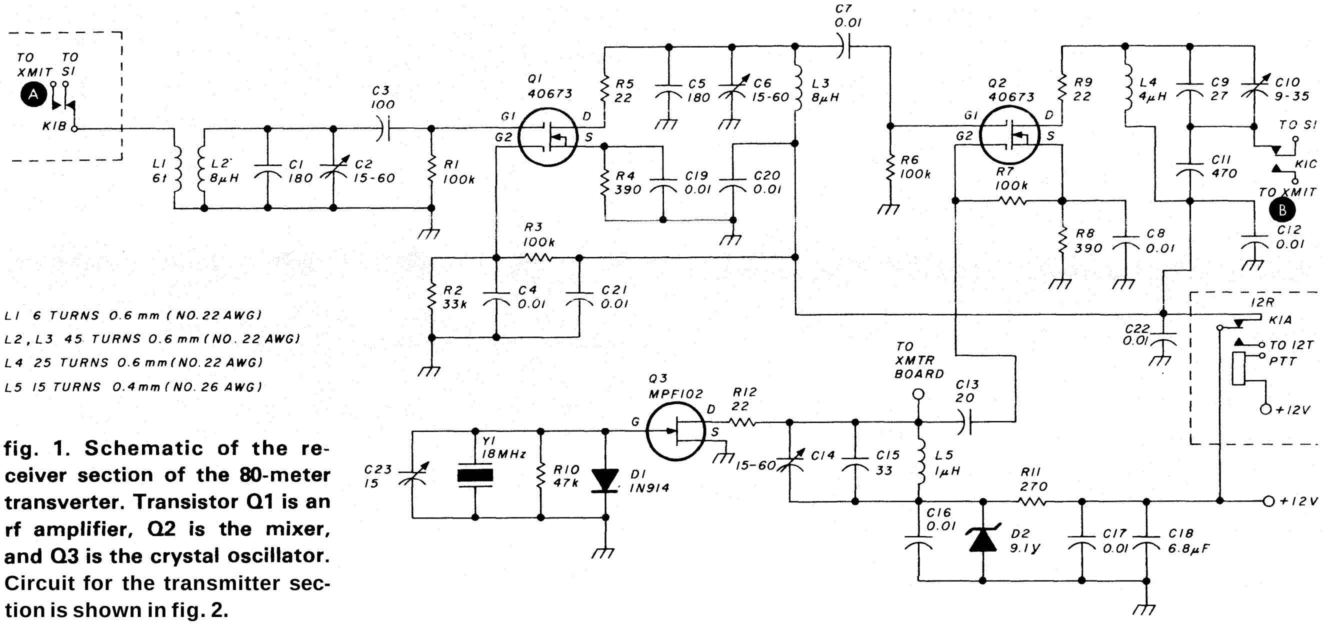


fig. 1. Schematic of the receiver section of the 80-meter transverter. Transistor Q1 is an rf amplifier, Q2 is the mixer, and Q3 is the crystal oscillator. Circuit for the transmitter section is shown in fig. 2.

ssb. Drive on CW, however, is not as critical. Some experimentation with Z1 may be necessary if you use ssb; *i.e.*, more or less attenuation may be needed with your rig. **Fig. 2** gives component values for the attenuation needed. The attenuator provides the additional bonus of a 50-ohm load to the driving transmitter, a near-must with solid-state rigs.

Liberal use of bypass capacitors on the dc lines helps to prevent oscillation in both transmit and receive modes.

construction

Transverter construction is easy because of PC-board layout and use. The receiver and transmitter sections are built onto separate boards. Circuit-board patterns for both the transmit and the receive board are shown. Double-sided, glass-epoxy board is used, with one side acting as a continuous ground plane. The copper around the component leads is drilled out using a large drill bit.

PC boards are definitely the answer to home construction. They provide clean layout, ease of construction, and, almost always, superior operation. Board layout is easy and provides a convenient way to make changes and experiment with your own ideas.

The two boards are wired together using RG-174/U cable. I mounted mine in an enclosure where

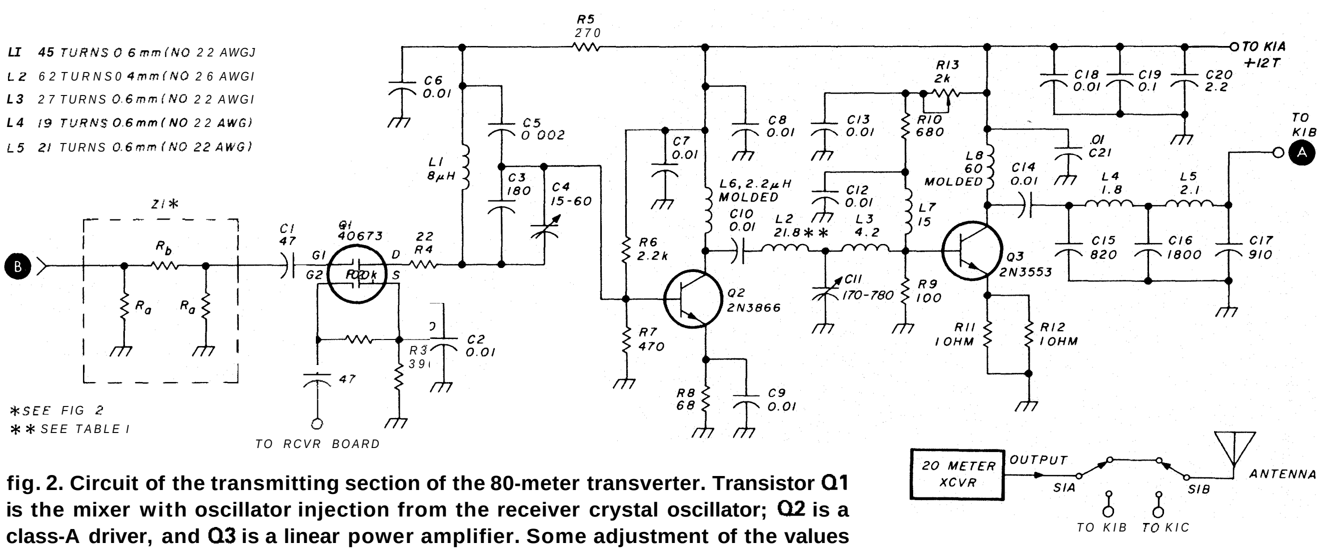


fig. 2. Circuit of the transmitting section of the 80-meter transverter. Transistor Q1 is the mixer with oscillator injection from the receiver crystal oscillator; Q2 is a class-A driver, and Q3 is a linear power amplifier. Some adjustment of the values of the input attenuator Z1 may be required for correct operation with different rigs.

table 1. Values for transverter components for 160, 40, 15, and meters.

receiver assembly

band	L1	L2(μH)	L3(μH)	C1, C5 (pF)	C2, C6 (pF)
160m	7t	16 53t/T68-2	8 45t/T68-2	430	15-60
40m	4t	4.3 29t/T50-2	4.3 29t/T50-2	82	15-60
15m	3t	1.5 22t/T37-6	1.5 22t/T37-6	18	9-35
10m	3t	1.1 19t/T37-6	1.1 19t/T37-6	5	9-35

band	C14(pF)	C15(pF)	L5(μH)	Y1(MHz)	bypass capacitor (μF)
160m	15-60	33	1.2	16	0.1
40m	15-60	15	1	21.5	0.005
15m	15-60	—	0.5	35.5	0.001
10m	9-35	—	0.5	42.3 42.8	0.001

transmitter assembly

band	L1	L2	L3	L4	L5(μH)
160	16 53t/T68-2	44 90t/T68-2	8.5 40t/T68-2	3.6 25t/T68-2	4.2 27t/T68-2
40	4.3 28t/T68-2	11.8 45t/T68-2	2.25 21t/T50-2	1 14t/T50-2	1.1 15t/T50-2
15	1.5 22t/T37-6	4 32t/T50-6	0.8 14t/T50-6	0.33 10t/T37-6	0.37 11t/T37-6
10	1.1 19t/T37-6	3 27t/T50-6	0.6 14t/T37-6	0.25 9t/T37-6	0.28 10t/T37-6

band	C3	C4	C5	C11	C15	C16	C17(pF)
160	430	15-60	5000	170-780*	1800	4000	2000
40	82	15-60	1000	50-380	430	1000	500
15	18	9-35	390	37-250	150	330	180
10	5	9-35	300	16-150	110	240	130

band	L6(μH)	L7(μH)	L8(μH)	bypass capacitors (μF)
160	4.7	33	150	0.1
40	1.5	8	33	0.005
15	0.47	2.7	12	0.001
10	0.27	2.2	10	0.001

*In parallel *with* 500

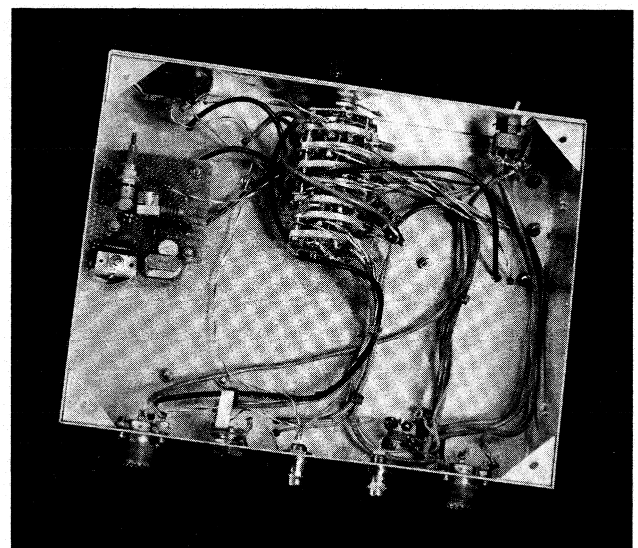
they are grouped with boards for 40, 15, and 10 meters to make an all-band transverter. A 3PDT relay switches +12V, the 20-meter transceiver, and the antenna between transmit and receive modes. S1 allows either the transverter to be used or the main rig to be operated straight through. Heat sinks should be used at Q2 and Q3.

alignment

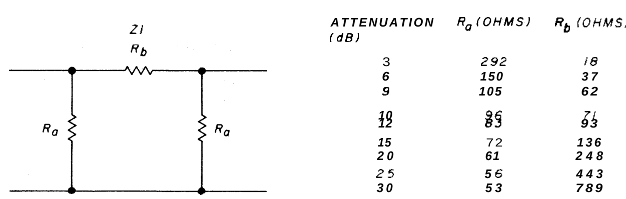
The first step in aligning the transverter is to obtain 18-MHz operation from Q3. Adjust C14 for maximum output using either an rf probe or a communications receiver. C23 can be used to adjust Y1 to exactly 18 MHz. The next step is to adjust C2, C6, and C10 for maximum signal strength while receiving an 80-meter signal.

Before applying drive on transmit, adjust R13 for minimum Q3 collector current. Next apply a low-level signal to Q1 (approximately 8V p-p at G1 of Q1). With a 50-ohm dummy load connected to the output, adjust C4 and C11 for maximum output.

Monitor the signal on a separate 80-meter receiver for best tone characteristics. For ssb adjust R13 for a collector current of 5 mA. Adjust C4 and C11 for maximum output. Monitor the ssb signal on an 80-



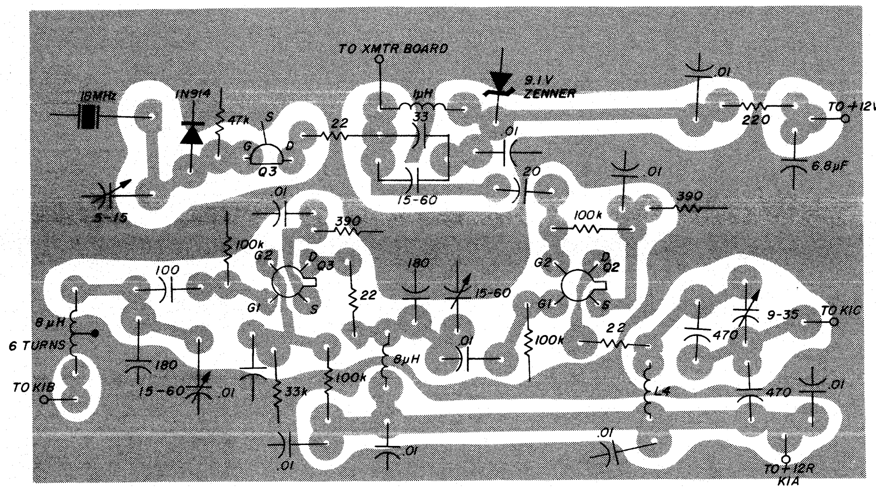
View of the chassis showing the bandswitch and internal wiring. The small board on the left is a WWV receiver converter.



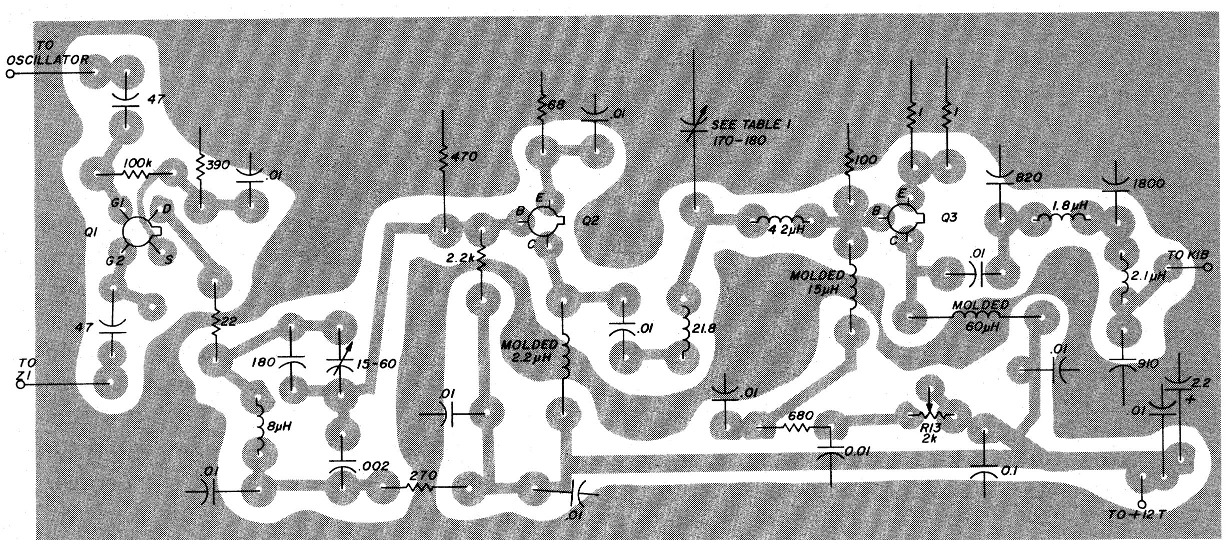
attenuation (dB)	R_a(ohms)	R_b(ohms)
3	292	18
6	150	37
9	105	62
10	96	71
12	83	93
15	72	136
20	61	248
25	56	443
30	53	789

fig. 3. Component values for different attenuation ratings of the pi attenuator, Z1, in fig. 2.

fig. 4. Component layouts for the printed circuit boards used in the low-band transverter.* Receiver board is shown on top, transmitter board at bottom.



RECEIVER BOARD



TRANSMITTER BOARD

meter receiver and readjust drive for no distortion (see photos). Readjust C4 and C11 for the best-sounding ssb signal. This completes alignment. The transverter should now be ready for on-the-air use.

operation

Because of the transistor used and the output network, the transverter ***must*** be operated into a 50-ohm load. Any departure from 50 ohms can cause the final stage to self-oscillate. Adding a 40V zener from Q3 collector to ground would be helpful in preventing damage to Q3 should oscillations occur. A transmatch is a great help in providing a nonreactive load for the transmitter.

With the output displayed on an HP-182 100-MHz scope, no harmonic energy could be observed. The CW and ssb outputs are shown in the photos.

Approximately two watts output was obtained on CW and one watt on ssb. The networks are sufficiently broadband to allow operation across the 80-meter band.

in conclusion

The transverter has provided good 80-meter coverage. Contacts have been made between the East and West Coast and into Mexico. Because QRP is my main interest, the one-watt output level is quite adequate. Under poor band conditions a linear amplifier can be added to increase output level.

reference

1. DeMaw and Rusgrove, "Learning to Work with Semiconductors," QST, April - September, 1975.

ham radio

*Full-size printed-circuit layouts are available from ham radio; please send self-addressed, stamped envelope with all requests.

lightning protection for the amateur station

A lightning strike can be devastating. This article provides a thorough treatment of how to avoid catastrophe in the ham shack

Lightning protection is a subject that should be of interest to every amateur and especially to those who have high antennas. It's a subject that's had little coverage previously in amateur publications despite the fact that there are some very effective protection techniques, which are widely known and applied to professional communications installations. Many of these techniques can be applied to the typical amateur station with minimal expense and effort. Some techniques, which involve more effort and expense, might be warranted only in situations where the probability of being struck by lightning is high.

Before protection techniques are discussed, it's desirable to have an understanding of exactly what causes lightning, the magnitude of a lightning stroke, and the probability of being struck.

The Earth's atmosphere is constantly being bombarded by cosmic rays. When these rays collide with gas molecules electrons are separated from some molecules, creating positive ions. Similarly, some molecules capture a free electron and become negative ions. It's estimated that our atmosphere contains about 4×10^3 ions per cm3 (6.5×10^4 per inch^3). At altitudes above 64 km (40 miles), the number of ions exceeds the number of neutral molecules; this is the region known as the ionosphere. At low altitudes, this number is insignificant compared with the number of neutral molecules. Still, the presence of these ions makes it possible for the air to conduct electricity to a small degree.[1]

Atmospheric makeup. The atmosphere taken as a whole has a net positive charge of about 106 coulombs. The Earth's surface has an equal negative charge, and a potential difference of about 3×10^5 volts exists between it and the electrosphere. The electrosphere is the region beginning at about 48 km (30 miles) up, in which the resistivity is sufficiently low so that there's no significant voltage gradient. (The ionosphere has the still lower resistivity necessary to reflect radio waves.) In fine weather, a constant flow of electrons occurs from the earth to the electrosphere, resulting in an electric field of about 100 volts per meter near ground. Lightning discharges return these electrons to earth at a rate sufficient to sustain a balanced dynamic system globally.[2,3]

For lightning to occur, a localized region of the atmosphere must attain sufficient electrical charge to produce a breakdown of the air molecules. The electric field near ground level rises to 500 volts per centimeter below a developing thundercloud and much higher still when lightning is about to strike.

By **John E. Becker, K9MM**, 201 E. Marion Street, Prospect Heights, Illinois 60070

The inside of a thundercloud is a turmoil of water, ice, and dust particles together with strong wind currents and temperature gradients. Although the mechanism is not totally understood, the result of this turmoil is a concentration of positively charged particles rising to the top of the cloud; negatively charged particles are concentrated in the lower areas of the cloud. This negatively charged region at the base of the cloud repels free electrons on the ground, resulting in an area beneath the cloud that's positively charged both with respect to the cloud and with respect to surrounding earth. Conditions now are correct for lightning to strike.

Evolution of a lightning strike. A lightning flash begins with a virtually invisible stepped leader, which travels down from the cloud toward the ground. Each step covers a distance of about 46 meters (150 feet) in less than one microsecond; the time between steps is about 50 microseconds. As this stepped leader progresses, it ionizes the air through which it passes, making it a good conductor. When the leader reaches within a few hundred meters of ground, ionized streamers begin to rise from the ground to meet it. Then the conductive path from cloud to ground is complete, and the visible portion of the bolt, known as the return stroke, begins.

It is the ***return*** stroke that has the destructive effects against which protection is needed. As soon as the conductive path is completed, electrons start flowing rapidly to ground. This action starts at the point of contact between the stepped leader and the rising streamer, and the greatly increased current causes the ionized path to glow brightly and get very hot – up to 2×10^4C (6.5×10^4F). The region of high current and brightness moves upward to the cloud at a speed of over 9.6×10^4 km (6×10^4 miles) per second, drawing electrons from higher and higher in the cloud. This contrasts with the speed of the stepped leader, which typically averages only 384 km (240 miles) per second. Although the region of high-current density and hence the visible flash moves upward, the actual flow of electrons is downward.

After the first return stroke usually enough charge remains in the cloud to initiate a second leader. This usually occurs within 70 milliseconds or less; because of remaining ionization, this leader usually follows the path of the previous stroke and travels directly to earth in one step of about one millisecond. For this reason it's called a ***dart leader,*** and, like the stepped leader, it's followed by a return stroke. Although the average bolt flashes only twice, about ten per cent will have as many as ten flashes; occasionally bolts will have up to twenty flashes over a period of about one second.

Electrical analysis. For all practical purposes a lightning bolt may be considered a constant-current source. That is to say, when it strikes an object protruding above ground, the current that flows through that object to ground will be the same, regardless of whether the object is a metal tower with a resistance of 1 ohm or a tree with a resistance of several hundred ohms. This is because the atmosphere's resistance is so high that the series resistance added to the path by the object struck is insignificant. The resistance of the object struck becomes important when the voltage drop across the object and the power dissipated in the object are considered.[4]

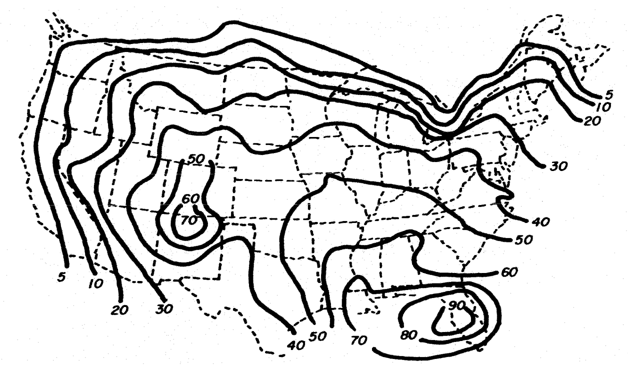


fig. 1. Map showing average annual number of thunderstorm days in the continental United States.

The magnitude and duration of the current in a lightning bolt will vary from stroke to stroke. The stroke-current waveform consists of a rapid rise to the peak current, followed by a more gradual decay. A waveform of this type is described with two numbers: the rise time to the peak value and the decay time to 50 per cent of the peak value. A typical lightning bolt would have a rise time of about 2 microseconds and a decay time of about 40 microseconds. The peak current in a stroke will exceed 1.7×10^4 amps in 50 per cent of all strokes. It will exceed 6×10^4 amps in 10 per cent of all strokes, 1.2×10^5 amps in 1 per cent of all strokes, and 2.4×10^5 amps in 0.01 per cent of all strokes.[5]

the probability of lightning damage in the U.S.

The probability that any given object will be struck by lightning depends on two things: the frequency and type of thunderstorms at the location of the object, and the object's height above average terrain. Observations made over a period of years have resulted in the compilation of an accurate map showing the average number of thunderstorm days per year in the United States (fig. 1). Note that this map does not take into account the possible occurrence of more than one thunderstorm in an area on any given day. Therefore it's somewhat conservative as an indicator of the total number of thunderstorms likely to occur in a year, particularly in areas having larger numbers of thunderstormdays.

Thunderstorms are classified either as convection storms or frontal storms. Convection storms, which account for the majority of thunderstorms, are local in extent and relatively short in duration. Frontal storms extend over greater areas and may last for several hours. Statistical data has been tabulated to predict the expected number of lightning strokes to

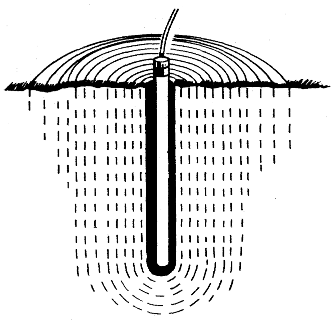


fig. 2. The concentric shell concept as an aid to understanding the resistance of the earth surrounding a ground rod.

ground per square mile (assuming flat terrain) per thunderstorm day. This number is called the ***stroke factor.*** The stroke factor for convection storms is 0.28; for frontal storms it is 0.37.[6]

In areas of flat terrain, the probability of lightning striking any given point is extremely low, but an object protruding above ground will attract lightning that would have otherwise struck all other points within a circular area surrounding the object. The radius of this area depends on both the height of the object and the intensity of the stroke. An object will attract a 2×10^4-amp stroke that would have struck anywhere within a radius of twice the height of the object had the object not been present. This radius increases to six times the height of the object for a 6×10^4-amp stroke and ten times the height of the object for a 1.35×10^5-amp stroke.[7]

the importance of proper grounding

Three distinct situations are associated with a lightning strike that can result in damage to electronic equipment. The most obvious and potentially destructive is the effect of a direct strike. Less obvious, but still capable of causing considerable damage, are two secondary effects: induced voltage and ground current.

A lightning bolt is surrounded by an intense magnetic field. The bolt itself may go harmlessly to ground, but its magnetic field will induce a voltage transient into any wires it encounters, such as antennas and feedlines, rotor-control cables, power lines, or telephone lines. When a lightning bolt goes to ground, current flows through the ground in all directions away from the point of the strike to dissipate the sudden excess accumulation of charge. This ground current can cause a significant potential difference to exist between different ground points. Improperly grounded equipment can end up in the path of some of this ground current. The lightning protection techniques to be described can prevent equipment damage from all three of these effects.

protecting your gear

Lightning protection comes in increments. An unprotected station will be wiped out by any kind of strike. A partially protected station will incur less damage and may completely escape damage from a small bolt. A carefully planned and executed lightning protection system will protect against just about anything.

There are three basic protection techniques. Everything described here will contribute to one or more of them. The first technique is to send as much as possible of the lightning stroke directly down your tower and into the ground. The second technique is to make it as hard as possible for any of the energy of the strike to get to the equipment. The third and final technique is to control the path of any energy that reaches the equipment so that it finds its way harmlessly to ground.

Grounding systems. The first technique, that of sending as much of the current as possible directly to ground, requires grounding your tower. This isn't always easily accomplished. There's more to it than just driving a rod into the ground and connecting it to the tower. Any ground system has a resistance, which can be measured. This resistance determines the voltage level to which the tower will rise when struck by lightning. The magnitude of this resistance depends on certain characteristics of the soil that determine its resistivity; *i.e.,* composition, temperature, moisture content, and salt content.

A ground rod driven into soil of uniform resistivity radiates current in all directions. To understand the resistance of the soil surrounding the ground rod, think of the rod as being surrounded by an infinite number of thin concentric shells of soil all of equal thickness (fig. 2). The greatest resistance is in the shell directly next to the ground rod, because it has the smallest cross-sectional area at right angles to the current flow. Each succeeding shell is larger in cross section and therefore has less resistance. At a small distance from the rod the area of each shell is so large that its resistance is negligible compared to the resistance of the shell directly next to the ground rod. The resistance varies inversely with the cross-sectional area. Measurements have shown that 90 per cent of the total resistance surrounding a ground rod is usually within a radius between 1.8-3 meters (6-10 feet) of the rod.

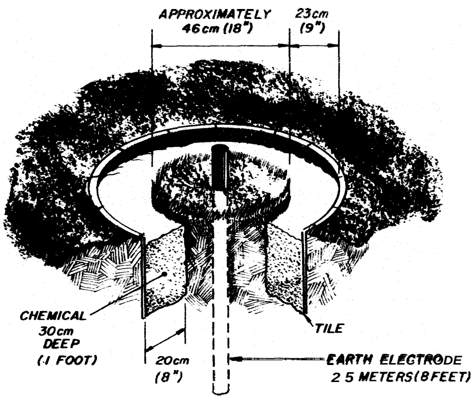
Soil considerations. The composition of the soil will depend on your location. Whatever soil you have you'll have to live with. Clay and loam are the most desirable soils for low resistance. Sand or gravel increase the resistivity in relation to their proportions, and soil that is mostly sand or gravel has fairly high resistance.

Whatever type of soil you have, a deep rather than a shallow ground is better for several reasons. Soil is seldom of uniform resistivity at different depths. The soil near the surface generally has higher resistivity than that at deeper levels because of wetting and drying out with seasonal variations. Deeper soil is more stable and less subject to such variations. Usually it has a higher moisture content than surface layers.

As the soil temperature decreases, its resistivity goes up. Frozen ground is a very poor conductor, as the negative temperature coefficient of soil resistivity increases sharply below freezing. Any ground rod should, at a minimum, extend several feet below the frost line in your area. Otherwise an early spring thunderstorm may occur while the ground is still frozen, and your ground system will be greatly reduced in effectiveness. A 2.5-meter (8-foot) ground rod is the absolute ***minimum*** that should be used.

reducing soil resistance

Soil resistivity may be reduced artificially by increasing its salt content. A doughnut-shaped trench can be dug surrounding the ground rod. Alternatively, a length of drain tile can be buried next to it. The trench or tile is filled with a chemical such as magnesium sulphate, copper sulphate, or ordinary rock salt. Rain and snow will dissolve the chemicals and wash them into the soil (fig. 3).



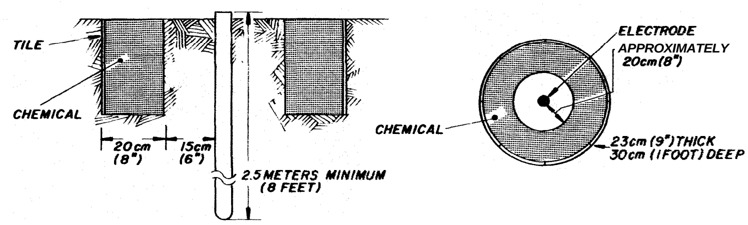


fig. 3. Method of reducing resistance of a ground connection by chemical means. The tile may be eliminated with little effect.

Deep-driven grounds are becoming the most popular and economical method for obtaining low-resistance ground connections. For ease of handling and driving, sectionalized ***Copperweld*** ground rods may be used. These rods are available in several diameters and lengths. They are threaded on both ends, and special threaded couplings are used to join them. A driving bolt is attached to the top of the section being driven to protect the threads from damage (fig. 4).

Another effective method of reducing the resistance of a ground connection is to parallel several ground rods. For this technique to be effective, the rods should be between 1.5-3 meters (5-10 feet) apart. With less than 1.5-meter (5-foot) spacing, the conducting paths from adjacent rods will overlap excessively. With more than 3 meters (10 feet) of spacing, the reactance of the connecting wires becomes a detriment.

If multiple ground rods are used around the tower base, each rod should be securely fastened to the tower base, and also to a buried ring of 6.5 mm (no. 2 AWG) copper wire surrounding it. Guyed towers should have at least one ground rod at each guy anchor point; these rods should also be tied back to the ground ring at the tower base. In areas of high soil resistivity, buried radials spaced every 60 degrees and up to 46 meters (150 feet) in length, if possible, may be connected to the ground ring to lower the ground system resistance.

There's no hard and fast rule as to how low the ground system resistance should be for good protection. A value considered acceptable by some electric power systems and protection codes is 10 ohms, so this is a good objective to shoot for. Where the soil is mostly clay or loam this value is achievable, but in areas with rocky or sandy soil it may be economically unreasonable to attain.

Accurate measurement of ground resistance requires the use of sophisticated equipment not usually found in the amateur station. An ohmmeter is totally useless due to voltage gradients present in the soil from stray power-system currents and electrochemical effects. However a method that can be used with reasonably good success requires only

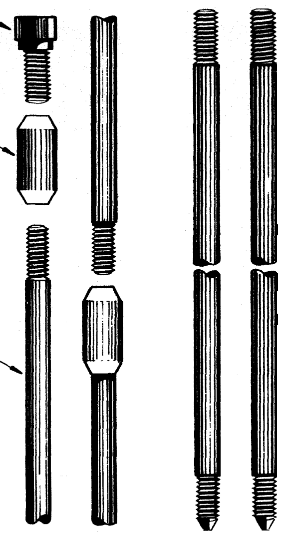


fig. 4. Copperweld sectionalized ground rods.

equipment that many hams would have or could borrow. It involves using an isolation transformer and a Variac (autotransformer) to pass 60-Hz ac through the ground between the ground rod to be tested and a known good ground, such as the power system neutral line. The voltage and current are measured, and the resistance is computed from Ohm's law. There's a very definite shock hazard associated with this method, so be careful! See fig. 5.

A question is often asked as to how large a ground conductor is necessary in a lightning-protection system. Often wire much heavier than necessary is used. A lightning stroke is of very short duration, and the heat produced in the wire by the current is limited. At a minimum, the wire must be large enough so that it's not heated to its melting point. A 2.6-mm (no. 10 AWG) copper wire can be sufficient to withstand a lightning stroke with 2.5×10^5 amps peak current and a 40-microsecond decay time. This amplitude is exceeded by only one stroke in 10^4. A 4.1-mm (no. 6 AWG) copper wire is recommended to provide an adequate safety factor in above-ground applications. For buried applications, 6.5 mm (no. 2 AWG) copper wire is recommended because of corrosion. Aluminum wire should have a cross-sectional area 1.6 times as large as the recommended copper gauges for the same current-handling capacity.

tower and rotator considerations

With the tower securely grounded at the base, the next step is to ensure that this good solid ground extends all the way to the top. Although you'd expect good continuity between sections in a fixed tower, the joints are designed for mechanical strength only, and the possibility always exists that oxidation and corrosion will cause increased resistance between sections. A good practice is to bridge each section joint on each leg with a short length of 4.1-mm (no. 6 AWG) copper wire secured with ground clamps. This is particularly important at the hinge joint of foldover towers.

Crankup towers are harder to deal with, which is unfortunate because they're also more likely to be a less-than-ideal path to ground for lightning – unless grounding straps are added. If the tower is lowered only occasionally for antenna work, the same techniques as those for a fixed tower can be used. The obvious inconvenience is that you'll have to climb the tower (make sure all safety stops are properly engaged first) and remove the grounding straps before the tower can be lowered. If the tower is raised and lowered frequently, about the only thing you can do is run grounding straps from the top of the top section all the way to the top of the bottom section. The straps should be pulled straight when the tower is fully raised, and of course they will hang down when it is lowered.

To protect your rotor, one or more bonding straps should be connected between the mast above the rotor and the tower legs. Leave no more slack than necessary to allow full rotation. Stranded wire, such as automotive battery cable, will stand up better to repeated flexing than will heavy solid wire.

The mast itself, or an air terminal clamped to the top of it, should extend far enough above the antenna so that its cone of protection includes the antenna and protects it from a direct hit. This mast extension should be at least half the turning radius of the antenna. The elements of an antenna are usually heavy enough to escape damage even if hit directly, and most antennas are at dc ground potential; but the path that provides this dc ground may not withstand the stroke current. Items such as ferrite baluns, traps, gamma matches, and element insulators are vulnerable and hard to protect in case of a direct hit on the antenna itself.

If the topmost point on the antenna system is the vertical element of a ground plane, make sure it is dc grounded at the base with an rf choke. This choke should have the lowest inductance possible for the frequency range of the antenna and should be made of the heaviest gauge wire practical.

If the tower is a wooden pole or other nonconducting support, it should be topped with an air terminal connected to the ground system with a 4.1-mm (no. 6 AWG) copper wire stapled to the pole on the side opposite the antenna feedline. A roof-mounted antenna system should be grounded similarly.

The shields of all coaxial feedlines should be bonded to the tower at the top and bottom, and at 15-meter (50-foot) intervals on taller towers. This is easily accomplished by using bulkhead coax connectors mounted on plates, bolted to the tower at the appropriate levels. Be sure to use waterproof connectors or take other suitable steps to keep moisture out of the coax.

residual lightning energy

All the techniques covered to this point have been concerned with shunting as much of the direct lightning energy as possible directly to ground. Now the emphasis will shift to the leftover energy that has found its way onto feedlines, rotor cables, etc., and to techniques for keeping it out of the rig. It all comes down to making these conductors look like a high impedance to the lightning. Remember that a lightning bolt is a series of pulses having rise times on the order of 2 microseconds. This makes it similar to a 500-kHz rf signal in terms of the effects of reactive elements it may encounter. Specifically, it's desirable to introduce ***series inductance*** in any conductors leading from the tower to the rig.

Obviously it's impractical to break the run of coax and place coils in series with the inner and outer conductors. While these might stop the lightning coming down the line, they would just as certainly also stop the flow of rf energy from the transmitter to the antenna. If the antenna is at dc ground as recommended earlier, the same potential will be impressed onto both the inner and outer coax conductors by a lightning strike. At every convenient opportunity, the coax should be bent at as sharp an angle as allowed by its minimum bending radius. The lightning pulse wants to go in as straight a line as possible, so these bends appear to it as an increased impedance. As far as an rf signal is concerned, as long as the dimensional relationship between coax inner and outer conductors is maintained, it makes no difference how often it is bent. Any excess feedline should be wound into a coil before entering the building, as this further increases the impedance it presents to lightning. Rotor cables and any other lines from the tower to the station should be treated similarly.

An additional technique that may be employed is to run the feedlines and control cables through a length of metal conduit. The magnetic field generated by lightning-stroke currents will perceive the conduit as one big shorted turn, resulting in another desirable increase in series impedance. This is particularly effective if the cables are run through conduit for 6 meters (20 feet) or more.

I'd like to emphasize that the isolation techniques just described are effective only if a good ground has been established at the tower. All the current is going to go to ground somewhere, and a low-resistance tower ground, together with a high impedance from the tower to the rig, only establishes how the current is divided up between these two paths.

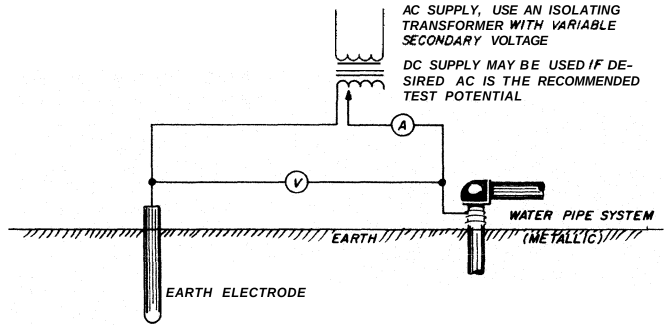


fig. 5. A simple method of measuring ground resistance.

station grounds

At this point, the protection job is two-thirds done. With good tower grounding and rig isolation, most of the stroke current will go straight down the tower to ground. The remaining task is to control the ground path followed by the remaining current which comes into the rig so that no equipment damage results.

It would be nice if it were possible to arrange things so that no ground path at all goes through the rig. In fact it's possible but seldom practical to do exactly that. It's worthwhile to examine what it would take to accomplish this, if for no other reason than to help in understanding the protection techniques to be followed in the more practical alternatives.

Let's assume that the only connection from the tower to the rig is a single feedline. Let's also assume that you have no other connection between the rig and the rest of the world; *i.e.*, no ac power lines, no telephone, nothing. The rig must be battery powered or have its own ac generator. The final assumption is that the entire rig, including its power source, is perfectly insulated from ground; no leakage, and sufficient clearance such that no arc-over to ground will occur.

Now if lightning strikes the tower, let's see what happens. If it is an "average" stroke of 1.7×10^4 peak amps, and if the tower ground resistance is 10 ohms, the tower, feedline, and rig will all rise to 1.7×10^5 volts above ground. This number is conservative because it ignores the effects of inductance in the grounding path. In many cases a significant inductive reactance will occur in series with the resistance, and the potential rise will be higher for a given stroke current. If the assumption is good that the rig is perfectly insulated from ground no current will flow, and no damage will occur.

practical considerations

It should be obvious at this point that this kind of situation is seldom likely to exist. In the real world, the rig is not perfectly insulated from ground and connections exist to power lines and to the telephone ground bus. This morass of ground interconnections may be fine in terms of signal distribution and ac safety, but it can spell disaster when lightning strikes.

Consider where the current is actually going to flow. For example, it may come down the coax-feed-line shield to the linear amplifier chassis, across this chassis to the input connector, along the input coax shield to the exciter, up the power cable ground wire to the power supply, through the power supply printed-circuit board to the ac line cord, and down the third wire to the ac conduit. A long and devious route such as this will look highly inductive to the lightning pulse. As a result significant voltage drops will occur along the way with a high probability of equipment damage.

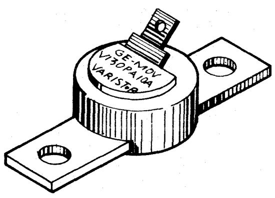
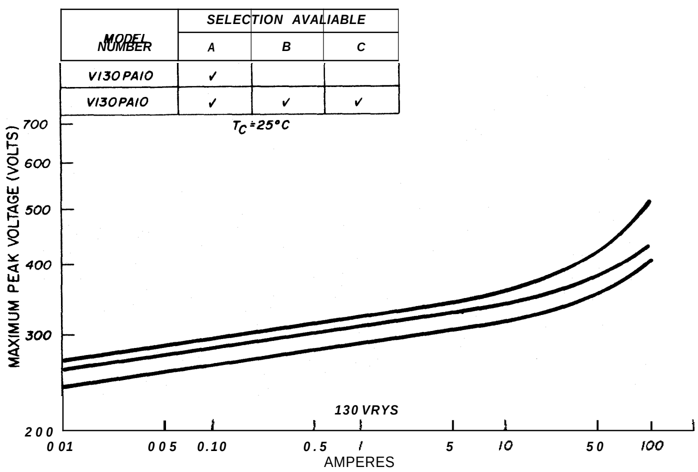
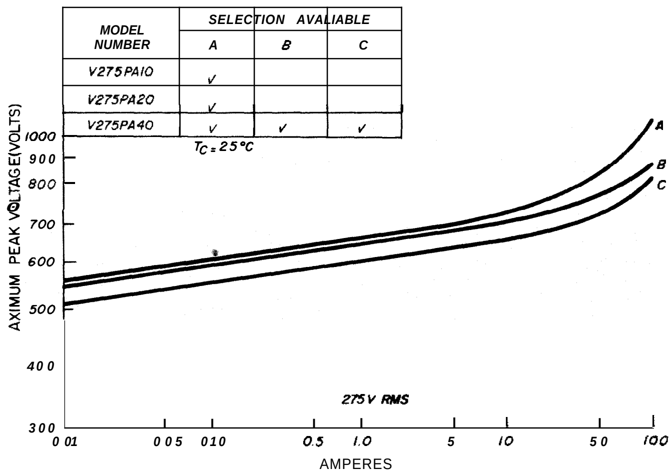


fig. 6. Ratings for a high-power zinc-oxide varistor. Note the sharply rising voltage characteristic above 10 amperes on parts rated for up to 2750 amperes peak current.





model number	rms applied voltage 50-60 Hz volts	recurrent peak applied voltage volts	dc applied voltage volts	energy joules	average power dissipation watts	peak current amperes	characteristics: varistor peak voltage at 1MA ac min. volts	max. volts	thermal resistance hot spot to case °C/watt
V130PA10				10	8	1200			6.8
20	130	184	170	20	15	2750	185	255	3.6
V275PA10				10	4	600			13.7
20	275	389	360	20	7	1200	390	523	7.8
40				40	13	2750			4.2

system. This brings us back to that small portion of the stroke current that will go to ground through whatever path it can find at the rig. It can follow two routes, and each requires attention if damage is to be prevented.

The first of these routes encompasses all direct grounding made at the rig. This grounding can include the shield of signal-carrying leads between different pieces of equipment, the third wire in ac line cords, and wires connected from the ground screws on various pieces of equipment to a main-station

Single-point grounds. The solution to this problem is single-point grounding of each piece of gear together with bonding between each item. Single-point grounding prevents ground current from flowing through a piece of equipment, while bonding prevents destructive voltage differentials from developing between different pieces of gear. The final ingredient necessary for this technique to be effective is a ***good station ground.***

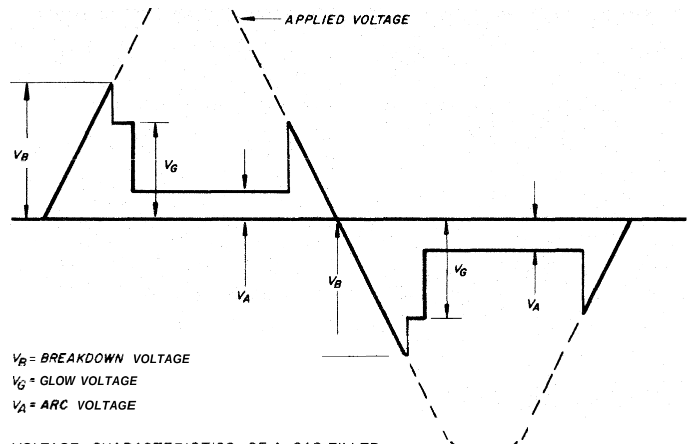
The station ground is second in importance only to the tower ground, and many of the techniques previously covered are applicable. The objective of the station ground is to provide a ground plane at uniform potential for the entire station.

Commercial installations. In commercial installations where an entire building is devoted to the equipmeni instaiiation, the recommended procedure is to install a buried ground ring around the outside perimeter of the building. This ring is supplemented with ground rods or radials. A second ground bus is run around the inside perimeter of the building. The two ground rings are interconnected at no fewer than four points and at intervals of no more than 15 meters (50 feet). This system is also connected to the power-system ground, and any extensive masses of metal such as water pipes and heating ducts. Finally, the building ground system is connected to the tower ground system. This wire must not be run through a conduit, nor should it be routed near feedlines or other cables from the tower, as it may induce transients into them.

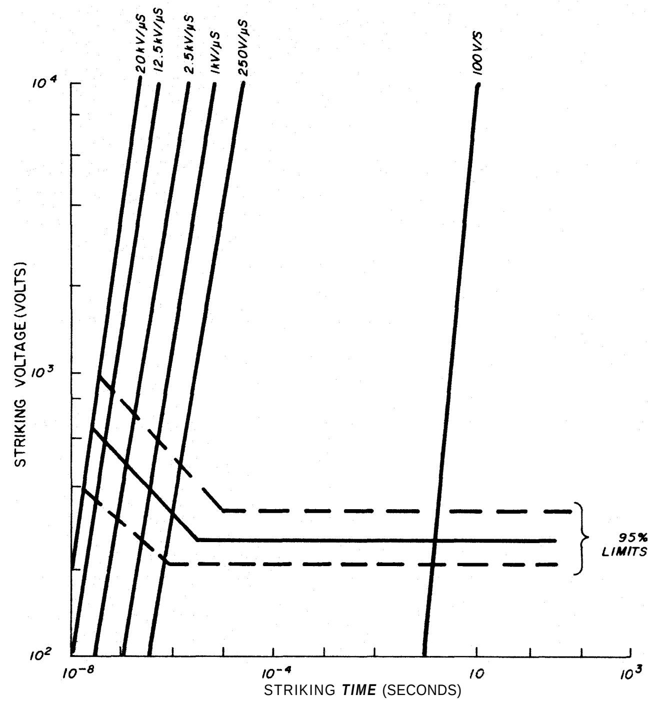
Amateur installations. For a typical amateur station in a home, a commercial-grade station ground would usually be impractical, since all the equipment is usually in one room and near to only a small portion of the building perimeter. The commercial procedures can be modified using common sense to fit the circumstances of any particular installation. The important thing to remember is that a single ground wire to the nearest water pipe is ***not*** adequate, and every extra bit of grounding adds something to the degree of protection obtained. Whatever modifications are made, the connections between the station ground, tower ground, power system ground, and the water pipes must not be eliminated.

Interconnections and bonding. Once a suitable station ground has been provided, the equipment must be properly connected to it. The shields of all coax feedlines should be connected to the station ground at the point where they enter the building. On each piece of equipment, a ground tiepoint should be selected. This point should be connected to the station ground and also to the ground tiepoint on physically adjacent pieces of equipment. These ground connections should be as short and straight as possible, using 2.6-mm (no. 10 AWG) or heavier copper wire.

It's particularly important on any piece of equipment where a coax feedline terminates from the tower that the point where the feedline connects be used as the ground tiepoint. The requirement for multiple connections of coax and other shielded wires to some pieces of equipment may make it impossible to eliminate secondary ground paths en-



VOLTAGE CHARACTERISTICS OF A GAS FILLED SURGE ARRESTER



maximum ratings	40,000A – 15 pulses of 8/20 waveform at 3-minute intervals 20A RMS ac or dc – for 1 second 10 times at 3-minute intervals
life	10,000 surges typical with 1011000 waveform at 1000 amps
dc arc voltage	30 vdc, maximum
glow to arc transfer current	0.25 to 1.0 amp
operate delay time	500 nanosecond, maximum
insulation resistance	10^9 ohms
capacitance	
end to end	1 picofarad
end to center	2.5 picofarads
dimensions	8 mm (5116 in.) dia. x 44 mm (1-3/4 in.) long

fig. 7. Ratings for a TII-16 gas-filled surge arrestor.

tirely, but careful grounding and bonding will make these secondary paths relatively unattractive to lightning current. The secondary ground path provided by these shielded cables can be made to look like an even higher impedance if the cables are made longer than necessary and the excess length is wound into a coil.

Threewire ac cords. The ground wire in three-wire ac line cords is fine for its intended purpose of preventing shocks resulting from power-line leakages, but it's detrimental as far as lightning protection is concerned. Not only is it neither short nor straight, but its proximity to the other wires in the power cord could couple unwanted surges into the equipment through the power line. When proper grounding and bonding techniques are applied, the ground wire in the three-wire power cord is a secondary ground and no longer needed for its original purpose. Therefore, it should be disconnected for optimum lightning protection.

Dielectric breakdown. The second route to ground at the rig is by dielectric breakdown to power or telephone lines. This is most likely to occur in installations where proper grounding and bonding practices have not been followed. But proper grounding and bonding practices do not completely eliminate the possibility of dielectric breakdown. This is because of the ever-present inductance in any ground wire, together with the fast rise time of the lightning pulse. As already noted, the chassis of all the equipment may rise momentarily to thousands of volts above ground when lightning strikes. The power-transformer primary in each piece of gear is connected to the ac power line, which is at a low impedance and always within a few hundred volts of ground. If the insulation between any of these primary windings and the transformer core, which presumably is chassis mounted, is insufficient to withstand this voltage, dielectric breakdown will occur and the transformer will be destroyed. The same thing can happen with the telephone lines, except that here the phone patch or telephone itself will be destroyed.

Surge suppressors. To prevent this sort of damage a form of bonding must be employed. Obviously the power and telephone lines can't be directly and permanently shorted to ground. Instead, some type of transient voltage surge suppressor must be used. This is a device that's normally an open circuit but which will momentarily break down and provide a low-impedance path across whatever it is connected to in case of a lightning strike. When the surge is over, the device opens the circuit again so that normal equipment operation is unimpaired.

Many types of transient voltage surge supressors are on the market. They offer varying degrees of protection and range in price from a few cents to hundreds of dollars. There are two devices which, when used together, will provide a high degree of protection at relatively low cost. These are the zinc-oxide varistor and the gas-tube surge arrestor. Each has certain advantages and certain limitations; this is why it's recommended that they be used together.8

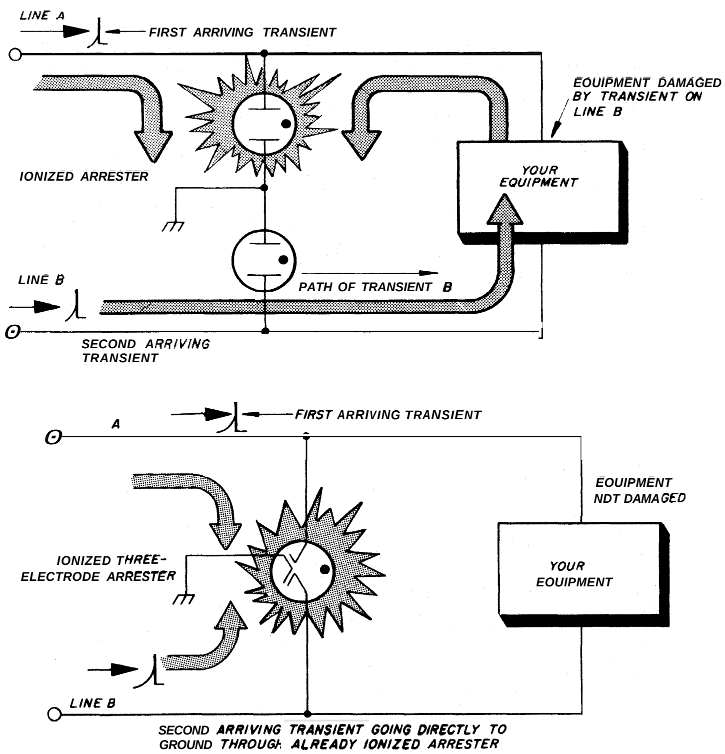


fig. 8. With two-electrode arresters (top), the first arrester to "see" the surge is ionized by that surge and provides an excellent path to ground. The second wire now uses the available ground, but, unfortunately, the path to ground goes right through the equipment to be protected. With three-electrode gas arresters (bottom), if either wire sees a surge, the arrester fires, ionizing the common gas chamber and instantly grounding both wires; there is no damaging potential difference across the equipment.

The zinc-oxide varistor is a device on the market for just a few years, but it has gained wide acceptance in the electronics industry as a transient suppressor. It's available from several manufacturers in a wide range of ratings. The MOV line from General Electric is typical of those available. These devices are designed to protect against transients on the ac power line from lightning and from inductive load switching. Its advantages are response time measured in nanoseconds and response to voltages only slightly above the normal operating voltage of the circuit. Disadvantages are limited maximum-current capability and high "let-through" voltage under high-current conditions **(fig. 6)**.

Gas-tube surge arrestors have a firing voltage that depends on the rise time of the transient waveform. Faster rise times result in higher firing voltages. For a lightning waveform having an 8-microsecond rise time, a gas tube can be expected to fire in about one microsecond **(fig. 7).** Once a gas tube has fired, the voltage drop across it is clamped to 30 volts. A gas tube can handle more peak transient current than a varistor, because the low voltage drop results in low power dissipation.

A gas tube will continue to conduct until the applied voltage drops below 30 volts. In an ac-power circuit, this means the tube wili conduct for a full half cycle, 8.3 milliseconds at 60 Hz, even though the transient that caused it to fire may have lasted only a few microseconds. This may exceed the long-term power handling capability of the tube, and for this reason gas tubes should always be installed on the *load* side of the fuse or circuit breaker in a power circuit.

Gas-tube surge arrestors are available from TII Corp., Joslyn Electronic Systems, and others. For power-line applications, a three-electrode gas tube is used, with the end electrodes connected to the two sides of the ac supply and the center electrode connected to ground. If the potential between the grounded electrode and either supply electrode becomes sufficient to fire the tube, the entire tube ionizes and all three electrodes are effectively shorted together for the duration of the transient **(fig. 8)**. The disadvantage of the gas tube is that it's slower to respond than the zinc-oxide varistor and requires a higher voltage relative to the normal voltage in the circuit before it will fire.

When both a zinc-oxide varistor and a gas-tube surge arrestor are used to protect a piece of equipment, the varistor will take care of short-duration, low-energy transients such as might result from a lightning strike on the power line at a considerable distance from your location. The gas tube will come into play when lightning strikes your tower or the close-by power line. If economic considerations dictate the use of only one of these two devices, the gas tube is the one that should be chosen.

Devices are available using gas tubes that can be inserted between an appliance and the power outlet. This may seem very convenient, but these devices should not be used. The reason is that they are designed to protect against transients coming in on the power line only. We are mainly concerned with protecting against a lightning strike on the antenna or tower. This requires locating the gas tube as close as possible to the power transformer of the equipment to be protected to eliminate the inductance of the line cord from the ground path for the surge. Installation of a gas tube at this location will give maximum protection against a transient either from the power line or antenna. A set of these suppressors should be installed in each piece of equipment connected to the power line for which protection is desired **(fig. 9)**.

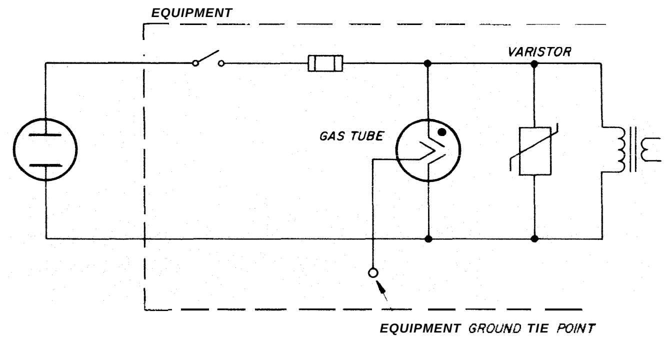


fig. 9. Use of both a gas tube and a varistor for maximum protection.

If your station includes a phone patch, a three-element gas tube should also be installed at the point where the telephone line connects to the patch.

concluding remarks

Lightning protection may appear to be a very complicated subject. But the average amateur station can be well protected without an unreasonable investment of time or money. Remember first and foremost that all of these techniques are nothing more than a means of controlling the path that the lightning bolt will take in its unstoppable search for ground. Work out the required protection system for your individual station with this in mind.

references*

1. R. E. Addy, "Lightning Phenomena," IMSA *Signal Magazine*, January-February 1974, page 13.
2. M. A. Uman, Lightning, McGraw-Hill, 1969, page 48.
3. M. A. Uman, Understanding Lightning, Bek Technical Publications, 1971, page 152.
4. A. K. Guthrie, "Learning to Live with Lightning," Communications, May 1977, page 53.
5. J. L. Marshall, Lightning Protection, John Wiley & Sons, 1973, page 30.
6. David Bodle, "Electrical Protection Guide for Land-Based Radio Facilities," Joslyn Electronic Systems, page 17.
7. R. H. Golde, "Protection of Structures Against Lightning," Proceedings IEE, October 1968, page 1523.
8. "High Speed Protection by Three Electrode Arrestors," TII Corporation, Section 3.4.3.

*A complete bibliography of material on lightning protection is available from ham radio and will be sent to interested readers upon receipt of a self-addressed, stamped envelope. **Editor**

ham radio

solar-powered repeater design

Why use antiquated methods to power your machine? Here's a strong case for using solar power — it's inexpensive and effective

In this article we provide the repeater-system designer with useful information about using solar power. The information is based on acceptable criteria for designing commercial solar-power systems and is supported by empirical data obtained in the actual operation of a repeater. The WR5ARO 19-79 solar-powered repeater is on Redondo Peak, New Mexico. It was built using the principles outlined below.

Solar systems to provide power for remote radio sites are not new. Recent advances in technology, and a national concern for conserving our energy resources, have brought recognition to solar power as a useful energy source. For remote sites, solar power is one of the few economically feasible power sources. As with other electronic devices, demand and popularity will reduce the price of solar-powered generators from its presently high level to one within the means of most amateurs. Based on projected technology, cost reductions of 50:1 are possible in the near future. This would put the price of a useful solar electric generator not far above the cost of a good, well-regulated bench-type power supply of equivalent capability. Solar power provides good mechanical reliability (no moving parts), good dependability (with proper array sizing), and an attractive price (considering the alternatives). First, some background.

photovoltaic cells

Silicon solar cells are P-N diodes whose photovoltaic characteristics (ability to produce electricity when exposed to light) have been optimized. Peak-current output occurs when the cell is exposed to direct, unrestricted sunlight that has an intensity of 100 mW per $cm.^2$

The rated output current of a solar array is directly proportional to light intensity. Therefore, at 50 mW/cm^2 the array output current is 50 per cent of the peak rated current. The output current capability of a single solar cell is a function of the cell cross-sectional area. Solar-cell output voltage is independent of cell size. Output voltage is constant from 10 mW/cm^2 to 100 mW/cm^2. The equivalent circuit to a solar cell is a constant-current generator with voltage limiting. A typical solar array I/V curve is shown in **fig. 1.**

array sizing

Solar cells are connected in series to provide the required voltage and are connected in parallel to provide the required current. Array specifications are given as a ***peak*** value. Since the manufacturer has no control over the amount of sunlight available, array specifications are relative to peak sunlight (100 mW/cm^2). It's the responsibility of the system designer to derate the peak specifications to an average value specific for the site.

Most solar arrays come from the manufacturer with enough cells connected in series to be optimum for charging batteries. Consequently, the system designer need be concerned only with the peak current available and the derating necessary to provide some average charging current. The amount of de-

By Ted Handel, WB5REA, and Paul Beauchamp, WB5RSN, 241 Rio Bravo, Los Alamos, New Mexico 87544.

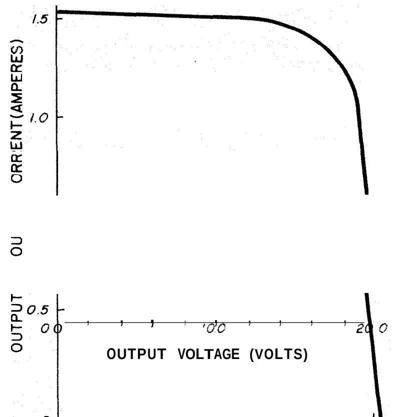


fig. 1. Typical I/V curve of nominal 12V solar array. The curve approximates that of a constant-current power supply with voltage limiting.

rating is a function of the solar-cell insolation at the site.

solar insolation

A number of variables influence the amount of sunlight that strikes the Earth's surface. Among the most important are elevation, cloud cover, atmospheric water content, pollutant level, the sun's incident angle, and the solar-day length. All these factors affect the amount of available sunlight. The amount of sunlight striking the Earth's surface is called insolation. Solar insolation data is given in many different quantities, such as Langleys and kilojoules per square centimeter. The mentioned quantities are usually given for one year; thus the quantities are units of energy. Since solar electric generation is an integrating process, it's permissible to use the average yearly insolation figure to size the solar array. This is true if the average yearly discharge rate of the repeater is also used.

Solar insolation data for Redondo Peak, New Mexico, was found to be 750 kJ/cm^2 per year. This gives an average solar intensity of about 23 mW/cm^2. The average current available at this site is, then, about 23 per cent of the peak current available from the solar array. **Fig. 2** shows average solar intensities for the continental United States.

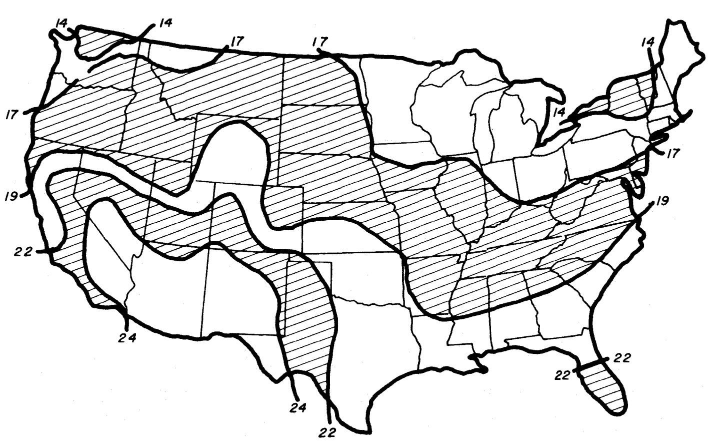


fig. 2. Map of the United States showing average solar intensity in mW/cm^2. Year-to-year variations are less than 10 percent.

storage devices

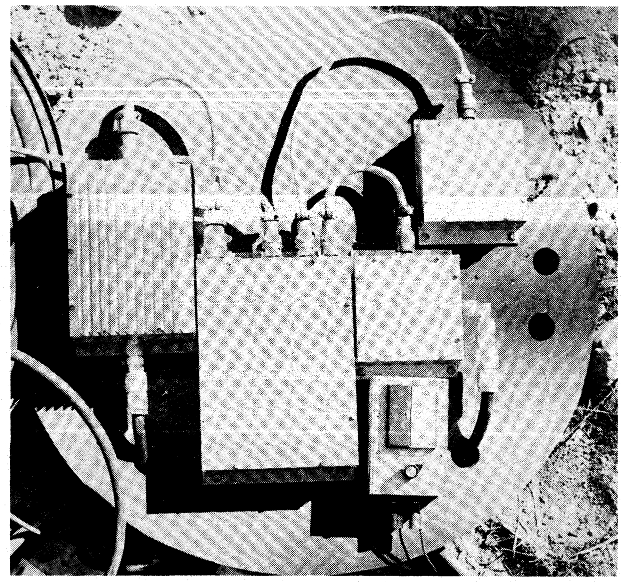
To make power available to the repeater during hours of darkness and foul weather, some type of storage device is used to hold energy collected during hours of daylight and good weather. A battery of some type is almost always used. The capacity of the battery must be large enough to carry the system through extended periods of poor weather and through the shorter days of winter. Battery capacity is relatively independent of array size. Generally, storage capacity is about ten days. The battery is the true power source for the repeater. It should be selected with care.

Co-author WB5RSN makes final angle adjustments to solar array that powers the 19-79 repeater on Redondo Peak, New Mexico (3433 meters, 11,254 feet). The array is mounted at 17 meters (55 feet) to prevent shading from nearby trees.

High-capacity, lead-acid automotive batteries should be avoided. This type of battery is designed to provide large amounts of current for short periods of time. To accomplish this, the battery must have a low internal impedance. High leakage currents occur in this type of design. A good-quality, electronic-grade rechargeable battery would be a better choice. Lead-calcium and gelled-electrolyte batteries, as well as telephone-type "wet" cells, are also good choices. Nickel-cadmiums aren't recommended because of their tempermental characteristics. Remember that most of the vagaries of these cells were discovered in space satellite power systems where solar cells were used to charge them.

voltage regulation

Solar generators normally provide more charging current than a fully charged battery can safely tolerate. To prevent damage to the battery, a voltage regulator is used to limit the charge voltage to a safe level. The circuit shown in *fig.* 3 is a simple shunt-type regulator. The series diode prevents the array from discharging the battery during hours of darkness. The diode is also a reverse-bias switch that allows the shunt resistor to absorb the excess power generated. The regulator receives its power directly from the array and therefore does not draw power from the battery. Some battery current is used by the regulator for voltage sampling, but this current is very low.



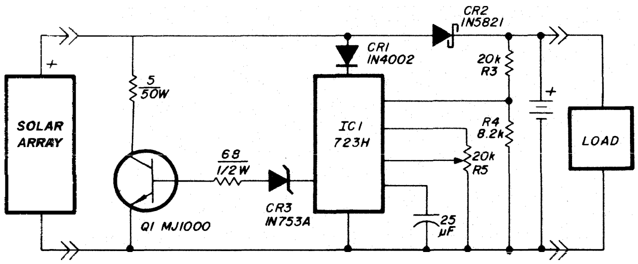
WR5ARO repeater is installed in a sealable 56-gallon barrel. The buried enclosure permits temperature stability. Mountain-top temperatures range from −34C in the winter to 38C in the summer.

Redondo Peak repeater installation. Solar array is mounted near the top of the tower to prevent shading. The repeater electronics are installed underground at the base of the tower.

One of the important characteristics of this type of regulator is its negative temperature characteristic. Simple zener regulators have a positive temperature coefficient, which causes the battery to overcharge in the summer and undercharge in the winter. The opposite characteristics are desirable. Failure modes have been arranged to cause an open circuit in the shunt element, thus permitting the array to charge the battery in the event of a regulator failure. Battery status can be monitored by occasional on-site check-ups or by telemetry.

array orientation

Proper array orientation is required to provide maximum power output during the year. Peak output occurs when the sun's rays are at normal incidence to the array plane. To obtain maximum output, the array is oriented true south (north in the Southern Hemisphere) and inclined from horizontal to an angle approximately equal to the latitude at which the site is located. This angle is then increased a few degrees to optimize the array for the winter months when the days are shorter and the sun is at a lower angle. Solar intensity is constant at all times of the year, but the



NOTES
1 RESISTANCE IN OHMS K=1000, 1/2 WATT UNLESS SPECIFIED OTHERWISE
2 R3, R4, ARE METAL FILM, R5 IS CERMET
3 CAPACITANCE IN µF.
4 R5 IS SET FOR APPROXIMATELY 14VdC WITH LOAD AND BATTERY DISCONNECTED
5 CR2 IS SPECIAL LOW FORWARD DROP DEVICE

fig. 3. Voltage regulator protects the battery from overcharging. Diode CR2 prevents the battery from discharging through the array during hours of darkness (Courtesy Solar Power Corporation).

solar day is shorter in the winter; hence, the net accumulated energy is lower. Fortunately, the solar array generates more power at lower temperatures, thus offsetting some of the loss.

A tracking array could be designed to follow the sun, but the additional power generated would probably be consumed by the tracking system. Reduced reliability would also be introduced into the system because of the mechanical components. In short, tracking sytems are not a good investment at this time.

equipment selection

The cost of solar systems requires that detailed attention be given to operating power requirements and system power overhead. Ideally, the system would draw no power during standby and would convert all current consumed by the transmitter into rf power. Of course, this isn't possible; therefore, the system designer must minimize repeater standby current and maximize transmitter efficiency. Obviously, vacuum-tube equipment can't be used. Surprisingly, most available solid-state, base-station equipment isn't sufficiently efficient to be considered. Solid-state mobile or portable equipment is a good choice, because it lacks many of the frills found in base-station equipment. Pilot lamps and similar amenities should be powered down or disconnected. Logic circuits should draw a minimum of power. CMOS devices can operate at high-voltage levels with amazingly low current consumption. The WR5ARO identifier is built with CMOS and draws about 50 µA. Likewise, the COR, control circuitry, supervisor, and timers all combined draw less than 1 mA.

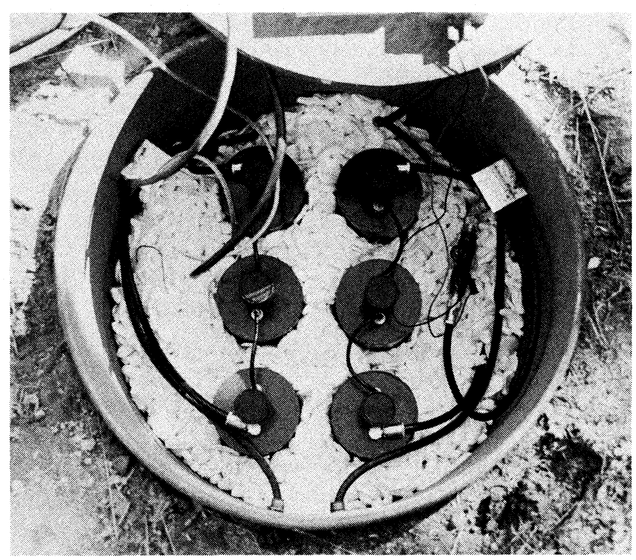
Reasonable numbers to achieve in equipment design or selection are idle currents (total) of 5-20 mA and transmitter efficiencies of 60 per cent. Under these circumstances, the size of the solar array will be a direct function of the transmitter output power.

If operation below 0°C is anticipated, extended temperature devices are required. It's wise to make sure that the circuitry will operate over the expected temperature extremes.

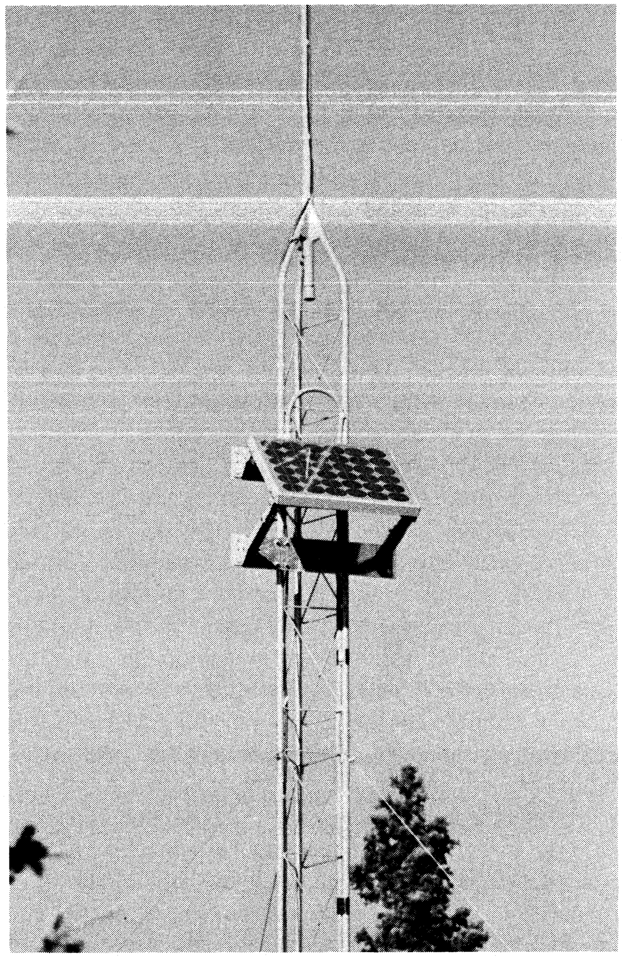
critical parameters

The most difficult data to obtain in designing a solar-powered repeater is the time the repeater is actually on the air. The time that the repeater is used varies from location to location and is also determined, to a great extent by the number of other repeaters in the area. The transmit time and transmitter power output will directly determine the size of the solar array. A mistake made in this estimate could be very costly; over-estimation is expensive; under-estimation is embarrassing. The best estimate can be obtained by timing the repeater use. The data should be accumulated for as long a period as possible. Ideally the period measured should be one year; this measurement will average out the concentrated operating times, seasonal variations, and other factors.

If the repeater is to cover an area not presently covered by an existing repeater, an arbitrary decision should be made regarding the length of time the transmitter will be in use. When the repeater has been in service for a period of time, the amount of its use can be determined. The transmitter output power can then be adjusted upward or downward to



Duplexer cavities are mounted in a barrel. Styrofoam packing chips slow rapid thermal changes. Cavities are copper-plated steel to minimize any detuning due to thermal stresses.



Solar panel is oriented true south and is inclined about 40 degrees for horizontal. Latitude of the repeater is approximately 36 degrees north. An additional four degrees of inclination optimizes the array output during winter when the sun is lower in the sky and days are shorter.

match the time use to the solar array. WR5ARO was designed to handle one-half the local traffic load that's presently divided between four local, wide-coverage repeaters. At its present power output, the repeater can comfortably provide 35 hours of operation per week.

Battery capacity is an important consideration in system design. The capacity figure will determine how long the repeater will operate when the solar array is not charging, or when the repeater is being used at a current rate greater than can be supplied by the array. It's necessary for the solar array to supply more current than the repeater will consume. It's not necessary for the array to fully charge the battery each day. The system can tolerate some deficit as long as the battery is not damaged by freezing or any peculiarities typically inherent to the type of battery used.

design procedure

Designing a solar power supply for a repeater isn't difficult. Remember that solar energy collection is a cumulative process. Its occurence is very regular and very predictable. Year-to-year variations are less than 10 per cent. Repeater use must be averaged to fit the collection criteria. The battery capacity is selected to be adequate to equalize the short-term variations in repeater use and local weather phenomena, which are highly unpredictable. The steps necessary for design are as follows:

1. Determine solar insolation for the proposed site. Source data can be obtained from sources listed at the end of this article. **Fig. 2** may be consulted directly.

2. Determine continuous idle current, multiply by 24 hours to determine the daily idle current ampere-hours.

3. Determine the transmitter current; multiply it by the transmitter on time to determine the daily average transmitter ampere hours.

4. Average the repeater load over a 24-hour period. Then, divide the idle ampere-hours plus the transmit ampere-hours by 24. This is the average load current. This number must be less than the average charge current as supplied by the array.

5. Determine the peak-panel output. The average solar intensity should be found (**step 1**) and divided into the average daily load current. Remember that if the average intensity is 22 mW/cm^2, then the average current available from a solar panel will be 22 per cent of the peak.

6. Calculate the "no-sun" storage requirement of the battery. Ten days of storage is an average number. Multiply the total ampere-hour load (**steps 2** and **3**) by ten to obtain the battery capacity.

Note that nothing has been said about battery-charging efficiency. Battery efficiency cancels, because the charging voltage is greater than the discharge voltage. The solar array provides the additional charging voltage required by design, with no sacrifice in performance. This assumption is valid if the internal leakage of the battery is not great (less than 3-5 per cent per month).

conclusions

Solar power is useful in providing adequate power to operate a radio repeater if care is used in designing the system. The designer has a wide latitude of options available. Enough considerations have been given to demonstrate that gross overdesign of a solar-power generator is not necessary. Attention to details and careful consideration of all available options will produce an economical design.

The Redondo Peak repeater has been in full solar-power operation since June 18, 1977. There has been no down time. A system checkout on December 16, 1977, showed that the solar array was generating its rated power output and the battery was fully charged.

The cost of the solar generator, when averaged over its 20-year life, comes to about $35 per year. This number compares quite favorably with the price charged to many mountain-top customers for similar power. As the price of solar power drops, so will the yearly cost for power generated by this means. In today's energy-cost spiral, solar power will become very attractive in the near future.

bibliography

1. "Solar Electric Generator Systems, Principles of Operation and Design Concepts," Solar Power Corporation, 5 Executive Park Drive, North Billerica, Massachusetts 01862.
2. "The National Atlas," U. S. Government Printing Office.
3. "Annual Solar Radiation in kJ/crn^2," map from Sensor Technology, Inc.

appendix

WR5ARO Specifications

idle current (mA)	12.
transmit current (A)	1.07.
operation	available 24 hours (open repeater).
design average daily use	2.5 hours per day.
transmitter power output (W)	9.5.
effective radiated power (W)	35.
solar array source	Solar Power Corporation E-01-369-1.5 (1.5 A peak output).
battery source	Globe Union Gel Cel 40 A-h (2 each GC12200).
environmental characteristics	elevation 3.43 km (11,254 feet). temperature −34°C to 38°C (−30F to 100°F). solar insolation 750 kJ/cm^2/year. average solar intensity 23.7 mW/cm^2. rainfall 46 cm (18 inches) per year. snowfall 91 cm (36 inches) per year.

Sample Calculations Using WR5ARO Data

Step 1 Solar insolation data = 750 kJ/cm^2/yr = 23.7 mW/cm^2.

Step 2 Continuous load = 0.012 A x 24 hrs = 0.288 A-h.

Step 3 Intermittent load = 1.07 A x 2.5 hrs = 2.675 A-h

Step 4 Daily average load = 2.963 A-h/24 hrs = 0.123 A.

Step 5 Peak panel output = 0.123 A + 10 per cent/(23.7 mW/cm^2/100) = 0.570 A peak.

Step6 Storage Capacity = 2.963 A-h x 10 days = 29.63 A-h battery. (Add some additional capacity to prevent freezing in the winter).

ham radio

universal digital readout

A universal digital readout system featuring reduced ambiguity, high input frequency, and low power consumption

This article describes a relatively simple design for a digital dial that evolved over several years of building and improving. For easy home duplication, it uses a minimum number of specially selected components. It is adaptable to virtually all types of shortwave equipment. Options are also described for low-power operation and reduced last-digit flicker.

The counter is connected to the vfo of the equipment and preset to the i-f, or the complement of the i-f. It is wired to count up or down depending on the internal frequency scheme of the equipment. It is even possible, without knowing these parameters, to set up the counter using only one calibration point and check whether the frequency indication moves in the right direction.

This counter has a 35-MHz capability, and thus covers the entire conventional shortwave range (3 to 30 MHz). It is therefore possible to measure the frequency that is generated by the premixing scheme in Drake equipment.

counter components

counter and readout. The basic four-digit counter consists of four low-power Schottky BCD counters, the 74LS190 (see **fig. 1**). The BCD outputs of the counters are connected to special LED readouts which contain an internal latch/decoder. The readouts, HP-type 7300, are somewhat expensive. However, for the home brewer they immensely simplify construction.

As mentioned, the counter reads only the four most significant digits, since the MHz digits have always been read from the band switch in the past. Plus, the complexity of added digits might make the job more than the average ham would want for a home project. The motto here is keep it simple.

time base. The time base is a very simple circuit, It consists of a single IC (CD4060) and a crystal, a trimmer capacitor, and a resistor. The IC contains the necessary amplifiers for a crystal oscillator and 14 divide-by-2 stages. At the output of the last stage, labeled Q_{14}, the oscillator frequency has been divided by a factor of 2^{14}. Starting with a 409.6-kHz crystal, the final output is a 25-Hz squarewave. This output, plus the 50-Hz squarewave from Q_{13}, are used to generate the necessary counter timing pulses.

counter timing. Operation of a counter generally requires various timing pulses to control the counter. in a conventional counter, the count gate provides a precisely timed interval which allows the number of counts admitted to be equal to the frequency of the signal. Since frequency is measured in terms of events per second, this gate is always a fraction of a second. Or, in this case, where we want to read tc hundreds of Hz, the gate is exactly 0.01 second long. Other pulses are required for presetting the counter to a fixed starting number (frequency) and for transferring the final count to the readouts.

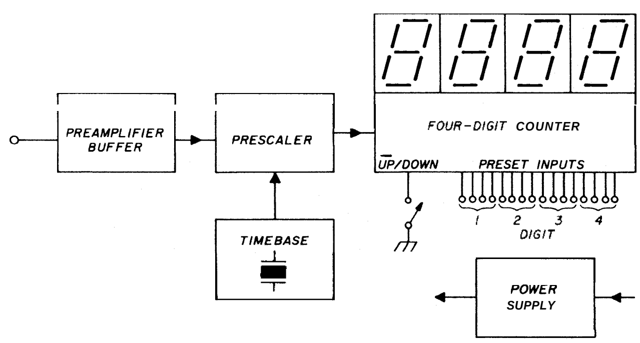


fig. 1. Simplified block diagram of the universal digital readout.

The count gate, the display, and the preset pulse are all derived from the two squarewaves provided by the time base (see **fig. 2**). The first two gates connected to the CD4060 buffer the CMOS outputs of the time base.

preamplifier. A preamplifier, though not always necessary, is a good idea. It not only increases sensi-

By Gerd Schrick, WB8IFM, 4741 Harlou Drive, Dayton, Ohio 45432

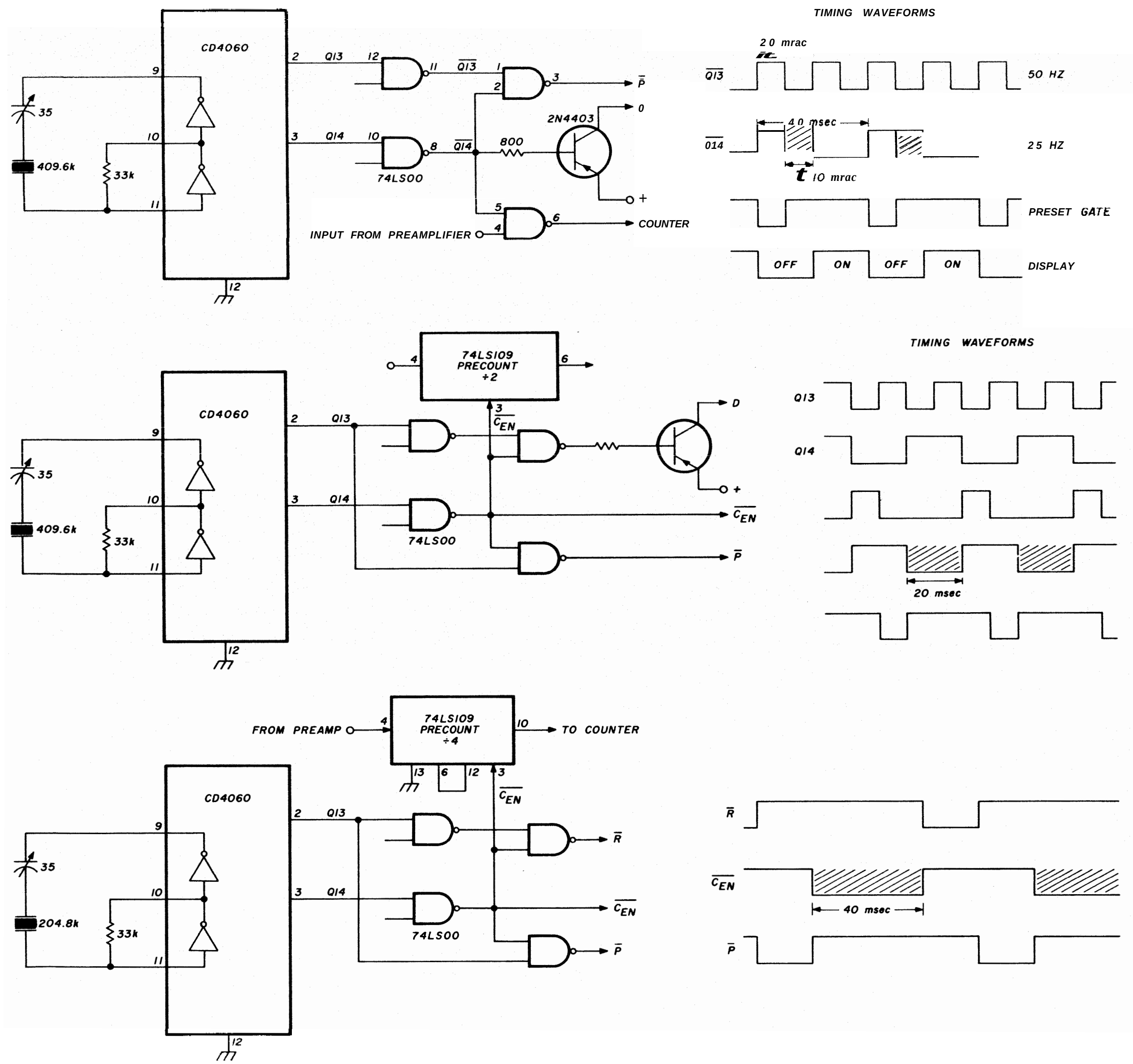


fig. 2. Schematic diagram of the three different timebase versions. In A, the count gate is actually enabled for 20 mS, but since the counters are held in the preset state for the first 10 mS, the time that the counters are allowed to be clocked is 10 mS. The low-power version, B, runs the display at a 25 per cent duty cycle. For the low-ambiguity version, C, the crystal is changed to 204.8 kHz, effectively quadrupling the count gate to40 mS.

tivity but also acts as a buffer, reducing possible spurious responses in the receiver generated by the timing pulses. A single transistor, as shown in fig. 3, is used in common-emitter configuration. The sensitivity is better than 50 mV RMS from 100 kHz to 30 MHz. The maximum voltage is about 1 volt RMS.

power supply. The digital dial, using mainly TTL-devices, requires a 5-volt dc power source. The current, depending on the desired version, will range from 170 mA for battery-powered equipment to 500 mA for the low-ambiguity 100 per cent display version.

last digit ambiguity

The problem last-digit ambiguity arises from the fact that the count gate, as generated from the crystal oscillator, is not synchronized to the incoming frequency. The gate will sometimes accept an additional count, changing the digit, for example, from a 5 to a 6. This is the well-publicized ±1 digit ambiguity that digital counters exhibit. One way of overcoming this ambiguity is to increase the number of digits counted, yet only display a limited number. This, in effect, is the same as simply covering the blinking digit.

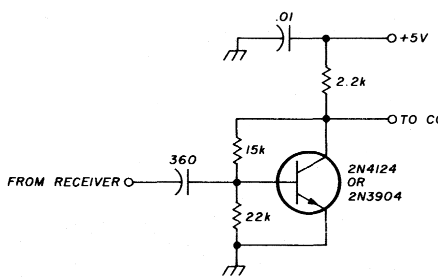


fig. 3. Schematic diagram of a simple bufferlpreamplifier which can be used between the counter and the receiver. Sensitivity is better than 50 mV RMS.

The conventional counter can achieve the additional counts only by lengthening the gate; if we consider another decade, the count gate would be 0.1 second, reducing the counter's final update rate to only 6-8 Hz. It is still possible, using the memory capability of the readout, to obtain a blink free and ***almost*** instantaneous update when turning the dial of the VFO.

However, it is not actually necessary to add a complete decade to the counter; any integer will do. How then is the ambiguity affected by the addition of the new counter? In reality, it never goes away. What does happen is that the probability that the ±1 ambiguity will occur is reduced by the reciprocal of the additional factor. For example, if you add a divide-by-two, the probability will be reduced by 112, or 50 per cent; for a divide-by-four it will be 114, or 25 per cent. However, for this reduced probability, there is a price that must be paid. The count gate will have to be ***lengthened*** by the same factor.

programming

To program the counters, the individual load lines, labeled D_A, D_B, D_C, D_D in **fig. 4,** are connected according to the required BCD code. To program a 5 into a particular counter, ground the data lines for D_B and D_D. The other data lines may remain open or connected to +5 volts.

A simple scheme using a single-pole, double-throw switch, as shown in **fig. 5,** can be used to preset the counter to two different starting frequencies. For more than two positions, a multiple-deck switch would be required.

counter options

standard version. This is the simplest form of the digital dial, with no precounting to reduce ambiguity. The time base generates a 0.01-second gate, giving a readout to the nearest 100 Hz. The display is updated at the rate of 25 Hz, with a display duty cycle of 50 per cent. Power requirements are 5 volts at 300 mA.

low-power version. In this version the time base output is slowed to 0.02 second, permitting the addition of a single divide-by-two counter which reduces the last digit flicker to 50 per cent. The display duty cycle is also reduced, to 25 per cent, giving a somewhat dimmer, but still quite visible, display. The update rate is still 25 times per second. Power consumption is under 1 watt (170 mA at 5 volts).

low-ambiguity version. For a little added complexity, this version is the most useful for fixed-station use. The time base is further slowed to 0.04 second by substituting a 204.8-kHz crystal for the 409.6 crystal. With this change, a divide-by-four prescaler, which reduces the last digit ambiguity to 25 per cent, can be used.

Since the update rate is now only 12.5 Hz, an intolerable flicker would occur if the display were switched at that rate. To eliminate the flicker, the latch in the

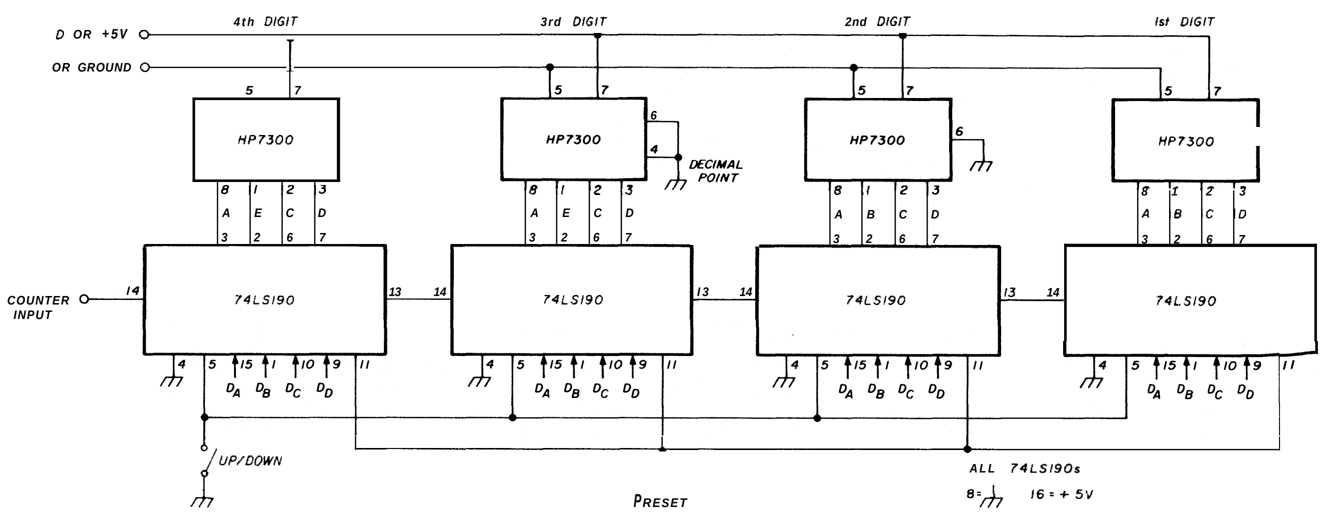


fig. 4. Schematic diagram of the four 74LS190 counters and the HP-7300 LEDs. The $\bar{R}$ line, connected to pin 5 of the 7300s, is used to strobe the latches when used in the low-ambiguity version. Pin 5 of the counters is taken low for up counting, and can be left open to count down.

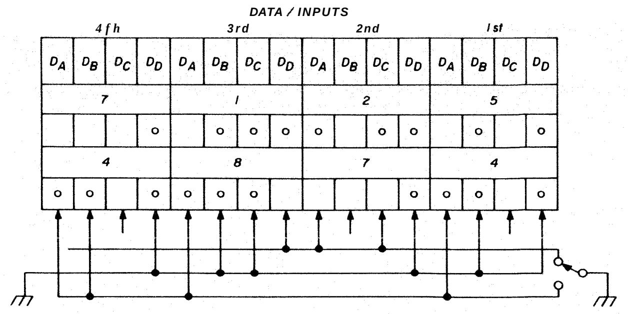


fig. 5. Example of using a single-pole double-throw switch to select different preset programming. Those lines that are always high are left open, while those that are always low are held low. The switch changes the level into the inputs depending upon the preset value.

7300s is also strobed, giving a 100 per cent display cycle. Even with the bright display power consumption is quite reasonable, 500 mA at 5 volts.

checkout and calibration

With an adequate power supply connected to the counter, checkout and calibration can be completed in a few simple steps:

1. Program the preset inputs according to the required BCD input. For a quick check, the number 7 can be programmed into the counter by grounding all pin 9s.

2. Apply a stable rf signal to the input of the preamplifier. The counter should count either up or down, depending upon the input to the control pin.

3. Next, apply a signal of known frequency (crystal calibrator, for instance). Check the displayed frequency against the input, the preset, and whether the counter was programmed to count up or down.

parts list

4 ea 74LS190
4 ea HP 7300 Numerical Indicators
1 ea CD 4060A
4 ea Resistors 2.2k, 15k, 22k, 33k, 1/4 W
5 ea Ceramic Cap 2x.1, 800p, 360p, 50p 50V
1 ea Trimmer Cap 35p
1 ea Rectifier Diode 1A, 50V PIV
1 ea LM 309 K IC
1 ea Electrolytic Cap 250 μF 15V
div. HW, sockets, chassis, etc.

Plus:

Standard Version
409.6 kHz crystal
PNP trans.
820Ω Res.

Low-Current Version
409.6 kHz crystal
PNP trans.
74LS109 IC
820Ω Res.

Low-Ambiguity Version
204.8 kHz crystal
74LS109 IC

If all readings are correct, the counter can then be permanently connected to the receiver.

bibliography

Gerd, Schrick, WB8IFM, "Digital Readout Variable-Frequency Oscillator," *ham* radio, January 1973, page 14.

ham radio

the Oscar Calcu-puter

If you are adventuresome enough to attempt automatic antenna tracking of Oscar, but don't have the necessary bucks to tie into a computer, this article should be just what you need. Even if mathematics isn't your strong suit, don't get discouraged, read on. The formulas for tracking Oscar are not that tough, especially if you apply one of the inexpensive hand calculators. Unfortunately, the main disadvantage of the hand-calculator method is the need to constantly manipulate the buttons, even for information for the Oscar pass. This article will explain my method of solving this problem — automating a small hand-held calculator. I'll even explain a few ideas for making a complete steering system. That way all you have to do is enter the equator crossing longitude, punch a button when Oscar crosses the equator, and from there on it's automated all the way.

program explanation

The terms I've used in the program (see **fig. 1**) have been summarized in **table 1.** In addition, I've assigned line numbers to each program step to make it easier to follow. Actually, the program is divided into six separate parts, each part solving one of the following equations:[1]

$$Lat(T) = \sin^{-1}[0.9790 \cdot \sin(3.1319T)] \quad (1)$$

$$Long(T) = \cos^{-1}\{\cos(3.1319T)/\cos[Lat(T)]\} + 0.25T + Lo \quad (2)$$

$$D = \cos^{-1}(\sin A \sin B + \cos A \cos B \cos L) \quad (3)$$

$$Az = \cos^{-1}[(\sin B - \sin A \cos D)/\cos A \sin D] \quad (4)$$

$$El = 90 - \tan^{-1}[4867 \sin D/(4867 \cos D - 3957)] \quad (5)$$

$$M = (4867 \cos D - 3957)/\cos(90 - El) \quad (6)$$

Steps 004 through 030 solve **eq. 1,** 031 to 058 **eq. 2,** 059 to 108 **eq.** 3, 109 to 148 **eq. 4,** 149 to 191 **eq. 5,** and steps 192 to 203 for ***eq*. 6.** Each step is actually a single key-stroke on the calculator. There are several steps that should be briefly explained. This might eliminate program questions as you follow the equations through the program.

Step 000 represents the unit being turned on. In addition, other circuitry resets the external logic back to a common starting point. The one-shot multivibrator which performs the reset function also enables the clock gate. Anytime the clock gate is enabled, the sequencer is allowed to advance to the next program step. If the gate is disabled, the program will stop on that particular step. This is an important feature, as I'll explain later on.

Steps 001 through 003 merely clear the calculator of any previous computations or stored answers. The next two steps shift the calculator into four decimal place readout. The calculations are actually done to the limit of the ICs involved, and the answers rounded. For all program steps you'll notice a listing for type of entry. This notation is explained in **table 2.**

Program steps 013 and 014 cause the time since the satellite crossed the equator to be entered into the calculator. This is entered as even minutes and results in a readout for antenna azimuth, elevation, and distance to the satellite for each minute of the pass. In theory, at least, if you provide the correct initial data and accurately enter the time, the calculator could provide information for tracking for the next pass, or even several later passes. This is limited only by the accuracy of the entries you make, and could easily be updated.

The last unusual steps are 038 and 039. In some calculations it is easier to find a denominator before

By Dave Brown, W9CGI, Route 5, Box 39, Noblesville, Indiana 46060

fig. 1. Program solved by the Calcu-puter. When broken into parts, this program will solve the six equations necessary to track Oscar.

Type of Entry	Program Step	Key	Type of Entry	Program Step	Key	Type of Entry	Program Step	Key	Type of Entry	Program Step	Key
*	000		S	051	EQX 1	Y	102	+	Me	154	STO
K	001	DPS	C	052	DP	Me	103	STO	Me	155	7
M	002	CLR	S	053	EQX .1	Me	104	2	Me	156	RCL
M	003	CLX	S	054	EQX .01	T	105	INV	Me	157	2
D	004	UP,	M	055	+	T	106	COS	M	158	X
D	005	4	Me	056	STO	Me	107	STO	Co	159	3
Co	006	3	Me	057	5	Me	108	3	Co	160	9
C	007	DP	M	058	CLX	Me	109	RCL	Co	161	3
Co	008	1	Cq	059	LAT 10	Me	110	7	Co	162	7
Co	009	3	Cq	060	LAT 1	M	111	ENT	M	163	
Co	010	I	C	061	DP	Me	112	RCL	Me	164	STO
Co	011	9	Cq	062	LAT .1	Me	113	2	Me	165	8
M	012	ENT	Cq	063	LAT .01	M	114	X	Me	166	RCL
Ti	013	Ttens	Me	064	STO	Me	115	STO	Me	167	7
Ti	014	Tones	Me	065	6	Me	lib	5	M	168	ENT
Me	015	STO	T	066	SIN	?le	117	RCL	Me	169	RCL
Me	016		Me	067	STO	Me	118	8	Me	170	6
M	017	X	Me	068	7	M	119	EN?		171	X
Me	018	STO	Me	069	RCL	Me	120	RCL	M	172	
Me	019	2	Me	070	3	Me	121	5	T	173	INV
T	020	SIN	T	071	SIN	M	122		T	174	TAN
C	021	DP	Me	072	STO	Me	123	STO	Me	175	STO
Co	022	9	Me	073	8	Me	124	5	Me	176	9
Co	023	7	M	074	X	M	125	CLX	Co	177	9
Co	024	Y	Me	075	STO	Me	126	RCL	Co	178	0
Co	025	0	Me	076	9	Me	127	1	M	179	ENT
M	026	X	M	077	CLX	M	128	ENT	Me	180	RCL
T	027	INV	Me	078	RCL	Me	129	RCL	Ye	181	9
T	028	SIN	Me	079	6	Me	130	3	Y	182	
Me	029	STO	T	080	COS	T	131	SIN	Me	183	STO
Me	030	3	Me	081	STO	Me	132	STO	Me	184	1
T	031	COS	Me	082	1	Me	133	6	D	185	DPS
Me	032	STO	M	083	ENT	M	134	X	D	186	0
Me	033	4	Me	084	RCL	Me	135	RCL	R	187	GoSUB2
Me	034	RCL	Me	085	4	Me	136	5	D	188	DPS
Me	035		M	086	ENT	Y	137		D	189	4
T	036	COS	Cq	087	LONG 100	K	138	DPS	Me	190	STO
M	037		Cq	088	LONG 10	K	139	1/X	Me	191	6
K	038	DPS	Cq	089	LONG 1	T	140	INV	Ye	192	RCL
K	039	1/X	C	090	DP		141	COS	Me	193	8
T	040	INV	Cq	091	LONG .1	D	142	DPS	M	194	ENT
T	041	COS	Cq	092	LONG .01	D	143	0	Me	195	RCL
C	042	DP	M	093	ENT	R	144	GoSUB1	Mc	196	1
Co	043	2	Me	094	RCL	D	145	DPS	T	197	COS
Co	044	5	Me	095	5	D	146	4	M	198	
Me	045	RCL	M	096		Me	147	STO	D	199	DPS
Me	046	1	T	097	COS	Me	148		D	200	0
M	047	X	M	098	X	M	149	CLX	R	201	GoSUB3
M	048	+	M	099	X	Co	150	4	Ye	202	STO
S	049	EQX 100	Me	100	RCL	Co	151	8	Me	203	5
S	050	EQX 10	Me	101	9	Co	152	6	**	204	STOP
						Co	153	7	**	205	END

its associated numerator. This leads to dividing the denominator (d) by the numerator (n), or d/n. To get the correct answer (n/d vs d/n), the reciprocal key ($1/X$) is used after the division answer has been obtained. On my calculator, I have to use the DPS key to access the $1/X$ function. This accounts for the use of two steps.

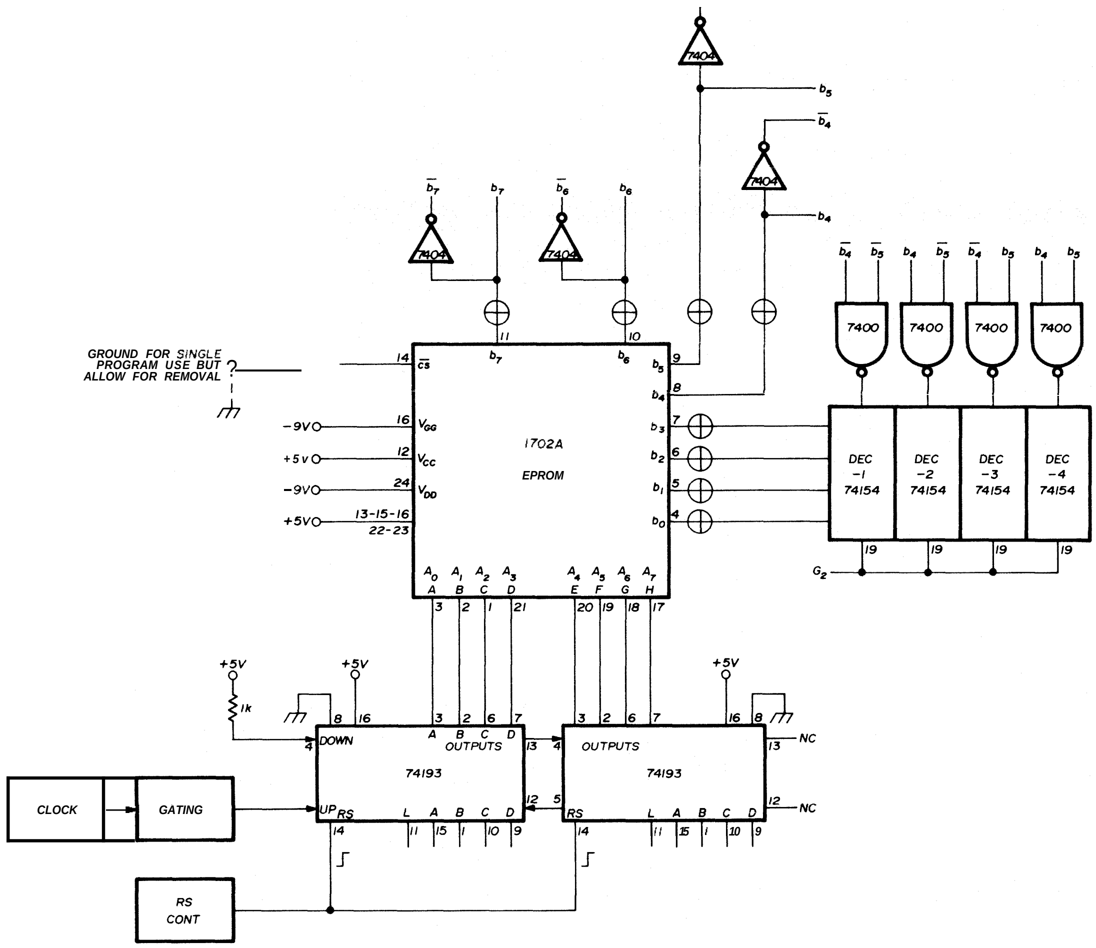
circuit description

The heart of the automating device is a 1702A EPROM (see **fig. 2).** The program, from **fig. 1,** is entered into the PROM such that the outputs, when decoded, will electrically press the appropriate keys on the calculator. For general use, a RAM would be more appropriate, but I wanted to solve one specific problem: the equations necessary to track Oscar.

The 1702A PROM has 256 distinct address locations, enough to handle the 205-step program. Each step of the program represents a sequential binary address in the PROM. As seen in the schematic diagram, 74193, 4-bit binary counters are used to sequentially address the 1702A. I decided to use the

table 1. Definition of terms used in the Calcu-puter program.

(T)	Time in minutes since EQX
Lat(T)	Satellite sub-point Latitude in degrees at (T)
EQX	Satellite equator crossing reference
Long(T)	Satellite sub-point Longitude in degrees at (T)
Lo	Satellite sub-point Longitude in degrees at EQX
D	Great Circle distance Stn to Sat in degrees
Stn	Station location (your QTH)
Sat	Satellite location
A	Latitude of Stn in degrees
B	Lat(T)
Ls	Longitude of Stn (QTH) in degrees
L	Ls - Long(T)
Az	Azimuth bearing for antennas (from true North)
E or El	Elevation bearing for antennas (from horizon upward)
M	Distance in statute miles from Stn to Sat (true position)



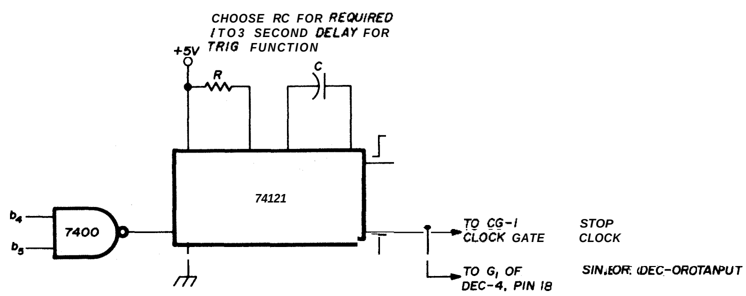


fig. 2. Schematic diagram of the input and output logic connected to the **1702A** EPROM. The preset inputs for the 74193s are normally connected to V_{cc} through 1000-ohm resistors: The clock and gating connect to the input of the first 74193 as shown in fig. **3. If** required, buffers can be inserted between the outputs of the PROM and any external devices. The actual pin numbers on the inverters and **NAND** gates have not been shown to allow flexibility in other systems. To incorporate the trigonometric delay feature, a 74121 is inserted between the decoder and the 7430 **NAND** gate.

74193 instead of the 7493, taking advantage of the preset capability. This means that the program can be started at any spot by simply entering the correct starting address into the data inputs and momentarily taking load line low. If this capability is not desired, the 7493 could be used.

Eight output lines are available on the **1702A.** For direct calculator control, I've only used six of the available outputs. The first four outputs, b_0 to b_3, are used as normal addresses for 74154 one-of-sixteen decoders. The outputs are simultaneously applied to all four decoders. The b_4 and b_5 outputs from the **1702A** are also decoded and used to select the appropriate 75154.

The final two outputs, b_6 and b_7, are used as a program stop and program halt. When step 204 is addressed, the output from the PROM will be 01000000. The high level from the b_6 output is detected and used to stop the program until a new "minutes" time is entered. This is one of the different means of disabling the clock gate. Output b_7 is programmed in a like manner to provide a high output at step 205. This output will stop the program, regardless of the minutes timer.

equator, the first flip-flop is set, which in turn sets the second flip-flop. Having both set will enable the clock gate.

I have tried to divide the decoders into a logical order, with DEC1 using the binary codes for number 0 through 9, to directly decode the number information. The keys for the four basic math functions are also included in DEC1. As seen in **fig. 4,** outputs from the decoders are used to drive open-collector buffers. The buffers then drive the reed relays which are connected across the calculator keys. It is imperative that the relays have a very high resistance across the open contacts, and also a very low closed-contact resistance.

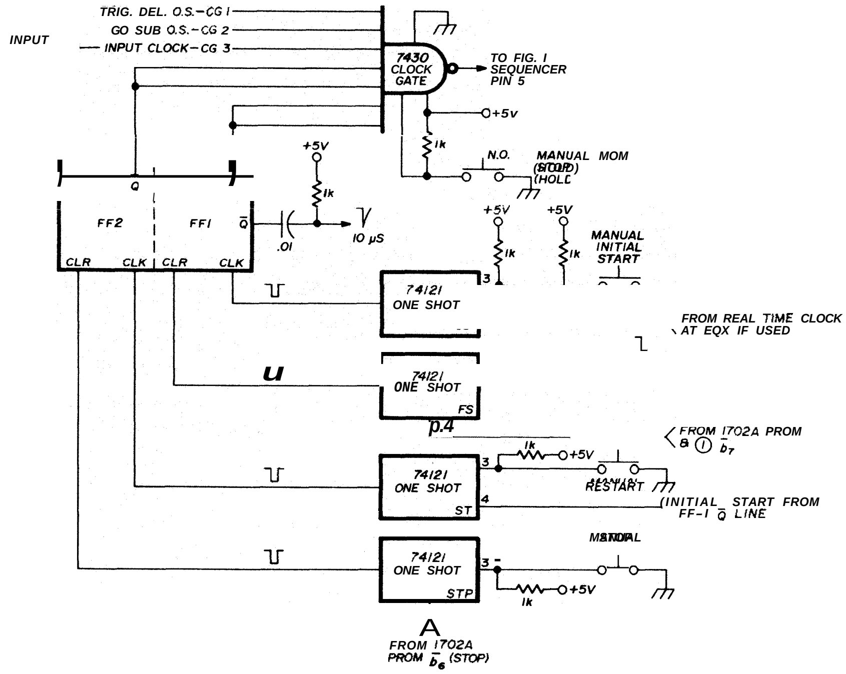


fig. 3. Schematic diagram of the clock gating logic. The actual input clock frequency will depend upon the speed at which your calculator can do the computations. For initial testing, it could be as slow as 1 pulse per second.

Two flip-flops are used for clock gating. As seen in **fig.** 3, the one-shot multivibrators receive the various start and stop commands. The pulses are then used to trigger the flip-flop into the desired states. In addition, provisions have been made to interface a real-time clock to signify the equator crossing. When you initiate the start command as Oscar crosses the

table 2. Type of entry notation used in the program.

*	Power on, reset timer (T) and program sequencer to 000, and all readouts to zero
M	Machine function – CLX, ENT, +, – etc.
D	Change of decimal point location
Co	Constants defined by Oscar and put into PROM
Cq	Constants defined by your QTH and put into only your PROM
Ti	Entry from timer output
Me	Storage or Recall function to/from memory and number
T	Trigonometric function (and added delay trigger)
K	Keyboard shift function (SHIFT that is not DPS)
S	EQX Longitude entry by switches
C	Constant (decimal point entry)
R	Readout Sub-routine function (external to calculator)
**	Operational system command (Stop, End)
STOP	Halts calculator until next timer period enters
END	Detects maximum period of pass elapsed – full stop

A complete listing of the respective decoder addresses is given in **table** 3. Note that the first address in DEC1 does not have an associated function. This is to prevent a problem when the step 204 and step 205 commands are initiated. If the address were used, you would have a simultaneous key closure in addition to either a stop or halt command.

You'll notice that the only functions in DEC4 are the trig functions. This was done for a very specific reason. In most calculators, depending upon the IC set used, a trigonometric operation will take longer to perform than a basic math function. This was also true in the calculator I used. To overcome the problem, I needed some method of momentarily stopping the program until the calculator had completed the trig operation. Otherwise, the program might have advanced several steps without correct data from the trig operation. By placing these operations in DEC4, detecting any 0011xxxx number will automatically indicate that a trig function is present. After the function has been detected, a one-shot disables the clock gate long enough for the calculator to do the computation.

Instead of using this program-delay technique, the time between steps could be made long enough to allow for a trigonometric operation, but this will considerably slow down the time necessary to perform all the calculations. Using the one-shot requires a few more parts, but the tradeoff is worthwhile, considering the time saved.

The GoSub routines, listed in DEC3, are used to output the data from the calculator to external readouts and other external processing. Gating for the GoSub routines is shown in **fig. 5,** with a quasi-schematic diagram of the readout system shown in **fig. 6.**

On the subject of the GoSub routines, you'll notice a DPS 0 step just before each GoSub step in the program. This truncates the display to eliminate any numbers to the right of the decimal point, and also shifts the answer to position the units digit on the extreme right of the display.

In the multiplexed displays (as used in my calculator), the same segments of each display are tied together, with a digit strobe activating the appropriate digit. The Calcu-puter, as I've aptly named it, is interfaced to external readouts by connecting the segment information lines and the data strobe lines to external latches. **Fig. 6** is not an absolute schematic diagram since the voltage levels and required interface, will differ between calculators.[2] You'll also notice the use of digital information to indicate the actual antenna position. This information, combined with the Calcu-puter information, nicely lends itself to completely automated antenna control.[3]

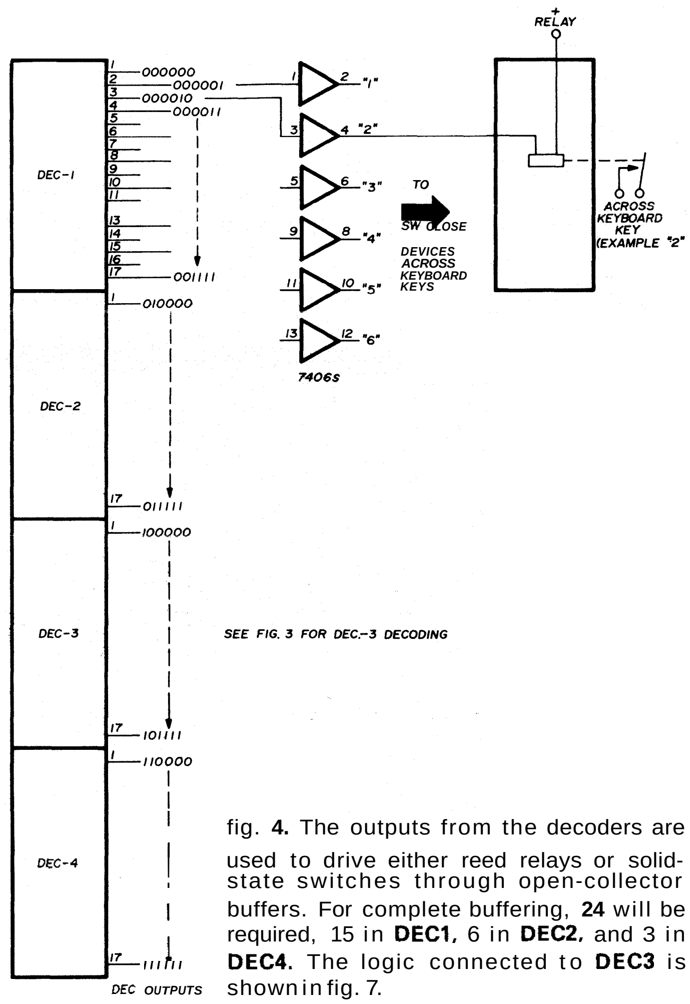


fig. **4.** The outputs from the decoders are used to drive either reed relays or solid-state switches through open-collector buffers. For complete buffering, **24** will be required, 15 in **DEC1,** 6 in **DEC2,** and 3 in **DEC4.** The logic connected to **DEC3** is shown in fig. 7.

DEC3 also decodes the commands necessary to enter the equator crossing longitude and time since crossing from the external BCD switches. As seen in ***fig.* 7,** the switches and outputs from the timers are OR-wired and used to feed a **7445** BCD-to-decimal

table 3. **DEC** output functions. In **DEC 2,** only the underlined functions are used in the program.

DEC 1 binary	DEC 1 decoded function		DEC 2 binary	DEC 2 decoded function		DEC 3 binary	DEC 3 decoded function	DEC 4 binary	DEC 4 decoded function
000000	not used		010000	not used		1000000	not used	110000	not used
000001	1		010001	SHIFT	(DPS)	100001	EQX 100	110001	sine
000010	2		010010	CLX	(clr)	100010	EQX 10	110010	cosine
000011	3		010011	ENT	(sci)	100011	EQX 1	110011	tangent
000100	4		010100	STO	(INV)	100100	EQX .1	110100	not used
000101	5		010101	RCL	(HYP)	100101	EQX.01	110101	not used
000110	6		010110	EE	1/x	100110	(T) 10 (minutes)	110110	not used
000111	7		010111	ln	(log)	100111	(T) 1 (minutes)	110111	not used
001000	8		011000	sigma+	(x,s)	101000	Go Sub 1	111000	not used
001001	9		011001	+/−	(x!)	101001	Go Sub 2	111001	not used
001010	Ø		011010	x↔y	(%)	101010	Go Sub 3	111010	not used
001011	D.P.		011011	roll x	(delta%)	101011	not used	111011	not used
001100	+	(add)	011100	y^x	(sqr x)	101100	not used	111700	not used
001101	−	(sub)	0111101	not used	not used	101101	not used	111101	not used
001110	x	(mult)	0111110	not used	not used	101110	not used	111110	not used
001111	÷	(div)	0111111	not used	not used	101111	not used	111111	not used

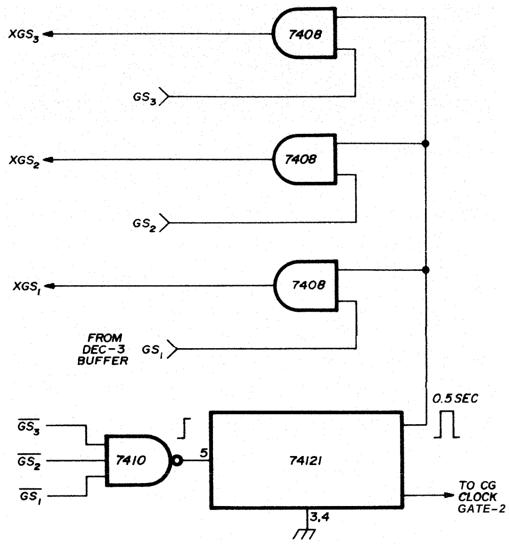


fig. 5. Details of the **GoSub** routine logic. The output pulse, in conjunction with the digit strobe, is used to enter the output data into the 7475 latches.

decoder. The output from the decoder also drives reed relays connected across the calculator keys.

Fig. 6 shows digital information indicating the actual azimuth and elevation of my antennas. In addition to being applied to the 7485, 4-bit comparators, the information is applied to BCD-to-7 segment decoders and readouts. Also, fig. 7 indicates where decoders could be connected to readout the time since equator crossing. With all the information converted for readouts, I have a panel that shows displays of: azimuth (beam), azimuth (calculated), elevation (beam), elevation (calculated), distance M (calculated), distance M (for later use), time, and sequencer location. The sequencer location was included as a troubleshooting aid should the program ever stop.

Limit switches have been included in my system to stop the antenna from going beyond the prescribed limits. If you run the program only during valid pass times, the program should never produce invalid commands. But, should this ever be a problem, the limit switches will prevent major damage. You can readily see from the readouts where the problems are if they occur.

Trigonometric functions near 0 or 90 degrees, and numbers which result in zero denominators ***can*** give the program fits, but there just isn't any easy way around this. I haven't found it to be a problem, however, except on way out, very short passes. A final note on PROM programming: in steps 059-063 and

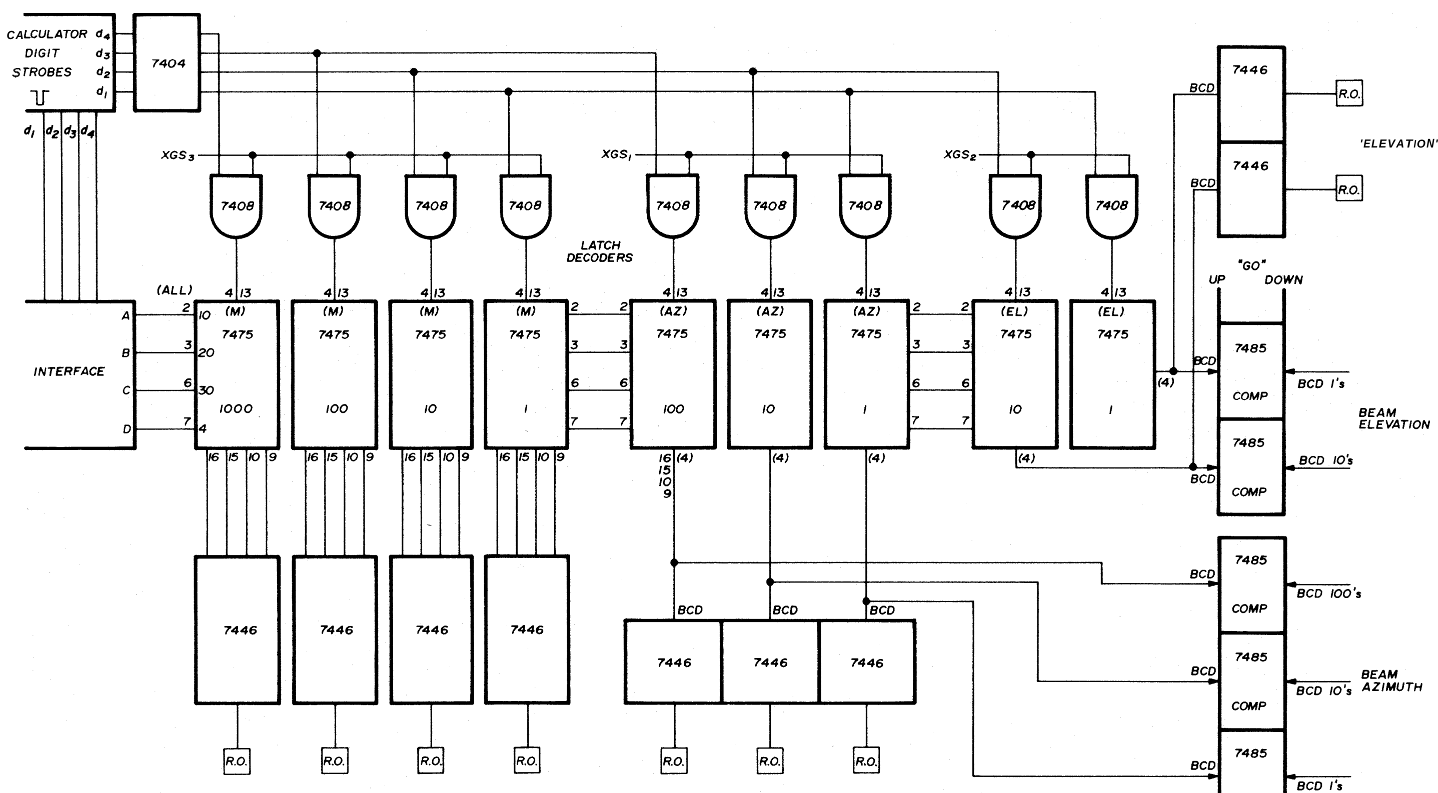


fig. 6. Quasi-schematic diagram of the readout system used with the Calcu-puter. It will be necessary to in- sert an interface device between the 7-segment information and the BCD format required by the 7475s. The author uses another **1702A** EPROM programmed to do the necessary conversion. An alternate method would be to use the National **74C915** to convert the data. Though not shown in this diagram, the author has also connected decoders and readouts to indicate the actual antenna position. The **7404** buffers between the calculator and the latches may have to be changed depending upon the type of strobe coming from the calculator.

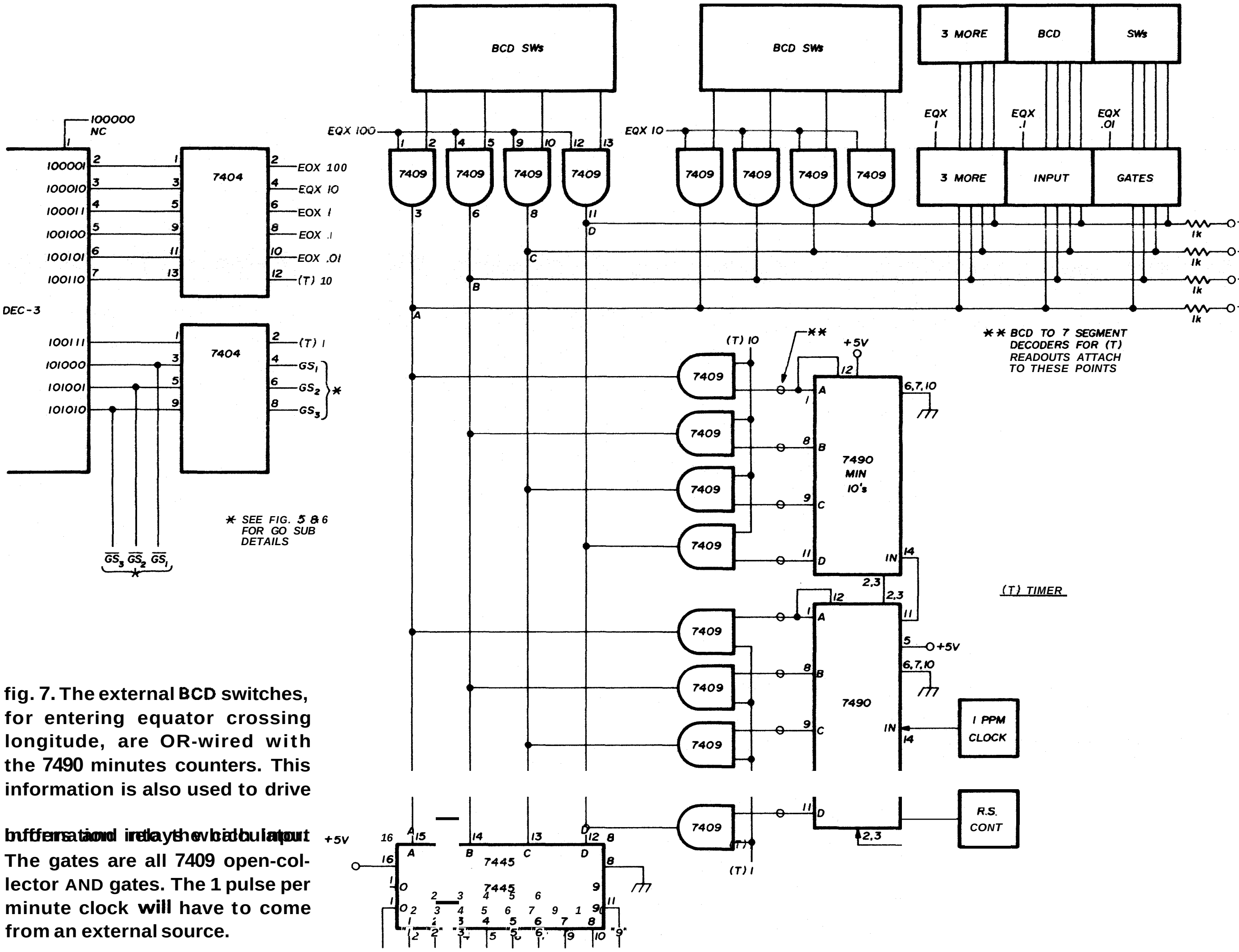


fig. 7. The external BCD switches, for entering equator crossing longitude, are OR-wired with the 7490 minutes counters. This information is also used to drive buffers and relays which output information to the calculator. The gates are all 7409 open-collector AND gates. The 1 pulse per minute clock will have to come from an external source.

087-092 be sure to enter the latitude and longitude for *your* location. This will be retained as permanent information in the PROM.

concluding comments

The primary message of this article has been to show you that a complete computer/microprocessor is not required to do simple math problems. The PROM is in a sense a simple BASIC language like no other. It has automated a calculator, providing both for inputing and outputing of data, much in the same way as a full-scale computer.

I did write a program in algebraic notation instead of RPN, but quickly discarded it when I couldn't find an inexpensive calculator with enough onboard memory. Lacking this capability meant dumping out the interim answers, performing more calculations, and retrieving the interim answers before the final numbers could be outputed. It generally amounted to a lot more hardware, fast approaching a full-blown computer, a mess that I wanted to avoid from the beginning. The APF 55 calculator I finally used was provided by a friend because some of the digit segments would not light. It was about as cheap, and definitely quicker, for him to buy a new calculator and give me his remains! Shop around because the price on some of the very sophisticated units is getting ridiculously low. For that matter, one of the many calculator shops around these days might part with some of their damaged returns, for the right price.

For anyone wishing additional information, a self-addressed, stamped, envelope will bring a quick reply; and any comments on improvements to the system will be welcomed.

references

1. Bob Henson, WB0JHS, "Computerized Satellite Tracking," 73, Februaty, 1977, page 72.
2. Bruce McNair, NZYK and Glenn Wiiliman, N2GW, "Digital Keyboard Entry Systems," ham radio, September, 1978, page 92.
3. David Brown, W9CGI, "Introducing Autotrak," 73, July, 1977, page 46.

ham radio

simple video display

Two projects
to get you started
in building a
video display unit
using readily
available devices

There has been much recent interest in video display units. They can be used as part of a video typewriter, for putting up displays on ATV and SSTV, and for decoding RTTY and Morse off the air. Most of the displays are complex and expensive. Even the available kits aren't suitable for those who haven't had much experience. For normal use you don't need whole screens full of characters, and the simple 32-character single-line display described here is an excellent beginning for those who would like to play around with an inexpensive video display unit.

description

The heart of the unit is the Fairchild 3258 dot-matrix character generator IC. Externally it's a 16-pin package (I hate to think what's inside it!), which accepts ASCII inputs and produces 64 characters on a 5 x 7 dot matrix. Apart from the inputs and outputs, the only other signal connections to the chip are inputs to a clock and a master reset. The chip has an internal clock and addressing system. After the master reset input operates and goes high, the information representing the first row of the character is available after the first clock pulse. Subsequent clock pulses select the next six rows in turn; after that, the outputs are clamped high and the character generator stops until another master reset pulse appears. Thus, if the character-generator clock is pulsed at line frequency, the character will appear on the screen.

experimental system: one character 32 times

Fig. 1 shows the logic diagram of the display unit. Only nine integrated circuits, including the character generator, are required. There is no reference crystal or dividing network. I used a monitor from a non-composite camera and monitor combination and simply fed the horizontal and vertical sync pulses into the VDU. If you wish to use a regular TV set, it's quite easy to add two 555 timers to provide horizontal and vertical sync pulses. **Fig.** 2 shows the connections for the monitor, which would be typical, and **fig.** 3 shows the circuit for the 555 timers required.

system operation

The second half of the 74123 feeds a gate, which feeds a second gate, which in turn feeds back into the 74123. This action sets up an oscillating circuit whose frequency is determined by the 5k pot in the +5-volt line. This frequency is used to step the 74195 shift registers and provides the basic video signal. The character-generator output is loaded into the 74195 shift registers then clocked out in a serial mode at the VIDEO OUT terminals. The J and $\overline{K}$ inputs of the first shift register cause highs to be entered as the data from the character generator is shifted along. Finally, when the six outputs to the 9007 gate are all high, a low is sent on the MOD 7 counter line, which reloads the shift registers and clocks the **horizontal character count** 7493 ICs.

By Roy Hartkopf, VK3AOH, 34, Toolangi Road, Aiphington, Victoria, Australia 3078

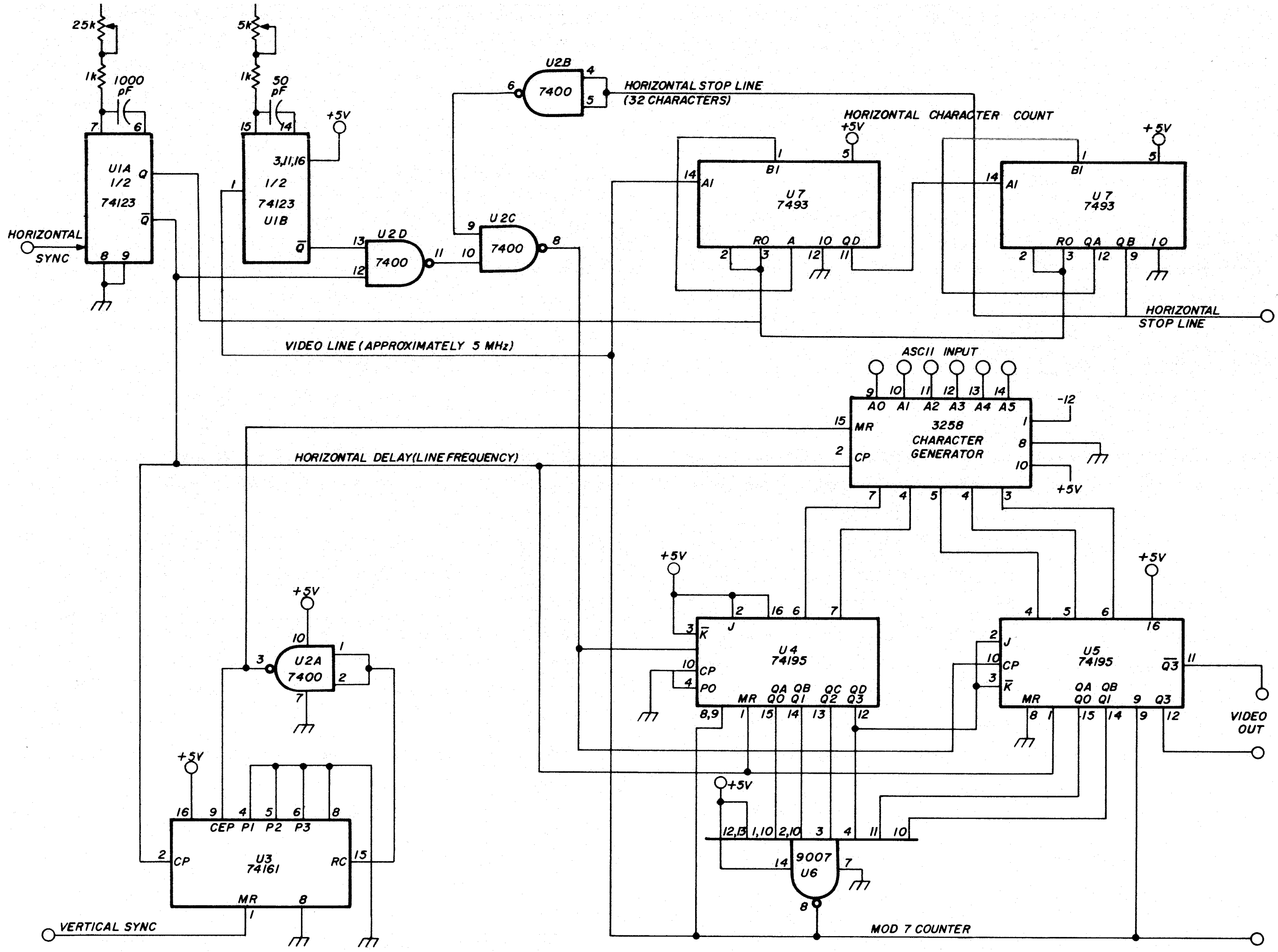


fig. 1. Logic diagram of the video display unit (VDU). Only nine ICs, including the character generator, are used.

When this action occurs 32 times, the 7493 provides an output to the HORIZONTAL STOP line, which inhibits the 5-MHz oscillator and stops the sequence.

This action occupies about two-thirds of a single horizontal line. When the end of the TV line is reached, a horizontal sync pulse operates the first half of the 74123, resetting the 7493 counters, providing a clock pulse to the character generator. This pulse acts to output the information for the next line and resets the 74195 shift registers. The sequence then repeats for the next line.

The character generator automatically blanks out after a complete row of characters has been sent, and if it's required to have more than one row, the 9316 will count the rows and reset the character generator. **A** vertical sync pulse resets the 9316, so the information is always at the same position on successive frames.

There are no critical adjustments in this circuit. The 25k pot in the first half of the 74123 positions the first character on the left-hand side of the screen, and the 5k pot in the second half opens up or closes the 32-character-length display so it can be spaced evenly across the screen.

The logic shown in **fig. 1** will produce a display of one character repeated 32 times across the screen. The character will be determined by the ASCII input to the character generator. For test purposes you can apply a combination of 5-volt and ground inputs as required. This can be treated as a project in itself, so that those who want to take a bit at a time can do this, then go on to the second half of the project.

32 different characters

The logic for the second half of the project is shown in **fig. 4** and again is quite simple. Here, the main device is a Fairchild **3349** hex 32-bit static shift register. Like the character generator, the 3349 has a

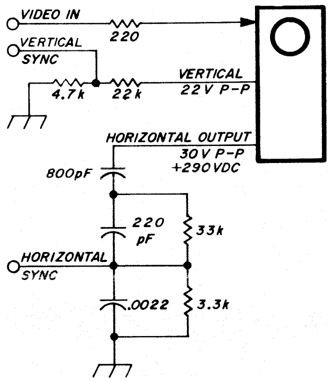
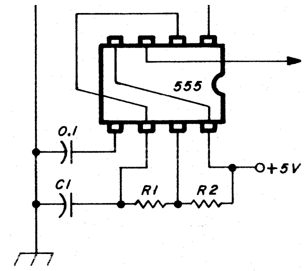


fig. 2. Connections for a typical monitor (Toshiba TMC2AX).



	R1	R2	C1
VERTICAL	4.7k	100k	0.047
HORIZONTAL	4.7k	22k	0.0022

fig. 3. Circuit using the NE-555 timer to provide horizontal and vertical sync pulses if a regular TV set is used as a monitor.

deceptively simple 16-pin package; and again, there is a complex integrated circuit inside.

Only two signal controls are needed: a clock input and a load/recirculate input. The clock pulse steps the 32 bits in each of the shift registers, which recirculate until the LOAD/RECIRCULATE input goes low. Then new data is accepted and the data at the other end is lost. When the LOAD/RECIRCULATE input goes high again, the 32 bits in each of the six shift registers at that time resume recirculating. To obtain 32 different characters across the screen, the shift registers must present the six new bits to the character generator each time the Mod 7 counter operates so that the shift-register clock is fed from the Mod 7 counter. This action would produce 32 characters, but they would be random characters that happened to come up when the display was first switched on. So we must have some way of putting in the characters we want. This is done by simuitaneousiy presenting the required ASCII code to the shift-register inputs and applying a negative key pulse to the set input of a flip-flop. This action sets the flip-flop output high and puts a high on the data input of a second flip-flop (both halves of a 7474). A pulse from the HORIZONTAL STOP line (at the end of the display of the 32 characters) clocks the second flip-flop. This allows a low to be put on the load input and also operates a gate, allowing an extra clock pulse (from the horizontal sync) to clock the new data into the shift registers. Then, at the end of the horizontal sync pulse, the flip-flops are reset and the 31 old characters and the one new one recirculate until another character is entered.

The two 74121s merely give **a** controlled-length

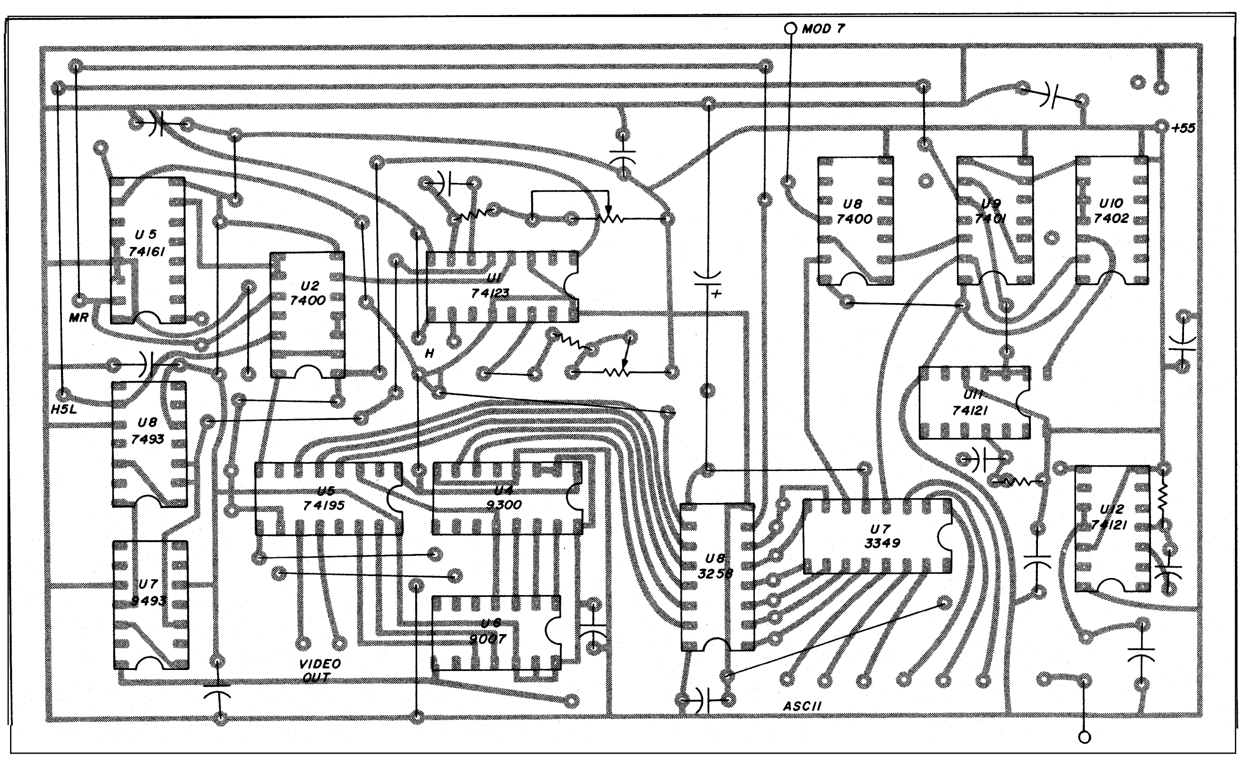


fig. 5. Full-size PC-board layout for both sections of the display unit. It has been tested by the author and works perfectly.

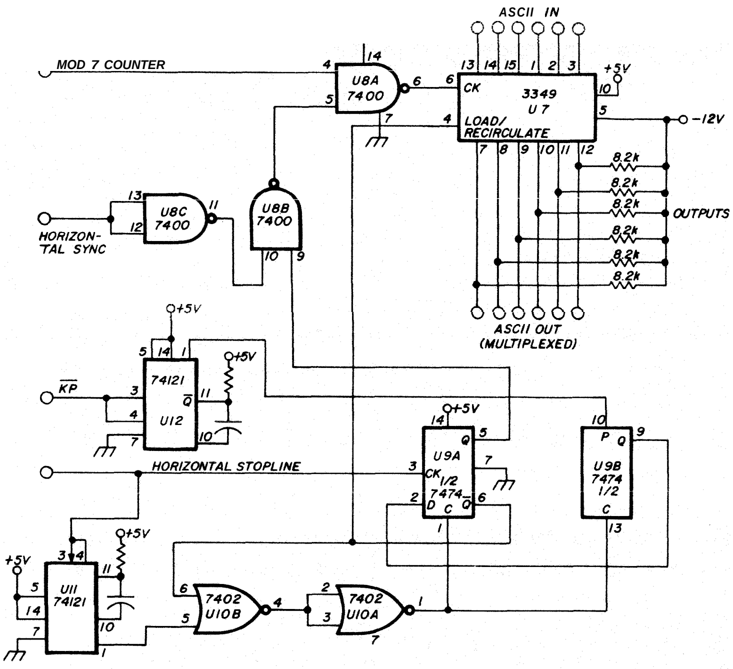


fig. 4. Video-display logic for producing 32 different on-line characters. The main device is a Fairchild 3349 hex 32-bit static shift register.

pulse and could be replaced with a resistor-capacitor combination, but they were used because the pulse length is more controllable. Apart from the shift register and flip-flops, the only other ICs are a couple of normal gates. The only other point worth mentioning is that the shift register outputs require external resistors (8.2k) from each output to the −12-volt supply.

construction

Two simple circuit boards about 3 inches (76mm) square will accommodate the entire system, or it can be built on a slightly larger board (**fig. 5**). It's a good idea to use sockets for the character generator and the shift register. Sockets for the other devices are a matter of personal preference.

When testing the circuit be very careful not to let the −12-volt supply get into any of the +5 volt TTL devices — it can have disastrous results!

final remarks

This project will give a beginner in this area an insight into the principles of VDUs and provide an excellent starting point for developing something more complex. To keep the project as simple as possible no attempt has been made to eliminate additional lines, so there will be several identical lines of 32 characters across the screen. In practice, they help rather than hinder reading the characters.

ham radio

updating the Collins 32S-1

Never being satisfied with the *status quo* when it comes to any radio equipment I've ever owned, I eventually succumbed to the urge to modify my recently acquired Collins 32S-1 transmitter. The modifications described here include the following:

1. BFO generation of the CW carrier
2. Voltage regulation of the PTO and HFO
3. Control of the keyed wave shape
4. A spotting switch (CW CAL)
5. The ability to monitor the final-amplifier plate (cathode) currents individually
6. Alterations to the tone oscillator

The modifications were made to bring the performance of the 32S-1 up to the standards of its successor, the 32S-3, without incurring an expenditure of some $300-$400 in the process.

Table 1 identifies the components involved in the modifications discussed here. Schematics and parts lists should be changed accordingly to reflect these changes, since removed components will have their identities transferred to newly installed pieces that correlate with those used in the 32S-3.

BFO CW generation

The 32S-1 generates its CW carrier with a tone fed from the tone oscillator through the mechanical filter (much like whistling into the mike or feeding AFSK into the mike jack on RTTY). The frequency of the tone used in the 32S-1 was chosen specifically so that its second harmonic falls well outside the mechanical-filter passband. However a weak residual signal still exists, and it has been heard on occasion at some distance.

The 32S-3, uses the BFO signal to generate the CW carrier, eliminating this residual signal. The resultant on-the-air signal is much cleaner and sounds much more like a true CW signal when compared with that of the 32S-1.

Installing this feature requires extra switching capabilities, which must be performed by the EMISSION switch, S8. The 32S-1 has four wafers on this switch, while the 32S-3 has five. Here are some ways in which this additional switching may be handled; a separate 4 PDT toggle switch may be used; S8 may be entirely replaced; or the existing switch may be disassembled and a new index and wafer added. Although the first possibility was initially pursued, I found it to be inconvenient. The most satisfactory arrangement was to replace the index assembly and add an additional wafer to S8.

The MIC GAIN pot and switch must also be replaced with a new unit using two pots commonly controlled and switch S14. The additional pot controls the cathode bias (CW DRIVE) on the rf amplifier, V6. Both parts are available from Collins; the switch is part no. CPN 259-1628-000 and the dual pot and switch is part no. CPN 376-2648-0000.

First, replace the existing MIC GAIN pot with the new dual unit. Note that space is at a premium, and the possibility of a shorted terminal strip lug exists next to V12. To avoid this, mount a two-lug terminal strip on the opposite side of the crystal board and secure it with the self-tapping screw that holds another two-lug strip. Remove the B+ ends of R60, L20, and the B+ feed wire (green/white) from their original location. Attach them to the new terminal strip. The now empty lug may be bent over to clear the pot and switch R8/S14.

Mount a single-lug terminal strip under the hardware securing the two ground lugs between V13 and V4. Lift C20 (0.01 μF) from ground and connect it to the strip. Route a length of RG-174/U cable from this junction to the vicinity of S8. Lift R39 (V6, pin 7) from ground and connect that end to a single-lug strip that has its ground lug straightened and soldered to the ground shield/barrier across V6. From this same point, run a wire to R8B and install a new R71 (68k/2W) between this lug and the terminal lug near V5 where R29 and R30 (4.7k/2W) connect to the +275-volt line (red/white wire).

From R8B run another wire to S8-B lugs 9 and 10, which are then connected in parallel. In the 32S-1

By Paul K. Pagel, N1FB, 4 Roberts Road, Enfield, Connecticut 06082

these two lugs are empty, as are those on the wafer to which R87 (470 ohms) is attached. Connect the empty lug of R8B to ground.

Connect a 33-ohm resistor (new R70) to V2A pin 9. Remove the BFO input cable. At this time S8 should be modified or replaced. Assuming the index and wafer are to be replaced and added, remove the switch and thread some bare wire through the rivet holes (which secure the switch contacts on wafers 1 through 4) at two points 180 degrees apart to prevent the spacers from separating from the wafers. Then the existing index may be removed, replaced, and the 5th wafer added with little effort.

Wire the switch as shown in fig. 1. Run a wire from S8B lug 11 (presently empty) to V10 pin 1 to prevent premature VOX relay dropout on CW.

ALC modification

Unlike the 32S-1, the 32S-3 does not use ALC in the CW position. During CW, switch selection S8G-5 grounds the midpoint of ALC capacitors C83 and C142. This change may be added to the modified

table 1. Component identification for the 32S-1 mods described in the text.

32S-3 part no.	original 32S-1 part no.	value	modified 32S-1 part no.	value	location
C81	not used		C81	0.005 μF	second mixer
C115	C115	0.01 μF	C115	0.33 μF	keying circuit
R17	R17	33k/1W	R17	5k/10W	voltage regulator
R70	R70	470k/1/2W	R70	33 ohm/1/2W	V2A
R71	R71	470k/1/2W	R71	68k/2W	B+

32S-1 by simply adding a jumper wire from S8G-1 and -2 to S8G-5 (fig. 2). Now, during CW operation, the GRID CURRENT position (instead of ALC) is monitored, and the MIC GAIN control is adjusted to obtain a grid current reading of 1 to 2 dB on the meter while sending a series of dots.

keying circuit and CW calibrate

The 32S-3 keying circuit provides some manual control of the keyed wave shape, fig. 3. The spotting feature (CW CAL) may be installed coincidentally. The CW CAL function switch should be front-panel mounted for ease of operation. The KEY SHAPE control, R123, may be located under the lid of the 32S-1 exciter on the bracket containing the VOX controls, or a separate bracket can be made and attached to the power-amplifier cage with self-tapping screws. Most of the other components are mounted on the terminal strips from which the 32S-1 keying circuit components will be removed. The addition of a single three-lug terminal strip (center ground) between K1 and V14 ensures that all components are securely mounted.

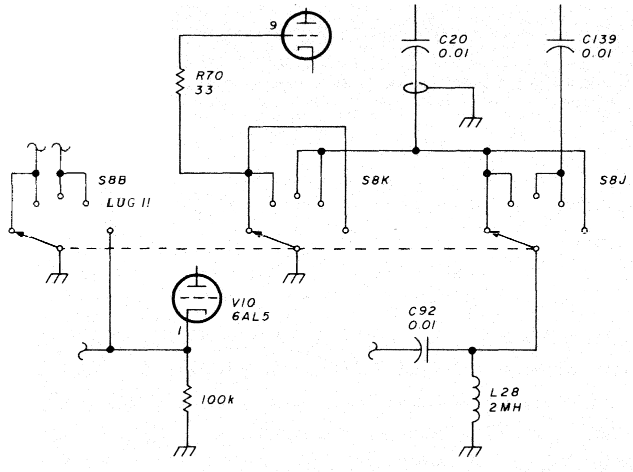


fig. 1. Schematic of the BFO generated CW showing modification of switch S8 to eliminate the weak residual signal in the 32S-1 when in the CW mode.

R70, R71, and R72 may be removed from the terminal strips at the bottom left of the chassis and R125 mounted in place of R72; R126 in place of R71; and R124 in place of R70. Remove relay Kl's lead and mount it onto the newly installed terminal strip.

Instead of using the multiple-leaf switch and 250k pot arrangement of the 32S-3 for the CW CAL function, a fixed resistor and three-pole rotary switch were used (fig. 4). The rotary (or toggle) switch has a more positive action and doesn't require constant depression to activate the desired function. A value of 68k resulted in a satisfactory over-all spotting level and this resistor was secured to the two innermost lugs of a 5-lug (center-ground) terminal strip mounted with its ground lug soldered to the ground lug of the strip behind K1 and at right angles to it. (The other lugs will be used in the regulated voltage modification.)

Mount the 3PDT switch (S13) on the front panel between the FREQUENCY CONTROL and MIC GAIN shafts. Center the holes 87 mm (3-7/16 inches) from the top of the panel. If done carefully it will appear to have been factory installed.

For ease of wiring and'installation I recommend that the FREQUENCY CONTROL switch be temporarily removed. Unsolder and tie back the green/white

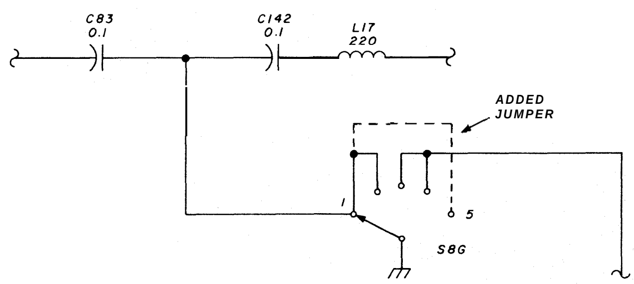


fig. 2. A jumper wire is added to switch S8-G to ground ALC voltage during CW operation.

wire at S9E-1. Wire the remaining circuit according to **fig. 4.**

In operation the transmitter must be properly tuned for CW operation for the CW CAL function to be enabled; it will not work on ssb.

The KEY SHAPE control (R123) should be adjusted to eliminate key clicks created by the rapid rise of the keyed signal. The effect of this control will be fully appreciated when the transmitted signal is monitored on an oscilloscope. The control should be adjusted to round the leading edge of the waveshape slightly.

Additional shaping of the waveform on the trailing edge may be accomplished by adding capacitance in two places: between the key line to ground and between the junction of R33/R37 and ground in the first mixer, V5. Some experimentation should provide a wave with the desired characteristics, with values of 0.025 μF (C115-A) and 0.005 μF (C81) being a good starting point in their respective positions. See **fig. 5.**

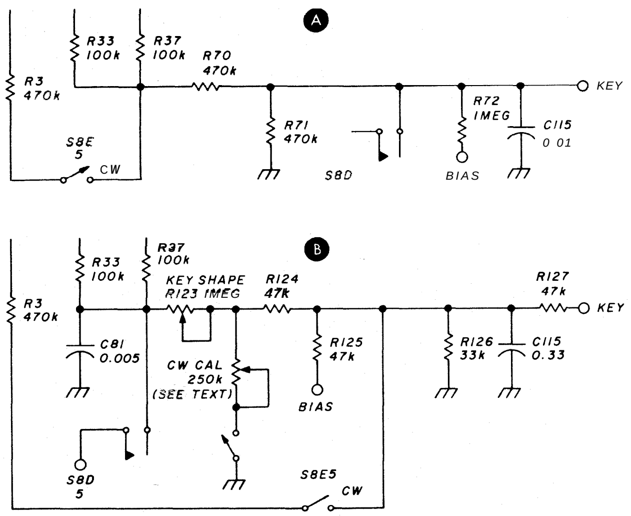


fig. 3. Keying circuits of the 32S-1 (A) and 32S-3 (B). By modifying the 32S-1 as described, you can have control of the keyed waveform within limits. The CW CAL feature is a handy addition. It won't work on ssb, however.

A difference will be noticed between on-the-air signals when using a transistor-output keyer versus a bug or relay-output keyer; the transistor provides a softer signal and you might use considerably more key-line capacitance with a bug or relay-output keyer, depending on personal preference and speed. Too much capacitance at high speeds tends to slur the code elements.

voltage regulation

In the 32S-3, the 6AL5 ALC rectifier was deleted and solid-state devices used in the ALC circuit. This freed socket V13, which was used to hold an OA2 tube to supply the regulated voltage for the oscillators. I found it simpler to use a 140-volt, 10-watt zener (1N3010A) for the regulator. They are inexpensive and eliminate the need to free V-13's socket, with the problems of rewiring the ALC circuit and finding space for more parts.

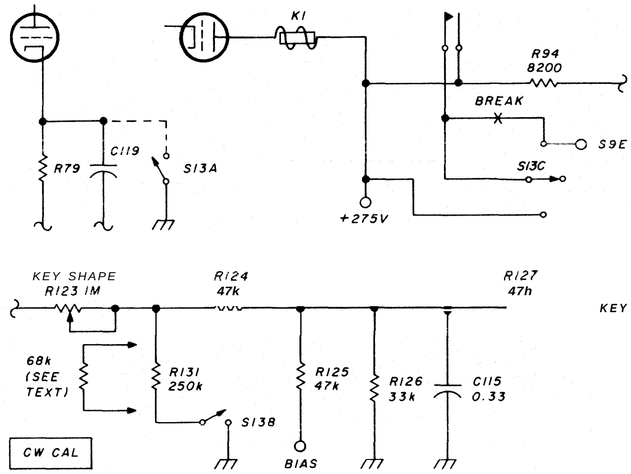


fig. 4. Modifications to the CW CAL circuit. A fixed resistor and a 3-pole rotary switch provide more positive action. It isn't necessary to hold down the switch to activate the desired function.

An advantage of the zener is its ease of mounting. Mount CR9 **(fig. 6)** on the perforated wall of the bottom side of the power-amplifier cage by enlarging one of the holes to accept the 10-32 threaded stud of CR9. Mount a dropping resistor (new R17, 5k/10W) on the terminal strip installed previously to the rear of K1. (Note: The original R17 must be removed according to the following steps.)

A convenient source of +275 volts is the terminal of C137 on the PA-cage wall; it has the 100-ohm/1/2-W resistor attached.

Modify the PTO and HFO circuit as follows. Remove the original R17 (33k/1W) and substitute

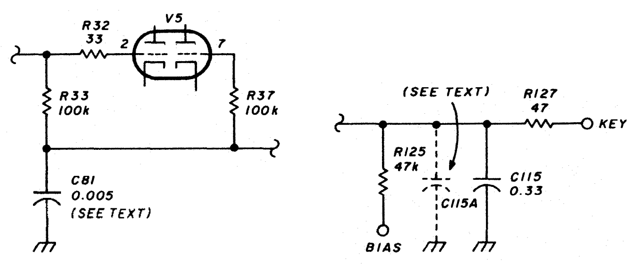


fig. 5. Improved keying wave is provided by this change. The added capacitances improve shaping of the signal trailing edge. Capacitances C115A and C81 (respectively 0.25 μF and 0.005 μF) are good starting points. Some experimentation might be needed to provide desired waveform characteristics.

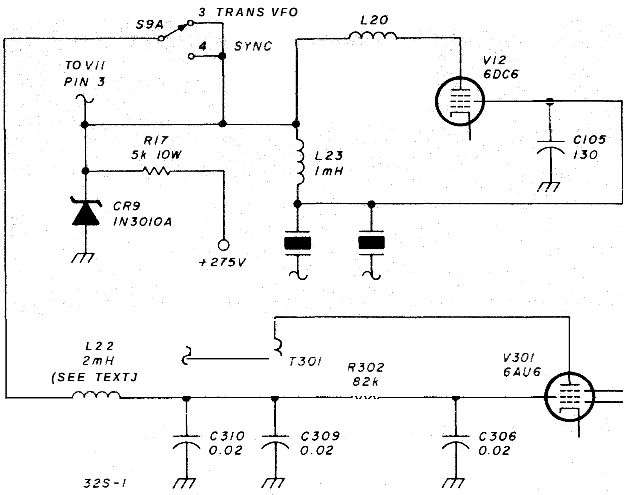


fig. 6. Regulated-voltage modifications. A 140-volt, 10-watt zener replaces the old OA2 regulator. This change is even simpler than that in the 32S-3 (see text).

L22 (2-mH). R17 is located close to C57 and the shield can. Run a wire from CR9 past the crystal board and up through the grommet to S9. At S9, locate the red/white/green/blue wire that connects to L22's B+ end. Cut the black jumper connecting the two S9 wafers (+275 volts) and attach the +140-volt line to the commoned lugs, 3 and 4, TRANS VFO and SYNC (fig. 6).

One of the two green/white wires on S9's rear wafer supplies +275 volts to the HFO, V12. Locate this wire, disconnect it at S9 rear, and move it to the +140-volt line on lugs 3 and 4. Disconnect R60 (47k) completely. Install L23 (1 mH) in its place. This completes this modification.

tone-oscillator changes

Both before and after the modifications described, an unwanted high-frequency oscillation was audible. I found it necessary to add 0.1 μF of capacitance between V11 screen and ground in parallel with C107.

Since the tone oscillator no longer supplies the on-the-air CW signal, its frequency may be altered to provide a more pleasant monitoring note. This note is purely a matter of personal preference, so some experimentation may be necessary. In my case, a 100-pF mica capacitor was paralleled with C110.

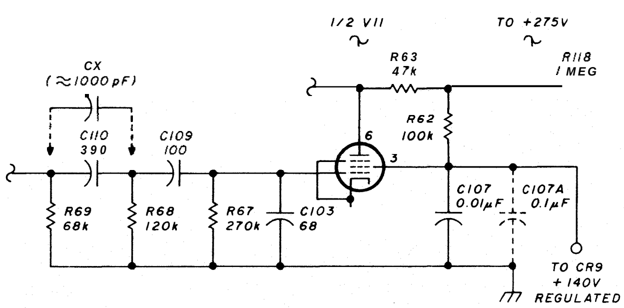


fig. 7. Tone oscillator changes. This change eliminates an unwanted high-frequency oscillation in the output. Capacitor CX in parallel with C110 provided a more pleasing monitoring note.

separate plate-current monitoring of the power amp

Unsolder R52 (1k). (Note: This value may differ from unit to unit.) R52 is attached to the copper strap joining the cathode pins of the two 6146s. Cut and remove the strap from between the tubes. Attach a length of hookup wire to each of the pins from which the strap was removed and route them toward the perforated wall of the PA cage. Mount a 4-lug terminal strip inside the enclosure toward the rear of the chassis with 4-40 (M3) hardware and wire as shown in fig. 8. Mount the DPDT switch, S15, directly beneath the meter. For ease of access the meter should be removed before drilling the mounting hole. Use a miniature toggle switch in this location, which is almost unnoticeable.

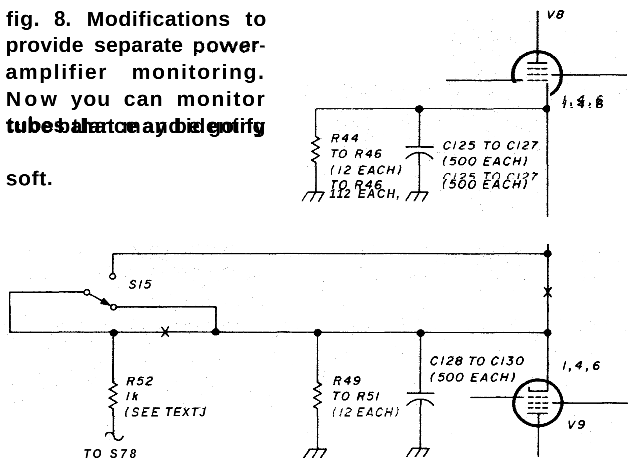


fig. 8. Modifications to provide separate power-amplifier monitoring. Now you can monitor tube balance and identify tubes that are going soft.

No interpolation of the readings is necessary since the cathode voltage/resistance ratios are unchanged. Tube balance, which is necessary in all parallel-tube amplifiers, is readily observed, and a soft tube may be easily spotted. The cathode currents of the individual tubes should track within +10 per cent to satisfy a balanced condition.

closing remarks

The incorporation of these mods into the Collins 32S-1 provided performance that rivals that of the more costly 32S-3. It's given a new lease on life to a veteran of some 18 years and has saved a couple of hundred dollars in the process!

ham radio

top-loaded delta loop antenna

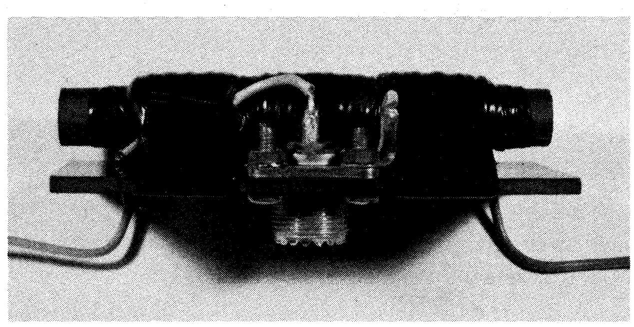
Design and construction of an efficient, low-frequency, vertically polarized antenna using wire elements

The vertically polarized full-wave loop has emerged as a popular antenna on the low-frequency bands. The most common form of this antenna is the triangular (delta) loop[1,2] with one of its vertices pointing skyward. Such an antenna can be suspended from a single point located on a tower or a tree.

The delta loop antenna is an interesting cousin of the popular inverted-V dipole. It has been around for quite a while and yet provides some pleasant surprises. For those interested in tracing its background I have provided references 1 and 2. Reference 1 is particularly informative and provides polar diagrams of the delta radiation pattern in three planes together with supporting mathematics. These references are available in most of the libraries in large cities.
Editor, W6NIF

On the 80- and 160-meter bands, height limitations can reduce the effectiveness of the delta loop. This article describes a method for reducing this problem by means of an easily implemented loading procedure. The case of a support height of 20 meters (65 feet) for an 80-meter antenna is shown in fig. 1. An interesting aspect of this comparison is that the top-loaded delta loop fig. 1B (TLDL) has more gain than a full delta loop. Experience since the end of 1976 at W1DTV has been that the antenna performs as well as an inverted V for short-range contacts and provides one to two S units better performance for DX contacts. In this article, I discuss the evolution of the TLDL and provide detailed design information for an 80-meter TLDL.

Two kinds of vertically polarized antennas are in



Homebrew matching transformer for the top-loaded delta loop antenna.

By **Frank J. Witt, W1DTV**, 20 Chatharn Road, Andover, Massachusetts 01810

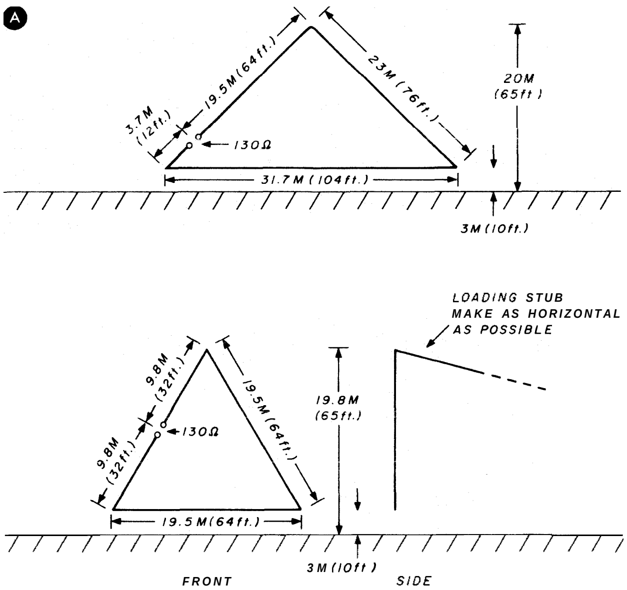


fig. 1. An 80-meter delta-loop antenna with apex at 20 meters (65 feet). Sketch A shows the classic delta loop for 3.825 MHz; B shows a top-loaded delta loop for the same resonant frequency. Loading-stub dimensions are discussed in the text.

common use on the low-frequency bands. One type is suspended above ground and fed directly; the other is erected from ground level and excitation occurs between ground or a simulated ground plane and the antenna. Both antennas would benefit from a highly conductive ground; but in the latter case, since ground resistance appears in series with the antenna at the drive point, efficiency is highly dependent on ground conditions. Therefore, the more successful monopole installations are those that use many radials. The TLDL is not fed against ground and hence ground plays only the role of a reflector. This is also true of full delta loops and sloping dipoles. Experience has shown that impressive performance may be obtained with such antennas without an elaborate system of radials.

evolution of the top-loaded delta loop

The signal at a distant point from a part of a transmitting antenna is proportional to the current in that part of the antenna. For a half-wave dipole, for instance, maximum radiation is received from the center of the dipole, where the current is greatest. The radiated contribution from the ends of the antenna is negligible.

The TLDL concept resulted from a recognition of the fact that for a conventional, vertically polarized delta loop, much of the antenna where high currents exist is horizontal and near ground. The objective of the TLDL design is to get these parts of the antenna away from ground and at least partly vertically oriented to increase antenna gain. **Fig. 2A** shows a typical vertically polarized conventional delta loop designed for 3.825 MHz. Actually, this antenna can only be said to be mostly vertically polarized because of the position of the feed point. True vertical polarization (in a direction perpendicular to the plane of the loop, *i.e.*, the direction of maximum gain) is obtained when the feed point is one-quarter wavelength away from the peak of the triangle as shown in **fig. 2B**. You can see that the polarization is vertical by noting the current flow: the vertical components from the currents in the two upper sides of the triangle add, while the horizontal components cancel.

The objective of the loading is to "lift" the current nodes higher in the vertical space available for the antenna and to make the vertically radiating sides of the antenna more vertical. Both actions will increase

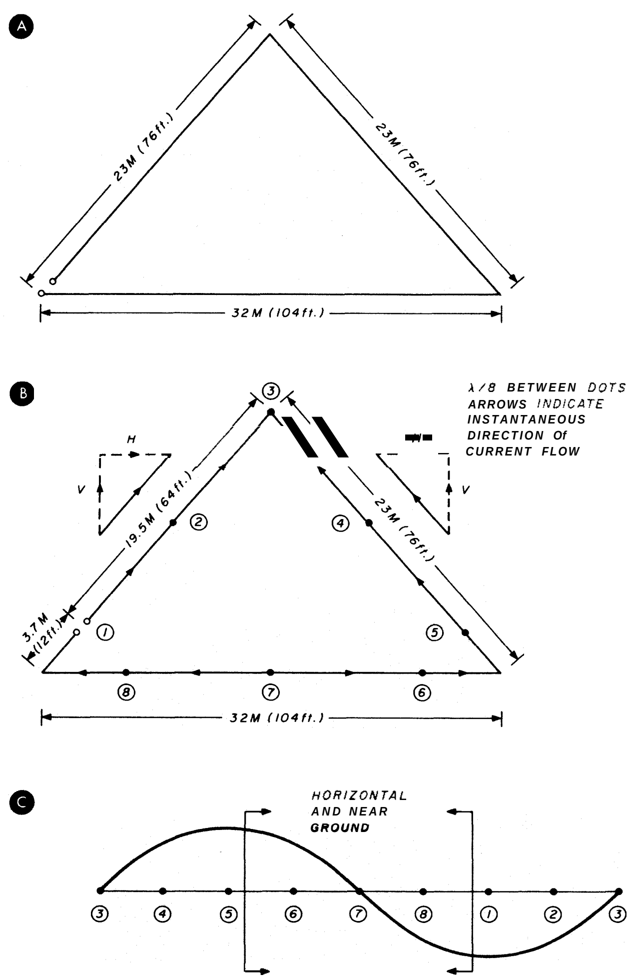


fig. 2. Physical dimensions of a typical corner-fed delta loop antenna (A). True vertical polarization occurs when the feedpoint is one-quarter wavelength from the apex (B). Sketch C shows current distribution.

the low-angle gain for vertically polarized signals. The derivation of the TLDL from a conventional delta loop is shown in **fig.** 3.

The feedpoint resistance for both a conventional delta loop and a TLDL has been measured at W1DTV to be i30 ohms. From this information and from the geometry of the two antennas (and if one assumes sinusoidal current distribution), the TLDL has a gain of 2.3 dB over a conventional delta loop. The dimensions of **fig. 1** have been assumed for this calculation. See reference 3 and **fig. 4** for an explanation of the methods used to arrive at this result.

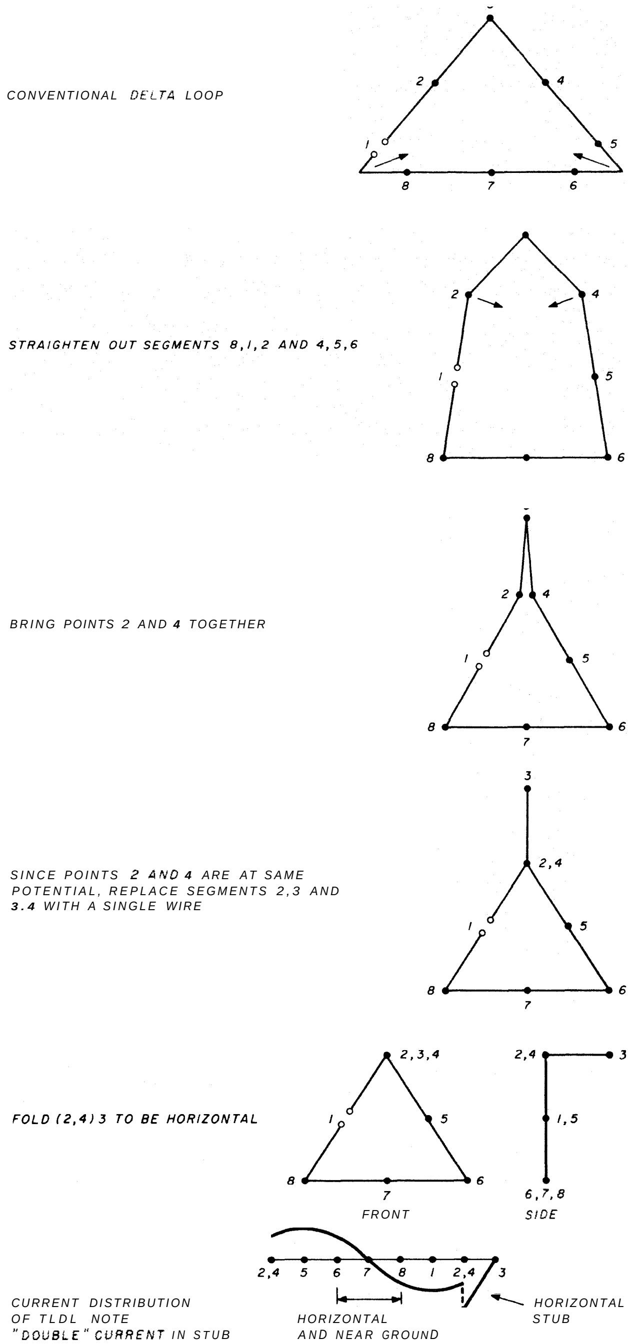


fig. 3. How the top-loaded delta loop (TLDL) is derived from the conventional delta loop. The top loading makes the current nodes higher with respect to ground and increases the vertical polarization from the antenna sides. This improves low-angle radiation.

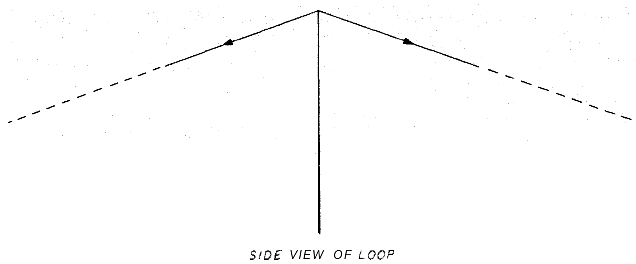


fig. 4. Method of adding the stubs on both sides of the delta loop to reduce horizontally polarized radiation.

The TLDL is truly vertically polarized in a direction perpendicular to the plane of the loop. **It** is mostly vertically polarized in other directions and exhibits an almost omnidirectional pattern.

loading stub

The loading (or matching) stub is shown in **fig.** 3 to be horizontal, but this is rarely possible. At W1DTV it runs to the farthest point on the property and makes about a 60-degree angle with the plane of the loop. The stub should be $\lambda/8$ or 9.8 meters (32 feet) at 3.825 MHz. However, it was necessary to lengthen it to 13 meters (43 feet) for resonance at that frequency. The probable reason for this is that the stub is severely folded back toward the loop; the consequent detuning is overcome by lengthening the stub. This effect is observed in inverted V antennas, where the length must be made longer than would be necessary for a straight dipole.

The stub could be added on both sides of the loop as shown in **fig. 4.** This would virtually eliminate the effect of the stub on the radiated pattern. This method hasn't been tried, and the practical effects are unknown.

The stub can be shortened considerably by means of a loading coil installed in series with the stub at the point where the stub is connected to the triangle apex. See **fig. 5.** The 13-meter (43 foot) stub was reduced to 4.9 meters (16 feet) by the use of a 32-pH loading coil. The loading coil reduces radiation from the stub, but it results in a reduction in antenna bandwidth. The loading coil is a B&W 3029/3905-1,* which is 63.5 mm (2 1/2 inches) diameter by 254 mm

*Barker and Williamson, Inc., Canal Street and Beaver Dam Road, Bristol, Pennsylvania 19007.

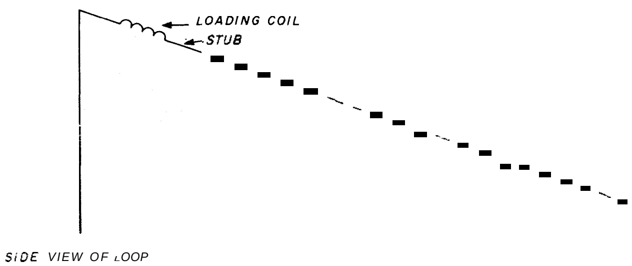


fig. 5. Using a loading coil to reduce stub radiation.

(10 inches) long (6 turns per 25 mm). This coil with the 4.9-meter (16-foot) stub allows the TLDL to resonate anywhere in the 80-meter band by changing the tap position.

matching methods

A common method for feeding delta-loop antennas is to use a quarter wavelength of 75-ohm transmission line between a 50-ohm transmission line and the feedpoint. For a feedpoint resistance of 130 ohms, the vswr at resonance would be

$$\frac{130}{75^2/50} = 1.16:1 \qquad (1)$$

which is quite acceptable. Since the conventional delta loop and the TLDL are essentially balanced antennas, it's desirable to use a 1:1 balun at the antenna to prevent antenna currents on the coax feedline.

At W1DTV, a transformer (shown in **fig. 6** and the photo) accomplishes both impedance matching and the unbalanced-to-balanced transformation; it handles the legal power limit quite satisfactorily. The transformer has been evaluated only on 80 meters, but the design could be trimmed to work over several bands. See reference 4 for details on optimizing such designs.

voltage standing-wave ratio

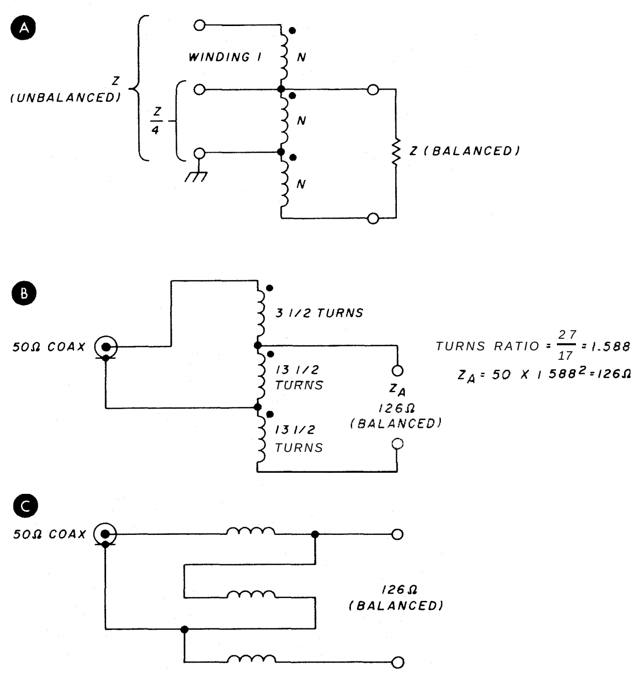
The vswr using the transformer of **fig. 6** and 50-ohm coax is shown in **fig. 7** for the conventional delta loop, the TLDL using a wire stub only, and the TLDL with a wire stub and loading coil. Note that an excellent midband match is obtained for all three cases, but the bandwidth depends on the configuration. The bandwidth of the worst TLDL case (using the loading coil) is substantially better (6.6 times wider) than that of a loaded 20-meter (66-foot), 80-meter sloping dipole,[5] which would require about the same mounting height. The vswr plot of the latter is also shown in **fig. 7** for comparison.

Note from **fig. 7** that the vswr of the TLDL with a loading coil is not unity. The reason is that the inductive loading modifies the current distribution at the top of the antenna, and the feedpoint resistance is changed. For the purist, a more optimum transformer for this case would be one where the 3-1/2-turn winding of **fig. 6** is reduced to 2-1/2 turns. In all cases, the installation involved a steel tower with the antennas supported 1.2 meters (4 feet) from the tower on a boom at the 19.8-meter (65-foot) level. All guy wires were broken with insulators to avoid resonance, and no guying was used above the 10.7-meter (35-foot) level.

concluding remarks

The TLDL antenna performs as well as other similar antennas requiring higher points of support. The design is based on the positioning of the high-current parts of the antenna so that they will provide a primarily vertically polarized radiated signal. The TLDL has substantially more bandwidth than its nearest low-height competitor, the λ/4 sloping dipole (loaded). Calculations indicate that the antenna should be a good performer, and on-the-air experience has substantiated these results.

A point of caution — if you try this antenna, or any new antenna, take steps to convince yourself that



WIRE 1 mm (AWG 18) HOOKUP WIRE

CORE AMIDON FERRITE ROD, 12 mm DIAMETER, 100 mm LONG, μ = 125 (AMIDON ASSOCIATES, 12033 OTSEGO STREET, NO HOLLYWOOD, CA, 91607)

CONSTRUCTION 13 1/2 TRIFILAR TURNS WITH 10 TURNS REMOVED FROM ONE WINDING WRAP WITH VINYL ELECTRICAL TAPE AND TAPE TO THE FEEDPOINT INSULATOR NO OTHER PROTECTION NECESSARY

fig. 6. Construction of the matching transformer for the top-loaded delta loop antenna. Sketch A shows the basic principle. Any impedance ratio between 1:1 and 1:4 may be obtained by tapping winding 1. The basic configuration is shown in B. Sketch C shows the schematic and winding logic.

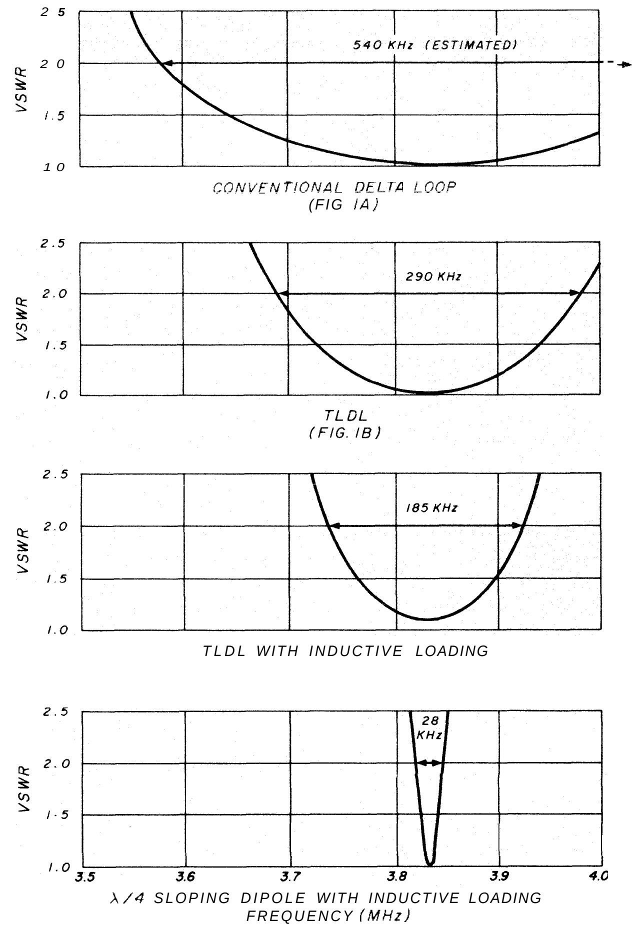


fig. 7. Voltage standing wave ratio and bandwidth of the conventional delta loop, the TLDL, the TLDL with inductive loading, and a quarter-wavelength sloping dipole antenna with inductive loading. Resonant frequency is 3.825 MHz in this model.

other nearby antennas are not significantly influencing its behavior. A considerable amount of interaction between a TLDL, an inverted V, and a sloping dipole, all supported by the same tower, has been observed. The data in this article were taken with the inverted V and sloping dipole removed from the tower.

references

1. H.E. Green, "On the Delta Aerial," IEEE Transactions on Antennas and Propagation, January **1963,** pages **98-100.**

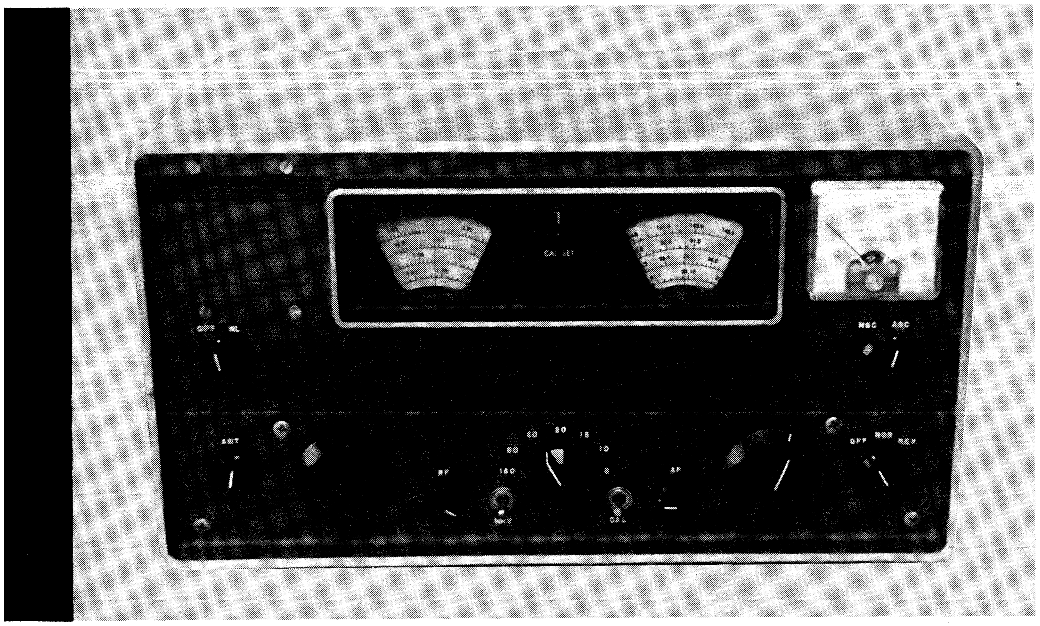
2. J.S. Hall, "Nonresonant Sloping Vee Aerial," Wireless Engineer, vol. XXX, September **1953,** pages **223-226**.

3. F. Witt, "Simplified Antenna Gain Calculations," ham radio, May **1978,** page **78.**

4. J. Sevick, "Broadband Matching Transformers Can Handle Many Kilowatts," Electronics, November 25, **1976.**

5. G. Hall, "Off-Center Loaded Dipole Antennas," QST, September **1974,** pages **28-34, 58**. (The antenna shown in fig. 5 of the referenced article is the one referred to as the λ/4 dipole. It's a popular radiator for sloping-dipole installations.)

ham radio



updating the vacuum-tube receiver

Rejuvenate that old vacuum-tube receiver by replacing the tubes with equivalent transistor stages

All of the communications receivers made in the 1940s and 1950s, and most of those made in the 1960s, were vacuum-tube designs. Receivers manufactured during those decades were intended for the amateur modes then popular — a-m and CW — and they usually performed quite well on a-m. Inclusion of the beat frequency oscillator made possible reception of ssb and CW, but, by today's standards, much was left to be desired. These older receivers can often be picked up at flea markets or swap meets for a small fraction of their original cost and, properly modernized, can be made to equal or even surpass the performance of new models selling at many times the price.

Probably the two worst shortcomings of older tube-type receivers are their lack of selectivity and their warm-up drift. The drift results mainly from temperature changes brought about by the large amount of heat generated by the tubes. Although much can be done with tube receivers to minimize temperature drift,[1] the best solution is to abandon tubes altogether and substitute transistors. Modern field-effect transistors have characteristics so similar to those of tubes that often a direct one-for-one replacement is possible with only minor circuit and supply voltage changes.

Any of several possible routes can be taken when transistorizing a tube receiver. Probably the most conservative (some would say cowardly) approach is to proceed one stage at a time; that is, substitute a transistor circuit for a vacuum tube stage and perfect its performance before moving on to the next stage. This strategy makes debugging easy, but, because of different impedance and signal levels, tubes and transistors do not always interface easily. Sometimes

By Fred Brown, W6HPH, 1169 Los Corderos, Lake San Marcos, California 92069

a transistor stage will have to be modified when a following or preceding stage is changed over from tube to transistor.

Next to frequency drift, the greatest shortcoming of these old a-m receivers is usually their poor skirt selectivity. The old single-pole crystal filters and *Q*-multipliers gave good selectivity at about 6 dB down, but the skirts were very wide at 30 to 60 dB down. The rectangular passband so necessary for today's ssb communications calls for a mechanical or multi-pole crystal filter.

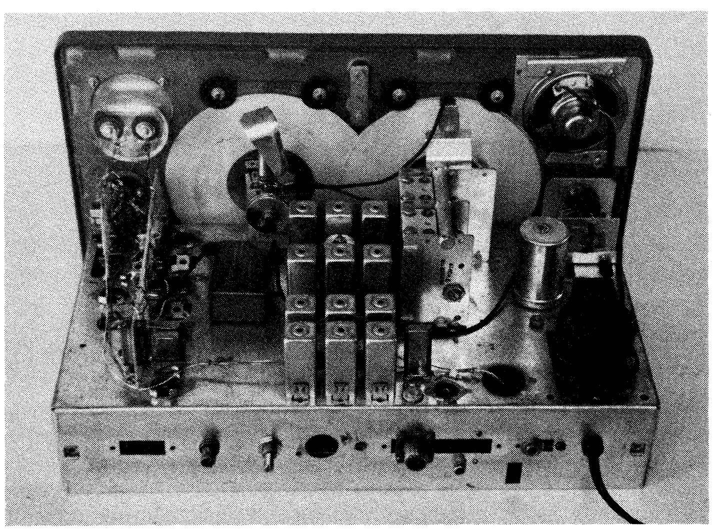
If your receiver is a single-conversion general-coverage type, you will probably have to stick with the original intermediate frequency. This means a mechanical filter if the i-f is below 1 MHz. A ham-band-only receiver, however, can be converted to practically any intermediate frequency so long as the i-f does not lie inside or near an amateur band. (It's very hard to make a superhet that will tune through its own intermediate frequency.)

Generally speaking, a high intermediate frequency is to be preferred, because it improves image rejection. Furthermore, if the local oscillator operates on the low side of the signal, a high i-f means a lower LO frequency, and this contributes to frequency stability. Today, the availability of high-frequency multi-pole crystal filters makes possible single conversion designs with high intermediate frequencies, thereby avoiding the spurious response, birdie, and cross-mod problems of dual-conversion designs.

modernizing the HQ110A

To start this project, I tore out all the old circuits and started with a clean slate, leaving intact only the front-end tuned circuits and bandswitch. Most of the tube sockets were left, as they make handy tie-points for the new wiring. There is something to be said for building from the ground up, but most of the work with sheet-metal and the mechanical drudgery is avoided if you rebuild a commercial receiver. Also, it is hard for the average amateur to duplicate the professional appearance and calibrated dials of a manufactured receiver.

Rear view of the HQ110A. The i-f amplifier, detector, first audio, and S-meter circuits are built on the circuit board that runs along the left side of the chassis.



The HQIIOA is a good example of how an intermediate frequency can be radically changed. Originally, the receiver was dual conversion on all bands above 80 meters, with a first i-f of 3035 kHz and a second i-f of 455 kHz. I happened to have a McCoy Golden Guardian 9-MHz ssb filter. I wanted to move the i-f to this frequency and make the receiver single-conversion on all bands. This took some doing; it meant all the original i-f transformers had to be scrapped, and all the local oscillator coils rewound. The result was worth it. The transistorized receiver has much better selectivity, frequency stability, and image rejection than the original. And the existing dial calibration holds on every band, even better than in the tube version.

Power supply and audio. A good place to start your conversion to solid state is in the power supply; you won't be able to try out any transistor circuits without low-voltage dc. Although the old power transformer will be less than ideal for a transistor power supply, it is possible to obtain low voltages from the 5.0- and 6.3- volt filament windings. For instance, a full-wave rectifier on the 6.3-volt winding will provide 9 volts dc, and a full-wave doubler about 18 volts dc. In this receiver, however, I replaced the old power transformer with a more appropriate one having a 16-volt center-tapped secondary.

Both positive and negative supplies were needed, since dual-gate MOSFETs, like vacuum tubes, normally require a negative AGC voltage. Although it is possible to use FETs with positive-only voltages, the AGC system is simplified if a negative supply is available.

The positive supply (see **fig. 1**) is filtered by a conventional L-C filter which uses an 88-mH toroid for a filter choke. The resulting unregulated 19 volts is used for the audio output stage and diode bias. A conventional emitter-follower regulator, Q1 and CR5, is used to regulate the positive supply to +12 volts dc for the remainder of the receiver.

Most designers would use a class B integrated circuit for the audio output, but I prefer the class A arrangement shown in **fig. 1.** This circuit has proven reliable in many different applications. It has excellent dc stability due to the negative-feedback biasing arrangement, and low distortion because of the negative-audio feedback around the output transformer.

Detector, i-f, and AGC systems. The 9-MHz i-f signal from the McCoy filter is applied to the first i-f stage, Q5. This common emitter stage is R-C coupled to the dual-gate MOSFET second i-f stage. R-C coupling is used for simplicity and to avoid the need for neutralization. The added gain that could be obtained from transformer coupling was no: needed.

The i-f transformers, T3, T4, and T5, are ordinary fm transistor radio transformers which are shunted with sufficient capacity to resonate at 9 rather than 10.7 MHz. The different resonating capacitors are due to the use of different brands of transformers.

The third i-f stage is an emitter-coupled pair, Q7 and Q8. The emitter of Q7 also drives the AGC amplifier, Q9 and 010. It was not possible to derive the AGC signal from the collector circuit of Q8 because the BFO signal is too strong at this point and would swamp the AGC. An audio-derived AGC system would have been easier, but in this receiver I wanted AGC that would work on a-m carriers.

The AGC signal is rectified by CR9, with the resulting dc applied to Q11. An emitter follower is used because its low output impedance allows C1 to quickly charge through CR10. In any single-sideband AGC system it is important to have a short attack time to avoid "popping" at the beginning of a word. One inherent limitation on attack time is the delay caused by the crystal filter. Envelope delay time in a bandpass filter is determined by the skirt steepness ratio: the steeper the skirts, the longer the delay. The McCoy filter, having exceptionally steep skirts, has an inordinately long delay time, and for this reason it was not possible to use as much AGC feedback around the filter as would have been desirable. The AGC control voltage to the r-f and mixer stages (AGC 2) is attenuated by the voltage divider, R4-R5. In addition, AGC 2 also controls the S-meter driver through the S9 adjustment pot R4.

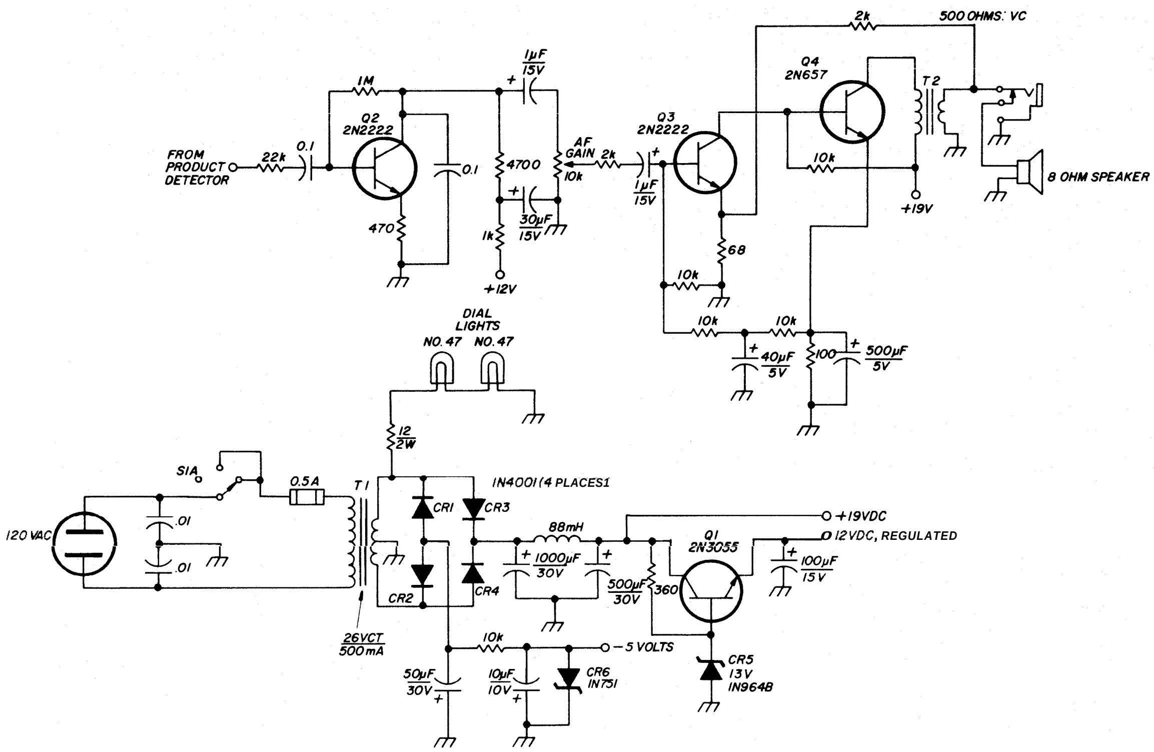


fig. 1. Schematic diagram of the revised power supply and audio circuit. Q1 and Q4 should be mounted on heatsinks. The audio stage will provide 100 mW of undistorted audio output. In addition, the use of Class-A transistors *vs.* an IC may prevent destruction of the devices if the output is shorted.

The product detector, CR7-CR8, requires push-pull BFO drive which is provided by the center-tapped secondary of T6. The BFO is switched by means of diode switches for upper or lower sideband.

Local oscillator. In the interest of best frequency stability, the local oscillator operates on the low frequency side of the signal on the 20-, 15-, 10-, and 6-meter bands. On 160, 80, and 40 meters, the oscillator is 9 MHz higher than the signal frequency. This arrangement automatically takes care of sideband switching; upper sideband is received on 20 meters

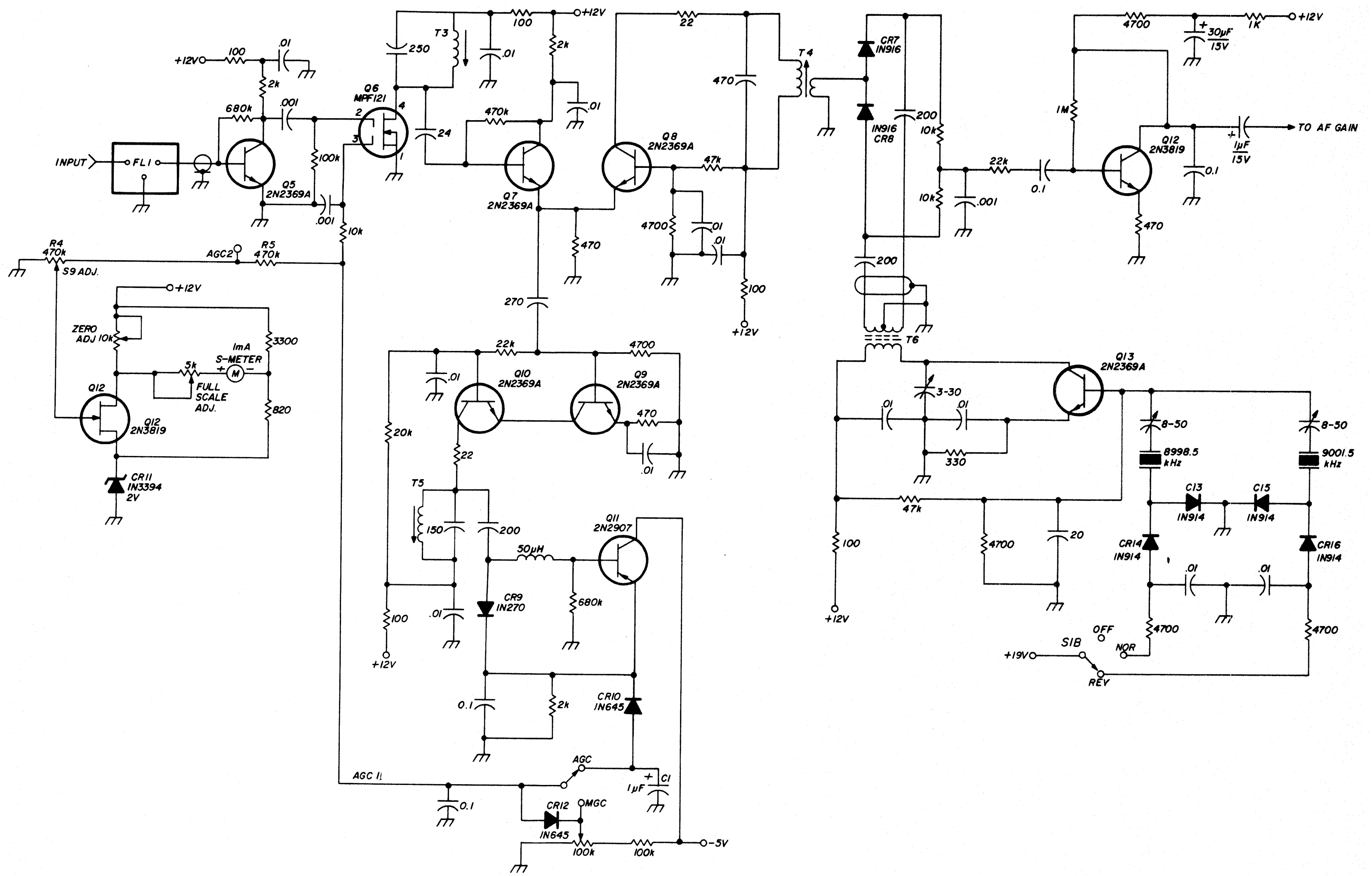
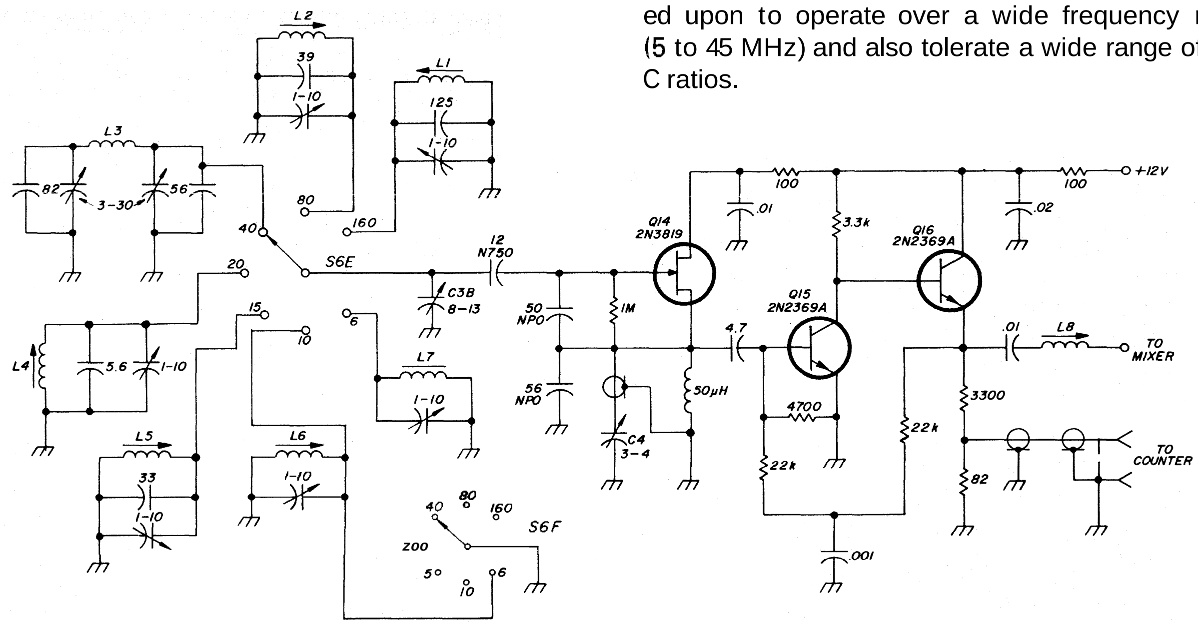


fig. 2. Diagram of the i-f amplifier, detector, BFO. AGC. and S-meter circuits. **FL1** is a McCoy Golden Guardian 9-MHz ssb filter. **T3, T4**, and T5 are 10.7 MHz i-f transformers from fm-style radios. In the case of **T3 and T5,** where only one winding is shown. the primary and secondary have been connected in series so as to provide maximum inductance. T6 is wound on a T-50-2 core with 33 turns of no. 30 **(0.25mm)** AWG on the primary and 8 turns of no. 30 **(0.25mm)** AWG on each side of the **center** tap for **the secondary.** Normally. R4 is used to set the S-meter for an **S9** indication with 100 μ**V** applied to the receiver.

and the higher bands and lower sideband on 40 meters and lower, as is the accepted standard. The BFO switch can therefore be labeled NORMAL and REVERSE without any reference to the bandswitch position.

Table 1 gives the required oscillator tuning range of each amateur band. On three bands – 20, 15, and 10 meters – the HQ110A dial calibration is wider than the amateur band limits. **Table 1** also gives the LO tuning range as a percentage of the lowest oscillator frequency on that band. Notice that the percentage tuning range varies from a maximum of 10.53 per cent on 10 meters to a minimum of 1.85 per cent on 160 meters. The oscillator percentage tuning range is determined by the amount of fixed capacity which shunts the tuning capacitor; the smaller this capacity, the greater the range.

To make the dial calibration come out right, both the inductance and shunt capacitance of the LO tank must be correctly chosen. In the HQ110A, the oscillator tuning capacitor (only one of two sections is used) has a minimum capacitance of 8 pF and a maximum of 13 pF, or a range of only 5 pF. This capacitor must be shunted with a total of about 134 pF to provide the 1.85 per cent tuning range for the 160-meter band (10.8 to 11.0 MHz). On ten meters, however, the tuning range expands to 10.53 per cent, and the total shunting capacitance must be reduced to a mere 23 pF. This 23 pF includes the 8 pF minimum of the tuning capacitor, and a few pF of stray and coil-distributed capacitance, and a few pF for the trimmer. When all this is added up, the shunting capacitance left for the remainder of the circuit will have to be limited to a maximum of about 12 pF in order not to exceed the 23-pF total. This is one restriction on the oscillator circuit. In addition, the oscillator is called upon to operate over a wide frequency range (5 to 45 MHz) and also tolerate a wide range of L to C ratios.



L1 14 turns no. 24 (0.5mm)	L4 60 turns no. 32 (0.2mm)	L7 7.5 turns no. 24 (0.5mm);
L2 17 turns no. 28 (0.3mm)	L5 17 turns no. 28 (0.3mm)	L8 17 turns no. 24 (0.5mm);
L3 20 turns no. 22 (0.6mm);	L6 16 turns no. 28 (0.3mm)	

fig. 3. In the local oscillator, C3B is one section of the main tuning capacitor. Inductors L1, L2, L4, L5, and L6 are wound on the original 6.5 mm (¼") slug-tuned form; L3 is wound on a T-68-6 toroid; L7 and L8 are wound on 6.5 mm (¼") slug-tuned forms. Most of the components can be mounted on a small circuit board.

Many different oscillator circuits were tried in search for one that would meet these stringent requirements and would also be stable in frequency. The circuit shown in **fig.** 3 worked best. It is an fet version of the Seiler oscillator,[2] followed by a two-stage buffer amplifier. **Table 1** shows the excellent frequency stability of this oscillator for supply voltage changes of one volt (from 11.0 to 12.0 volts). Since the 12-volt supply is fairly well regulated, it's possible to vary the line voltage from 80 to 130 volts ac with no noticeable change in beat note. Try to do that with your tube receiver!

Theoretically, this circuit will oscillate with any L to C ratio in the tuned circuit. The only limitation on L-to-C ratio is Q; the lower the ratio, the higher the tank circuit Q must be to sustain oscillation. The lowest L-to-C ratios occur on those bands with the smallest percentage tuning range, 160 and 40 meters. On 40 meters, it was not possible to get enough

Q with the original 6.5-mm (0.25-inch) slug-tuned coil form. Therefore, a toroidal coil was used for this band. Since toroids are not adjustable, the circuit arrangement shown in the schematic was adopted. The capacitor in series with the coil (C2) effectively trims the inductance.

On 160 meters, it was possible to wind the coil on the original coil form and still get enough Q, but just barely. All the other coils, except 6 meters, are wound on the original forms and mounted in the original shield cans.

Notice in **fig.** 3 that the 10-meter coil is shorted by S6F when the bandswitch is in the 6-meter position. This is because the ten-meter coil happened to resonate near 42 MHz, which, unfortunately, was within the oscillator tuning range on the 6-meter band and caused a back "suck-out" as the oscillator was tuned past this resonant frequency. Also, unfortunately, there was no unused contact on the bandswitch in the 6-meter position, so one had to be added to do this shorting job. This, however, was the only modification that was necessary to the original bandswitch.

The untuned buffer amplifier is made up of a common-emitter stage (Q15), for amplification, directly coupled to an emitter follower (Q16) for low output impedance. As with any amplifier, there is an upper-frequency limit where gain begins to fall off. In this case, output is down about 3 dB at 19 MHz (10 meters), but on 41 MHz (6 meters), it is down about 12 dB. For this reason, a compensating coil (L8) is placed in series with the output. This inductance forms an L network in conjunction with the mixer input capacitance, thereby boosting the 41-MHz LO voltage at the mixer gate. On the lower bands it has little effect.

The 82-ohm resistor in series with the emitter resistor of Q16 delivers a local oscillator signal to a phone jack located on the rear apron of the chassis. By plugging a frequency counter into this jack, you can have a poor-man's digital readout. Of course, it is necessary to add or subtract 9 MHz from the counter reading, but in practice it is easier to simply ignore the digits to the left of the decimal point and mentally substitute the appropriate digits for the band being received. For example, the receiver is on forty meters and the counter reads 16.238. Obviously, this means 7.238 MHz. Remember also, that the counter reading corresponds to the center of the passband, which is typically 1500 Hz higher or lower than the carrier frequency, depending on which sideband is being received.

The tuning rate of the HQ110A was a bit fast for ssb, especially on 10 and 6 meters. For this reason, a clarifier, or ultra-fine tuning control (C4 in **fig. 3**) was added to the LO circuit. This variable capacitor is mounted in the hole formerly occupied by the function switch and is turned by the large, function-switch knob, which is ideal for the purpose.

A very small capacitance range is needed for the clarifier capacitor. It could have been made from a small variable by removing all but two plates, but in my case it was fabricated from the remains of an old wire-wound pot. The resistance element and wiper arm were removed and a semi-circular plate was soldered to the rotor shaft. Another semi-circular plate was fastened to the Bakelite case to form the stator. The resulting capacitor has a range of only 1 pF and produced the frequency ranges given in the fifth column of **Table 1.**

table 1. Frequency ranges for the new local oscillator in the rebuilt HQ110A.

band MHz	oscillator range MHz	oscillator percentage range	oscillator stability Hz/volt	clarifier range Hz
1.8 - 2.0	10.8 - 11.0	1.852	169	420
3.5 - 4.0	12.5 - 13.0	4.0	225	870
7.0 - 7.3	16.0 - 16.3	1.875	165	510
14.0 - 14.4	5.0 - 5.4	8.0	340	1100
21.0 - 21.6	12.0 - 12.6	5.0	257	1000
28.0 - 30.0	19.0 - 21.0	10.53	400	3200
50.0 - 54.0	41.0 - 45.0	9.76	280	7300

rf and mixer stages. Because dual-gate MOSFETs make such ideal replacements for pentode vacuum tubes, no radical changes were necessary in the receiver rf and mixer stages shown in **fig. 4.** None of the antenna or rf stage coils were changed, with the exception of the tap on the six-meter antenna coil: it was moved up to three turns above ground. Both rf and mixer stages, 017 and 018, are RCA40673 dual-gate MOSFETs. For simplicity, coils for only the 6-, 10-, 15-, and 20-meter bands are shown in **fig. 4.** Experimentally, it was found that receiver sensitivity on these four bands could be optimized for a 50-ohm antenna by placing a proper size capacitor in series with the antenna input.

When the receiver is tuned to 28.0 MHz, the image frequency happens to be 10.0 MHz, a fact that is exploited with a minimal increase in complexity to provide 10-MHz reception of WWV. For WWV, S5 in **fig. 4** shunts the 10-meter rf stage coil with trimmer C5. This capacitor resonates the 28-MHz coil down to 10 MHz, with the WWV signal from the antenna coupled into the mixer through the series-tuned circuit, L9-C6. The rf stage is not used in this mode; WWV is normally so strong that an rf stage is not really needed.

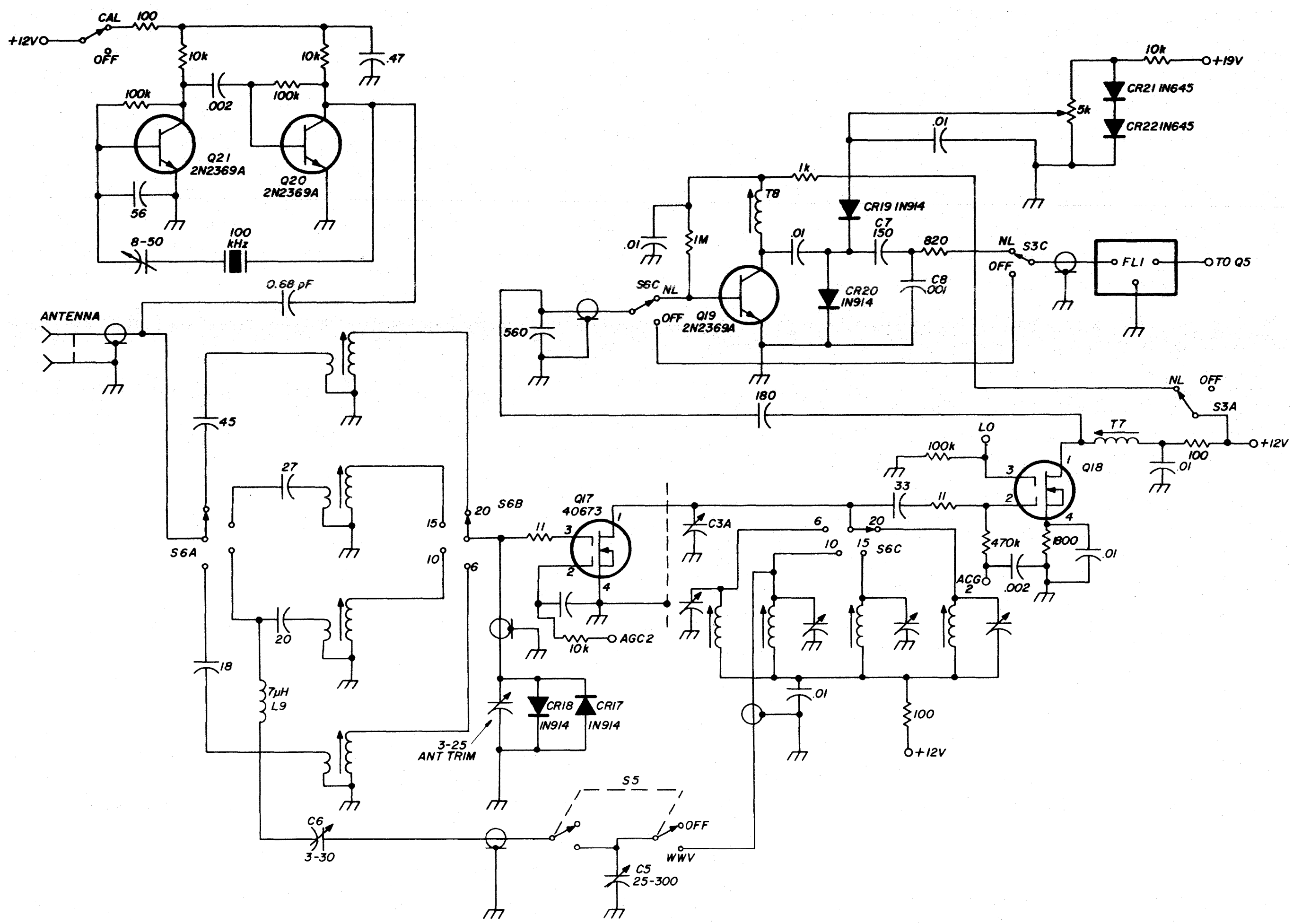


fig. 4. Schematic diagram of the rf, mixer, noise limiter, and crystal calibrator circuits. The antenna and rf stage coils and trimmers are unchanged from the original receiver. For simplicity, only four bands are shown. L9 is 7 μH, made by close winding 30 turns of no. 24 (0.5mm) AWG wire on a 4-mm (5/32-inch) diameter ferrite rod.

Noise limiter and crystal calibrator. I wanted some kind of first-line defense against impulse noise without resorting to all the complexity of a noise blanker. Thus, the simple noise limiter in **fig. 4** was included. In order to be effective, the limiter should clip the noise pulses before they become stretched out by the crystal filter.

The limiter was placed between the mixer and sideband filter and has the option of being switched in or out. The common-emitter stage, 019, amplifies the noise pulses up to a level where they can be clipped by the shunt limiter, CR19-CR20. These silicon diodes are biased by R6 to a point where they are just beginning to conduct. After being clipped, the signal is brought back down to its original level by means of the capacitive voltage divider, C7-C8.

I make no claim that this limiter can compete with a good noise blanker, but it can make the difference between copy and no-copy. Only high-amplitude noise pulses are clipped. It takes about 50 microvolts at the antenna before clipping begins to occur. This means weak noise pulses are unaffected, but the weak pulses are not normally a problem because they are reduced in amplitude after being stretched out by the i-f filter. For best results, the AGC should be turned off when using the noise limiter.

Of course, the greatest shortcoming of this type of limiter is that it is highly susceptible to cross-mod from strong in-band signals. If you have both high noise and strong signals at the same time, you must choose between cross-mod and impulse noise, depending on which is worse.

The 100-kHz crystal calibrator, Q19 and 020, uses the original 100-kHz crystal and trimmer capacitor. The circuit is patterned after one in the Atlas 210X; it provides plenty of harmonic strength right up through 54 MHz.

references

1. Fred Brown, W6HPH, "Souping Up the Old Receiver," CQ, January, 1970, page 15.
2. E. O. Seiler, W8PK, "A Low-C Electron-Coupled Oscillator," QST, November, 1941, page 26.

ham radio

double-stub tuner

for 1296 MHz

Feedline and matching problems at 1296 MHz are easily overcome by using this simple but effective double-stub tuner

Shortly after the military **APX-6** became available on the surplus market in the 1950s, a number of magazine articles were published which described conversion of the unit to the 1215-1300 MHz amateur band. For a time there was considerable interest in this band; here in the San Fernando Valley there were about seven stations active for a year or more. Eventually interest declined, until there were only two of us left – K6BV and myself. We had 85 MHz of amateur frequency spectrum with almost no takers. That's too bad, because this band is ideal for repeaters, wide-band television, satellite communications, and amateur experimentation.

Why is the use of this band so meager? Probably the main reason is that there is no equipment available for the band. Also, it requires the application of techniques quite different from those used on the lower frequencies; of necessity, striplines and cavities must be used. These take some getting used to by newcomers to the amateur bands above 1000 MHz.

Another aspect that has not received too much attention is transmission lines. Losses in coaxial cable are tremendous at this frequency. Standing waves are a disaster if they appear on the line. The best RG-8/U coax has a loss of 8 dB per 30 meters (100 feet) at 1296 MHz. Any reflection at the antenna is power lost in all but the shortest lines, and putting a radiator at the top of a tower does not promote a short feedline. On the low-frequency amateur Bands a reasonable reflection of power at the antenna does not represent any appreciable loss in radiated power; if nothing gets hot, and the transmitter loads, the power has to radiate. Reflected power makes a round trip at low frequencies. At 1296 MHz: the reflected power suffers the 8-dB loss on the trip back to the transmitter and again on the second trip to the antenna. The net result is that reflected power is power lost, and power is hard to come by at 1296 MHz.

Measurement of power and standing waves on the feedline was a serious problem at my station for several years. A field-strength meter is easy to make for 1296 MHz, and setting it up in front of the antenna gives a good indication of relative power levels, but it usually turned out that a slotted line was still showing a considerable standing wave on the feedline, even after everything was tweaked for hours on end. Just taking the slotted line out of the line was apt to be a disaster if only because of the change in length of the line. In addition to this problem, the field-strength meter didn't give a true reading of the actual power being radiated. It was strictly a relative measurement, and I could not tell whether the transmitter was running at 5 per cent efficiency or the hoped for 30 to 40 per cent.

Uncertainty over the power output of the transmitter and the unhappy situation with standing waves on the feedline led to several not-so-cheap purchases and also a rather close-tolerance construction project. The first purchase was a Bird model 43 wattmeter with a plug-in covering 1.1 to 1.8 GHz. The second purchase was a Sierra Model 160B-300 dummy load (I couldn't build a suitable dummy load and gave up after a number of attempts). The way to rationalize these purchases is to say that the test gear will not wear out and with much cheaper plug-in elements the Bird 43 wattmeter will work on other amateur bands.

With the wattmeter and the dummy load I could read transmitter power output and pick up reflected

By George Hatherell, K6LK, 10160 Maude Avenue, Sunland, California 91040

power when the **feedline** was moved from the dummy load to the antenna radiator. Things began to fall into place. It became possible to trim the driving dipole for a reasonable match to the line. Single-helix antennas were stubborn, and responded only half-heartedly to matching devices; a twin-helix antenna was much better; a quad helix never seemed to work just right, regardless of the matching scheme I tried.

A nasty development surfaced when the dipole in front of my 2.5-meter (8-foot) dish was moved for field-strength tests and beamwidth measurements. Any change of antenna configuration forced me to retune the driver. The answer, of course, was to find some kind of impedance-matching device that could be put into the line at the antenna and at the transmitter. The double-stub impedance transformer was the logical solution. But, where to find one? The surplus houses I trade with never had a single one that could possibly have worked at 1296 **MHz.** I had to build one; in fact, I had to make four of them. (Anybody working 1296 is bound to have more than one antenna.)

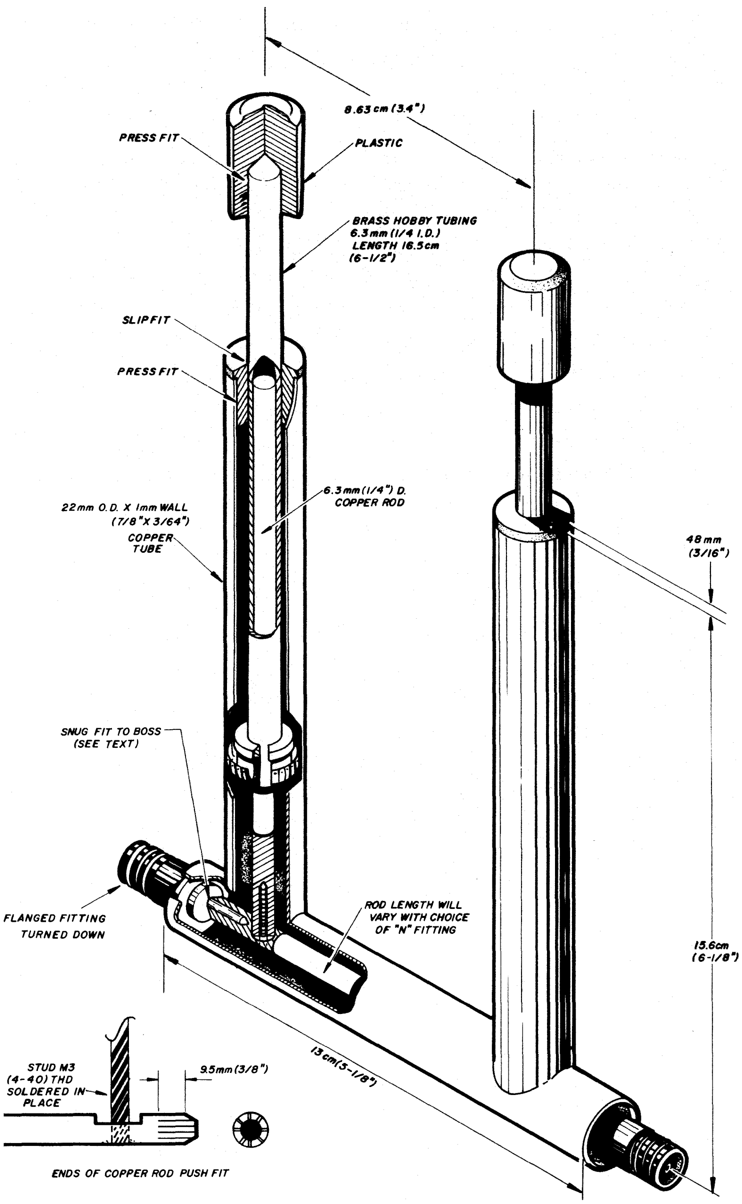


fig. 1. Mechanical diagram of the double-stub tuner. As mentioned in the text, the length of the center conductor of the main line is determined by the type of N connector which is used. Matching is accomplished by the two stubs with sliding shorts which provide adjustable amounts of reactance.

The impedance-matching range of the double-stub impedance-matching transformer is astonishing. For example, it is no problem at all to bring the feedlines from two dipole antennas together at a T and match the resulting 25-ohm impedance to a 50-ohm line.

tuner construction

The stubs of a double-stub tuner are normally 318 wavelength apart and mounted at right angles to a 50-ohm line section. Insertion loss is not readable on a 0- to 25-watt rf power meter running at half scale, There is just one problem with the tuners: they are hard to make because mechanical tolerances must be held tightly so that the tuning plungers do not bind when they are adjusted. It is not necessary to have a mill to build a double-stub tuner, although there is one minor place where it produces a more professional job. A lathe should be available.

construction

Referring to the cross-sectional sketch (see **fig. 1),** tubing sections used for the stubs can be sawed to length; remember that they must be long enough to reach halfway around the mainline section. Cut and file one end of each stub tube to fit snugly around the main section line. Scribe the ***outside*** contour of one of the stub tubes onto the main line, **43** mm (1.7 inches) from the center of the main line; scribe a similar contour **43** mm (1.7 inches) from the center toward the other end. These outlines must be parallel along the main line tube. Remember, these scribed contours are the ***outside*** of the stub tubes. Cut and file openings in the main line tube to match the inside diameter of the stub sections. Final fit must be made with the file strokes at right angles to the line tube and parallel to the axis of the stub tubes, as they will solder in place. This may sound difficult, but it's not, just a bit tedious.

Assembly of the two stubs to the main line must be done accurately. I found the simplest way was to lay the three parts on a piece of 1.9-cm (314-inch) plywood and drive nails on both sides of all three tubes to hold them in place. A square will indicate when the stubs are at right angles to the main line. The plywood will keep all three parts in the same plane. Put something heavy on each of the tubes to keep it from shifting during the soldering operation. The stubs are soldered to the main line section with a substantial bead halfway around the joint. The assembly is then turned over in the jig to complete the second half of the soldering job. If the half-soldered

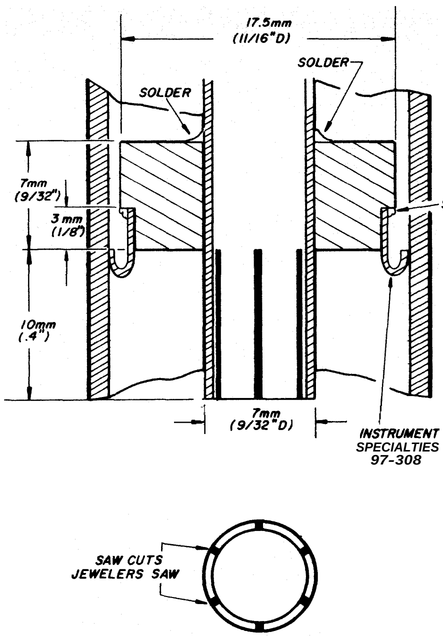


fig. 2. Close-up view of the fingerstock assembly. If brass shim stock is used instead of fingerstock, the size of the brass ring can be changed to compensate for the size difference.

assembly, when flipped over, does not drop into the nail jig without binding, the right-angle alignment of the jig is faulty. If this happens, it is best to start over.

In **fig. 1,** the dimension of the inner conductor of the main line has not been shown. This is because flanged N connectors vary somewhat in design, and the builder must determine this length to match the available fittings. For final assembly, the outer face of both N flanges must be recessed into the main line tube approximately 1.5 mm (1/16 inch). The normally square flanges are turned on a lathe to a diameter allowing a slip fit into the copper tube. The center-conductor's length is set to fit snugly up against the boss at the base of the conductor extension of each N fitting. This is not critical, but recessing of the N connector flanges is necessary to permit a substantial solder bead for maximum strength under torque load from connected equipment.

Once the inner conductor length is established, the ends of the rod are tapered and an 11 mm (7116 inch) deep hole is drilled in each end. The diameter of the hole is selected to provide a slip fit on the center pin of the connector. With a jeweler's saw, three longitudinal cuts are made on each end of the center-conductor rod, giving six segments at each end. The segments are given a set toward the center to establish a press fit over the center pin of the N conductor (no soldering is possible at final assembly). The longitudinal cuts are 9.5 mm (3/8 inch) deep, If the N conductor pins are more than 6.5 mm (1/4 inch) long, cut them back to 6.5 mm (1/4 inch).

There is one very critical dimension in the inner conductor assembly, the spacing between the points at which the stubs connect. This spacing must match, as closely as possible, the center spacing between the outer tubes already assembled. Measure the distance between the stub centers accurately and transfer this measurement to the center conductor of the main line. (Center-to-center measurement is most easily done by measuring from the left-hand side of one tube to the left-hand side of the other tube, or conversely, from the right-hand side of one tube to the right-hand side of the other tube.) This measurement is transferred to the center conductor of the main line so that the distance from one end of the line to the first stub is equal to the distance from the other end of the line to the second stub.

Clamp the rod in a vise and file flat spots 6.5 mm (1/4 inch) wide over the stub locations. These should be filed one-third of the way through the rod. The faces must be flat and parallel to each other since they affect final alignment of the stubs. Drill and tap each flat for a 4-40 (M3) machine screw, ensuring that the distance between the drilled holes is exactly the same as the distance between the centers of the two stub tubes already assembled. Insert brass 4-40 (M3) machine screws from the bottom so that the threaded end of the screws extends 9.5 mm (3/8 inch) above the face of the flats. The above procedure is for those who don't have access to a mill. (With a mill, sink a 6.5-mm [1/4-inch] end mill 3 mm [1/8 inch] deep, then drill and tap for the screws.)

The center-stub conductors are squared in the lathe, drilled and tapped for the screws already in place in the main line. Screw the stubs onto the 4-40 (M3) studs and check for perpendicular projection from the line and for parallel alignment. If alignment is correct, the screws can be soldered to the main line, after which the excess should be sawed off and smoothed to maintain the shape of the line. The stubs must be removed from the mounting studs for final assernbly.

The plungers are made from 6.5-mm (1/4-inch) ID brass hobby tubing, which is a slip fit on the center conductor stubs. The tubing is also split with a jeweler's saw as indicated in **fig. 2.** The segments are pressed inward to give a good sliding fit on the center conductor. The brass ring which supports the finger-stock is bored, turned and soldered in place. The indicated fingerstock is ideal for the job, but brass shim stock may be substituted if the fingerstock is not available. When using shim stock, wrap it around the ring, slit in segments, and solder in place. Give the segments an outward bias and contour the tips to slide rather than dig into the outer conductor.

with the plungers completely in. It is good insurance to wrap the base of the stub tuner lines, where they solder to the main line, with a damp cloth while soldering the N fittings into place.

With these three pieces of gear, directional watt meter dummy load, and stub tuners, 1215 MHz to 1300 MHz becomes one of the easiest of the amateur bands to work on, rather than one of the hardest.

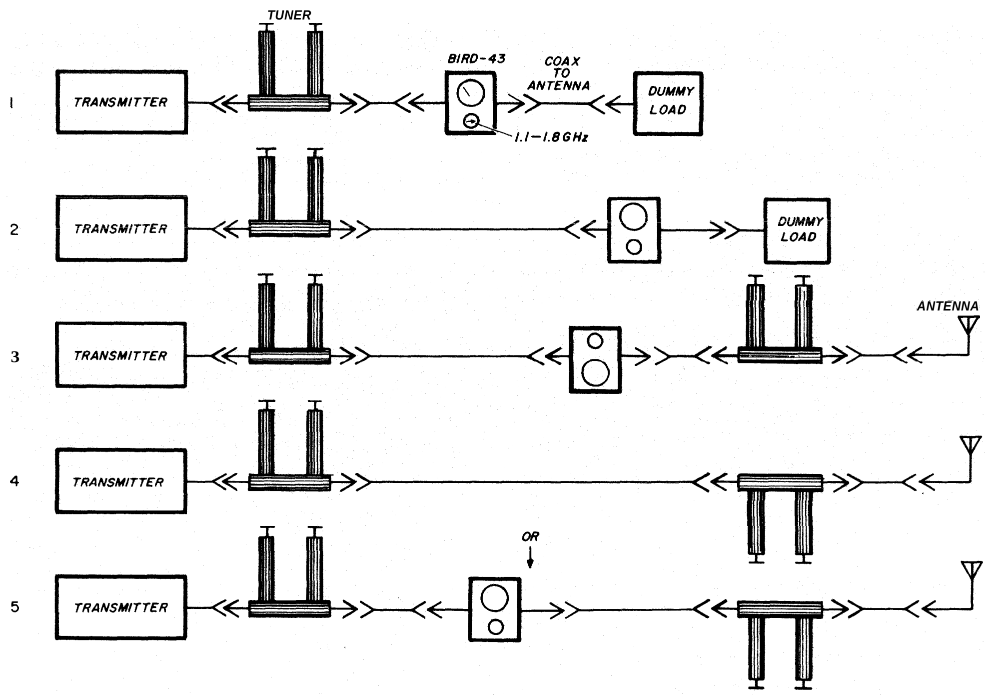


fig. 3. Different configurations for using one or two double-stub tuners.

The plastic fitting at the end of each stub is drilled for a slip fit on the plunger tubing and pressed onto the end of the stub. The plunger knobs may be any type of plastic and are also press fitted onto the plunger tubing.

final assembly

The center conductor of the main line is slipped into the tube, the center conductor stub lines are screwed into place on the **4-40 (M3)** screws and set tight with a pair of heavy pliers. The N fittings are pressed into the ends of the main line center conductor and the plungers are slid down over the stub center lines. The plastic guide sleeves are pressed into the open end of the stub tubes and the plungers are run up and down to check for any binding. There should be none, though a slight difference in resistance may be noticed at first as the plungers move full stroke. If no problems show up, the flanges of the N fittings may be soldered with the plungers both fully inserted. This is important, as any misalignment will be less troublesome if the N fittings are soldered

The guesswork disappears in checking cavity or stripline efficiency, antenna match, and line reflections. Comparing one antenna to another becomes routine, rather than the long, drawn-out process of trying to match power into an antenna system in the face of all the line and feed problems that constantly disturb the results.

Different tuner configurations are illustrated in **fig.** 3. As a simple indication of what can be accomplished with one of these tuners, a test set-up was arranged on the bench. Eight watts of power was fed into a dipole that had been trimmed to give a good match to a 50-ohm line when used to drive a 2.5-meter **(8-**foot) parabolic dish. The wattmeter showed that two watts were being reflected. With a double-stub tuner into the line at the dipole, it took less than ten seconds to wipe out the reflected power by adjusting the stub plungers. This was a 25 per cent increase in radiated power without having to touch the antenna. If aggravation has a price tag, these three pieces of gear are a bargain.

ham radio

1.5 GHz divide-by-four prescaler

Extending the range of your counter is made easy by this simple divide-by-four 1500-MHz prescaler

In the last few years frequency counters have developed from bulky, slow, costly units affordable only by well-heeled laboratories, into small, high-frequency instruments at a price easily within the reach of the average ham. Quite a few manufacturers now offer counters with 1-Hertz resolution to 50 or *60* MHz, low power drain, and a price in the neighborhood of $100. In addition, many articles have appeared in the amateur magazines describing the construction of frequency counters and prescalers; recently, an entire issue (February 1978) of ham *radio* was devoted to the subject of frequency counters.

The upper frequency limit of every counter is set by the maximum operating speed of the input divider stage. This first divider is usually a fast prescaler — a fixed, asynchronous divider operating with the main divider chain. The most popular low-cost prescalers use the 11C90 digital divider, which has a typical maximum operating frequency of about 600 MHz. IC prescalers with higher operating speeds are available, but their high prices have kept them from finding much amateur use.

However it is now possible to inexpensively extend a counter's operating frequency range to well beyond 1 GHz, with a recently introduced ECL (emitter-coupled logic) divide-by-four prescaler. This device is the Motorola MC1697, a very fast ECL prescaler which has a typical operating frequency of 1600 Mhz. Thus, a prescaler using this circuit will extend the range of a 400-MHz counter to above 1.5 GHz. It will operate with signals as low as 1 mW, requires only a single dc supply, and will drive 50-ohm loads. The MC1697 is a low-cost, plastic-packaged device that sells for less than $18.

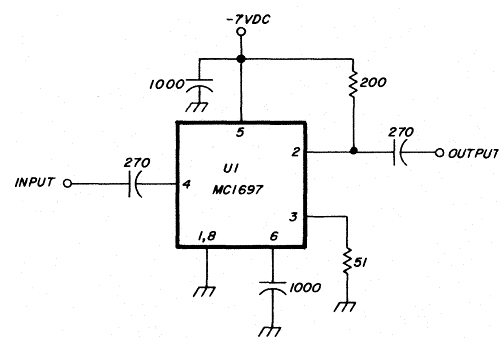


fig. 1. Schematic diagram of the divide-by-four prescaler using the Motorola MC1697. Chip capacitors are preferred for the input and output coupling capacitors, though silver micas may be used if the other types cannot be obtained. If the IC should free-run, the oscillations can be stopped by connecting a 10k resistor across pins 4 and 5.

By Jerry Hinshaw, N6JH, 2500 Medallion #193, Union City, California 94587

circuit description

The schematic diagram of the prescaler's circuit is shown in **fig. 1.** The signal to be counted is fed through C1 to the MC1697, which contains two divide-by-two stages in cascade. The input to the first counter is internally biased to accept zero-mean signals so that no external set-up bias is needed. The second flip-flop has complementary outputs. One of these is fed to the output connector through C2, the other is terminated in a 51-ohm resistor. The 200-ohm resistor terminates the open emitter of the output emitter follower.

A single dc power supply is required to provide −7 volts at approximately 60 mA. A simple power supply for this prescaler module can be built using an adjustable voltage regulator. The design of power supplies using this type of regulator has been discussed by K5VKQ.[1]

construction

The prescaler is built on the small printed circuit board shown in **fig. 2.*** Double-clad, 1.6-mm (1/16-inch) glass-epoxy board is used. One side forms a ground plane, which provides for low-impedance grounding and permits the use of microstrip lines for the input and output.

All components, with the exception of C1 and C2, are mounted on the ground plane side of the board. If chip capacitors are used for C1 and C2, they should be soldered directly across the gaps in the microstrip lines. If silver mica capacitors are to be used, they may be mounted on the ground plane side of the board with the rest of the components. The use of chip capacitors is preferable, as they will provide lower loss and less inductance than capacitors with wire leads.

Be sure to leave enough clearance under the bodies of the two 1000-pF bypass capacitors so that the lead can be soldered to the ground plane at the points indicated by Xs on **fig. 2C.** One end of the 51-ohm resistor is connected to the top side of the board.

If a socket is used to mount the MC1697, it must be the type that can be soldered to both sides of the board so that pins 1 and 8 can be properly grounded. Molex or swage-in pins will work here, or the IC can be soldered in the circuit board. Soldering the IC is the best choice in terms of electrical and thermal performance, but it does make changing the device difficult. If you are going to purchase only one IC (rather than buy several to select the fastest), you may as well solder it directly to the board.

*An etched, drilled, and plated glass-epoxy circuit board is available for $6.00 postpaid, from H&M Microwave, Post Office Box 185, Montrose, California 91020. California residents please add 6 per cent sales tax.

The board was designed to fit into a Bud CU124 cast aluminum box, although it will also fit into a 5.1 x 7.6 x 10.2-cm (2 x 3 x 4-inch) sheet-metal box. The cast box provides better grounding and rf shielding. The board fits into the box as shown in **fig.** 3. The component side is mounted facing down; this is done so that the BNC input and output connectors can be directly soldered to the microstrip lines after the board has been installed in the box.

The board itself is mounted in the box with four 9.5-mm (3/8-inch) standoffs located at the four mounting pads near the corners of the board. The standoffs should be soldered to the ground plane and then securely clamped to the box with screws.

If an external power supply is used, the lead may

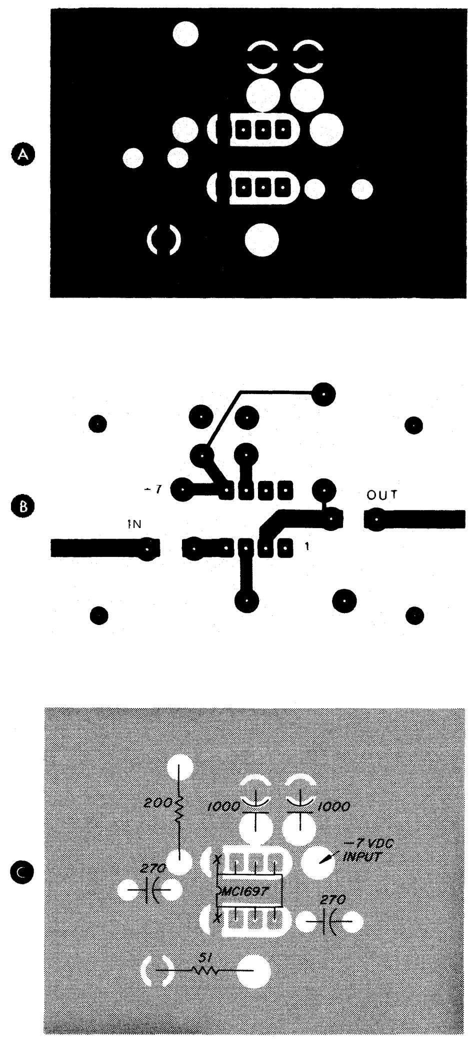


fig. 2. The top and bottom sides of the printed circuit board are shown in A and B respectively. In the parts placement diagram, C, C1, and C2 are installed on the top of the board if glass or mica capacitors are used; if chip ceramics are used, they are mounted on the foil side. The leads marked with an X are to be soldered on the ground plane side.

be brought into the box with a feedthrough bypass-type capacitor. Alternatively, a small power supply can be built in the box containing the prescaler.

operation

This prescaler module has no adjustable components and should work, Murphy willing, when it is first turned on. However, if it does not, there are several points to check. First, see that the − 7 volts is present at pin 5 on the MC1697. If you have an oscilloscope, check to see that the IC is not oscillating. (A cure for sscillations is to connect a 10k-ohm resistor from pin 4 to V_{ee}.) Also check the grounds on the two bypass capacitors, pins 1 and 8 on the MC1697, and on the 51-ohm resistor. All of these ground points must be soldered to the top of the printed circuit board.

Once the prescaler is built and operating you will naturally want to check it to find the maximum operating frequency. One way of doing this involves finding a signal generator or source that covers the 1- to 2-GHz range. Attach the prescaler and counter to the generator. Start with the generator set near 1 GHz and verify that the counter reads approximately 250 MHz.

Slowly increase the frequency of the signal generator and watch the counter display. Somewhere between 1500 and 1700 MHz (about 375 to 425 MHz on the display) the indicated frequency will suddenly jump downward. This shows that the divider is no longer able to operate normally. The point at which the display jumps to a radically different value indicates the maximum input frequency at which your prescaler will operate.

There is another test that will help locate the critical upper frequency. While slowly increasing the input signal frequency, watch the least significant digits on the display of the counter. As the maximum operating frequency is approached these digits will become erratic, indicating that the phase noise on the prescaler's output is becoming significant. This flickering occurs a few megahertz below the cutoff frequency.

If you wish to vary the power supply voltage slightly, you may be able to find a point at which the maximum operating frequency is increased. However, I found that presetting the supply to − 7 volts is nearly optimal, and that adjusting the voltage never raised the upper frequency limit by more than a small amount.

Another good test of the prescaler is to count a signal whose frequency is near the maximum limit for the counter itself, note the frequency displayed, and then count the same signal using the prescaler. If the counts (after multiplying by four) are very nearly the same, the prescaler is probably operating correctly.

The most serious shortcoming of this prescaler module is its limited dynamic range. Typically, the input signal must be in the range of 200 to 1600 mV p-p for proper operation. The dynamic range could be extended by the use of an input stage of amplification and limiting. The problem with this approach is that the amplifier stage must be broadband, covering at least 600 to 1500 MHz; such amplifiers are not easily built. Avantek[2] has published an application note which covers the construction of a 40-dB dynamic range preamplifier covering 25 to 1000 MHz. Using thin-film hybrid amplifier modules to achieve its performance, its principal shortcoming lies in the cost − more than $120.

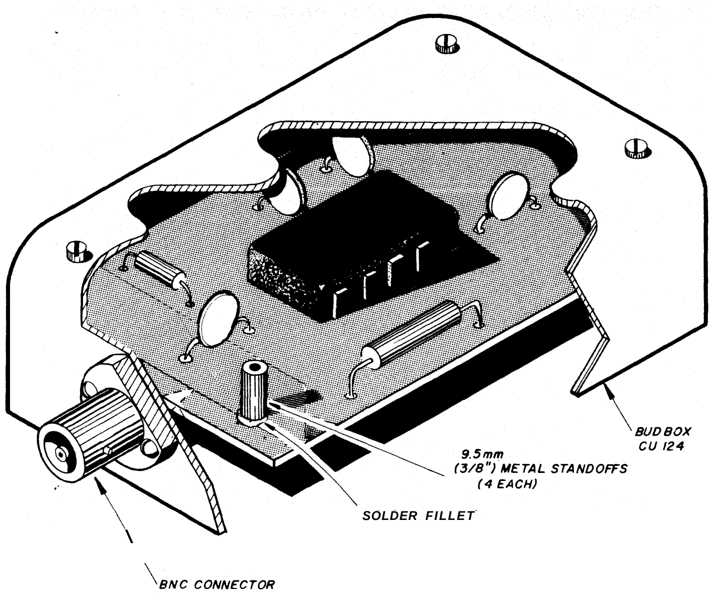


fig. 3. Mounting details for the divide-by-four prescaler. The center pin of the BNC connectors is soldered to the trace on the printed circuit board.

The simple prescaler described here, even without a preamplifier, is a useful test instrument. If the input signal's amplitude is outside the operating range, the prescaler does not produce spurious outputs − it simply stops operating. As long as the input signal source can provide at least 1 mW, some attenuator level can be found which will permit operation of the scaler. Several versions of this prescaler module have been built. All functioned without problems and all had a maximum operating frequency of at least 1500 MHz.

references

1. Chris Cogburn, K5VKQ, "How to Design Regulated Power Supplies," ham radio, September 1977, page 58.
2. "A 1-GHz Prescaler Using GPD Series Thin-Film Amplifier Modules," Avantek Application Note ATP 1036, Avantek Corporation, 3175 Bowers Avenue, Santa Clara, California 95051.

ham radio

ham radio cumulative index 1969-1978

antennas and transmission lines

general

high-frequency antennas

Selective antenna system minimizes unwanted signals
W5TRS p. 28, May 76
Selective receiving antennas
W5TRS p. 20, May 78
Shunt-feed systems for grounded vertical radiators, how to design
W400 p. 34. May 75
Simple Bntennas for 40 and 80
W5RUB p. 16. **Dec** 72
Sloping dipoles
W5RUB p. **19, Dec** 72
Performance (letter) p. 76, May 73
Small-loop antennas
W4YOT p. 36. May 72
Stressed quad **(HN)**
W5TIU p. 40. Sep 78
Stub bandswitched antennas
W2EEY p. 50. Jul 69
Suitcase antenna, high-frequency
VK5BI p. 61. May 73
Tailoring your antenna, how to
KH6HDM p. 34. May 73
Telephone-wire antenna **(HN)**
K9TBD p. 70. May 76
Triangle antennas
W3FQJ p. 56. Aug 71
Triangle antennas
W6KIW p. 58. May 72
Triangle antennas (letter)
K4ZZV p. 72. Nov 71
Triangle beams
W3FQJ p. 70. **Dec** 71
Tuning aid for the sightless **(HN)**
W6VX p. 83. Sep 76
Unidirectional antenna for the low-frequency bands
GW3NJY p. 61. Jan 70
Vertical antenna radiation patterns
W7LR p. 50. Apr 74
Vertical antenna, low-band
W4IYB p. 70, Jul 72
Vertical antenna, portable
WA8NWL p. 48. Jun 78
Vertical antenna, three-band
W9BQE p. 44. May 74
Vertical antennas, improving **performance** of
K6FD p. 54. **Dec** 74
Vertical antennas, performance characteristics
W7LR p. 34, Mar 74
Vertical beam antenna. 80 meter
VE1TG p. 26. May 70
Vertical dipole, gamma-loop-fed
W6SAI p. 19. May 72
Vertical for 80 meters, top-loaded
W2MB p. 20. Sep 71
Vertical radiators
W4OQ o. 16. **Apr** 73
Vertical, top-loaded 80 meter
VE1TG p. 48, Jun 69
Vertical-tower antenna system
W4OQ p. 56. May 73
Windom antenna, four-band
W4VUO p. 62. Jan 74
Correction (letter) p. 74. Sep 74
Windom antennas
K4KJ p. 10, May 78
Windom antenna (letter)
K6KA p. 6, Nov 78
Zepp antenna, extended
W6QVI p. 48, **Dec** 73
ZL special antenna, understanding the
WA6TKT p. 38. May 76
3.5-MHz phased horizontal array
K4JC p. 56, May 77
3.5 MHz tree-mounted ground-plane
K2INA p. 48. May 78
7-MHz antenna array
K7CW p. 30. **Aug** 78
7 MHz rotary beam
W7DI p. 34, Nov 78
7-MHz short vertical antenna
W8TYX p. 60. Jun 77
14-MHz delta-loop array
N2GW p. 16, Sep 78
160-meter loop, receiving
K6HTM p. 46, May 74
160-meter vertical, shortened **(HN)**
W6VX p. 72. May 76
160 meters with 40-meter vertical
W2IMB p. 34, Oct 72

vhf antennas

Antennas for satellite communications, simple
K4GSX p. 24. May 74
Circularly-polarized ground-plane antenna for satellite communications
K4GSX o. 28. **Dec** 74
Collinear antenna for two meters. nine-element
W6RJO p. 12. May 72
Collinear antenna (letter)
W6SAI p. 70. Oct 71
Collinear array for two meters. 4-element
WB6KGF p. **6,** May 71
Collinear antenna, four element 440-MHz
WA6HTP p. 38, May 73
Collinear. six meter
K4ERO p. 59. Nov 69
Converting low-band mobile antenna to 144-MHz **(HN)**
K7ARR p. 90. May 77
Corner reflector antenna. 432 MHz
WA2FSQ p. 24. Nov 71
Cubical quad, economy six-meter
W6DOR p. 50. Apr 69
Feed horn, cylindrical, for parabolic reflectors
WA9HUV p. 16, May 76
Ground plane. 2-meter. **0.7** wavelength
W3WZA p. 40. Mar 69
Ground plane, portable vhf **(HN)**
K9DHD p. 71, May 73
Log-periodic, **yagi** beam
K6RIL, W6SAI p. **8,** Jul 69
Correction p. 68. Feb 70
Magnet-mount antenna, portable **(HN)**
WB2YYU p. 67, May 76
Magnetic mount for mobile antennas
WØHK p. 52. Nov 78
Matching techniques for **vhf/uhf** antennas
W1JAA p. 50. **Jul** 76
Microwave antenna, Low-cost
K6HIJ p. 52, Nov 69
Mobile antenna. magnet-mount
W1HCI p. 54. Sep 75
Mobile antenna, six-meter **(HN)**
W4PSJ p. 77, Oct 70
Mobile antennas, vhf, comparison of
W4MNW p. 52, May 77
Moonbounce antenna, practical 144-MHz
K6HCP p. 52. May 70
Multiband J antenna
WB6JPI p. 74. Jul 78
Oscar antenna, mobile **(HN)**
W6OAL p. 67. May 76
Oscar az-el antenna system
WA1NXP p. 70. May 78
Parabolic reflector antennas
VK3ATN p. **12,** May 74
Parabolic reflector element spacing
WA9HUV p. 28. May 75
Parabolic reflector gain
W2TQK p. 50. Jul 75
Parabolic reflectors. **finding** the focal **length (HN)**
WA4WDL p. **57, Mar 74**
Parabolic reflector. 16-foot **homebrew**
WB6IOM **p. 8, Aug 69**
Quad-yagi arrays. 432- and 1296-MHz
W3AED p. 20. May **73**
Short circuit p. 58. **Dec 73**
Simple antennas. 144-MHz
WA3NFW p. 30. May **73**
Switch, antenna for 2 meters, solid-state
K2ZSQ p. 48. May **69**
Two-meter fm antenna **(HN)**
WB6KYE p. 64. May **71**
Vertical antennas. truth about **⅝-wavelength**
KØDOK p. 48. May **74**
Added note (letter) p. 54. Jan **75**
Whip. **5/8-wave,** 144 MHz **(HN)**
VE3DDD p. 70, Apr **73**
Yagi antennas, how to design
W1JR p. 22, **Aug 77**
Yagi. 1296-MHz
W2CQH p. 24, May 72
7-MHz attic antenna **(HN)**
W2ISL p. 68, May 76
10-GHz dielectric antenna **(HN)**
WA4WDL p. 80. May 75
144-MHz vertical. **⅝-wavelength**
K6KLO p. 40. Jul 74
144-MHz antenna, **⅝-wavelength** built from **CB** mobile whip **(HN)**
WB4WSU p. 67, Jun 74
144-MHz colinear uses PVC pipe mast **(HN)**
K8LLZ p. 66, May 76
144-MHz mobile antenna **(HN)**
W2EUQ p. 80. Mar 77
144-MHz vertical mobile antennas, ¼ and ⅝ wavelength. test data on
W2LTJ, W2CQH p. 46, May 76
144-MHz. **⅝-wavelength** vertical
W1RHN p. 50. Mar 76
144-MHz. **⅝-wavelength,** vertical antenna for mobile
K4LPQ p. 42. May 76
432-MHz high-gain Yagi
K6HCP p. 46, Jan 76
Comments. WØPW p. 63. May 76
432-MHz OSCAR antenna **(HN)**
W1JAA p. 58. Jul 75
1296-MHz antenna, high-gain
W3AED p. 74. May 78
1296-MHz Yagi array
W3AED p. 40. May 75

matching and tuning

Antenna bridge calculations
Anderson, Leonard H. p. 34. May 78
Antenna bridge calculations (letter)
W5QJR p. 6. Aug 78
Antenna coupler for three-band beams
ZS6BT p. 42. May 72
Antenna coupler, six-meter
K1RAK p. 44. **Jul** 71
Antenna impedance transformer for receivers **(HN)**
W6NIF p. 70. Jan 70
Antenna instrumentation, simple. (repair bench)
K4IPV p. 71. Jul 77
Antenna matcher, one-man
W4SD p. 24. Jun 71
Antenna tuner adjustment **(HN)**
WA4MTH p. 53, **Dec** 75
Antenna tuner, automatic
WAØAQC p. 36. Nov 72
Antenna tuner, medium-power toroidal
WB2ZSH p. 58. Jan 74
Antenna tuner for optimum power transfer
W2WLR p. 28. May 70
Antenna tuners
W3FQJ p. 58. **Dec** 72
Antenna tuning units
W3FQJ p. 58, Jan 73
Balun. adjustable for yagi antennas
W6SAI p. 14. May 71
Balun. Simplified **(HN)**
WAØKKC p. 73, Oct 69
Broadband balun, simple and efficient
WIJR p. 12. Sep 78
Couplers. random-length antenna
W2EEY p. 32. Jan 70
Dummy loads
W4MB p. 40, Mar 76
Feeding and matching techniques for **vhf/uhf** antennas
WIJAA p. 54. May 76
Gamma-match capacitor. remotely controlled
K2BT p. 74, May 75
Gamma-matching networks, how to design
W7ITB p. 46. May 73
Impedance bridge, low-cost RX
W8YFB p. 6, May 73
Impedance-matching baluns, open-wire
W6MUR p. 46. Nov 73
Impedance-matching systems, designing
W7CSD p. 58. Jul 73
Loads, affect of mismatched transmitter
W5JJ p. 60. Sep 69
Matching, antenna, two-band with stubs
W6MUR p. 18. Oct 73
Matching system. two-capacitor
W6MUR p. 58. Sep 73
Matching transformers, multiple quarter-wave
K3BY p. 44. Nov 78
Measuring complex impedance with swr bridge
WB4KSS p. 46, May 75
Mobile transmitter, loading
W4YB p. 46. May 72
RX noise bridge, improvements to
W6BXI, W6NKU p. 10, Feb 77
Comments p. 100. Sep 77
Noise bridge construction (fetter)
OH2ZAZ **p. 8,** Sep 78
Noise bridge, antenna
WB2EGZ **p.** 18. **Dec** 70
Noise bridge, antenna **(HN)**
K8EEG p. **71,** May 74
Noise bridge calculations with TI 58/59 calculators
WD4GRI p. 45. May 78
Noise bridge for impedance measurements
YA1GJM p. 62, Jan 73
Added notes p. 66. May 74; p. 60. Mar 75
Omega-matching networks, design of
W7ITB p. 54, May 78
Phase meter, **rf**
VE2AYU, Korth p. 28. Apr 73
Quadrifilar toroid **(HN)**
W9LL p. 52. **Dec** 75
Stub-switched, stub-matched antennas
W2EEY p. 34. Jan 69
Swr alarm circuits
W2EEY p. 73. Apr 70
Swr bridge
WB2ZSH p. 55. Oct 71
Swr bridge and power meter, integrated
W6DOB p. 40, May 70
Swr bridge readings **(HN)**
W6FPO p. 63, Aug 73
Swr indicator, aural, for the **visually** handicapped
K6HTM p. 52, May 76
Swr meter
WB6AFT p. 68. Nov 78
Swr meter
W6VSV p. 6, Oct 70
Swr meter, improving **(HN)**
W5NPD p. 68, May 76
T-Network impedance matching to coaxial feedlines
W6EBY p. 22. Sep 78
Transmatch, five-to-one
W7IV p. 54. May **74**
Transmission lines, grid dipping **(HN)**
W2OLU p. 72, Feb **71**
Transmission lines, uhf
WA2VTR p. 36. May **71**
Uhf coax connectors **(HN)**
WØLCP p. 70, Sep **72**

towers and rotators

transmission lines

audio

commercial equipment

construction techniques

features and fiction

fm and repeaters

integrated circuits

keying and control

measurements and test equipment

microprocessors, computers and calculators

miscellaneous technical

novice reading

operating

oscillators

power supplies

propagation

receivers and converters

general

high-frequency receivers

vhf receivers and converters

receivers and converters, test and troubleshooting

RTTY

satellites

semiconductors

single sideband

television

transmitters and power amplifiers

general

high-frequency transmitters

vhf and uhf transmitters

transmitters and power amplifiers, test and troubleshooting

troubleshooting

vhf and microwave

general

vhf and microwave antennas

vhf and microwave receivers and converters

vhf and microwave transmitters