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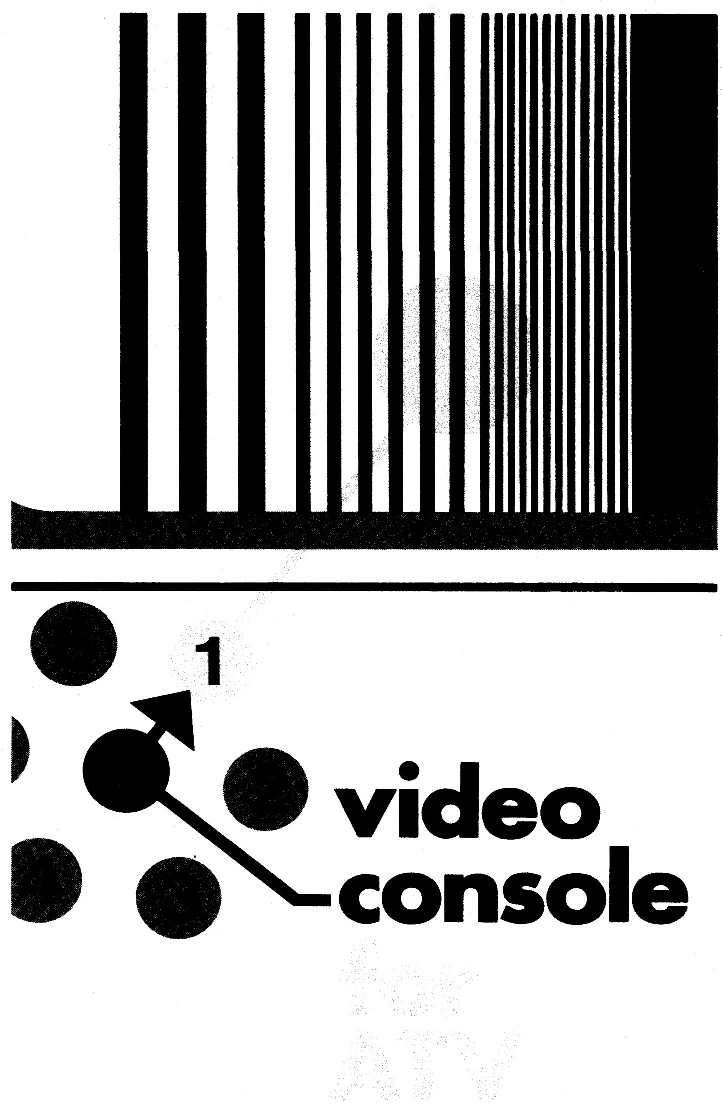
JANUARY 1980



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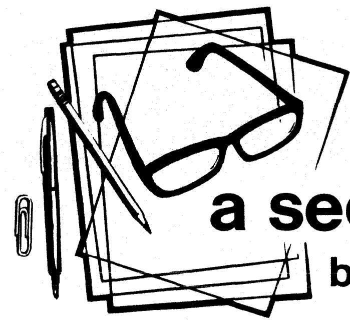


a second look

by Jim Fisk

Are you thinking about buying one of those new solid-state, phased-locked rigs with digital readout? If you are, be sure you buy a rig with a good solid ***factory-backed*** warranty! As some buyers have sadly discovered, there are ***some*** Amateur Radio equipment manufacturers who require their dealers to provide the warranty as part of their "dealer agreement;" and in some cases the manufacturer or supplier does not reimburse the dealer for his warranty work. Under such arrangements it doesn't take much imagination to understand why the dealers aren't particularly enthusiastic about fixing faulty equipment. The dealer is further hampered by the unavailability of up-to-date service information – his technicians often have little more available to them than whatever information is packed with the equipment.

I won't dispute the fact that if you had a problem with your equipment you would be able to get it repaired. You probably could; but it might be a long drawn out affair, depending on the dealer you're working with and how much interest he has in doing the factory's warranty repair work with no reimbursement for his parts.

In this day and age of sophisticated equipment the dealers and repair technicians must stock many, many different components for each item they are called on to service. If the dealer is not reimbursed for his time and labor, it's obviously not in his best interest to tie up vast amounts of cash in spare parts – he simply orders the repair parts as he needs them, and you have to wait weeks – perhaps months – before you can get your transceiver back on the air.

Under such circumstances shipping your rig back to the factory for repair may be the best move you can make. They have all the parts and expertise to do it right, and in many cases they can give you better "turn around" than your local dealer. This is not a criticism of your dealer's repair facilities; his technicians must be familiar with many different types of equipment, whereas the factory technicians are able to specialize and hence can often do the job faster.

As a final note, any equipment manufacturer who does not offer factory repair facilities is suspect – they are simply passing the buck down the line to the dealer. And you know who suffers in the long run!

Most of the Amateur Radio manufacturers, fortunately, offer full factory-backed warranties, but those who do not should be avoided like the plague. Keep that in mind before you plunk down your hard-earned dollars for a new rig.

Jim Fisk, W1HR
editor-in-chief

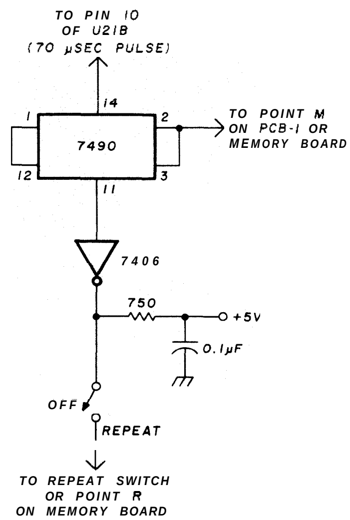
comments

W7BBX memory keyer

Dear HR:

I have used the WBBX memory keyer* with great success. Compared with the WB4VVF memory keyer (built by SP5JC) and other designs described in the European ham magazines, the WBBX circuit is definitely the best. I would like to tell your readers of a small addition to the existing circuit which is especially valuable in meteor-scatter (MS) communications.

One inconvenience in the unmodified keyer is the necessity to completely fill the memory, because any empty memory space will show up as breaks in transmissions with repeated messages; such transmissions are used a great deal in MS operations. I used to try to completely fill the memory, but with some callsigns and reports used in MS it was not always possible. Then you would either lose valuable time by sending nothing, or push the REPEAT button every few seconds, depending on the speed. The last method leaves no time for other aspects of MS work, such as listening through the recorded tapes at reduced speed while your keyer does the transmission sequence. I decided to add a circuit which senses the end of a message stored in the keyer memory and automatically REPEATS the message. The circuit is very simple and can be built on a small add-on PC board and wired into the existing boards either above or between them. One additional switch on the front panel is necessary for switching the circuit in or out of operation, as shown below.



I also added a switch with a bank of multiturn precision adjustable potentiometers, each for a preset high-speed value. Using the original 2.2-μFd tantalum capacitor for C4 and 180 ohms for R6, I can use any speed up to 1000 letters per minute (LPM) simply by choosing the position of the speed switch, which replaces the original speed pot. One speed switch position switches in a multiturn pot with dial for "normal" variable-speed operation at conventional speeds. There are no keyer problems with transmission speeds up to 1000 LPM (higher speeds were not tried), but the time constants in the transmitter keying circuit should be carefully checked to ensure proper operation on the air. I'm using 5k and 20k pots for the preset speed trimmers; in Europe we use speeds around 500-700 LPM for MS, depending on the available speed reduction in the recorders at the other receiving end.

Another valuable feature of the WBBX keyer is the remote start facility; I'm now using a pulse from the station electronic clock, which gives a pulse every 15 seconds, 30 seconds, 1 minute, or 5 minutes (the latter is still the most common transmit/receive sequence in European MS operation to handle the timekeeping). I now have plenty of time during a MS QSO, whereas before I had to be constantly on the aiert, checking tapes, and looking at the clock.

Wes Wysocki, SP2DX
Gdansk, Poland

*Howard Batie, W7BBX, "Programmable Contest Keyer," ham radio, April, 1976. page 10.

low-power wattmeter

Dear HR:

I recently built my own version of the low-power wattmeter described by WA4ZRP in the December, 1977, issue of *ham* ***radio.*** The electronics perform very well — I eliminated the need for carefully measuring the calibration resistors to 0.001 ohm by using 500-ohm trimpots with parallel and/or series resistors to make up the required ranges.

The key to the design, however, is the subminiature lamps. I was unable to find the correct voltage and current ranges for the stated ***generic*** types T 3/4 and CM units listed by the author. For those who have built similar units and have had difficulty finding the barreter lamps, the Sylvania 28ES will do (rated at least 28 volts and 40 mA with a resistance of 80 ohms); however, these lamps are somewhat large, as they're rated at 1.2 watt.

It may be possible to locate some of the subminiature 3-volt, 30-mA lamps, but you must first know the rated value of dc resistance (it is not voltage divided by current). I have sought this information from Chicago Miniature Inc. (who make the CM units mentioned in the article), but have had no response to date.

Gene Shapiro, WØDLQ
Leawood, Kansas

video console for ATV

A project for Amateurs interested in improving their fast-scan television stations — three video generators, a special-effects generator, and a simplified form of Gen-Lock are described

The video console project began after I'd completed a solid-state ATV transmitter. I had no way of knowing how it was performing except by having other ATVers try to "talk me in" on alignment. I soon learned this was almost impossible.

I discussed my problem with Bill Parker, W8DMR, another ATV ham, and he suggested that I needed a few test generators. So I consulted *A5* magazine and found articles on two generators.[1,2] After running some tests and reading a book by Tektronix,[3] I found a need for one more generator so that I could check 99 per cent of my new transmitter. I then designed a pulse and bar generator. After this I thought, "Boy, wouldn't it be nice to have some special effects, such as a video switcher." After a short time I had it operational.

the video console

The video console consists of three function generators, a video switcher, and a Glen-Lock circuit. All can be locked to either externai or internai sync or to a composite video signal. Everything is contained on five PC boardsand powered by a single 12-volt supply.

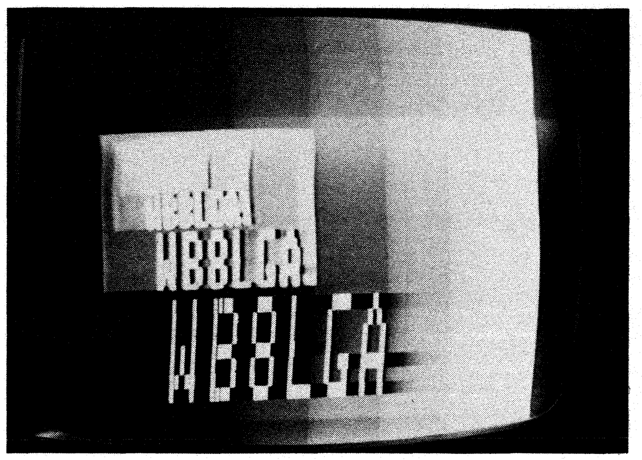
The function generators are multiburst, gray scale, and pulse and bar. The multiburst generator creates a frequency-burst pattern on the TV screen of white to the extreme left followed by a low-frequency signal of 0.5 MHz, then 1.7 MHz and 3.4 MHz signal. The extreme right of the TV screen shows a 4.5-MHz burst frequency (see photo **D**).

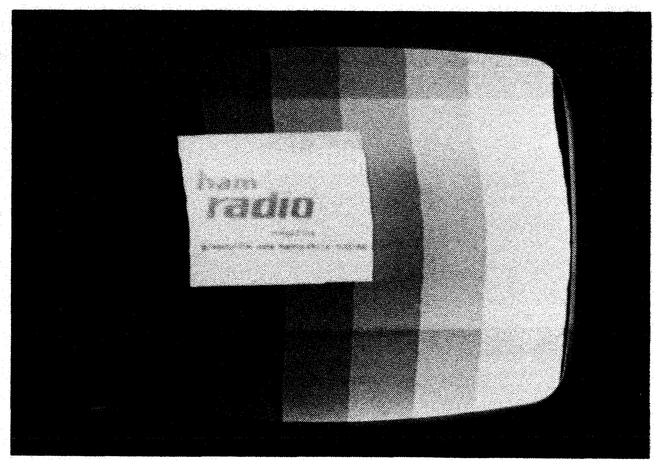
The multiburst generator can be used to check bandwidth of video equipment such as amplifiers, TV transmitters and tape recorders. It can also be used with oscilloscopes. Just insert the multiburst generator output into the input of the equipment under test and look at the results on the TV screen or on an oscilloscope (see photo **E**). Photo **F** shows high-frequency rolloff, and photo ***G*** shows the middle frequencies with too much gain. These test patterns depict only a few of the results, but photo E is the best example.

The second generator is a gray-scale (staircase) generator. It produces a pattern on the TV screen, as seen in photo H from left to right, in seven shades of gray. The staircase generator is used to check your video-system linearity. It shows if the white or black video is being compressed (see examples in photos **I** and **J**). If you insert the gray-scale generator into a video modulator or other video input, you should see a gray scale as in photo H. If you look at the gray-scale generator output on the oscilloscope, you'll see a staircase pattern as in photo K.

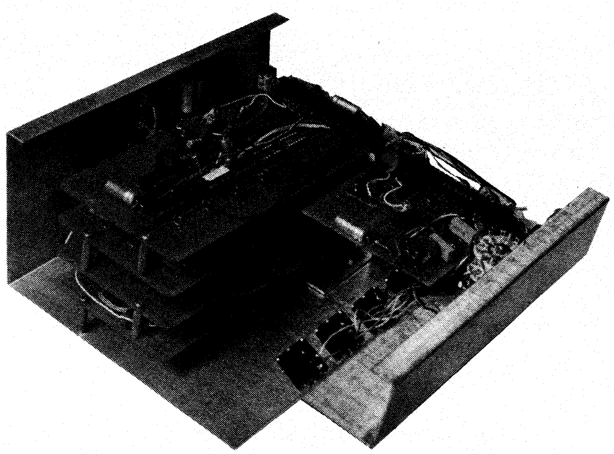
The third generator is called a "pulse and bar." It produces a white pulse on the left-half, and a white box in the right-half, of the TV screen. Its function is to check for high- and low-frequency bandpass characteristics. The pulse and bar generator output as seen on the TV screen is shown in photo L. On an

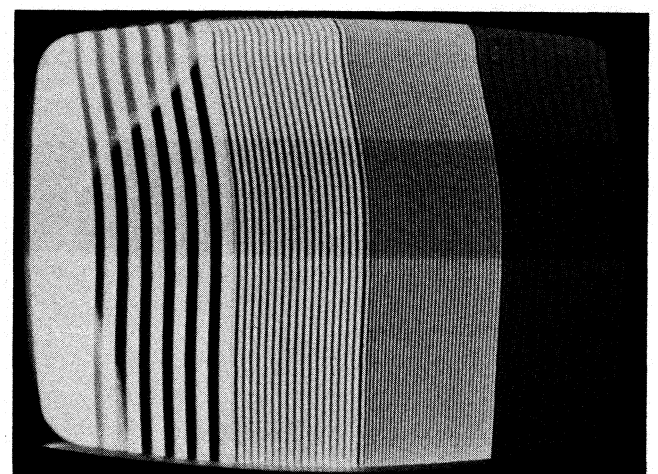
By Charles A. Beener, WB8LGA, 2548 SR 61, Marengo, Ohio 43334





Photos A and B: Special effects that can be generated by the video console. Photos by WB8HXR.





Photos C and D: Inside view of the video console, left. At right, multiburst generator output as seen on TV screen.

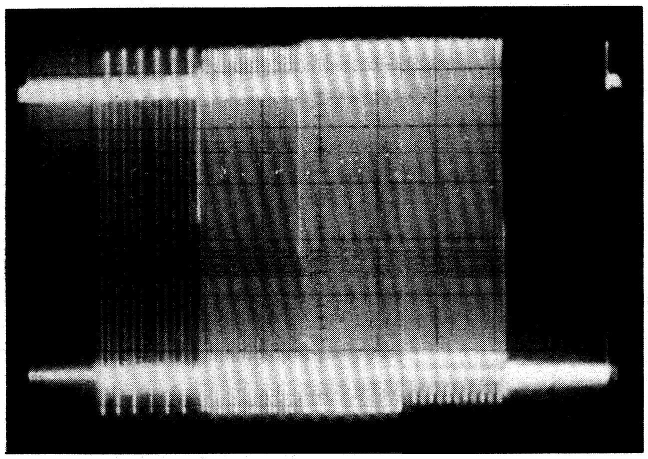
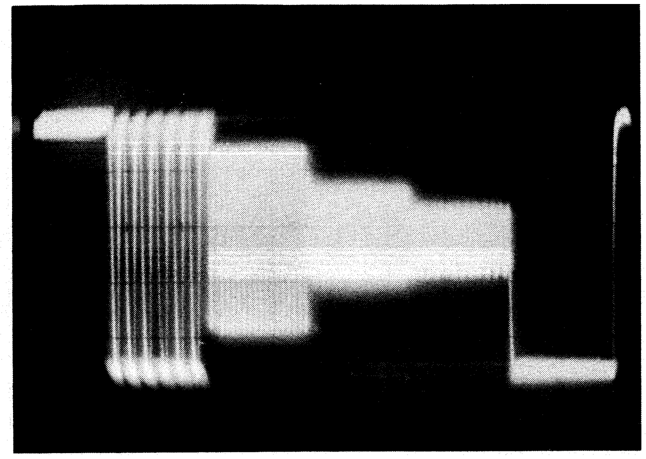
oscilloscope, you see a wave form as shown in photos M and N.

circuit description

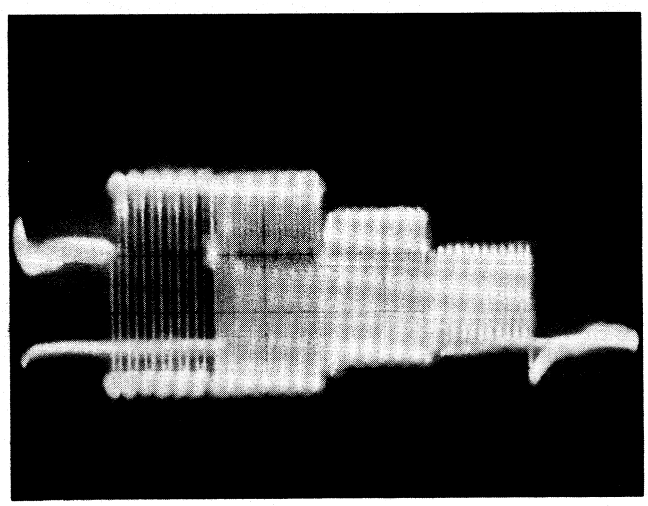
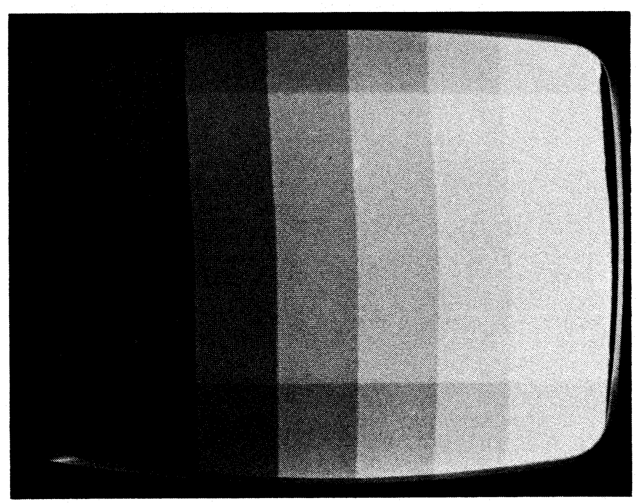
The following is a description of the five circuits that compose the video console. They consist of the multiburst generator (A1 board), gray-scale generator (A2 board), pulse and bar generator (A3 board), video switcher (A4 board), and Gen-Lock (A5 board).

Multiburst generator (**A1** board). This circuit is shown in fig. 1. Referring to fig. 1, U9 is a free-running oscillator whose frequency is 3.402 MHz. U9 drives U1 and U2, a divide-by-three counter with an output frequency of 1.134 MHz. U2A drives U2B, a divide-by-two counter whose output is 0.567 MHz. U2B output drives a divide-by-six counter, U3A, U3B, and U4, whose output frequency is 94.5 kHz. This signal drives a six-stage shift register that shifts the gating from left to right on the screen. The five H signals (fig. 1) drive the five NAND gates, U8A, B, C, D, and U9C. One section of U1A is fed to U1B, a divide-by-two counter with an output of 1.7 MHz. The same signal is fed to U8D. The output of U2A, 1.134 MHz, feeds U9A, which drives U9C. All U8 and U9C outputs are NANDed into U12. They are controlled by **H1**, H2, **H3, H4** and H5, and are turned on, one at a time, by the 6-stage shift register. The sum results in an output from U12 of the first high signal followed by a 0.567-MHz signal, 1.7 MHz-signal, 3.4-MHz signal, then a 4.5-MHz signa. This occurs in 54μs. This time, plus that of the horizontal sync pulse, is equal to 63.5 μs, which is the time required for one horizontal sync line on the TV screen.

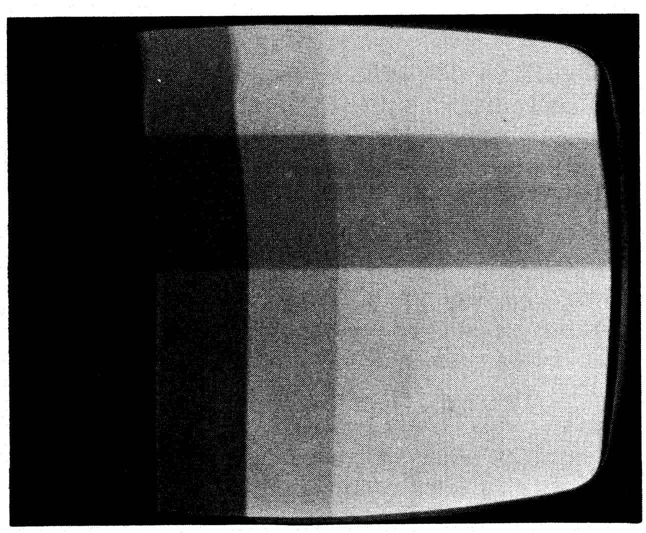
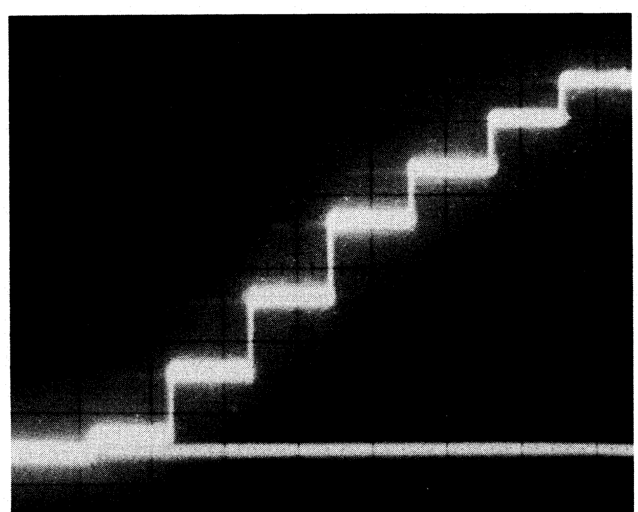
The sequence repeats until a vertical sync pulse appears on U10, pin 1. U10 is a one-shot. U10 provides a 1-ms pulse, which resets the 6-stage shift register at U5A, pin 4. U10 Q output (pin 1) is used for vertical blanking. The output of U7B, pin 6, is a horizontal timing pulse. It is NANDed at U10 pin 1 and is the vertical timing pulse for the composite sync. The

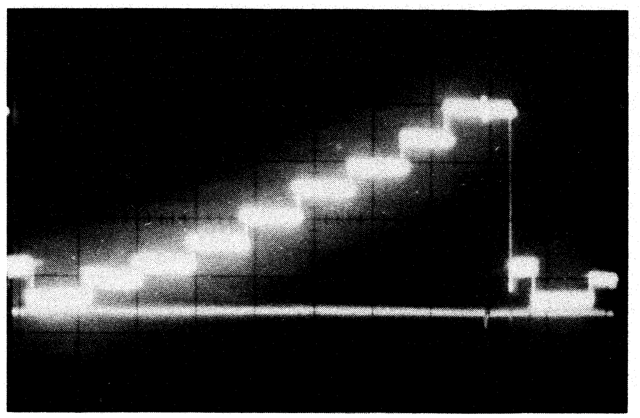
Photos E and F: Oscilloscope test pattern showing output of the multiburst generator, left. At right, multiburst generator test pattern showing high-frequency **rolloff.**

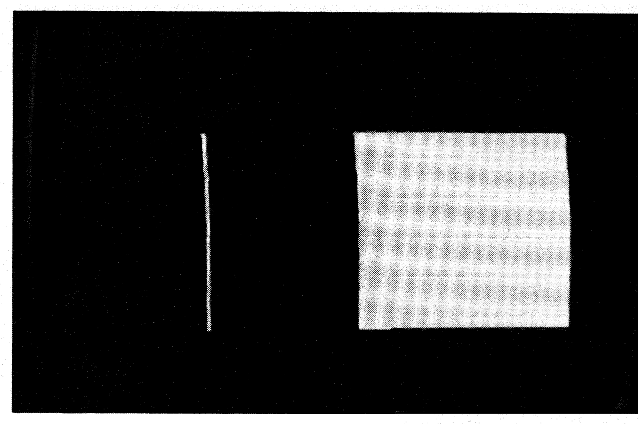



Photos G and H: Another test pattern of the multiburst generator showing excessive gain at the middle frequencies, left. At right, output of the gray-scale (staircase) generator as seen on the TV screen.

Photos I and J: Gray-scale generator output showing white video being compressed on TV screen, left. At right, oscilloscope test pattern showing gray-scale generator output. White video compressed.





Photos K and L: Gray-scale generator output shown on *oscilloscope. left. At:* right, pulse and bar generator output as seen on the TV screen.

composite video outputs appear at R15.

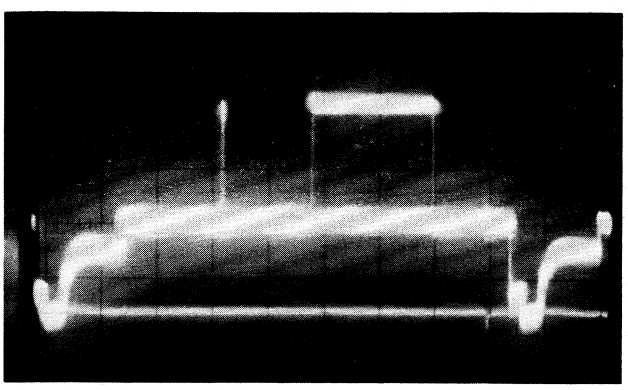
Gray-scale generator (A2 **board).** The main oscillator is U1 (**fig. 2**). Its frequency, 126 kHz, is determined by C1, R1. U1 output is fed to U2, a divide-by-eight counter. Counter output is NANDed at U1A, which drives U1B. This stabilizes the oscillator by the horizontal sync pulses on U2 pins 2 and 3. (U2 is a decade counter.) Each time the horizontal sync appears on U2 pins 2 and 3, the counter is reset to all zeroes. The three outputs from U2 also drive two hex inverters, whose output provides a staircase at the output. (See **photo K.**)

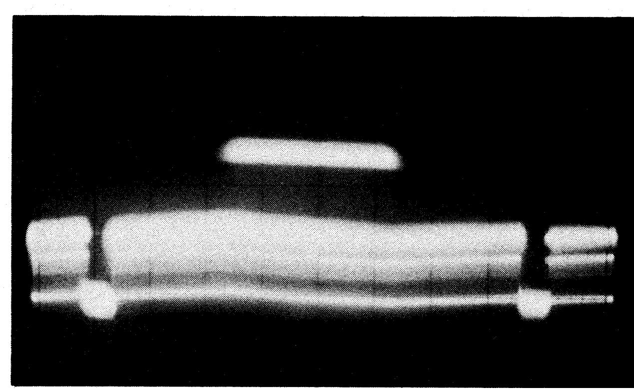
From pin 8 U2 drives U4, pins 3 and 4. U4 is a delay trigger, which drives U5. U5 provides an output 11-μs wide of horizontal sync. The sync and video are mixed by R9 and by R10, which is the video-output pot. The A2 board also contains the oscillators for the horizontal and vertical sync. These are common NE-555 timers, U6, and U7.

Pulse and bar generator (A3 board). The generator works on the principle of the one-shot delay, using 74121s. The horizontal sync feeds U3, U4, U5 **(fig. 2).** The time **delay is** from the horizontal **sync** pulse. The amount of delay is determined by the one-shot RC time constant. U3 provides the start of the bar; U4 provides the stop of the bar. U5 positions the pulse on the screen. U3 and U4 outputs drive U1A and U1B, a set and reset latch. This latch output is equal to the bar width.

The vertical sync triggers U6 and U7. U6 and U7 drive a set and reset latch U1C and U1D. U6 delay determines where the top of the pulse and bar signals start. The bottom of the pulse and bar signal is determined by U7. The output of U1D, pin 11, are NANDed together through U2A and U2C. The pulse and bar signals are mixed in R17. Output is obtained through R19. The horizontal sync signal is 4 μs wide and is resistor mixed to make the horizontal sync tips. The same input drives U8, a 11-μs-wide pulse for the horizontal sync front and back porch. It is also resistor mixed through R20.

Video switcher (A4 board). The switcher (fig. 3) works almost on the same principle as the pulse and bar generator. U1, U2, U3, and U4 perform the same job as U3, U4, U1, U6, and U7 in the pulse and bar generator. The outputs of U5, pins 6 and 11, are NANDed in U6A. U6A output drives U7B and U6B. U6B is an inverter that drives U7A. In other words,





Photos M and N: Pulse and bar generator output shown in one horizontal time period, left. At right, pulse and bar generator output shown in one vertical time period.

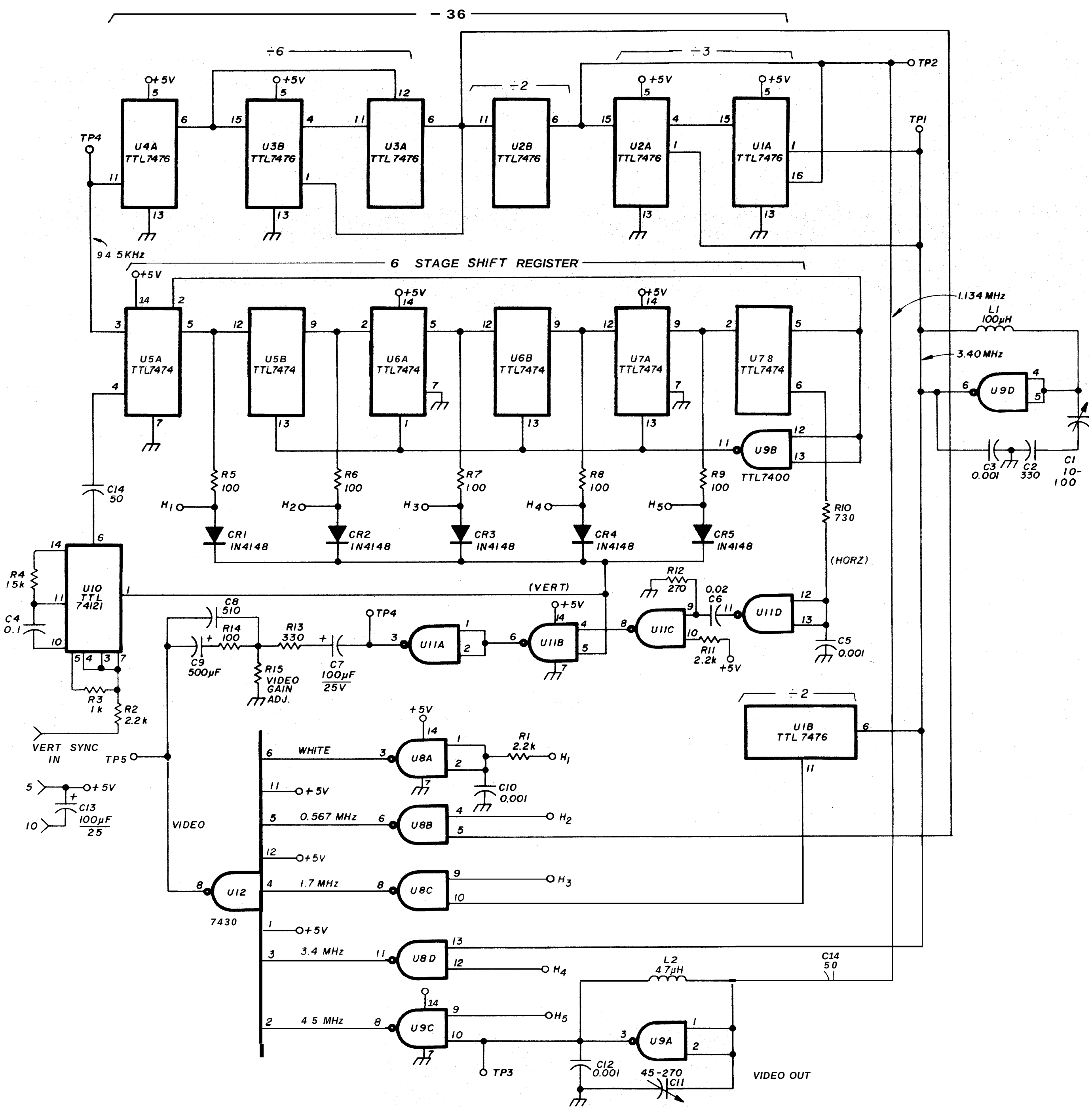
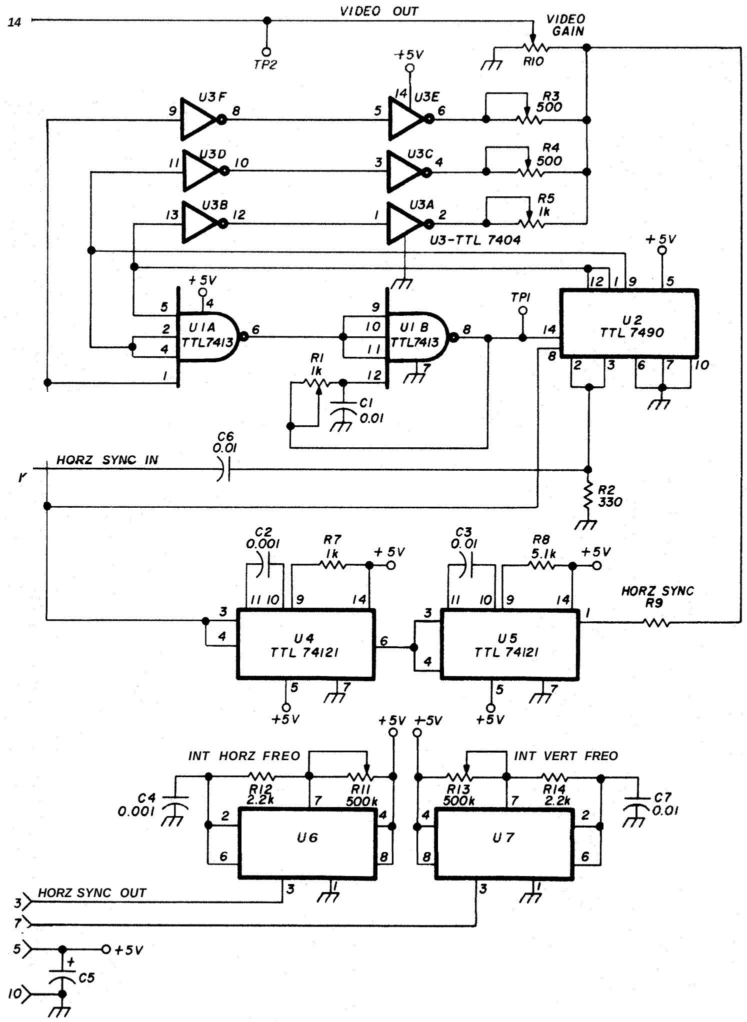


fig. 1. Multiburst generator schematic (A1 board).

only one section of U7 is on at any one time. Window-positioning pots R1-R4 control the vertical and horizontal start and stopping position of the switched video.

The video switcher, U7, is a CMOS IC. A low on pin 5 turns off the switcher between pins 3 and 4. If you put a high signal on pin 5 the switch turns on. Now it acts as if a 300-ohm resistor were placed between pins 3 and 4. (This CMOS IC is a CD 4016, but a CD 4066 can be used.) Switcher U7 outputs are tied together and run into a 74C04 inverter used in linear mode. The output of U7 is positive video to be amplified, so it was necessary to run it through two inverters. The output of a CMOS IC doesn't like a 100-ohm load, so it's buffered with an emitter follower, Q1.



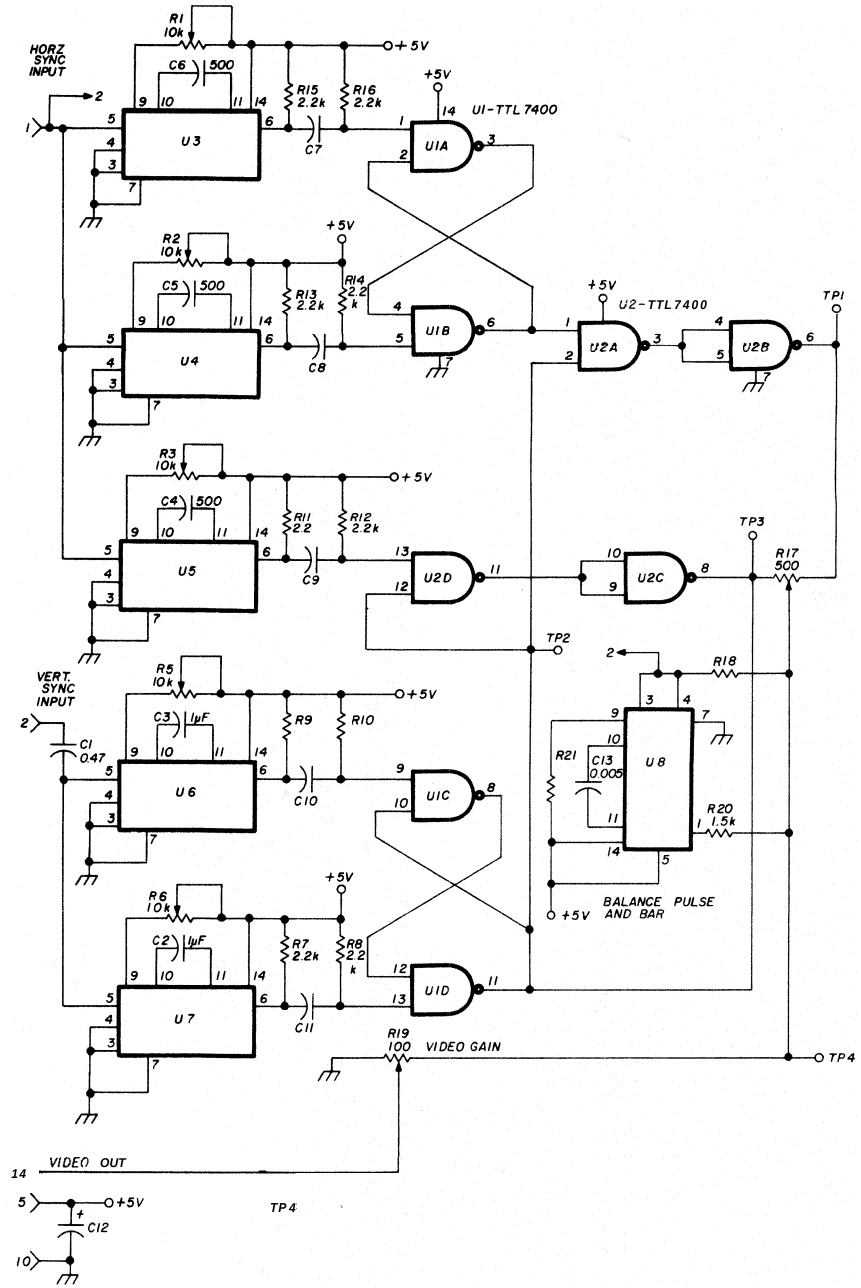


fig. 2. Above: Gray-scale generator schematic (A2 board). At right: Pulse and bar generator schematic (A3 board).

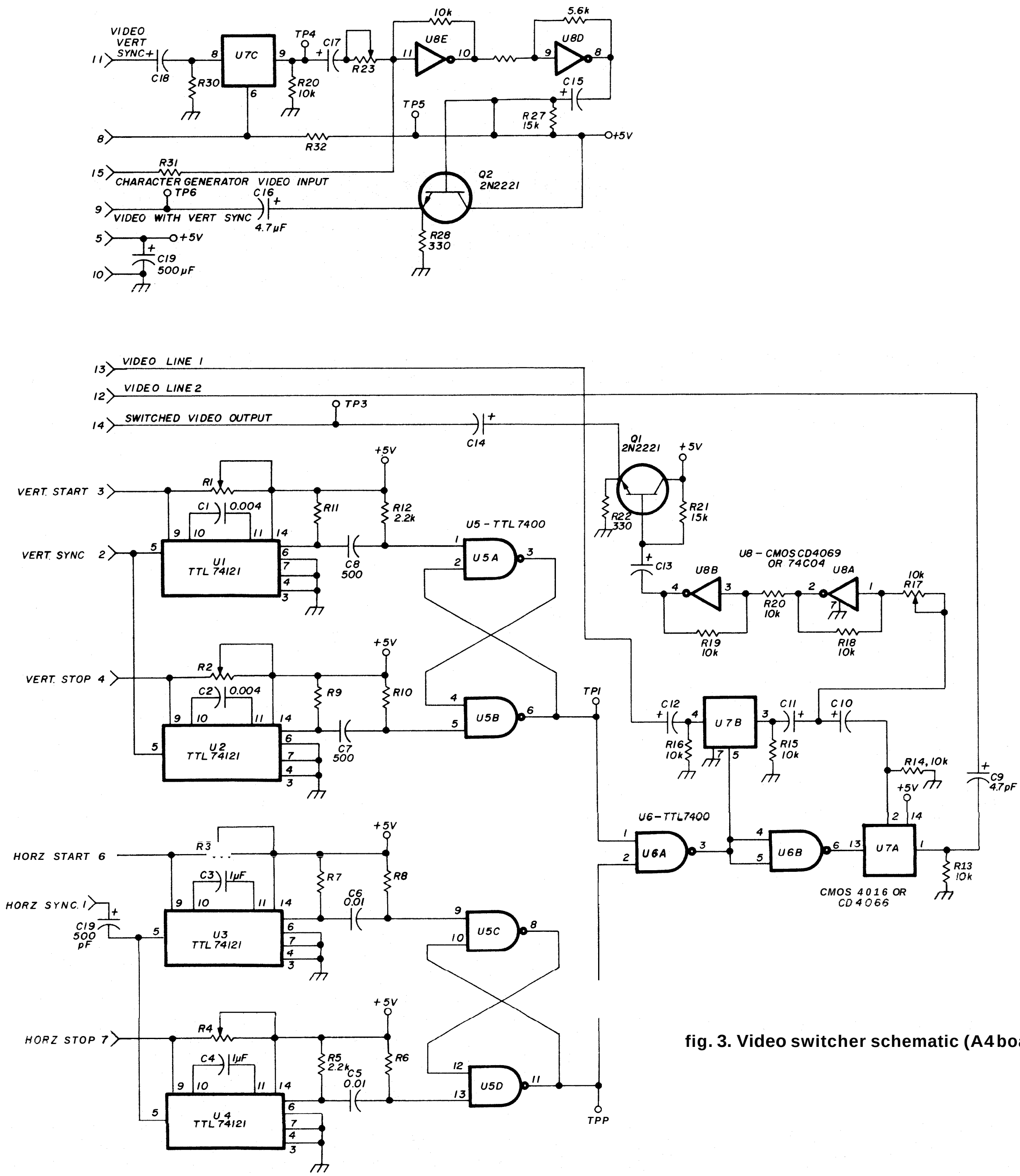


fig. 3. Video switcher schematic (A4 board).

The vertical sync blanker and dc restorer circuit are on the video switcher board. **U7C** is a switch controlled by the vertical sync pulse. Any video coming in at A4 pin 11 will be blanked during the vertical sync interval. After this, the video is reamplified through **U8** and is applied to emitter follower Q2. At the base of Q2, the vertical sync pulse is reinserted to take the vertical base line down to the correct reference. The character generator video input is inserted at **U8** pin 11, through R31.

Gen-Lock (A5 board). The video input enters on A5 pin 1 (see **fig. 4**). It drives 01, a sync stripper. 01 removes all the video, and only composite horizontal and vertical sync signals remain at the Q1 collector.

The sync signals are run through RC filter R4, C3, C4, R5. The vertical sync drives Q2, an amplifier, whose output is applied to the sync selector on the front panel.

The horizontal sync is applied to U1, a phase-locked loop. The IC is a PLL567, which is tuned for 15,750 kHz. This IC is a tone decoder. When it's locked onto an incoming signal, it has a dc output at pin 8. There's also a signal at U1 pin 5. This signal has the same frequency as the incoming signal during lock. U1 pin 5 output is then used as the horizontal sync in the generator and goes to the sync selector switch.

The sync-selector switch output is fed to amplifier Q3 for vertical; Q5 for horizontal. U2 is a Schmitt trigger for fast trigger timing. Q4 is a buffer for the vertical sync. Q4 output drives all the vertical and/or horizontal inputs of boards A1-A5.

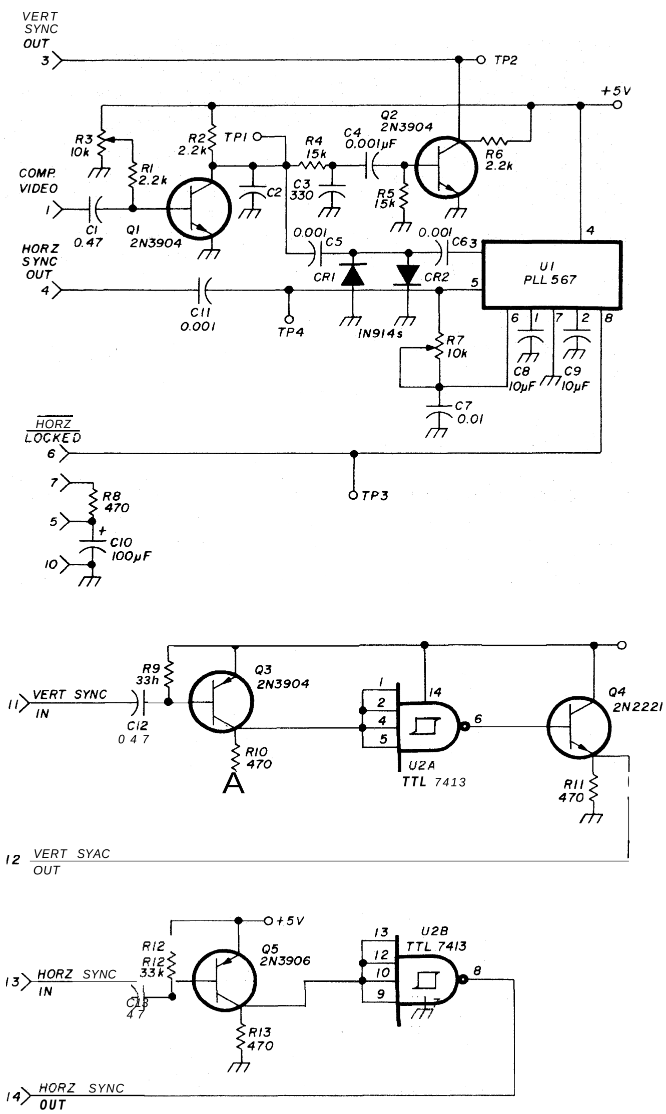


fig. 4. Gen-Lock schematic (A5 board).

interconnections

Fig. 5, the main board, shows how to tie it all together. **A** shows sync-selector switch connections. B shows how to interconnect the rotary function-generator switches to the five PC boards. **C** shows the interface between the window-position controls and video switcher/Gen-Lock boards.

adjustments

Muitiburst generator (A1 board).

1. Adjust C1 for an output frequency of 3.402 MHz at U9 pin 6 or for horizontal locked sync on TV when video is fed into the TV.

2. Adjust C11 for an output of 4.5 MHz on U9 pin 3.

3. Adjust R15 for video gain. It should be I-volt p-p at AI, pin 14.

Gray-scale generator (A2 board).

1. Adjust R1 for 126 kHz at test point 1, or for seven shades of gray on the TV screen.

2. Adjust R10 video gain for 1 volt p-p video at test point 2.

3. Adjust R I1 for 15,750 kHz at U6 pin 3.

4. Adjust R13 for 60 Hz at U7 pin 3.

5. Adjust R3 for approximately 75 ohms.

6. Adjust R4 for approximately 250 ohms.

7. Adjust R5 for approximately 600 ohms.

8. Adjust R3 = R5 for the best linearity of output video.

Pulse and bar generator (A3 board).

1. Adjust R1 for vertical position of left side of bar on TV screen.

2. Adjust R2 for vertical position of right side of bar on TV screen.

3. Adjust R3 for vertical position of pulse (see **photo L**).

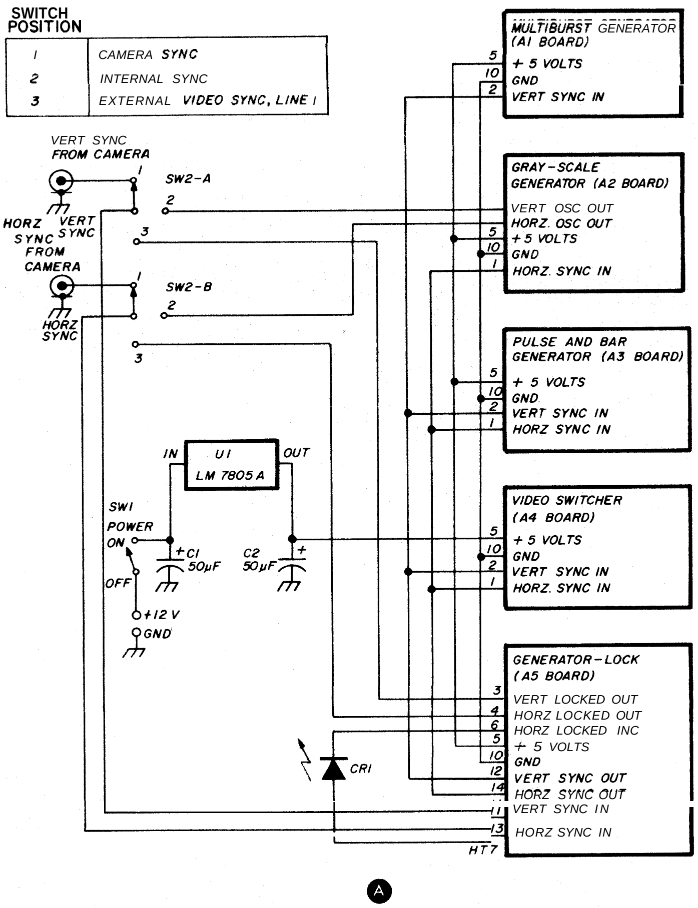
4. Adjust R5 for horizontal position of top part of pulse and bar.

5. Adjust R6 for horizontal position of top part of pulse and bar.

Video switcher (A4 board).

1. Adjust R23 for 1.5 volts p-p video at output of test point 6 when you have 1 volt at input.

2. Adjust R17 for 1 volt at test point 3. R1-R4 are front-panel controls but could be preset pots on board.



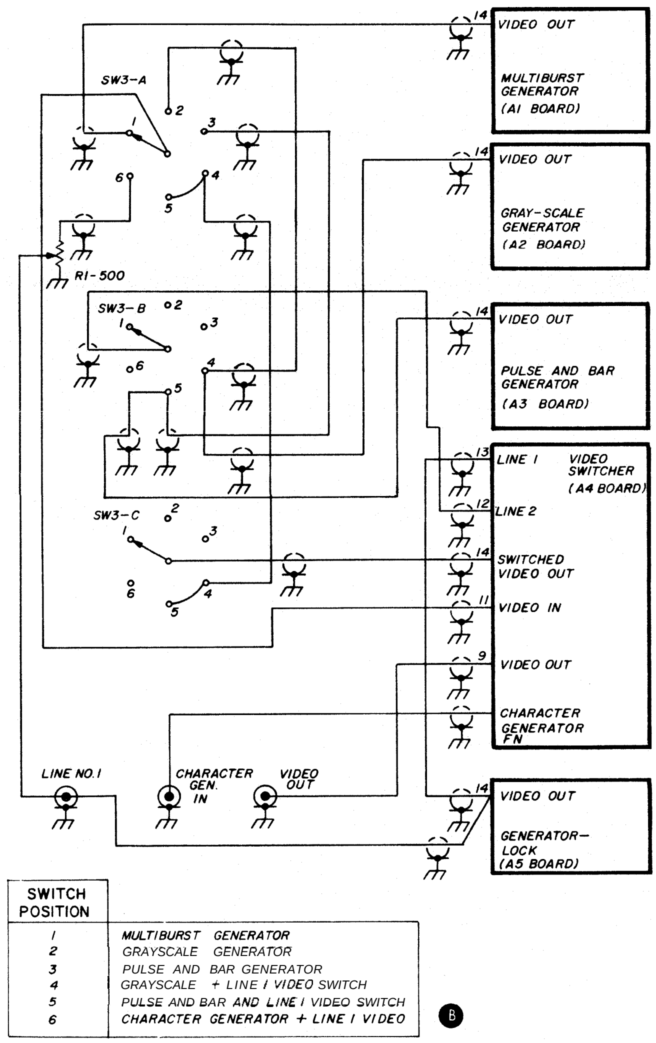
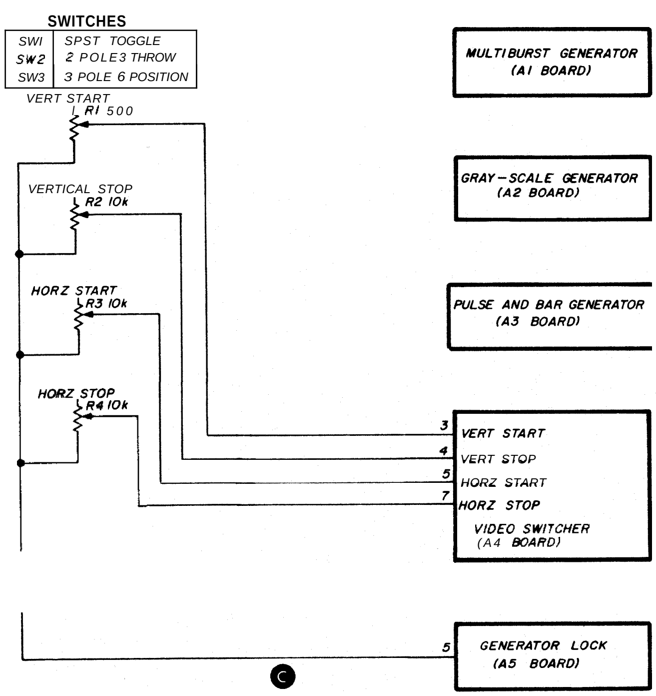


fig. 5. Circuit-board interconnections. (A) shows switching between horizontal and vertical sync signals from camera and the five PC boards. Function-generator switch logic is shown in (B). Interface between window-position controls and video switcher-Gen-Lock inputs is shown in (C).



Gen-Lock (A5 board).

1. Adjust R3 for clipping of video at test point 1

2. Adjust R7 for 15,750 kHz at test point 4 with no video input at A5 pin 1.

This video console was constructed on five PC boards, one for each generator. It was done this way for ease of construction and system checkout. You can use any one of the generators or switches by itself. Just insert the vertical and horizontal sync signals and the +5 volts to the board you want to use. I've had very good results with this console and a lot of fun in its construction and use.

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ham radio

Yagi antenna design:

performance calculations

Development of a mathematical model of the Yagi antenna for computing antenna gain, front-to-back ratio, and operational bandwidth

Since its invention by H. Yagi and S. Uda of Tohoku University in 1926,[1] the Yagi-Uda antenna – commonly referred to as the ***Yagi*** – has received a great deal of theoretical and experimental attention. The Yagi has also become the most widely used Amateur Radio communications antenna, not only because of its excellent performance characteristics over the rather narrow frequency bands occupied by Amateurs, but also because of its remarkable tolerance to construction variations and even construction faults; it is an antenna that "wants to work."

Those readers who are interested in the theoretical basis of Yagi linear array antennas should take a look at the excellent book published in 1954 by S. Uda and Y. Mushiake;[2] this book is highly recommended and contains sixty-six relevant references. There are several other excellent reference works on the Yagi including those by Kraus,[3] King,[4] and Walkinshaw;[5] for those interested in the experimental side of Yagi arrays, attention is drawn to Ehrenspeck and Poehler,[6] Lindsay,[7] Greenblum,[8] and Viezbicke.[9] Reference 9, National Bureau of Standards ***Technical Note*** 688, describes the results of a decade-long NBS experimental investigation of Yagi antennas. Although this study represents one of the most complete and relevant sources of information on the Yagi, the report is flawed by inconsistencies, incomplete experimental information, and above all, measurement techniques which are sensitive to the effective height of ground midway between the transmitter and receiver. Nevertheless, the rich range of experimental results in NBS ***Technical Note*** 688 provide an excellent area to test the validity of theoretical ideas.

It appears to me that designers of a high-performance Yagi array are faced with four facts:

1. Accurate antenna experiments are very difficult to make, especially if the antennas are designed to be used over earth. There are many variables, and it is difficult (if not impossible) to avoid unwanted reflections. Also, accurate instrumentation is simply not available for many of the quantities to be measured (current distribution in the parasitic elements, for example).

2. Conceptual design ideas are often misleading: *e.g.*, the concept of "optimum element spacing." Optimum, indeed, but with respect to what?

3. Practical design of real antenna components in some cases does not exist; a physically "tapered" element, for example, must be significantly longer than an equivalent cylindrical element, but by how much?

4. A good theoretical basis for design is uncertain. First of all, a real antenna must be simulated by a simplified physical model; the accuracy of the final result depends crucially on the excellence of this physical simulation or model. Secondly, the physical model, in conjunction with accepted physical laws and a modern computer, is used to compute the electrical performance of the Yagi model, *i.e.*, the parasitic element (parasite) currents first, and the spatial radiation pattern second. Since both computations involve simplifying approximations, overall accuracy depends directly on the excellence of physical modeling and the accuracy of the necessary mathematical approximations which are made.

I shall attempt to address each of these four items over the next several months. I have broken the subject down into six major categories, listed below, which will be covered in separate articles.

By James L. Lawson, W2PV, 2532 Troy Road, Schenectady, New York 12309

1. Modeling and computational methodology

2. Simple Yagis; free-space performance

3. Earth (ground) effects

4. Preferred designs

5. Scaling and construction corrections; specific designs

6. Stacking, screening, and scattering

I shall address the subject primarily in terms of antenna arrays which are most useful in the frequency range from 5 to 30 MHz, including designs ranging in size up to the largest practical levels. "Conventional" Yagi components will generally be used, but there will be some discussion of quads and of the quad-Yagi hybrid known as the *quagi.* For convenience, I will use horizontal polarization unless otherwise stated.

antenna properties

Before beginning an investigation of antenna characteristics it is necessary to define design criteria; what antenna properties are important and how can those properties be defined in quantitative terms? The antenna *user* is concerned with several properties:

1. Antenna gain G or directivity

2. Pattern (including front-to-back ratio, F/B)

3. Bandwidth

4. Feedline matching or standing-wave ratio

5. Cost

6. Longevity (wind survival, corrosion resistance, etc.)

Of these, the first four are *electrical* properties; the last two depend basically on construction engineering and will not be discussed further. Feed-line match (item 4) is controlled, at one frequency at least, by the matching system which transforms the antenna driving-point impedance to the transmission line's characteristic impedance. We are interested in the *inherent* driving point bandwidth of the antenna, specifically the actual driving-point impedance and its behavior with frequency.

Antenna gain, pattern, and bandwidth (items 1, 2, and 3) must be defined rather carefully. The gain (and directivity) is clearly of paramount importance; I shall use the definition of isotropic gain for all situations, *i.e.*, the ratio of *maximum* radiated energy flux density (at the "best" elevation angle) to the average radiated energy flux density (averaged over the 4π solid angle). This is equivalent to using an isotropic reference antenna in free space.

While the antenna's complete spatial pattern is of concern to the user, it is basically not practical to measure it or to specify it. The pattern characteristics of primary interest are the angular widths (horizontal and vertical) of the main beam and the amount of back radiation. The beam widths are rather accurately and simply determined by the antenna gain or directivity (see Kraus, page 25[3]).

Radiation in the rear hemisphere is usually variable and consists of one or more lobes. Perhaps the single most meaningful measure of rear radiation is the front-to-back ratio, or F/B, but we must define where the front wave is to be measured (elevation angle), where the back wave is measured, and what property of the waves is used to obtain the ratio. One popular concept is to plot the field strength, E, pattern of the antenna in free space and compute the ratio of front (maximum) field strength, E_F, to that found in the reverse (back) direction, E_R. The ratio can be also expressed in decibels as $(20 \log_{10} E_F/E_R)$.

What is the front-to-back ratio of a *real* antenna over earth or ground? I define this quantity as the ratio of maximum front power or energy flux density, W_F, (at an elevation angle where this maximum occurs) to the rear power or energy flux density, W_R, at the *same* rear elevation angle. This can also be expressed in decibels as $10 \log_{10} W_F/W_R$.

Note that this F/B ratio is only one parameter of the complete antenna pattern. It is perfectly possible, in principle, to have an antenna with a large F/B ratio defined in this way, but which has serious backward (but not directly back) lobes. Nevertheless, in the interest of simplicity I shall retain this simple notion of F/B ratio; it is perhaps the most important single index of pattern behavior.

The last parameter is the frequency bandwidth, but there are at least three important bandwidths: 1) the bandwidth of the driving-point impedance (electrical input), for example; 2) the bandwidth over which the gain remains high; and 3) the bandwidth over which the F/B ratio remains high. All three are important, so it's necessary to measure or calculate all three. This is best done by observing a quantitative frequency-swept plot of them all.

computation vs experiment

Antenna characteristics can be determined either by experimental measurements on a physical model or by calculations from a mathematical representation of the physical model. Which should be used? Experimental measurements are laborious and it is difficult to ensure accuracy. By contrast, computer calculations are fast and, in principle, can be made with great accuracy. Using a modern computer, a large number of antenna configurations can be investigated in a few days; a number it would take a lifetime to explore experimentally! Moreover, because of the inherent accuracy of calculations, subtle changes and radiative coupling effects can be explored. Therefore, it appears that if a computational procedure is believable, it can be used *very* effectively. Careful experimental tests *are* needed, however, to validate the computation methods.

modeling

A real Yagi antenna can be represented physically by a set of parallel cylindrical conducting elements, each of which has space coordinates at its center of ***X,*** Y, and *Z*. As shown in **fig. 1,** the Yagi will be oriented so that the eiements lie paraiiei to the Z axis and the boom is parallel to the ***X*** axis. In free space, the origin of this coordinate system can be placed anywhere, but when modeling a Yagi over the earth it will be advantageous to locate the origin on the conducting earth plane.

The elements themselves approximate a half-wave in length; the reflector, ***R,*** is usually somewhat longer than a half-wavelength, the drive element, ***DR,*** is normally about a half-wavelength, and the directors, *D1* and *D2* (or more), are somewhat shorter than a half-wavelength. For convenience, and to make the representation independent of wavelength, all coordinates, lengths, and dimensions will be expressed in wavelengths at some chosen design center frequency, f_o.

This model is clearly a simplification of a real Yagi. A real Yagi, for example, does not ordinarily have strictly cylindrical elements; real elements usually have telescoping diameters with connecting hardware clamps; moreover, the elements are mounted at their centers with plates or brackets to the boom, which is usually a conductor.

In this mathematical model I have totally neglected

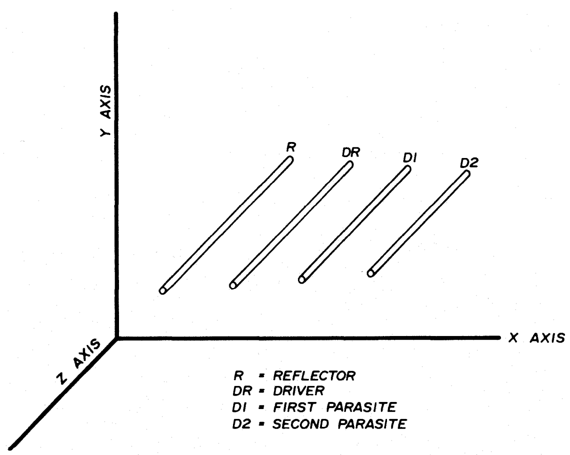

fig. 1. Basic model of the Yagi antenna showing the relationship of the elements, which are parallel with the Z axis: the boom is parallel to the *X* axis.

the conducting boom; this is justified only if the real Yagi is completely symmetric around the boom. Symmetry guarantees the electrical potential of each parasitic element at its center to be zero and guarantees no mutual coupling to the boom – hence no current will flow along the boom, and the boom can be neglected.

We also assume the driven element to be open at its center and driven from a baianced source; this eliminates current along the boom. The real Yagi driven element, if fed from a balanced source, will also induce no boom current. However, unbalanced feed systems can, in principle, produce currents aiong the boom which are not considered in the model. Unbalanced feed systems such as the gamma match are usually not especially troublesome in this respect, because if the driving point impedance of the element is relatively low, as it usually is, the voltage impressed on the boom is also low and boom current will be correspondingly low. Moreover, the loaded *Q* of the driven element is usually high enough to insure reasonable symmetry of element currents; this also makes for low induced boom currents. Incidentally, it is possible to construct the real Yagi with insulating element-to-boom supports; this helps to ensure negligible boom currents.

The clamping and mounting hardware used in a practical Yagi all amount to corrections in the *actual length* of the element to an *equivalent length.* Similarly, the telescoping (radius tapering) element will act like a cylindrical element of the same *average* radius, but with a different *equivalent* length! The way in which the actual element dimensions can be converted to a cylindrical element of the same average radius and equivalent length will be discussed later, as will corrections to length caused by mounting hardware. For the moment, note that the real Yagi element dimensions can be converted to equivalent cylinder dimensions for use in the model. As a side note, the element radius taper corrections to convert actual lengths to equivalent cylinder lengths are substantial; this is often overlooked by builders who try to use someone else's element lengths with different element diameters on tapering.

In the mathematical model you can arrange any number of parasitic elements and any number of drivers. It will be shown later how to use the model to approximate a quad or a quagi, or even a "broadband" drive (as in the KLM antennas) where there is a main driver but one or more dependent drivers which are connected through a transmission line to the main driver.

computational methodology

With this mathematical model we are in a position to compute the performance of the Yagi array — to

table 1. Element reactance for different length-to-diameter ratios, K. $X_{0.50}$ is the reactance of an element 0.5λ long; $X_{0.45}$ denotes the reactance of an element 0.45λ long; ΔX is the reactance change when shortening the element from 0.5 to 0.45 wavelength. As shown in fig. 2, a plot of ΔX vs K falls on a straight line defined closely by eq. 1.

K	$X_{0.50}$	$X_{0.45}$	ΔX	eq. 1
10	34.1799	21,6094	– 12.5705	– 11.030
30	36.7352	4.6906	– 31.0446	– 31.561
50	37.4666	– 3.8791	– 41.3457	–41.107
100	38.2110	– 15.8869	– 54.0979	– 54.060
200	38.7673	– 28.1857	– 66.9530	– 67,013
10^3	39.6375	– 57.3954	– 97.0329	– 97.090
10^4	40.3603	–99.9997	140.3600	–140.120
10^5	40.7961	– 143.0558	– 183.8519	– 183.150
10^6	41.0873	–186.3402	–227.4275	–226.180

find out how the performance of the array varies with frequency f around the design f_o. The first task is to compute the complex currents (or current magnitudes and phases) which flow in all elements as a result of driver excitation; to do this, it is necessary to determine both the self and mutual impedances of all elements.

I shall start with the behavior of a single nearly half-wavelength element in free space. The self impedance of such an element has been calculated by many authors; an excellent comparison of the various methods is given by Kraus[3] (pages 272-276). Uda and Mushiake have used the method originated by Hallen (boundary-value problem) and presented an approximate equation and a table (pages 23-24) which show the self impedance of a cylindrical nearly half-wavelength doublet as functions of radius and length. It is apparent from this table that a half-wavelength doublet has an impedance of about 73 + j40 ohms, so a somewhat shortened antenna is needed to resonate (to show zero reactance). The required

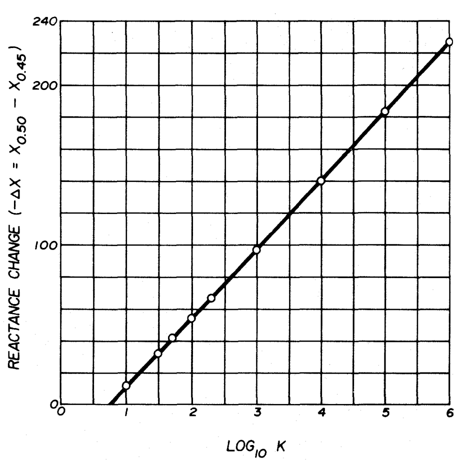

fig. 2. Graph showing the relationship between the length-to-diameter ratio, K, of a half-wavelength element and the reactance change. ΔX, as element length is reduced to 0.45λ from 0.51.

shortening is basically only a function of K, which is defined as the ratio of (central) wavelength to element radius. In **table 1**, I have extended Uda and Mushiake's data to a wider range of K values and have calculated the reactance $X_{0.50}$ (in ohms) of the full half-wave dipole and $X_{0.45}$ of an element 0.45 wavelength long; ΔX is the reactance *change* in ohms when shortening the element from 0.5 to 0.45 wavelength.

If you plot ΔX against log_{10} **K**, you will find, remarkably, that the points fall in a straight line (fig. **2**). This suggests that the reactance change ΔX can be expressed with rather good accuracy as:

$$\Delta X = -43.03 \log_{10} K + 32 \quad \textbf{(1)}$$

The accuracy of this empiricai relationship is remarkable over the range of K of real interest ($100 < K < 10,000$).

Note that a simple series-resonant circuit displays an input reactance of:

$$X = 2RQ\,(F/FR - 1) \quad \textbf{(2)}$$

where Q is the electrical Qfactor ($Q = 2\pi f_o L/R$)

F is the frequency relative to f_o

FR is the resonant frequency of the circuit relative to f_o

R is the series resistance

Equations 1 and **2** can be used to derive:

$$RQ = (215.15 \log_{10} K - 160) \quad \textbf{(3)}$$

In other words, the dipole element behaves electrically like a series-resonant circuit of resistance R and Qgiven by the empirical relationship of **eq.** 3.

You may also use **eq. 1** to derive the resonant length (zero reactance), LER, of the dipole element in units of wavelength:

$$\text{LER} = [1 - (10.7575 \log_{10} K - 8.00)^{-1}] \quad \textbf{(4)}$$

It is interesting to note that the ratio of the resonant length to a true half-wavelength depends only on K! Some representative values computed from **eq.** 4 are shown in **table 2** and graphically illustrated in fig. 3.

table 2. Element resonant length $2 \cdot LER$ (in units of $\lambda/2$) for different length-to-diameter ratios K. These data are plotted in fig. 3.

K	2•LER	K	2•LER
30	0.8733	10^3	0.9588
50	0.9027	10^4	0.9715
100	0.9260	10^5	0.9782
200	0.9403	10^6	0.9823

Note that greater shortening is needed for thick cylinders than for thin wires, but even for very thin wires appreciable shortening is required to achieve resonance. Thus, if the actual length of an element is *LE* in terms of wavelengths at central design frequency f_o, and its resonant length is *LER,* its resonant frequency, *FR* (in terms off,), may be written as:

$$FR = LER/LE \qquad (5)$$

The self impedance of such an element is then:

$$R+jX = 73+j(430.30 \log_{10} K - 320)(F/FR - 1) \text{ ohms} \qquad (6)$$

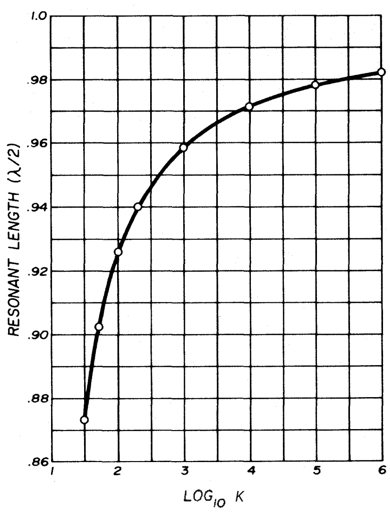

fig. 3. Resonant length of a half-wavelength element as a function of length-to-diameter ratio, *K*. Note that greater shortening is needed for thick elements (low *K*) but even very thin wires require appreciable shortening to achieve resonance.

The accuracy of this expression should be good to a very few per cent for elements within a few per cent of a half-wavelength long.

mutual impedance

Now must be considered the mutual impedance between two nearly half-wavelength elements separated by a distance *s* measured in wavelength λ. Good calculations have been made by several authors of both the real and imaginary components of this complex quantity, but only for the limiting case of infinitely thin, half-wavelength elements; equations, plots, and tables are shown in Kraus[3] (pages 265-268) and Uda-Mushiake* (pages 69-70). Kraus (page 266) also shows calculations by Tai for two cases of thicker, half-wavelength dipole elements. Tai's calculations suggest that inaccuracies caused by using the limiting thin case will not be large; therefore, for convenience, I use it for all calculations. I have extended the table of Uda-Mushiake, and in table 3 show values in ohms for the real part of the mutual impedance (*RMUT*) and for the imaginary part (*XMUT*), as a function of element separation, *s*, in wavelength. For separations greater than $s = 1.5$, a reasonable approximation can be derived from:

$$RMUT = \frac{19.06}{s} \sin(2\pi s) \text{ for } s > 1.5 \qquad (7)$$

$$XMUT = \frac{19.06}{s} \cos(2\pi s) \text{ for } s > 1.5 \qquad (8)$$

Although some caveats are necessary, you now have the necessary tools to calculate the parasitic element currents. Recall that the physical model of the Yagi is a good representation only to the extent that proper corrections can be made for element hardware variances (clamping and mounting hardware and element radius taper). These corrections will be detailed later. Also recall that the computation of self and mutual impedances (**eqs. 4, 5, 6, 7, 8,** and table **3**) are approximations! Though they are probably good approximations, and should give reasonably accurate results, you should not rely on them for accuracy better than a few per cent.*

computations

The first step is to calculate the complex currents (or magnitudes and phases) of all parasitic elements given the currents or voltages applied to all drivers. The method is uncomplicated, following the technique of **P.S.** Carter shown in Kraus,[3] (page 302). For simpiicity, i wiii iiiustrate how this is done using one driver and three parasitic elements; extension to any number of elements will be obvious. For each element add all voltages and equate to the terminal voltage V_n:

$$\begin{aligned} I_1Z_{11}+I_2Z_{12}+I_3Z_{13}+I_4Z_{14} &= V_1 \\ I_1Z_{21}+I_2Z_{22}+I_3Z_{23}+I_4Z_{24} &= V_2 \\ I_1Z_{31}+I_2Z_{32}+I_3Z_{33}+I_4Z_{34} &= V_3 \\ I_1Z_{41}+I_2Z_{42}+I_3Z_{43}+I_4Z_{44} &= V_4 \end{aligned} \qquad (9A)$$

"Any improvement in these approximations will require a great amount of theoretical work through a rigorous examination of the boundary value problem with attention to:

1. Current distributions along driven and parasitic elements (they are somewhat different in principle)
2. Complete numerical solutions to both real and imaginary components of element self and mutual impedance; it will be necessary to distinguish mutual coupling coefficients between elements of different function, *i.e.,* driven to driven, driven to parasite, and parasite to parasite. In principle, all coefficients will depend on each affected element length and radius.

Assume in this example that the first three elements are parasites, *i.e.*, $V_1 = V_2 = V_3 = 0$ and that the fourth element is driven with the complex voltage V_4; Z_{nn} is recognized as the complex self impedance of the nth element, and Z_{jk} (which is the same as Z_{kj}) is the mutual impedance between the jth and kth element. Thus, all of these impedances can be calculated once the positions (and hence separations) of the elements are specified. Since there are four linear equations with four unknowns, I_1, I_2, I_3,

table 3. Complex mutual impedance as a function of element spacing. s (wavelength). The real part is designated R_{MUT} and the imaginary part as X_{MUT} (both in ohms)

s(λ)	R_{MUT}	X_{MUT}	s(λ)	R_{MUT}	X_{MUT}
0.	73.13	42.54	0.80	−18.49	12.26
0.05	71.66	24.27	0.85	−13.32	16.29
0.10	67.33	7.54	0.90	−7.49	18.55
0.15	60.43	−7.10	0.95	−1.55	18.99
0.20	51.40	−19.17	1.00	4.01	17.74
0.25	40.79	−28.35	1.05	8.75	15.04
0.30	29.26	−34.44	1.10	12.32	11.22
0.35	17.50	−37.42	1.15	14.52	6.71
0.40	6.22	−37.43	1.20	15.25	1.94
0.45	−3.97	−34.78	1.25	14.56	−2.66
0.50	−12.53	−29.93	1.30	12.59	−6.70
0.55	−19.06	−23.42	1.35	9.62	−9.84
0.60	−23.31	−15.87	1.40	5.97	−11.88
0.65	−25.21	−7.94	1.45	2.01	−12.70
0.70	−24.86	−0.25	1.50	−1.89	−12.30
0.75	−22.50	6.63			

and I_4, the easiest way to solve this array is by a matrix inversion. In matrix notation **eq. 9A** is represented by:

$$\begin{vmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} \\ Z_{21} & Z_{22} & Z_{23} & Z_{24} \\ Z_{31} & Z_{32} & Z_{33} & Z_{34} \\ Z_{41} & Z_{42} & Z_{43} & Z_{44} \end{vmatrix} \times \begin{vmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{vmatrix} = \begin{vmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{vmatrix} \qquad \textbf{(9B)}$$

or: Z (matrix) x I (vector) = V (vector) where all terms are complex.

The solution is I (vector) $= Z^{-1}$ (inverted matrix) $\times$ V (vector) where Z x $Z^{-1} = 1$.

The process of matrix inversion is readily accomplished with a computer using matrices having complex terms, *e.g.*, with a program usually called CLINEQ under FORTRAN IV. Although the actual solution is usually done through a mathematical process known as Gaussian elimination, the result is equivalent to matrix inversion. With this technique, a corn-puter will provide solutions quickly for very large arrays of fifty elements or more.

If you wish to specify the driven element *current*, I_4, instead of voltage, V_4, rewrite the first three parasitic equations (**eq.** 9) as:

$$\begin{aligned} I_1Z_{11} + I_2Z_{12} + I_3Z_{13} &= -I_4Z_{14} \\ I_1Z_{21} + I_2Z_{22} + I_3Z_{23} &= -I_4Z_{24} \\ I_1Z_{31} + I_2Z_{32} + I_3Z_{33} &= -I_4Z_{34} \end{aligned} \qquad \textbf{(10)}$$

This (smaller) array can be solved in the same way for I_1, I_2, and I_3 and the results used in the fourth equation:

$$I_1Z_{41} + I_2Z_{42} + I_3Z_{43} + I_4Z_{44} = V_4 \qquad \textbf{(11)}$$

to solve for V_4. The driving point impedance for either calculation is simply:

$$Z_4 = V_4/I_4 \qquad \textbf{(12)}$$

This procedure is best done on a computer, and it is not difficult to write a suitable program. Although I have written a number of such programs in FORTRAN IV, I would prefer not to supply them. It has been my experience that those who understand programming can easily write their own; those who are not capable of programming invariably need substantial assistance, which I am unwilling to supply.

summary

In this article I have discussed the basic properties of the Yagi antenna, and the construction of a mathematical model which can be used for computer analysis. In the next article of this series I will outline the computer programs which accomplish this analysis, and will confirm, using published experimental information, that calculated Yagi performance is in close agreement with that realized in practice.

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ham radio

remote synthesized fm transceiver for 2 meters

Construction details for a remote synthesized 2-meter transceiver with 220-MHz control

Activity on the Amateur 220-MHz band has always been on the sparse side, although in recent years more and more fm operators in the urban areas have been moving up to 220 to avoid the congestion of 2 meters. For a number of years we were seriously threatened with the loss of 220-MHz to the CB interests, but that threat seems to have passed for the moment; other services still covet the 220-MHz allocation, however, and one of the best ways to protect the band is to increase the activity level.

With a band as good as 220 MHz, I used to wonder why it was not being used to its fullest; I concluded that few Amateurs really "know" about 220, so there is little demand for good equipment, and, consequently, there is a limited selection. Just by way of introduction, the 220-MHz band has a number of characteristics to recommend it: it provides propagation similar to that of 144 MHz, but the noise levels are considerably lower; 220 is wider than 2 meters by 1 MHz, and 220-MHz antennas require less space for the same amount of gain.

After I had sold myself on the advantages of 220-MHz, I decided to see what I could do to generate more interest in the Atlanta area. I first put up a 220-MHz repeater; no one, however, seemed very interested. I had considered adding an autopatch when I got an idea for a remote base. Not just a fixed or few selectable channel remote base, but one which was fully synthesized and could be user programmed to any frequency in the 2-meter band.

features and limitations

I decided to make use of the excellent transmitter, receiver, and synthesizer kits manufactured by VHF Engineering and to develop only the necessary control circuits to operate the synthesizer and associated circuits. I felt it should be possible to operate any frequency combination between 144.105 and 147.995 MHz; it would be necessary to dial up the receive freqency prior to the transmit frequency so a listening watch could be made prior to transmitting (later it would be possible to dial up receive frequencies without transmit frequencies so the band can be scanned manually if desired).

theory of operation

Fig. 1 is a functional diagram of the overall system starting with a basic 220-MHz repeater. The audio from the repeater receiver is connected to a tone-pad decoder which feeds a BCD data bus; this bus carries data from the operator to three circuit boards.

The first, the access and control board senses the

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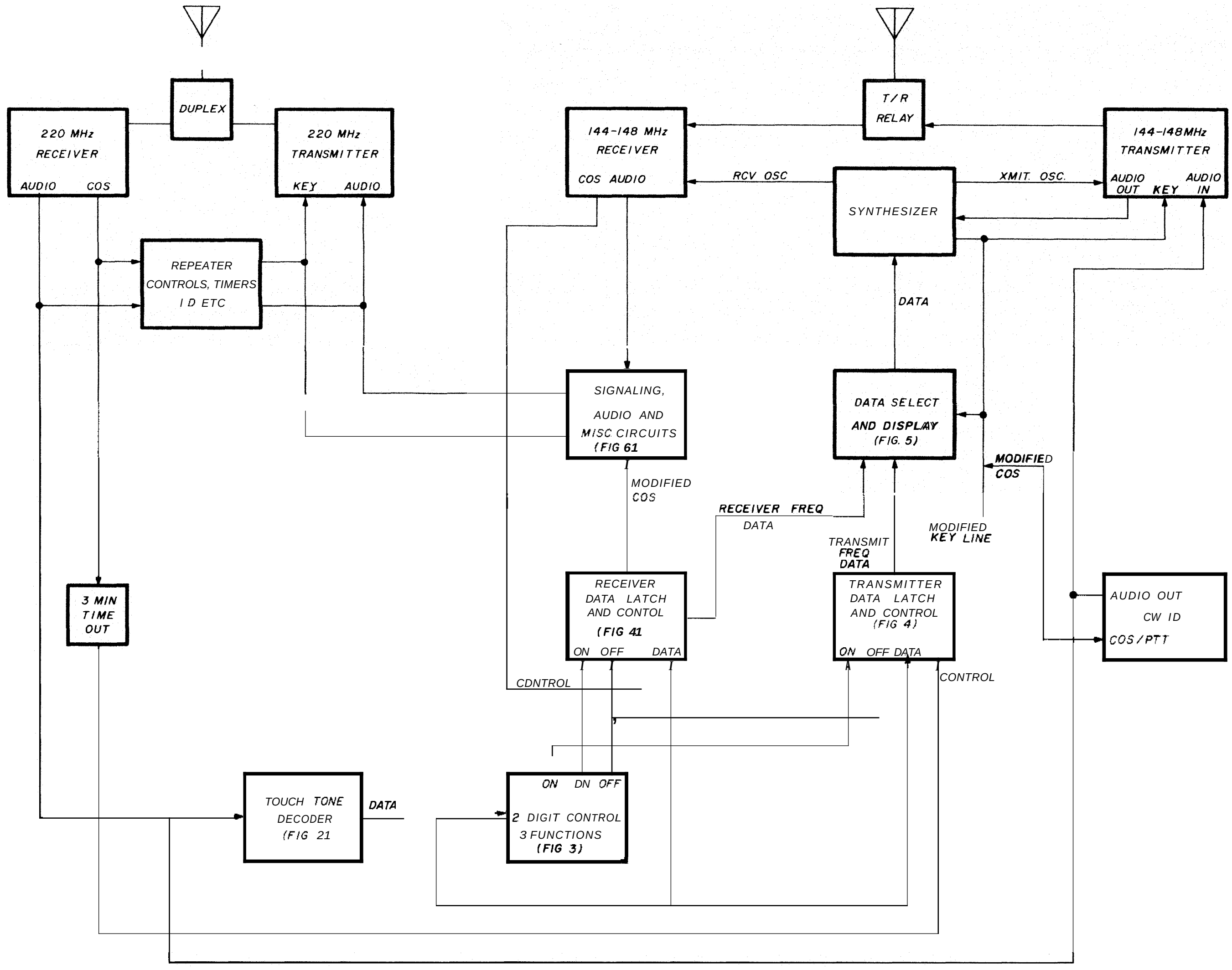


fig. 1. Block diagram of the basic remote system. A 220-MHz transceiver provides the control for the synthesized 2-meter transceiver. Circuit boards and parts kits are available from Creative Electronics, Box 7054, Marietta, Georgia 30605.

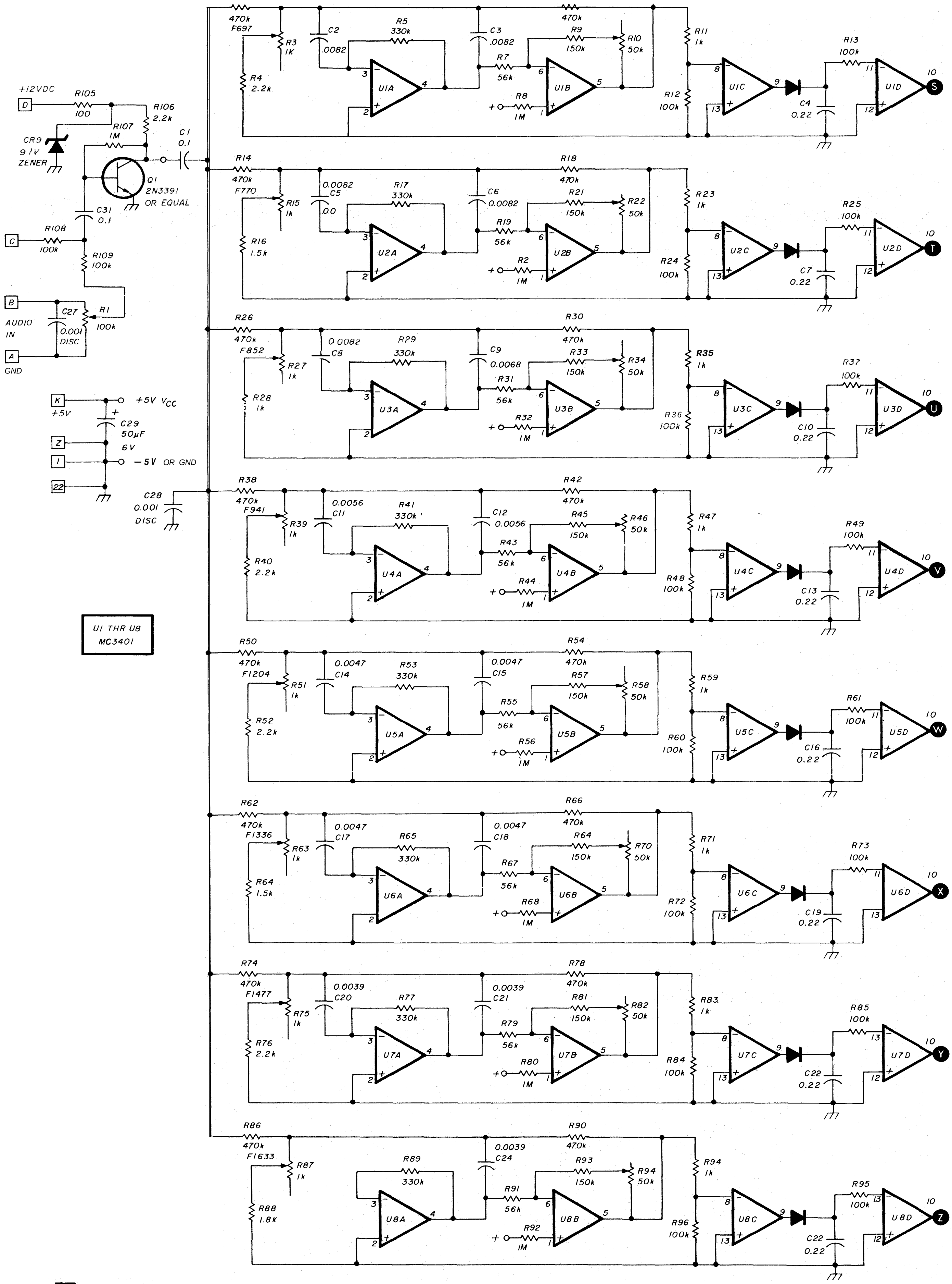
first two digits, which alert either of the other two boards (both identical) to accept the information following on the data bus. The access and control board will also sense two digits which signal the remote base to shut down. The two data latch boards, when enabled by the access and control board, store the subsequent four digits as frequency information. This board also senses whether or not a valid (inband) frequency has been received.

The data outputs from these two boards are connected to the data select and display board. This board contains the necessary circuitry to select whether the receiver or transmit data latch board will control the synthesizer frequency; it also displays the desired frequency and retains the last transmit frequency for the final identification. The data output from this board is connected to the synthesizer, which controls the receiver and transmitter. The audio and COS lines from both the repeater and remote base are processed on the miscellaneous circuits board.

tone-pad decoder

The tone-pad decoder **(fig. 2)** consists of eight active filters, U1 through U8, each of which is tuned to a standard ***Touch-Tone**** frequency. Except for the frequency-determining components, the following description of the 697-Hz filter is applicable to all others: Audio from the repeater receiver is coupled into the detector through a level adjust control, R1. The input on pin C allows the tones from a local pad to be applied to the decoder. The output from the active filter, U1A and U1B, is coupled through R11 to reduce the loading effects of U1C on the active filter. U1C normally is in a conducting state (output near

*"**Touch-Tone** is a registered trademark of the American Telephone and Telegraph Company.



+12VDC
D
R105
100
R106
2.2k
R107
1M
CR9
9.1V
ZENER
C1
0.1
Q1
2N3391
OR EQUAL
C31
0.1
C
R108
100k
R109
100k
B
AUDIO
IN
C27
0.001
DISC
R1
100k
A
GND
K
+5V
+5V V_{CC}
C29
50µF
6V
Z
1
−5V OR GND
22
C28
0.001
DISC
U1 THR U8
MC3401
470k
F697
R3
1K
R4
2.2k
C2
.0082
R5
330k
U1A
C3
.0082
470k
R9
150k
R10
50k
R7
56k
U1B
R8
1M
R11
1k
R12
100k
U1C
R13
100k
C4
0.22
U1D
S
R14
470k
F770
R15
1k
R16
1.5k
0.0082
C5
R17
330k
U2A
C6
0.0082
R18
470k
R21
150k
R22
50k
R19
56k
U2B
R2
1M
R23
1k
R24
100k
U2C
R25
100k
C7
0.22
U2D
T
R26
470k
F852
R27
1k
R28
1k
0.0082
C8
R29
330k
U3A
C9
0.0068
R30
470k
R33
150k
R34
50k
R31
56k
U3B
R32
1M
R35
1k
R36
100k
U3C
R37
100k
C10
0.22
U3D
U
R38
470k
F941
R39
1k
R40
2.2k
0.0056
C11
R41
330k
U4A
C12
0.0056
R42
470k
R45
150k
R46
50k
R43
56k
U4B
R44
1M
R47
1k
R48
100k
U4C
R49
100k
C13
0.22
U4D
V
R50
470k
F1204
R51
1k
R52
2.2k
0.0047
C14
R53
330k
U5A
0.0047
C15
R54
470k
R57
150k
R58
50k
R55
56k
U5B
R56
1M
R59
1k
R60
100k
U5C
R61
100k
C16
0.22
U5D
W
R62
470k
F1336
R63
1k
R64
1.5k
0.0047
C17
R65
330k
U6A
0.0047
C18
R66
470k
R64
150k
R70
50k
R67
56k
U6B
R68
1M
R71
1k
R72
100k
U6C
R73
100k
C19
0.22
U6D
X
R74
470k
F1477
R75
1k
R76
2.2k
0.0039
C20
R77
330k
U7A
0.0039
C21
R78
470k
R81
150k
R82
50k
R79
56k
U7B
R80
1M
R83
1k
R84
100k
U7C
R85
100k
C22
0.22
U7D
Y
R86
470k
F1633
R87
1k
R88
1.8k
R89
330k
U8A
0.0039
C24
R90
470k
R93
150k
R94
50k
R91
56k
U8B
R92
1M
R94
1k
R96
100k
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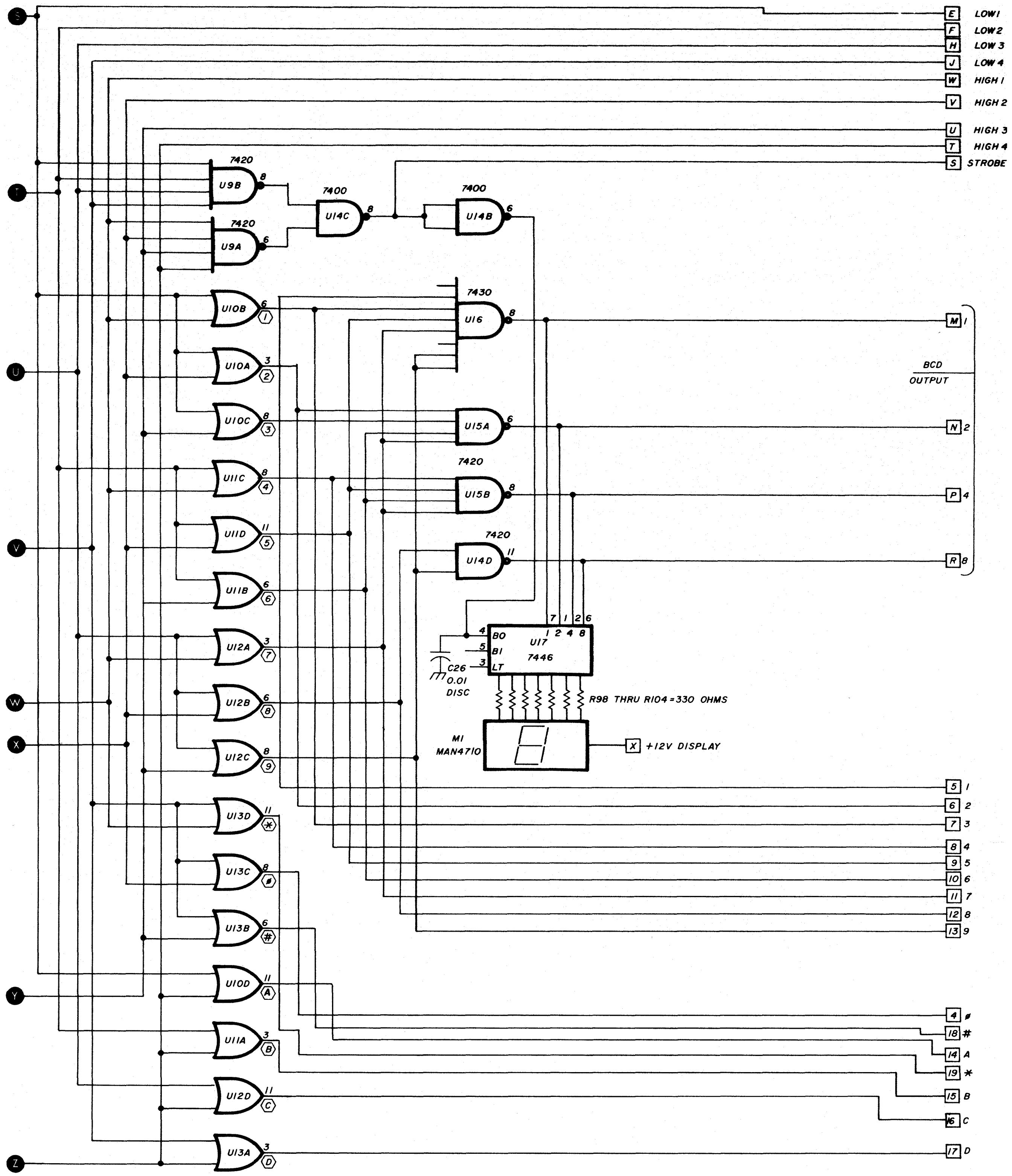
C22
0.22
U8D
Z



groups	desired frequency	frequency adjustment	Q adjust
Low 1	697 Hz	R3	R10
Low 2	770 Hz	R15	R22
Low 3	852 Hz	R27	R34
Low 4	941 Hz	R39	R46

tone pad frequency table

groups	desired frequency	frequency adjustment	Q adjust
High 1	1209 Hz	R51	R58
High 2	1336 Hz	R63	R70
High 3	1477 Hz	R75	R82
High 4	1633 Hz	R87	R94

fig. 2. Schematic of the tone decoders. All resistors are 1/4 watt, 5 per cent; all capacitors are Mylar unless otherwise specified.

zero volt). When a signal appears on its input; however, the negative-going peaks cause the output to go positive. This positive voltage is coupled through CR1 across C4, resulting in a fast attack. U1D operates as an open gain amplifier; the output is zero for a valid tone group.

The outputs from the four low-group tones are each connected to OR gates U10 through U13. The other gate input is connected to the appropriate output of the high-group filter detector. The outputs of each OR gate provide a zero logic state for the digit being received. These outputs are then connected to a series of logic gates, U14D, U15D, and U16D, which convert them to a BCD format. Included in this output format is a negative-going strobe, generated by U9 and U14C, which signals the other circuits that the data is valid.

In addition to the BCD data and strobe, other outputs are provided for each individual tone group, digits including # and *, and the extra right hand column, A, B, C, and D. These outputs all go to a low state when valid and may be used, if desired, for control functions.

access and control board

The access and control board, fig. 3, accepts the information from the data bus or from the extra outputs of the tone-pad decoder and converts this data to the necessary system control signals. When the strobe goes low, it starts a timer, U7, which is connected to the other input of U9D. When the timer's output returns to a low state, and if a valid digit is still present, the output of U9D will go to high. This reduces the effects of voicing.

When U5 goes to zero, it triggers timers U8A and U8B. If a valid digit is recognized, as represented by zero output at U5, timing capacitor C20 discharges. When the strobe returns high, the capacitor starts to recharge. If another digit is received, the strobe again discharges C20. So long as a steady stream of digits is received by the tone pad decoder, the timer will not reset, allowing U10 to count the U8B output. When the digit stream ceases, U10 is reset to zero.

U8B act as a one shot, providing a pulse for each digit received from the tone-pad decoder. The single-shot pulse is coupled through gate U6D, divided by 10 through U10, and then applied to decimal decoder U11. The decoded BCD digit information, plus the U11 output form a programming matrix. When more than nine digits are received, an output from U11 stops the counter by inhibiting gate U6D; this prevents the counter from recycling.

On the other side of the matrix are three sets of circuits, all the same, each used for one of the control functions. Following is a description of the receiver access portion; the transmitter access portion and the disable portion are identical except that only the first and second digits can be used on the disable circuit. One U1B input, for the first digit to be recognized, is connected to the U11 matrix outputs. The other input is connected through the matrix to the desired, preset digit. Therefore, when a received digit corresponds to the desired sequence, U1B output goes low and is inverted by U5C. When the digit is released, the negative-going transition starts timer U3B, providing a window through NAND gate U6B when the second control digit must be received and recognized. If this occurs, the U6B output goes to a zero, sending a control signal to enable the receive mode data latch and control board. The other control functions enable the transmitter data latch and control board or can disable both boards.

data latch and control boards

Two data latch and control boards, fig. **4,** are required: one for the receive frequency and one for transmit. When the access and control board signals receipt of the proper access code, it provides a zero-going pulse to the alert line (pin C) on the data latch board. This zero-going pulse performs several functions: It toggles U3A (which enables the counter U4, OR gate U12D, and NAND gate U13C), and U3B (which disables either the transmitter or receiver through OR gates U12B and U12C); it also provides a start pulse to U2B through U13D and U13C. When enabled, U4 counts the number of zero-going pulses on the strobe line.

The strobe pulses are conditioned by U2A to provide a single, short positive pulse regardless of how long the digit is held. These output pulses are also used to enable data latch chips before advancing the counter; this protects the currently stored digit and prepares for the next. The pulses from U2A are inverted by U16F and applied to one input of NOR gates U6A, U6B, U6C, and U6D.

The other gate input comes from BCD-to-decimal decoder U5. When more than five pulses have been counted by U4 and U5, a signal from U5 inhibits U1B, preventing the data latches from accidentally being recycled. As the counter advances from the starting point through three, each of the U6 NOR gates is allowed to pass the pulse generated by U2A; the following latches are sequenced to sample and store the information from the data bus.

In addition to decoding the data bus, the latch outputs are connected to a series of BCD-to-decimal decoders for sensing dialing errors, and/or the selection of out-of-band frequencies. The outputs of U17, U18, U19, and U20 are connected to a series of gates that check for specific frequencies. The U22C output goes low when out-of-band operation is sensed, sending a pulse through U15A, U15B, and U15C to

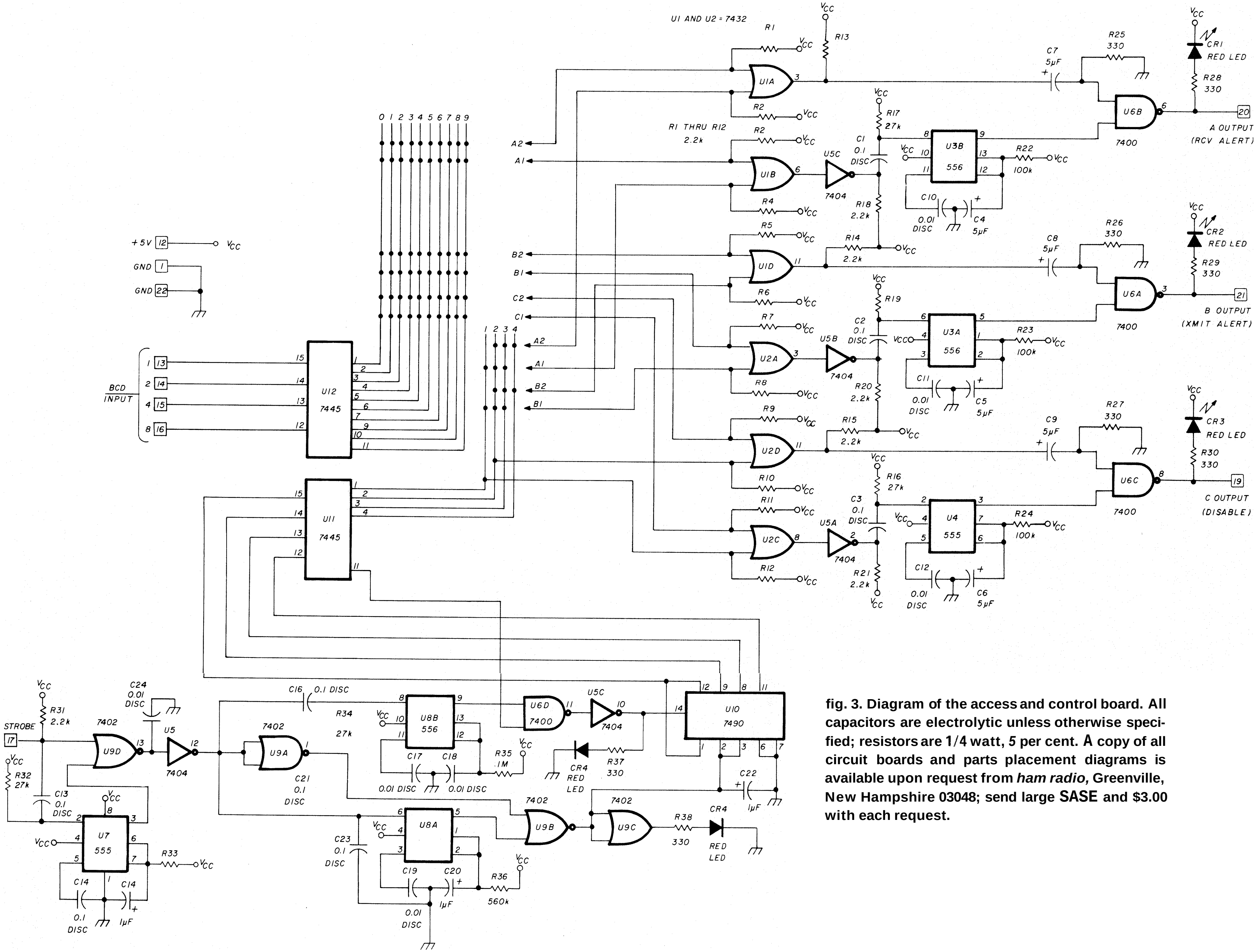


fig. 3. Diagram of the access and control board. All capacitors are electrolytic unless otherwise specified; resistors are 1/4 watt, 5 per cent. A copy of all circuit boards and parts placement diagrams is available upon request from *ham radio,* Greenville, New Hampshire 03048; send large SASE and $3.00 with each request.

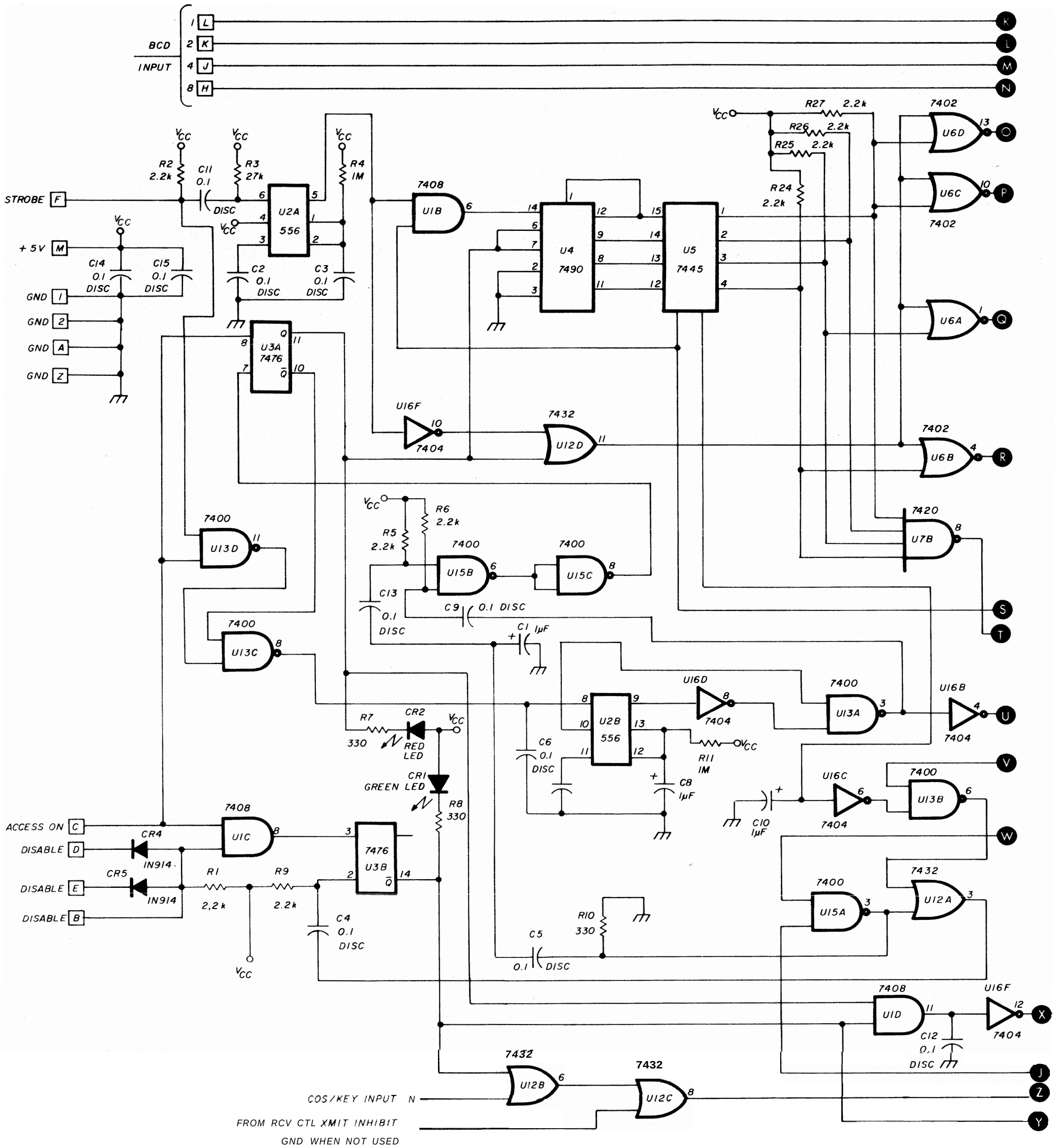
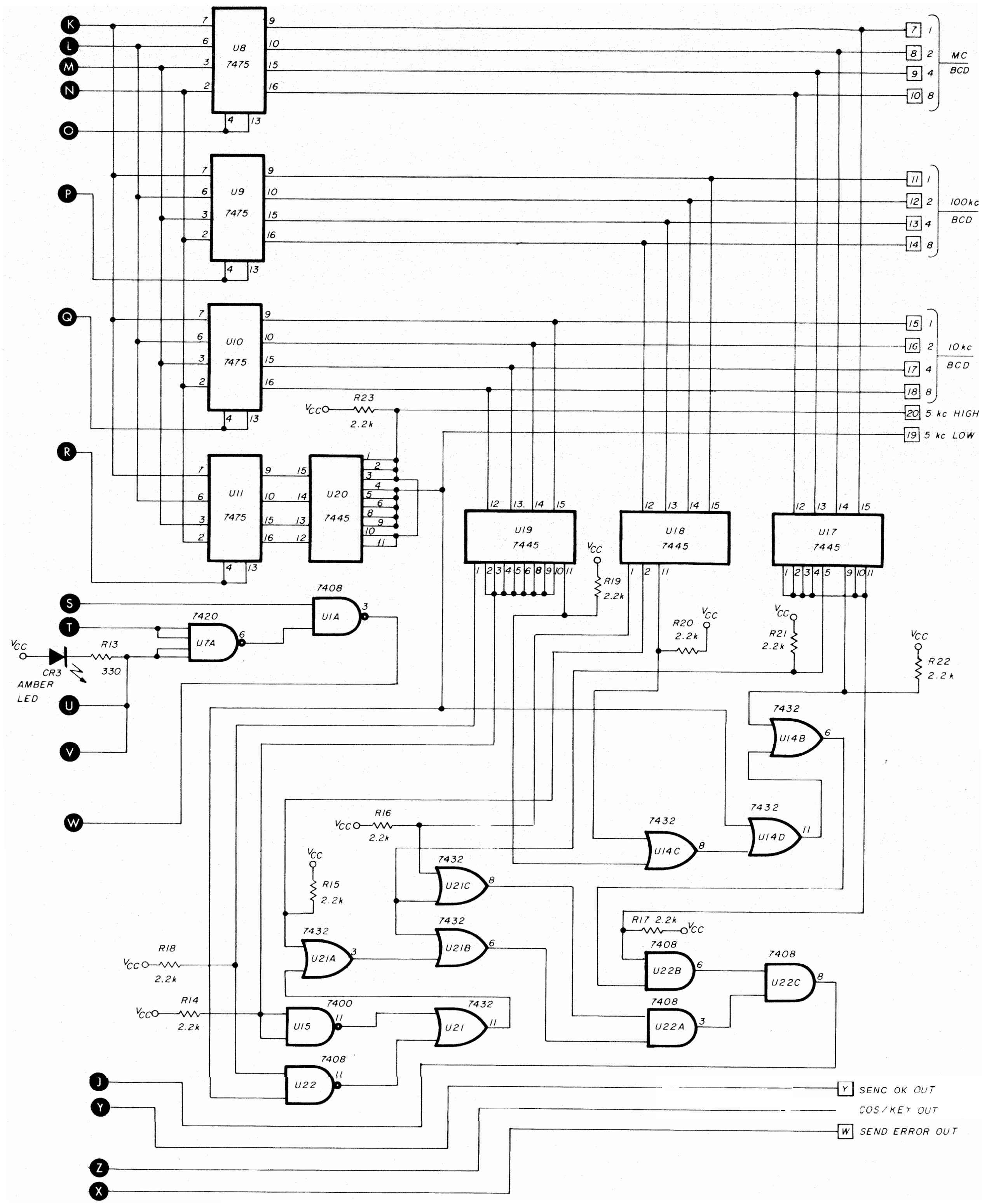


fig. 4. Circuit for the data latch and control board. All resistors and capacitors are as specified in fig. 3.

reset U3A. U12A blocks the reset pulse during the dialing time.

When the user has finished dialing, U2B is allowed to reset; both U13A inputs go high, passing a pulse through C9. This pulse is subsequently inverted by U15C and used to reset U3A and counter U4. U13A output is also connected to inverter U16B, which enables the circuits for testing number of digits dialed. If the tests are passed, U12A resets U3B, enabling U12B to pass the COR/COS input on pin N of the card edge connector.

If the wrong number of digits is received or an out-of-band signal is sensed, the reset signal is blocked by U12A, and U3A is reset by the output from U15A; U3B is not reset, leaving either the transmitter and/or the receiver disabled.



The main function of the data select and display board, **fig. 5,** is to determine whether the receiver or transmitter data latch and control board controls the synthesizer frequency. It also displays the operating frequency and the mode (receive or transmit). This board also senses when the transmitter is moved to a new frequency and instructs the identifier to send an ID; the old frequency is stored until the ID has been sent and then released.

The data selectfunction is performed by **U14, U15,** and U16. The outputs are connected to three data latches (U11, **U12,** and U13) which normally have

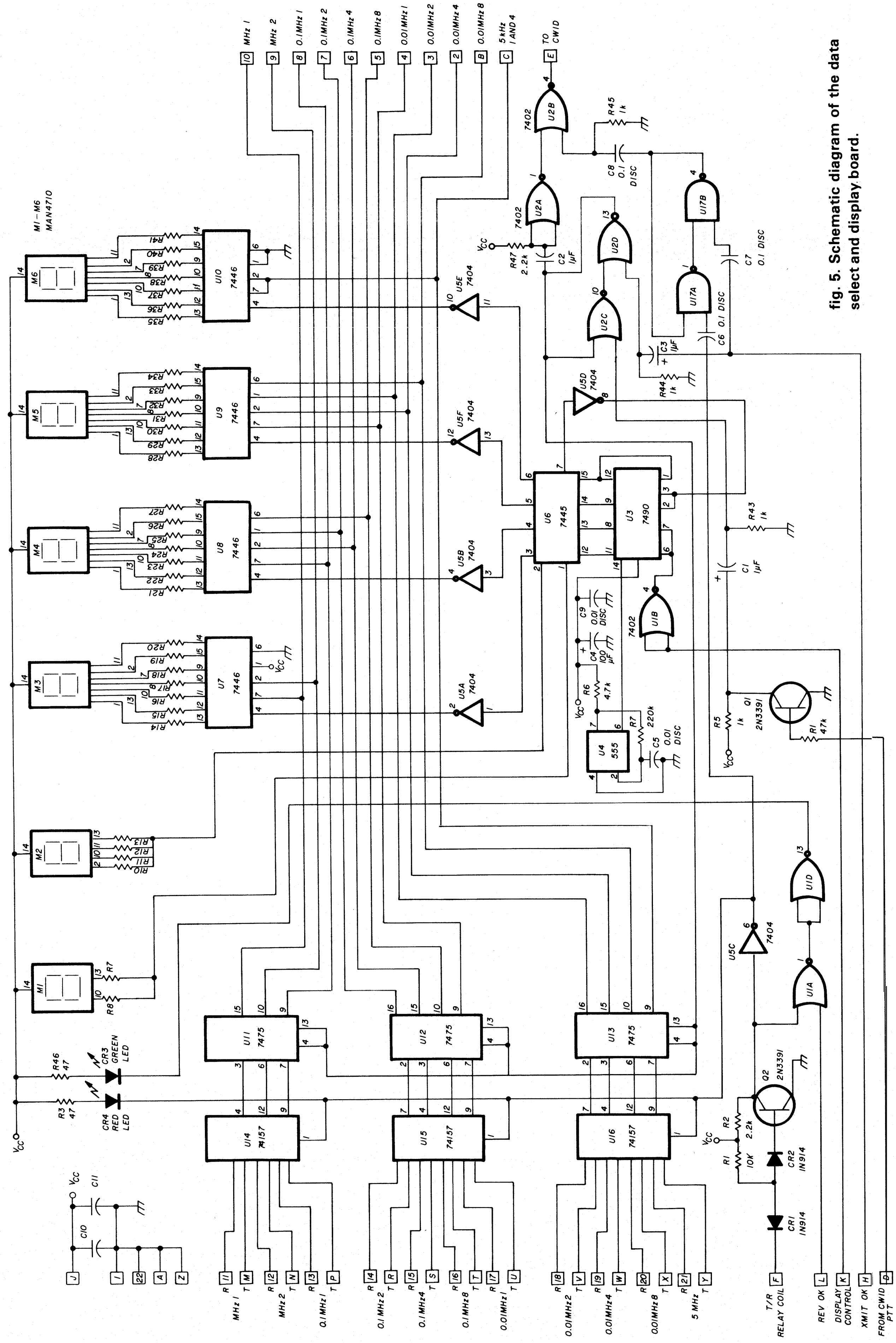


fig. 5. Schematic diagram of the data select and display board.

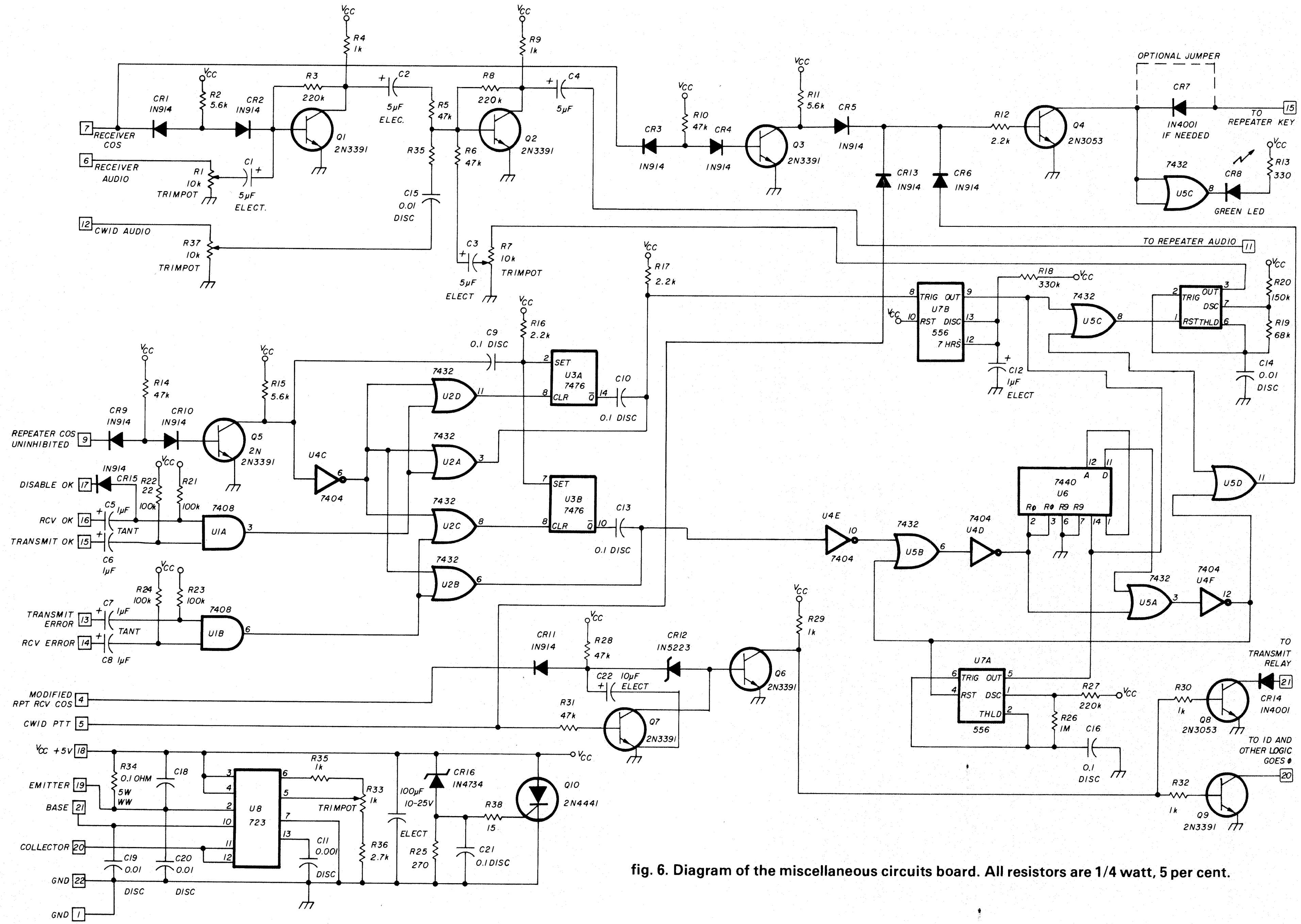
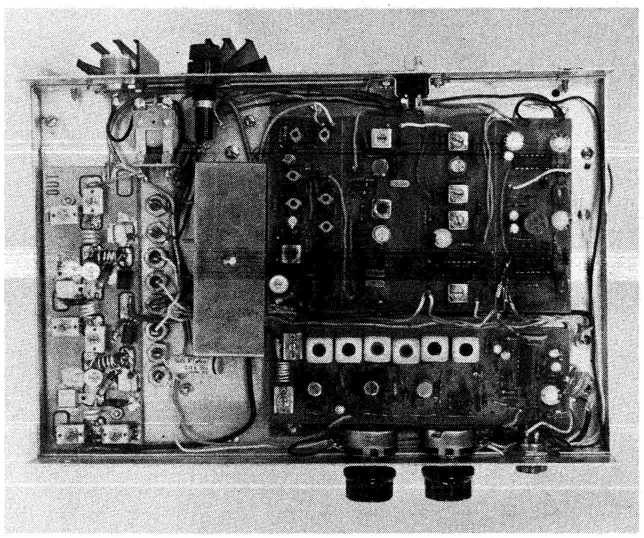


fig. 6. Diagram of the miscellaneous circuits board. All resistors are 1/4 watt, 5 per cent.



Layout of the bottom of the remote synthesized base. At the far left is the 15-watt 144-MHz rf power amplifier; the heatsink is on top of the chassis; enclosure at left center contains the 144-MHz filter. The four circuit boards at top right are, beginning at the left: 144-MHz receiver, 10.7-MHz i-f, 455-kHz i-f, and audio; lower right is the VHF Engineering 144-MHz transmitter strip.

their enable inputs held high by U2C and U2D so the outputs follow the input. Whenever the transmitter is disabled, the okay to transmit (XMT OK) signal from the transmit board goes high, setting U2C and U2D, which causes the data latch enable line to go zero, holding the data at that time. The frequency will be displayed and held and used by the synthesizer as the transmit frequency. The setting of U2C and U2D provides a pulse through U2A and U2B which then triggers the CW ID.

When the ID finishes, it causes a reset pulse to be generated from 01 through C1 to reset U2C. On an initial transmission, the CW identifier is triggered by a positive pulse generated when U17 toggles. It is set by a pulse generated through C7 when the XMT OK line goes to zero, then resets through C6 the next time the remote base key line is engaged.

The display portion of the board consists of six, 7-segment LED displays. Digits M1 and M2 are hardwired so they always display one and four; the remaining displays are driven through current-limiting resistors and decoder drivers connected directly to the same data bus as the synthesizer. To reduce power consumption, the display is strobed.

To control the data select ICs, the key line to the transmitter is also connected to the input of Q2. When the transmitter is keyed, the input to inverter Q2 goes low; the output is inverted by U5C, which controls the data select chips U14, 15, and 16. When the key is released, 02 conducts, the green LED (receive) is illuminated, the red **LED** goes out, and the data select outputs return to the selected receive frequency. The green receive LED is inhibited whenever a valid frequency has not been accessed.

The RCV OK signal is connected to one input of U1A while the other input of U1A is connected to the collector of Q2. The output of U1A, therefore, is held low whenever the RCV OK signal is high (disabled), regardless of the Q2 output.

miscellaneous circuits board

This board, fig. **6,** contains the regulator chip and the associated components to control the 5-volt pass transistor. Also on this board is the keying control for the remote transceiver, repeater keying control, repeater audio inputs, and signaling generating circuits.

The remote base is keyed by the repeater receiver COS signal after it has passed through the transmitter data latch and control board. The other keying input is the CWID PTT output. The CWID gets it COR input from the output of 69, which allows it to ID during a series of transmissions. A slight delay in dropping of the remote transmitter is provided by C22 and CR12 to prevent the transmitter from dropping out because of flutter on the 220-MHz input.

Repeater keying is provided from three sources: the remote receiver (COS after it has passed through the receiver data latch and control board), the signaling circuits, and CWID. The remote receiver COS is connected to the inputs of 01 and 03 after it has gone through the receiver data latch and control board. When a signal is being received by the remote base receiver, this COS input goes low, Q3 is cut off, and 04 is driven into saturation.

The CWID PTT is coupled through CR13 and R12 to 04. When the ID is running, a positive voltage keeps the repeater on the air and allows the ID audio to be heard over the repeater.

The other input to Q4 is from the signaling circuitry through CR6. Whenever one of the control functions has been activated, a signal is returned to the operator through the repeater to let him know the status of his controlling commands. These outputs are connected to U1A for the OK signal, and U1B for the ERROR signal.

When an OK signal is sensed, a short pulse is developed across either C5 or C6 which results in a short zero-going pulse at U1A output. The pulse is connected to gates U2A and U2D. If there is an input on the repeater receiver, it is assumed that the operator is still transmitting, causing the output of 05 to be high, blocking the pulse output of U2A. Since the output of 05 is inverted by U4C, the other input to U2D will be low, so when the signaling pulse arrives, it is allowed to pass through U2D and set U3A. When the receiver COS is released, the Q5 output returns low, sending a short pulse through C9 to reset U3A.

When U3A resets it sends a pulse through C10 to trigger U7B.

If there was no input to the repeater receiver when the signaling pulse appears on the output of U1A, the Q5 output would be low, allowing this pulse to pass through U2A and not through U2D, triggering U7B. U7B is a one-shot timer which provides a pulse to operate the tone generator U9, and key the repeater transmitter through U5D and CR6. This results in a short continuous tone to signal that either the receiver or transmitter had been properly accessed.

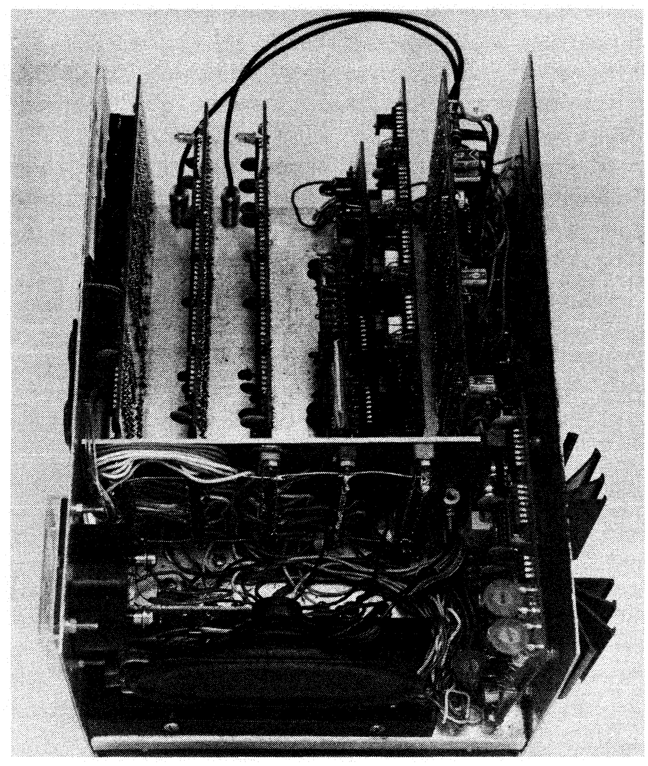
If a valid frequency had not been accessed, however, an ERROR signal would be received on one of the inputs. If there is no input to the repeater, U2B is allowed to pass this puise to U4E, while U2C wiii pass the pulse to U3B. U3B will then reset when the repeater is no longer receiving a signal, sending a pulse through C13 to U4E, U5B, and U4D.

When U4D output goes low, U6 counts the pulses from U7A; this output is also coupled through U5C to the tone generator U9 to provide a series of dits. Once started, the BCD 8 output on the counter will go to zero. This level is connected to one of the inputs of U5A; the other U5A input is from U4D which went zero with the signal-starting pulse. With both inputs at zero, the U5A output is zero. This is inverted in U4F and connected to the other U5B input. This allows the counter to cycle through to the eighth count where the output again goes high, causing U5A output to go high, eventually stopping counter U6. The U4F output is also used to stop the pulse generator U7A and key the repeater transmitter through U5D, CR6, and Q4.

Audio from the remote receiver is connected through R1 to Q1, which is controlled by the receiver COS. This blocks the receiver audio when an invalid frequency has been selected or the receiver has been disabled. The Q1 output is coupled through C2 and R5 into Q2. The output of the signalling generator U9 is also connected through R7, C3, R6 to *02,* while the CWID is coupled through C15 and R35 and R36 to Q2. This allows independent adjustment of the receiver, ID, and signalling audio levels.

construction

The remote base was built into an LMB cabinet (W-2J). The receiver, a VHF Engineering RX144B, was assembled according to the manufacturer's instructions, with the exception of a modification to make the COS output go low upon the receipt of a signal. This strip is mounted on the underside of the chassis as shown in **fig.** 7. Two small potentiometers were added to the VHF Engineering TX144B transmitter for the line and mike inputs so that both the levels and the deviation limit could be set. A VHF Engineering PA144-15 power amplifier was used,



Top of the synthesized base showing circuit board placement; boards, left to right, are data select and display, transmit data latch, receive data latch, two-digit access. CW ID, touch pad decoder, and VHF Engineering synthesizer. Miscellaneous circuits board (fig. 6) is at lower right next to the heatsink (on rear panel) for the 5-volt regulator. Heatsink for the power amplifier stage is hidden by the speaker in foreground.

although the PA144-25 will also work. The amplifier was mounted with its heatsink on top of the chassis and the printed-circuit board with the components below the chassis (see **fig. 7**). The double-pole, double-throw relay (Magnetcraft W88X-7) mounted next to the antenna connector is used to switch the antenna and the 12-volt supply between the receiver and transmitter.

The receive and transmit oscillator signals are fed through phono connectors located on the opposite side of the receiver. The volume and receiver squelch controls are mounted between the exciter and front panel.

The tone-pad decoder board, two-digit access and control board, data latch boards, data select and display board, miscellaneous circuit board, and the CWID and synthesizer are all mounted on the top side of the chassis. A heatsink (Wakefield 641-Z) with the 2N3055 5-volt regulator transistor is mounted on the rear panel behind the miscellaneous circuit board; the 9-volt and 5-volt regulators for the synthesizer are mounted without heatsinks on the other side of the rear panel.

The VHF Engineering SYN II synthesizer board was assembled according to the manufacturer's instructions with the exception that resistors R1 through R12 and R14 were not installed. The offset-programming diodes, CR3 through CR8 are not needed.

The eight feedthrough capacitors are used for the following items (listed in order from the front to the rear):

1. Local mike PTT
2. Transmitlreceive relay
3. Transmit audio
4. 13.6-Vdc supply
5. Receiver audio
6. Receiver COS
7. Receiver speaker
8. Optional S-meter

When stuffing the two-digit control board (TDC 203), the control codes should be chosen and programmed; I used 71 (R1) for receive, 81 (T1) for transmit, and 73 for disabling.

installation and interfacing

The 2-meter base was connected to a VHF Engineering 220-MHz repeater. The repeater COS (low state on) from the COR-2 board (PTT), along with the audio from the repeater speaker line, are connected to the audio and COS input lines of the remote base. Initially the installation was made at a site which contained one 2-meter repeater and another 220-MHz repeater. In addition, there is a 2-meter repeater located approximately 0.4 km (114 mile) from the remote base. The major problem is with the 2-meter repeater located at the same site as the remote base, although some interaction (desense) also occurs when the other 2-meter repeater is in use. Based on these experiences, I recommend that the remote base not be co-located with a 2-meter repeater. I also recommend that the 220-MHz and 2-meter antennas be located one above the other on the tower.

operation

For maximum versatility the system was set up to allow separate dialing of the receive and transmit frequencies. To dial the receive frequency, dial R (7 on the tone pad) and the frequency desired, R147.340 for example. After the mike PTT has been released, a short continuous tone will be heard, indicating that the frequency has been successfully accessed (a short burst of dots indicates the operator was unsuccessful). When a valid receive frequency had been accessed, any signals appearing on the remote receiver input will be heard on the 220-repeater output.

Once a frequency has been selected, then the appropriate transmit frequency, simplex or repeater input, can be dialed starting with T (digit 8) and the frequency, T147.940 for example. Again, shortly after the mike key has been released, a signalling tone should be heard which indicates either success or failure. If it's success, any signals appearing on the input to the 220-MHz repeater will be retransmitted on the selected 2-meter transmitter frequency.

When the transmitter is used on a new frequency, it will transmit its ID on both the 2-meter remote base as well as the 220-MHz repeater. The ID will reoccur periodically during a series of transmissions at a rate set by the control operator.

After a QSO is complete, only the transmitter can be dropped, allowing the operator to monitor any last comments from others by dialing the transmit access code T1 (81), but no additional digits. When this is done, the transmitter starts to reprogram, but since a valid complete frequency is not received, a *misdial* (error) is sensed and an ERROR signal is returned; the transmitter identifies on the last frequency it used and is then disabled, leaving the receiver on. The receiver can be dropped by dialing the access code R1 (71) and no further digits. The code 73 will drop both the receiver and transmitter and return an OK signal indicating successful receipt of that command; this causes the remote transmitter to identify a final time on the last frequency it used.

current result and the future

The prime purpose of this project was to encourage the use of 220 MHz by Amateurs in the Atlanta area. Though the selection of 220-MHz rigs available is currently limited, I hope the demand for equipment will stimulate an abundance of gear with the many bells and whistles that are now available on 2-meter rigs.

Over a year ago there were two 220-MHz repeaters on the air in the Atlanta area and only four people using them; six months ago two more operators became regular users but I was unable to convince others to invest in 220-MHz equipment. Since the remote base has been on the air, about fifteen new operators have been added to the list, and this was before demonstrations were given to the Kennehoochee Radio Club and the Atlanta Radio Club.

acknowledgments

I would like to thank the following Radio Amateurs who helped assemble the prototype remote base, and who provided valuable suggestions and help in preparing this article: W. C. Clonninger, W4NX; Dave Rogers, AA4DR; Boyd Cone, AG4X; and Jack Sanders, WB4CDP. Also to Judi, my very best friend, who not only assembled most of the remote base but also typed the manuscript for this article.

ham radio

portable monoband shortwave receiver with electronic digital frequency readout

I have previously expressed my views on the advantages of monoband receivers.[1,2] This article describes my latest effort in this regard. A portable receiver design is shown with digital frequency readout. The front end covers 2.3-6.5 MHz without bandswitching. Frequency coverage can be extended by changing the rf and mixer coils. Two frequency-display options are offered depending on the i-f chosen; these are dubbed GP-58 and GP-59. The receiver weighs about 3.4 kg (8 lbs.), uses about $250 worth of parts, and required about 450 hours to build.

general considerations

My sets are designed for portable operation, so size and power consumption are of prime importance. Although frequency readout with mechanical counters is attractive, much design and construction difficulty occurs. Judging from reader response to earlier designs,[3] many prospective builders lacked facilities to homebrew a set with mechanical digital frequency readout within the standards set forth.

With the advent of electronic frequency counters designed around CMOS ICs,[4] it's possible to read frequencies with an extremely low current drain, which makes this method of frequency measurement suitable for battery operation. The physical size of CMOS counters is also quite small, and they require fewer components than the older TTL devices.

designs

If you're interested in receiving signals below 10 MHz (as in my case), frequency readout is extremely easy to implement by using a 9-MHz i-f. The MHz digit is read on the VFO bandswitch knob and the kHz numerals are read on the counter display (fig. 1). The VFO frequency (12.015 MHz) is set equal to the received frequency plus the i-f.

Note that, using this design option (the GP-58), the readout is entirely digital yet only three LED displays are used. Counter current drain is reduced to a bare minimum. The 9-MHz crystal filter is available for about $50.00.

An alternative design (the GP-59) employs electronic digital frequency readout by using a 10-MHz i-f (fig. 2). This option offers more convenient frequency readout but at the price of more battery drain and a custom-made 10-MHz filter. The 10-MHz crystal filter is available in the U.S. on special order for about $80.00. It can be home built.[5,6] The filter in reference 6, which I've tried, is particularly attractive. It uses standard 10-MHz crystals.

A word of caution if you're planning to use the 10-MHz i-f: Some performance degradation may occur because of WWV signals feeding into the 10-MHz i-f strip. The receivers are well shielded and use a trap tuned to the i-f in the front end, but isolation may not be sufficient where the WWV 10-MHz signal is particularly strong. In this case more elaborate shielding may be required.

Despite the fact that the VFO operates in the 10-20-MHz region, extremely stable operation occurs. (Sta-

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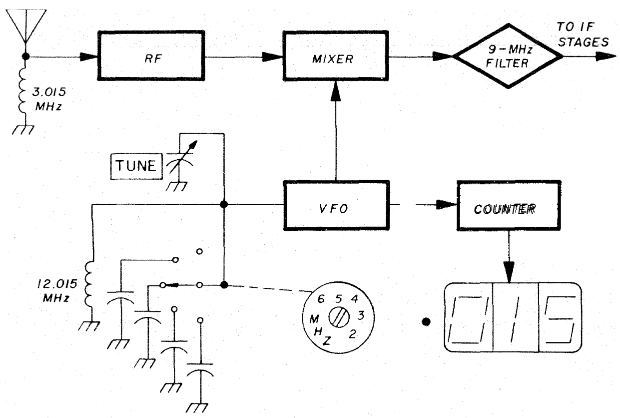


fig. **1**. Electronic digital frequency readout using a **9-kHz** i-f. Received signal is **3.015 kHz.**

bility, sensitivity, and avc response are shown below.)

construction

The photos show the GP-58 receiver and give details for its duplication. A schematic is provided in **fig.** 3.

Cabinet dimensions are 140 x 65 x 160 mm (5.5 x 2.5 x 6.5 inches). Cabinet material is 16-gauge stainless steel. The cabinet was formed on a steel press. The main chassis and back panel are of 3.2-mm (1/8-inch) aluminum. All shields are of 0.8-mm (1/32-inch) thick aluminum. The PC boards are mounted on 6.4-mm diameter (114-inch) pillars fastened with M3 (4-40) allen screws.

coil*	description	frequency range (MHz)
L1	18t on 5-mm (3116-in) slug-tuned ceramic form	9.0
L2, L3	6t and 55t respectively (Amidon T37-2 toroid)	2.3-6.5
L4	55t (Amidon T37-2 toroid)	2.3-6.5
L5	18t on 6.5 mm (1/4-in.) slug-tuned nylon coil form	9.0
L6	42t on 6.5-mm (114-in.) slug-tuned nylon coil form	9.0
L7	18t on 6.5-mm (1/4-in.) slug-tuned nylon coil form	9.0
L8	18t on 6.5-mm (114-in.) slug-tuned nylon coil form	9.0
L9	6t on 6.5-mm (114-in.) slug-tuned nylon coil form	9.0
L10	10t on 5.0-mm (3116-in.) slug-tuned ceramic form	11.3-15.5

***all** coils are wound with 0.3 m m (**no.** 28) enameled wire

Knobs are by National (Raytheon makes a similar design). The antibacklash gear reducer is a British import with a 1:100 ratio; the manufacturer is Muffett, Ltd. The VFO capacitor is also a British import (Windgrove and Rogers). The S-meter is a Japanese import with a 1-mA movement. (The original dial was replaced to show S-meter readings.)

The 9-MHz crystal filter is available from Yaesu dealers. Its bandwidth is 2.4 kHz at – 6 dB. The i-f coils are made in Brasil and are about 15 x 15 mm (0.6 x 0.6 inch) with a 6.4-mm (114-inch) internal coil form.

power supply

The power supply accepts both ac and dc. It has been designed for portable operation, in which the display can be switched off to reduce battery drain. You can use a 12-volt ac supply or a 12-13 volt dc source rated at 100 mA. Current consumption at 12 volts dc is 85 mA with the counter on and about 45 mA with the counter off. For portabie use i operate this set with a calculator power supply. The 13-volt zener is a surplus item and, being of nonstandard voltage, may be difficult to find. A 12-volt zener could be used.

devices

The digital displays are Fairchild FND-357. ICs 7207 and 7208 are by Intersil (described in reference 4). Note that the 7207A is *not* a replacement for the 7207. The 74LS90 is a standard low-power Schottky device available from many sources. Transistors BF-255, BD-135, and BD-136 are Siemens; BF-115 is Phillips. All other transistors are of U.S. manufacture.

performance

With the filter shown (bandwidth 2.4 kHz at 6 dB down), overall sensitivity was measured at slightly over 122 dB (while receiver tuned to 4.9 MHz):

front end and mixer	14 dB (including filter loss)
i-f strip	70 dB
audio section	38 dB

Operation is stable for the input-voltage range of 12-8

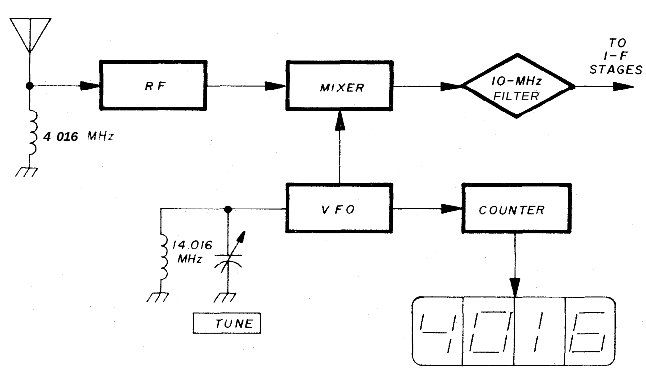


fig. 2. Electronic digital frequency readout using a **10-kHz** i-f. Received signal is **4.016 kHz.**

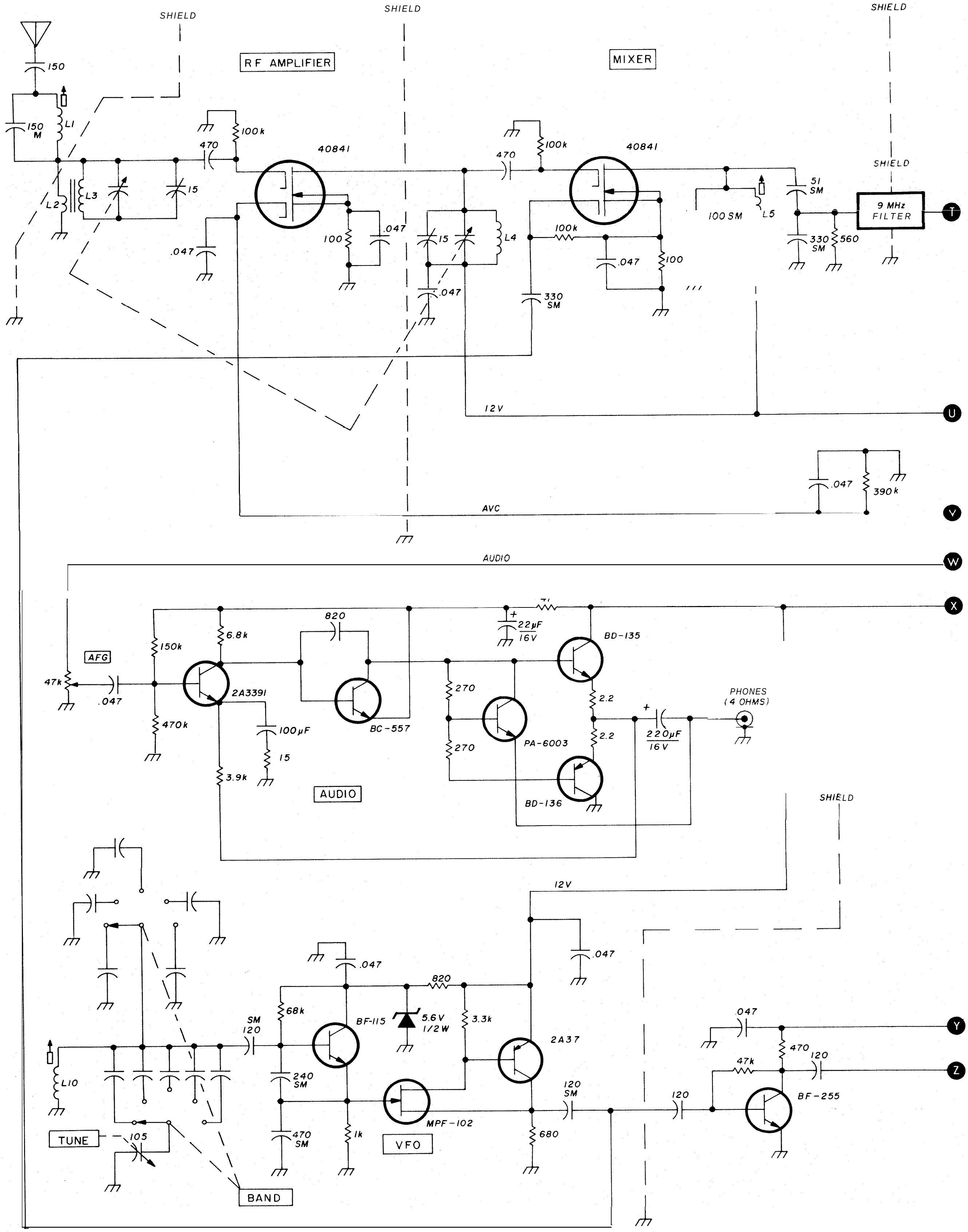
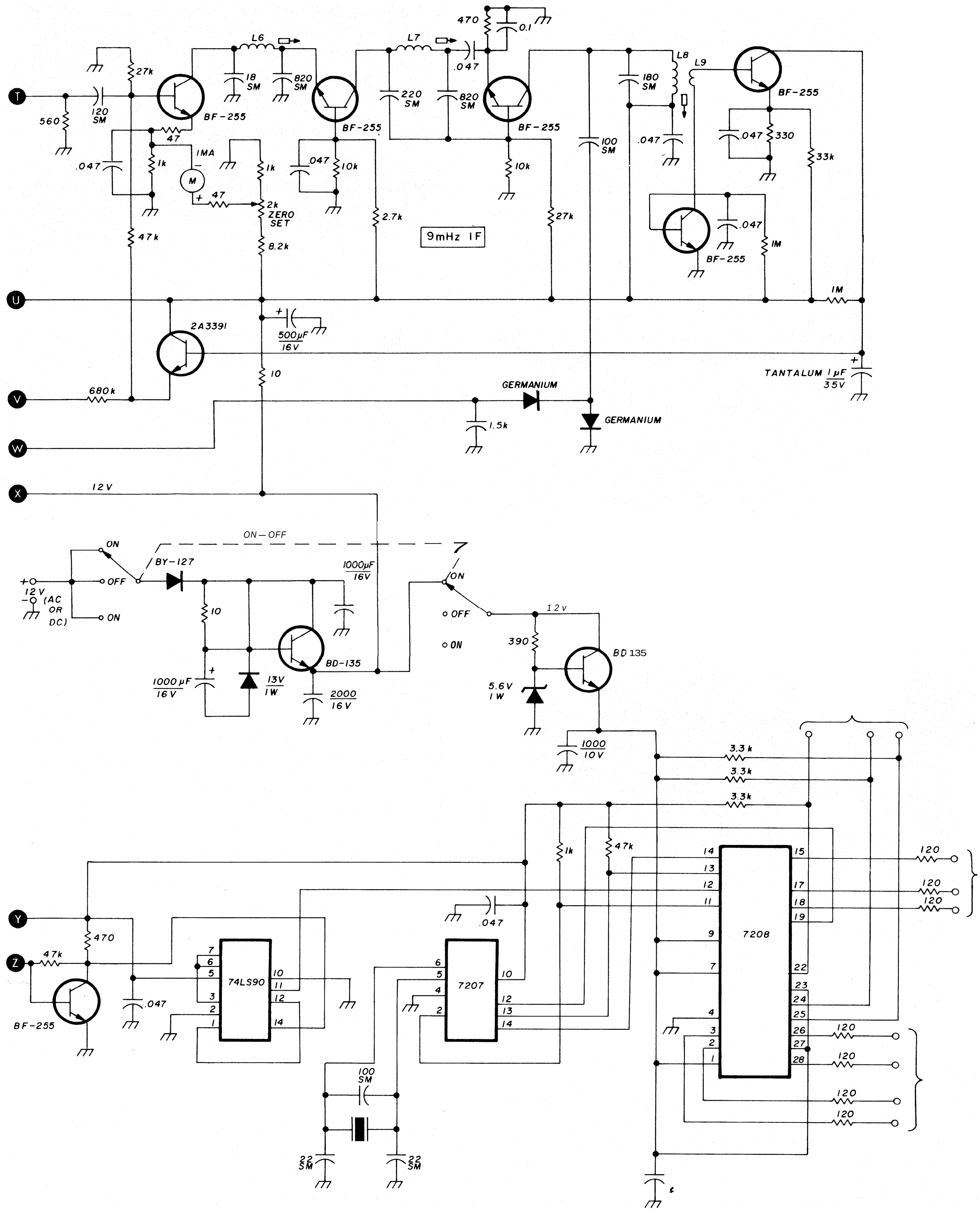


fig. 3. Schematic for the portable shortwave receiver covering 2.3-6.5 MHz. (See text for other frequency ranges.)



L6
L7
470
0.1
L8
L9
27k
BF-255
18 SM
820 SM
120 SM
560
47
.047
1k
IMA
1k
2k ZERO SET
8.2k
220 SM
820 SM
.047
10k
2.7k
9mHz IF
27k
180 SM
100 SM
.047
330
33k
IM
47k
2A3391
500μF 16V
10
680k
GERMANIUM
1.5k
TANTALUM 1μF 35V
12V
ON—OFF
BY-127
1000μF 16V
ON
OFF
12V (AC OR DC)
1000μF 16V
13V 1W
BD-135
2000 16V
390
BD 135
5.6V 1W
1000 10V
3.3k
74LS90
7207
7208
470
.047
100 SM
22 SM
120

volts. At about 7.5 volts the counter becomes unstable. Desensitization becomes noticeable below 9 volts, with overall gain dropping to about 100 dB.

Worthy of mention is the fast-attack, slow-release avc response (also measured at 4.9 MHz):

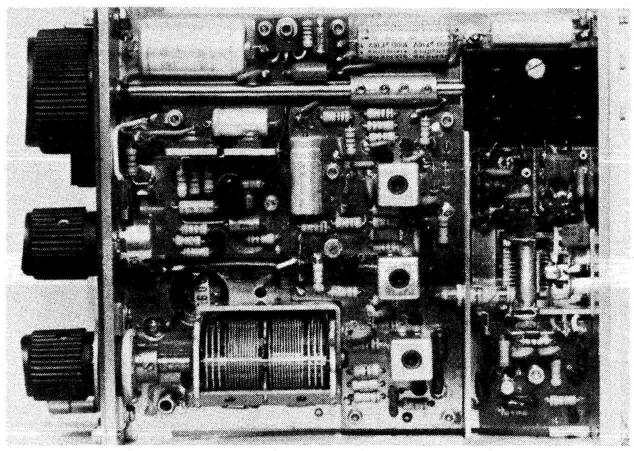
input signal (mV)	output (mV)
125	275
20	275
2	200

Counter frequency stability is excellent. Operation to 18-19 kHz was stable and reliable. (These are my qualitative observations; they were not checked specifically.)

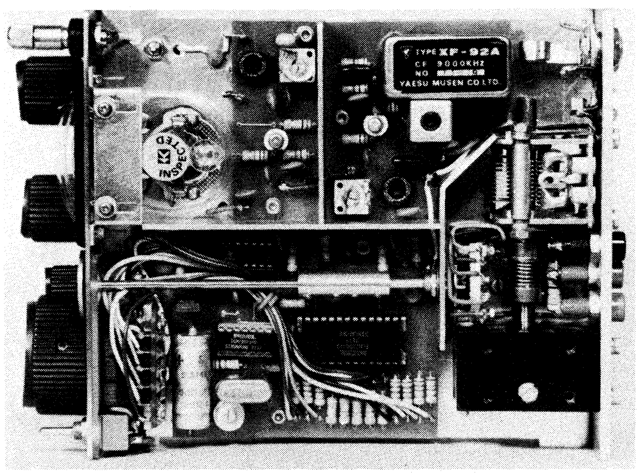
Overall stability was measured with a laboratory frequency meter. After a 5-minute warmup (radio tuned to 6.2 MHz), the frequency did not vary more than ± 25 Hz during a 15-minute period. The key to such performance is to operate the VFO transistor with V_{cc} at only 5.6 volts while selecting the best capacitive feedback ratio (see fig. 3).

conversion to other frequencies

This set was designed for shortwave broadcast listening between 2.3 and 6.5 MHz; the Amateur 80-meter band is therefore included. The Amateur 40-meter band could be covered by appropriate changes to the rf and mixer coils. For higher-frequency Amateur bands the VFO range will probably exceed the counter upper-frequency response. This problem could be cured either by adding a prescaler to the counter or by running the VFO below the input frequency. Note that VFO operation above 10 MHz requires the first two digits to be shown on the bandswitch knob rather than only one digit as shown in **fig. 1.**



Bottom view. The 12-volt regulated power supply is at top above the tuning shaft. The af amplifier is at center left. The rf 2-gang tuning capacitor is at bottom left. Below is the **9-kHz** input trap coil. The i-f double-sided PC board is in the center. The tuning gear reducer is at top right; below it are the padder and trimmer fixed capacitors, the **VFO** variable capacitor and VFO coil, and the VFO board (bottom right). A shield separates the **VFO** from the i-f board.



Top view. At top left, near the S-meter, is the rf amplifier board; at top right is the mixer board with the **9-MHz** crystal filter and matching transformer. The VFO variable capacitor is at center right. Toroid coils are seen for the rf and mixer stages. Below the VFO **capacitor** is the **VFO** bandswitch and the tuning gear reducer. At bottom left is the CMOS frequency counter, which is completely shielded from all other stages. On the frequency-counter board above the bandswitch shaft are the **74LS90 IC** preceded by the two amplifier stages **(BF-255).** The 5-volt regulated power supply is at top left near the **1000-μF** electrolytic capacitor, which is partially hidden by the cable harness.

final remarks

I used particular care in this design to maintain low current drain without sacrificing performance. The LED displays are visible using a 120-ohm series resistor as shown in **fig.** 3. The af strip was adjusted for headphone operation with low quiescent current (about 12 mA), yet it can deliver peaks of over 2 watts into a 4-ohm load.

acknowledgement

I am deeply indebted to Maiso, PY2GP, for his support and to Fernando, PY2DQU, who helped with much design work and measurement equipment.

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ham radio

solid-state
vhf linear amplifiers

With the increasing use of single sideband on 2 meters and above, more solid-state linear amplifiers are being built. The power range of these amplifiers has been extended into the 80 to 100 watt per transistor package class. However, a major difficulty occurs since the existing market is almost exclusively class C iland-mobile fm), with no linearity requirement; Amateurs want the ruggedness of the land-mobile components, with the additional features of linearity and 4-MHz bandwidth.

distortion products in linear amplifiers

Amplifiers have been in use since the inception of Amateur Radio. The term "linear" is a precise term describing the output waveform as being a constant multiplied by the input. Since only "approximately linear" amplifiers are available to us, we generally limit the term ***linear*** to amplifiers which have odd-order mixing (intermodulation) products 30 dB or more down from peak power. This figure was derived in the early sixties by power tube engineers at Eimac; they determined that with intermodulation products down 30 dB, an Amateur would cause minimal disturbance to an adjacent channel; the industry currently specifies most high-frequency equipment near this standard. However, in the vhf range, many solid-state amplifiers exhibit intermodulation distortion products similar to that of a class-C amplifier. In fact, one amplifier tested showed near identical intermodulation performance with or without forward bias.

causes of nonlinearity

The sources of nonlinearity can be related to three major areas in bipolar semiconductors: the input, the transfer function, and the output. The input characteristics are affected by two components: the base-emitter diode and the base resistance. The base-emitter diode will function as a detector or mixer. This nearly-ideal diode responds as predicted by the diode equation with a full set of harmonics and mixer products when driven by a sinusoidal input. The base resistance also changes in a nonlinear fashion because of base-width modulation and resistivity changes by heavy injection of minority carriers. However, these changes are masked somewhat by the effects of the base-emitter diode.

The second problem area is the transfer function, the variation of h_{fe} with current. This variation is dependent upon the device-doping profiles which are fixed by the semiconductor manufacturer. High-frequency (2-30 MHz) linear components show less change in h_{fe} than do the class-C parts we must use on 2 meters.

The output characteristics of the transistor are a third source of intermodulation products. These relate to the junction capacitances and the voltage across the junctions. Fortunately, the output characteristic contributions are minor when compared with the total problem.

Fig. 1 shows a transistor model that I borrowed from Amprex which illustrates the effects and contributions of the various components. Note that I have added an input T network similar to the input of a JØ packaged rf power device.

optimization of linearity

The two predominant areas, within the design, that control linearity are choice of operating point and the matching networks. The base-emitter diode is not practical because of series impedances within the transistor package; high emitter currents reduce distortion by reducing the applied signal voltage across the diode. The emitter resistance, R_e, will contribute to this effect as does the package inductance. Increasing the collector current excessively will slide the operating area into the saturation region where the nonlinear transfer effects (h_{fe}) will predominate. I have found that close attention to this detail and presenting the proper load impedance to the transistor will provide acceptable IMD products.

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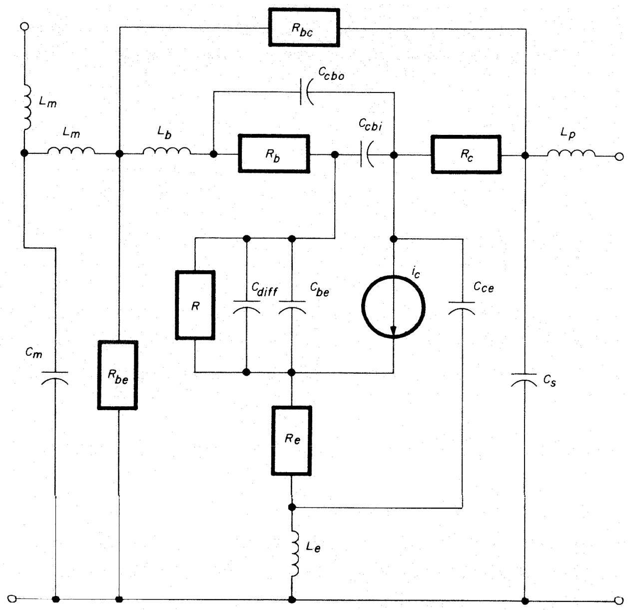


fig. 1. Equivalent circuit of a JØ packaged rf transistor in the common-emitter configuration. In addition to the basic model described by Arnperex, the author has added a T network at the input to approximate the JØ package.

biasing

The biasing of vhf power devices should be the same as that of high-frequency devices. The characteristics of this bias source should include a thermal cutback, to prevent a shift of the operating point when the transistor die heats up, and a low-impedance, constant-voltage source. These seemingly contradictory requirements are created by the drop in base-emitter voltage with temperature and the rectification of drive in the base-emitter junction. A typical 80-watt high-frequency device might experience a 200-mA change in base-current drive requirements because of this effect. These changes make the paralleling of devices even more difficult than at high frequencies. If you must parallel devices, match V_{BE} (on) and h_{fe}.

practical bias circuits

There are many bias circuits which meet the following two requirements: the dc base voltage is constant with rf drive level variations, and the dc base voltage decreases as die temperature rises within the package. The diode clamp and Byistor are two extremely similar circuits which meet these requirements. The Byistor offers superior protection from thermal runaway because the increase in voltage drop across the silicon resistor lowers the bias voltage. The diode clamp, without the series resistor, offers a lower source impedance. The Byistor has an output impedance of about one ohm. The current required for the diode clamp does not change appreciably over the operating cycle (see **fig. 2A**). The base current requirement can be in the 200 to 400 mA area.

A reduction in current requirements may be obtained by using a series, rather than a shunt, regulator. The simplest is a diode clamp circuit feeding the base of a pass transistor (see **fig. 2B**). A second series diode is added to compensate for the base emitter drop of the emitter follower. This reduces the current required from the diode source by $1/h_{fe}$. Single-transistor amplifiers work well, as demonstrated by the circuit from Amprex shown in **fig. 2C.** The variable resistor is used to set the output voltage. This circuit has an output impedance of approximately 0.1 ohm for a 360-mA current variation.

Operational amplifier circuits provide even lower output impedances and may feature current limiting and an even greater range of operating points (see **fig. 2D**). If you are interested in superior performance, this type of circuit is appropriate.

bias circuits precautions

Always begin experimentation with minimum bias voltage. The h_{fe} of rf devices designed for class-C

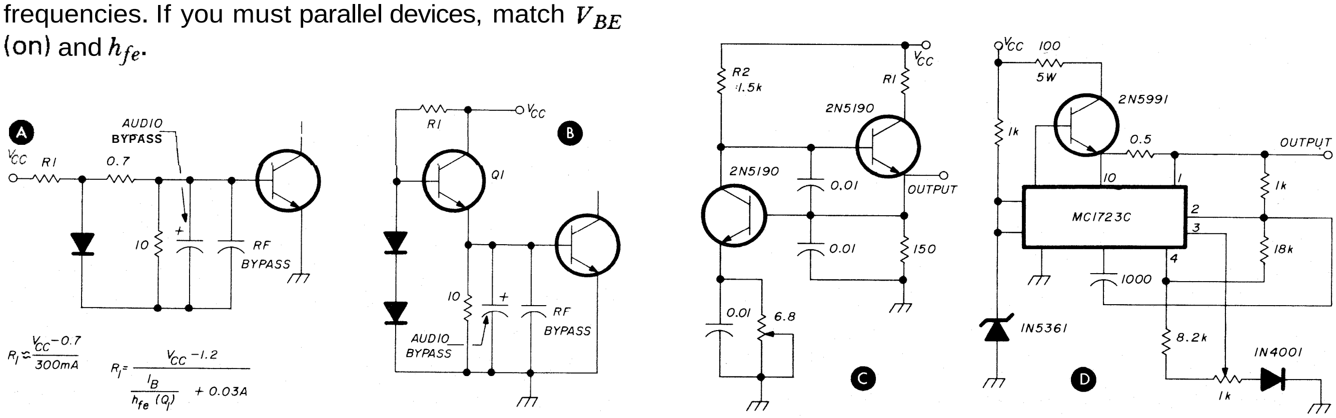


fig. 2. Four different methods for supplying base current. (A) illustrates a simple diode clamp/Byistor circuit. The Byistor provides superior thermal runaway performance. By changing to the series regulator (B) instead of shunt regulators, the current requirement is reduced by $1/h_{fe}$. (C) shows another version which incorporates a simple transistor amplifier. This circuit, along with (D), is an active bias supply which features high current capability and low output impedance.

operation varies widely, unless specified at a certain value. The bias voltage may be set at a fixed value once the desired current is determined for a specific transistor.

Mount the temperature sensor to the rf transistor's package for maximum protection against thermal runaway. The second best is to mount the sensor on the heatsink near the rf transistor. ***Never*** expect protection against thermal runaway if the diode or transistor sensor is floating with its leads in the air.

L-network matching

Input and output impedance matching can be performed by using a series of L-matching sections. The L match (see **fig. 3**) is a minimum Q match and is

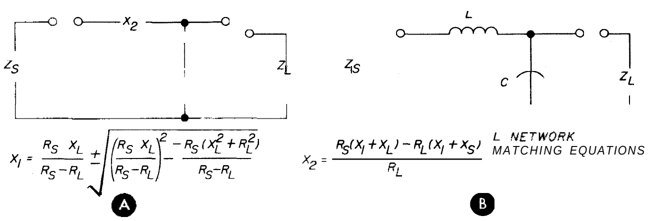
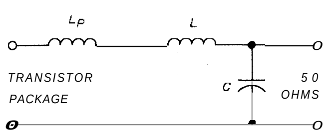


fig. 3. Schematic diagram of an L network and the necessary design equations. The lowpass prototype at the output of the transistor provides additional harmonic filtering which may make additional filters unnecessary.

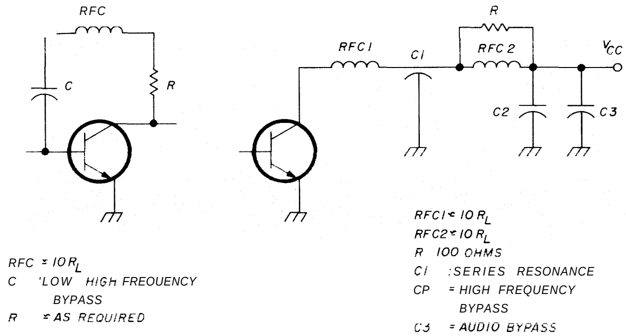
ideal for matching across an entire Amateur vhf band. The maximum impedance transformation should be less than 5:1, if possible. A 10:1 transformation, however, may be possible with real world components, if care is used.

When matching the output of the transistor to 50 ohms, a lowpass matching section reduces the required filtering at the output of the unit. My experience has shown that harmonic levels will be approximately 40 dB down, without additional filtering. The design of the first L-matching section is performed with the package inductance in mind.



low-frequency stability

Low-frequency oscillations can occur between several hundred kHz and several MHz if suitable precautions are not taken. These oscillations are quite troublesome, and the high low-frequency gain of the device and the mismatch presented to the collector can cause thermal runaway and destruction of the transistor before the thermal foldback can prevent runaway. Therefore, design must guard against this possibility by either bypassing the collector in a suitable fashion or using a feedback network that kills gain at the low frequencies



The networks shown above will cure most problems caused by the classical modes of oscillation. As an aid, here are some basic rules of thumb that can be used to rate different rf transistors and the low-inherent frequency stability:

1. Low frequency gain should be as low as possible (low h_{fe}).

2. Emitter resistance or inductance decreases gain and therefore increases low-frequency stability.

3. High collector resistivity lowers gain.

This sums up the low-frequency stability problems; other new problems, however, crop up with the increase in gain to linear circuits — vhf parasitics. The parasitics are due to excessive emitter inductance and feedback from the collector forming a grounded-base oscillator. These oscillations may be killed by a further increase in emitter inductance, with the resulting loss of gain, or by a reduction in emitter inductance and a gain increase. It is wise to note that 3.2-mm (118-inch) increases on all four emitter leads of a JØ package will cause a 3-8 dB decrease in gain; it is possible to get only 8-10 dB at best. Obviously, decreasing the emitter inductance is a superior course of action. This can be done by

1. Using copperfoil to strap the grounds of the printed circuit board together, or using eyelets to do the same.

2. Mounting the base-to-emitter and emitter-to-collector capacitors on top of the transistor leads.

This mode of oscillation is supported by the output matching network and from computer aided analysis seems to be present in all the matching networks I've compared. One way to determine if this is the case in your amplifier is to add additional capacitors to the base circuit; if you still draw excessive collector current, the chances are you have a common-base oscillator.

ham radio

expanded memory for the Autek MK-1 programmable keyer

Additional 2120AL RAM chips and switching logic allow memory capacity increase

Many on-the-air contacts confirm the popularity of the Autek Research MK-1 memory keyer. It's a superb performer and not too expensive. Recently a good friend, Mr. Meyer Minchen (AG5G), asked for my help in expanding the memory capability of his MK-1. What follows is a description of a simple and inexpensive addition to the keyer that can increase its memory capacity by as little or as much as desired.

The heart of the MK-1 memory is the versatile 2102 random-access memory (RAM) chip. It's capable of ctoring up to 1024 bits of data and comes in a 16-pin, ...al-inline package. One very important feature of this chip is its TRI-STATE™ outputs, which allow any number of 2102s to be paralleled as long as only one memory is enabled at a time. The logic state of pin 13 determines whether the chip is active. By switch selecting one – and only one – chip at a time, the total memory capacity can be increased to virtually any amount.

construction

Fig. 1 shows how simple this system actually is. In AG5G's case, five additional 2102s were added to the original memory for a total capacity of 6144 bits, a substantial increase.

AG5G's method of constructing the memory is unique and bears serious consideration by those interested in this modification. The original 2102 (Q12 on the Autek schematic) was removed and replaced with a 16-pin DIP header purchased from the local Radio Shack store (part 276-1980). A single 2102 was soldered to the exposed pins of the DIP header with the exception of pin 13 (the chip enable pin), which was bent out at a right angle. A second 2102 was then soldered directly to the first chip, piggyback style, again with the exception of pin 13. This process was continued until the last chip had been soldered to the stack.

Be very careful not to allow one or more of the chips to be turned around during this part of the

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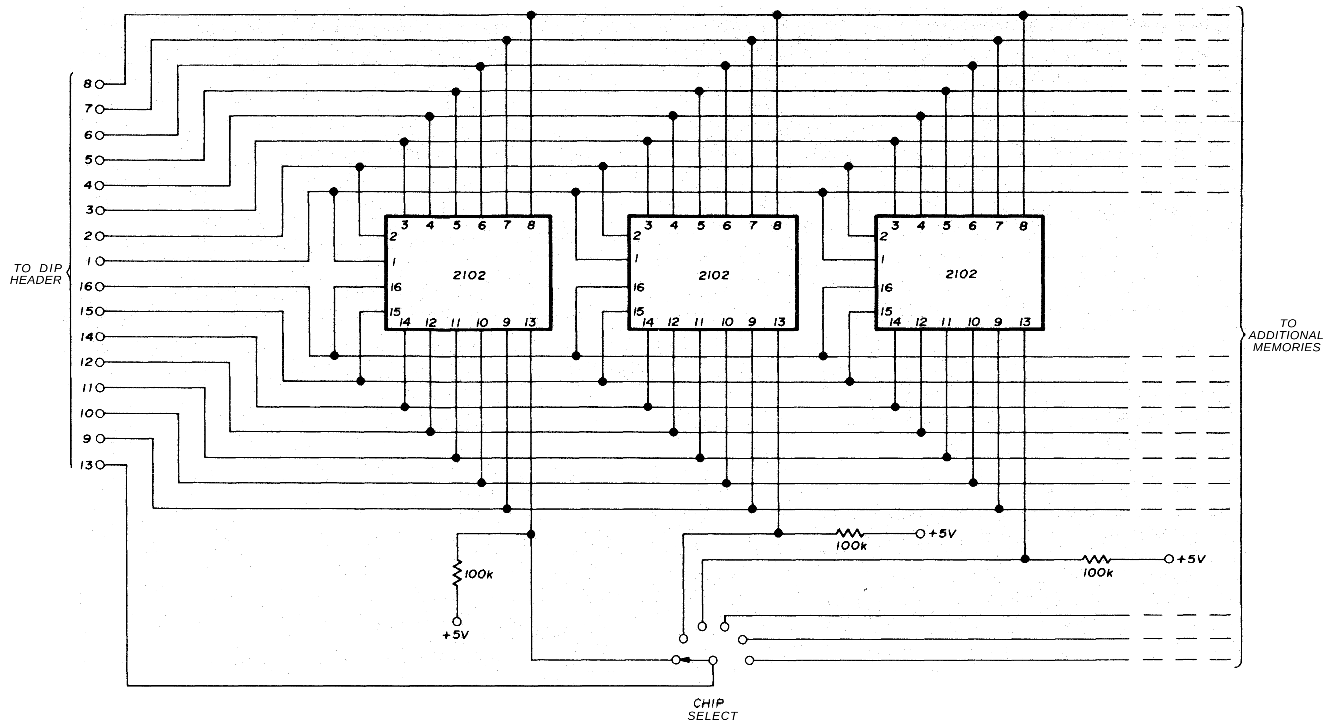

fig. 1. Expanded memory for the Autek MK-1 programmable keyer.

assembly. Use a low-wattage soldering iron, and make every effort not to get the chips too hot. A piece of ribbon cable or individual wires are then soldered from pins 13 to the rotary switch (Radio Shack 275-1386) as shown in **fig. 1.** In addition, a 100k resistor is soldered from each pin 13 to +5 Vdc, which disables the chip when not being used. Finally, a single wire is soldered from the common terminal of the rotary switch to pin 13 of the DIP header. AG5G says that there's more than enough space on the keyer front panel to mount the rotary switch.

Although five of the six memory chips are always disabled, the information stored in them is not lost or altered in any way. Normal operation of the MK-1 is not changed by this modification. There are still four separate message areas per chip. Messages C and D may still be combined at will; it's just that now you can switch between any of the desired chips, each one operating independently of the others.

power considerations

Anyone contemplating this modification to the MK-1 keyer should be aware that the built-in power supply was designed to operate the keyer and ***one*** memory chip. The supply isn't capable of supplying thecurrent that five more 2102s require. AG5G solved this problem by substituting a Radio Shack 2-amp transformer (part 273-1512) for the existing unit. He mounted the transformer on the inside of the cover and used small in-line cable connectors as disconnects to facilitate separation of case and cover.

In addition, he replaced the 1N5231B 5.1-volt zener and its 100-ohm current-limiting resistor (D15 and R32 respectively) with a three-terminal, 5-volt regulator similar to the 7805 because he thought that the zener might not handle the additional current demands. The 7805 is also available at your local Radio Shack store and the cost is under two dollars. With minimal heat-sinking the regulator runs very cool, as it's required to supply less than 100 mA when using six low-power versions of the 2102 (2102-AL).

comments from AG5G

The cost of doubling the memory capacity of the unit by adding only one memory (2102-AL) chip and a mini SPDT toggle switch is less than $4.00. My experiments indicate that the memory capacity can be doubled within the present transformer capability by adding only one 2102-AL piggyback, following the procedure covered in this article, and using a mini SPDT toggle switch in place of the 6-position rotary switch.

The cost of adding five 2102-AL chips, the 6-position switch, transformer, and voltage regulator (7805) with heatsink is less than $15.00. I used six 2102-AL chips, replacing the higher-current 2102 chip that was in the unit.

The time required for doubling the memory is less than one hour. It takes less than three hours to install the transformer, voltage regulator (7805), six chips, and switch.

It's important to note, from my experiments and tests, that all memory functions and capabilities of the MK-1 are multiplied by the number of 2102-AL chips added without any impairment of the operations of the unit, provided *the voltage is regulated and current is amply supplied.*

In building my stack of six chips piggyback, I left about 1.5 mm (1/16 inch) between each chip as I added them vertically. (I don't think the space I left for cooiing is necessary.) The transformer mounted easily, centered inside the top of the cabinet with ample space remaining to clear other parts when the top was replaced on the unit. With the unit's original transformer removed, there's some space available for mounting other items.

After installing the transformer to meet the increased requirements for current, to ensure better voltage regulation I removed the output of the two diodes (D12 and D13) from the board, joined their output together as input to a 7812 voltage regulator, and wired the output of the 7812 to where I had removed the output of D12 and D13. (The transformer that came in the unit could not handle the current requirements of the six 2102-AL chips but could meet the requirement for current for one additional 2102-AL.)

A 24-36 VCT transformer with 500-600 mA capacity will handle the requirements of the unit after adding six 2102-AL chips. The total (maximum) current requirements of the MK-1 as measured (after installing the transformer, 7812 and 7805 voltage regulators, and six 2102-AL chips) between the output of the 7812 and the load was 225 mA.

I believe that, thanks to N9AKT's creativity, this has been the easiest, lowest cost, most beneficial and rewarding project I've built in the past forty years. It substantially increases the capability and versatility of the MK-1 and makes this already excellent piece of Amateur Radio equipment even more useful.

closing notes

The simplicity of this project definitely puts it in the "weekend" category. Both the casual operator and the contester should benefit from the added memory capacity this modification affords. It's expandable one step at a time, as desired, and, best of all, it won't strain the family budget. Try it and see how much easier operating can be with an expanded memory for the Autek MK-1 programmable keyer.

ham radio

More Details? CHECK—OFF Page 110

observations on the speed of light hrough the metric system

A look at the phase velocity of EM waves in nonionized atmosphere and space

Upsetting apple carts is not the way to treat the fruits of one's labor, and I'm becoming very disturbed with people who are trying to destroy my faith in the metric system of units. After all, why should I use an EM wave phase velocity of 186,363.6 miles per second when I can use a nice even 3.0 x 10^8 meters per second?

background

Some inconsiderate person came along and concluded that this phase velocity should be 186,300 miles per second, which is about 2.998766 x 10^8 meters per second. Upon further examination it was discovered that this wasn't right either. It was concluded that the phase velocity should be 186,287 miles per second, which is about 2.998756 x 10^8 meters per second.

In 1953, Kraus got us back into the right (?) units.[1] But he concluded that this phase velocity is 2.998 x 10^8 meters per second, which is about 186,239.39 miles per second. (Rounded off, it's 186,240 mps. Take your choice.)

Then in 1962 Corson and Lorraine concluded that this phase velocity should be 2.9979 x 10^8 meters per second,[2] which is about 186,233.18 miles per second. It appears that they were either lazy or sloppy in their work, or else they were victims of round-off errors.

In 1970, the ITT staff concluded that this phase velocity should be 2.99793 x 10^8 meters per second,[3] which is about 186,235.05 miles per second, give or take a few mm. Then, less than a year ago, I saw somewhere that this phase velocity should be 2.997925 x 10^8 meters per second, which is about 186,234.73 miles per second.

Now, authorities agree that this velocity should be 2.99792456 x 10^8 meters per second,[4] which is about 186,234.71 miles per second.

In analyzing these data, I could conclude that the last twenty years of observations have shown that the EM wave phase velocity decreases at rate of about

$$V_d \approx \frac{1}{3.0(1.284)^{yr}} \text{ miles/second/year}$$

On this basis we would eventually arrive at an EM wave phase velocity of about 186,234.40 miles per second, or about 2.99791961 x 10^8 meters per second. Since neither of these figures are neat round numbers, I now wonder if those in EM-wave work should develop a new set of units, which are not related to a British King or the size of the earth. Meanwhile, if I want to stay with the MKS system of units on a *current* basis, the above observations suggest that I should reconcile my wishes and reality and round off the EM wave phase velocity to 2.99792 x 10^8 meters per second.

At the same time, I question the phase velocity of 299,792,456 meters per second given by the "authorities" cited in reference 4. First, what decimal fractional part was rounded off to the whole number of 6? Second, is the EM wave velocity in space or is it near a mass — the same as that in an atmosphere? I doubt it, and we'll have to be patient until better answers are forthcoming.

Since these observations were noted, I've read somewhere that this phase velocity is now 186,000 miles per second, or about 2.99338 x 10^8 meters per second. This is very disturbing because it suggests that we'll not be able to propagate EM waves at all by the 21,000th century, and everything propagated beforehand will be returning to us thereafter.

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3. ITT Staff, *Reference Data for Radio Engineers,* Howard W. Sarns & Co., Inc., fifth edition, March, 1970, pages3-11.
4. J. Reisert, "Matching Techniques for VHF/UHF Antennas" *ham radio,* July, 1976, page 55.

ham radio

By Harold F. Tolles, W7ITB, P.O. Box 232, Sonoita, Arizona 85637

log-periodic antennas
for the high-frequency Amateur bands

Design data for
two LP wire beams:
one for 10, 15, and 20 meters,
and one for
future ham bands
in the 10-30 MHz region

In a recent article[1] W6PYK illustrated the design of a log periodic (LP) antenna covering the 10-, 15-, and 20-meter Amateur bands using a simplified approach. The antenna design parameters were $\tau = 0.875$ (the taper factor) and $\sigma = 0.13$ (the spacing factor). This design provides a ten-element LP with a boom length of 15.6 meters (51 feet) and a gain of 6.7 dB over a dipole. All element lengths and element spacing dimensions were provided for those wishing to build a good three-band fixed-wire beam with a fair amount of gain.

measurement accuracy

You'll note that W6PYK[1] rounds off the values for element lengths and spacing distances. Some authors describing LP-antenna designs for Amateurs often show these measurements with decimals to the fourth, fifth, or even sixth place! Probably many Amateurs have been discouraged after seeing this amount of measurement precision. No doubt some of this showmanship is to impress the reader that the author has access to a computer. But as Paul, W6PYK, says, "The LP is very forgiving of construction and design tolerances." I've found this to be true in my LP antenna designs.

Over the past eight years I've assembled and tested more than thirty high-frequency LP antennas for various frequency ranges and gains including monobanders and two- and three-banders. Using a four-function calculator, lengths and spacings can be shown to four or five decimal places, but I always round off to no more than two places for ease in measuring, usually converting the measurements to feet and inches (meters and centimeters) for the lower-frequency bands and certainly no closer than 6 mm (0.25 inch) for the higher-frequency bands, which is plenty close.

By George E. Smith, W4AEO, in collaboration with **Paul A. Scholz, W6PYK.** Mr. Smith's address is 1816 Brevard Place, Camden, South Carolina 29020. Mr. Scholz's address is 12731 Jimeno Avenue, Granada Hills, California 91344

calculation example

Let's run through some LP design calculations using W6PYK's data (reference 1). We'll use his **table 1** and antenna 3 ($B = 2$), with fourteen elements and $\ell/\lambda = 1.37$.

table 1. 14-element LP designed to $\tau = 0.917$ **and** $\sigma = 0.17$. $\ell/\lambda = 1.37$; *gain* = 8 *dBd*

element	length meters	(feet)		spacing distance meters	(feet)
1	10.0	(33.4)	S1	3.7	(12.0)
2	9.4	(30.7)	S2	3.4	(11.0)
3	8.5	(28.0)	S3	3.0	(10.0)
4	7.9	(25.8)	S4	2.8	(9.2)
5	7.2	(23.6)	S5	2.6	(8.5)
6	6.6	(21.7)	S6	2.4	(7.8)
7	6.0	(19.9)	S7	2.2	(7.1)
8	5.6	(18.2)	S8	2.0	(6.5)
9	5.1	(16.7)	S9	1.8	(6.0)
10	4.7	(15.3)	S10	1.6	(5.4)
11	4.3	(14.0)	S11	1.5	(5.0)
12	3.9	(12.9)	S12	1.4	(4.6)
13	3.6	(11.8)	S13	1.3	(4.2)
14	3.3	(10.8)	total:	29.7	(97.3)

Since we start with 14 MHz as our low-end cutoff frequency, f_L, the free-space wavelength, λ_0, will equal 984114 or 21.4 meters (70.3 feet). The boom length will be 29 meters (94.9 feet). If these dimensions aren't too large for the available space, we proceed as follows:

1. As our $f_L = 14$ **MHz,** the length of the E1 rear element will be $468/14 = 10$ *meters (33.4 feet)*. Under the τ column of the table note that $\tau = 0.917$ and $\sigma = 0.17$. These parameters determine the taper factor, τ, and spacing factor, σ, for the remaining calculations.

2. Next we calculate the spacing distance between E1 and E2. $S1 = \sigma \times \lambda_0 = 0.17 \times 70.3 = 11.96$ *feet*. Make it 3.7 meters (12 feet).

3. Next we calculate the other element lengths, E2, E3 . . . En, where En is the *nth* element. From **step 1** $E1 = 10$ *meters (33.4 feet)*. $E2 = E1 \times \tau = 9.4$ *meters (30.7 feet)*. The remaining element lengths are calculated similarly.

4. The remaining spacing distance, S2, S3 . . . Sn are calculated thus: $S2 = S1 \times \tau = 3.3$ *meters (11 feet)*. $S3 = S2 \times \tau = 3$ *meters (10 feet)* and so on for S4. . . Sn.

design aid

For those who don't wish to compute an LP using W6PYK's easy design method, complete dimensions are given in **tables 1** and **2** for two 14-29-MHz LPs for 20, 15, and 10 meters.

about cost

Some hams with whom I've discussed LPs feel that the LP requires quite a bit of wire for the number of bands covered, especially one covering a 2:1 ($B = 2$) frequency range (one octave). This is true for a large LP covering 80 and 40 or even 40 and 20 meters; however, since an LP covering 14-29 MHz includes three bands (20, 15, and 10) you get more for the money. My antenna for these bands cost about $35.00 to $50.00 for wire, nylon line, and lucite insulators (coax, baluns, and towers not included). For these reasons I feel that LPs for 80 or 40 meters should be limited to a monoband LP ($B = 1$) since the range between 4.0 and 7.0 MHz or 7.3 and 14 MHz isn't needed.

The least expensive, or rather the most "cost effective," LPs built here have been those designed for 14-21.5 MHz ($B = 1.5$) for 20 and 15 meters only. The first LP put up here was a seven-element array with a boom length of 11 meters (37.5 feet). It was described in reference 2.

table 2. 19-element LP designed to $\tau = 0.943$ **and** $\sigma = 0.175$. $\ell/\lambda = 1.97$; *gain* = 8.9 *dBd*.

element	length meters	(feet)		spacing distance meters	(feet)
1	10.0	(33.4)	S1	3.7	(12.3)
2	9.6	(31.5)	S2	3.5	(11.6)
3	9.1	(29.7)	S3	3.3	(10.9)
4	8.5	(28.0)	S4	3.1	(10.3)
5	8.1	(26.4)	S5	3.0	(9.7)
6	7.6	(24.9)	S6	2.8	(9.2)
7	7.2	(23.5)	S7	2.7	(8.7)
8	6.8	(22.2)	S8	2.5	(8.2)
9	6.4	(20.9)	S9	2.3	(7.7)
10	6.0	(19.7)	S10	2.2	(7.3)
11	5.7	(18.6)	S11	2.1	(6.8)
12	5.3	(17.5)	S12	2.0	(6.5)
13	5.0	(16.5)	S13	1.9	(6.1)
14	4.8	(15.6)	S14	1.7	(5.7)
15	4.5	(14.7)	S15	1.6	(5.4)
16	4.2	(13.9)	S16	1.5	(5.1)
17	4.0	(13.1)	S17	1.4	(4.8)
18	3.7	(12.2)	S18	1.3	(4.5)
19	3.5	(11.6)	total:	43.0	(140.8)

Fig. 1 is a drawing of the 14-29-MHz LP suggested by W6PYK for 20, 15, and 10 meters with an array length of 15.6 meters (51 feet). Note how the three cells for 20, 15, and 10 meters overlap in a fairly short (array length) LP. All elements are used (no waste of wire), as compared with an LP for 80 and 40 or 40 and 20 meters.

a 10-30 MHz-LP for future Amateur bands

An LP designed to cover 10-30 MHz may be of future interest, as W6PYK mentioned in his article.[1] Hopefully, we'll be awarded the proposed ham

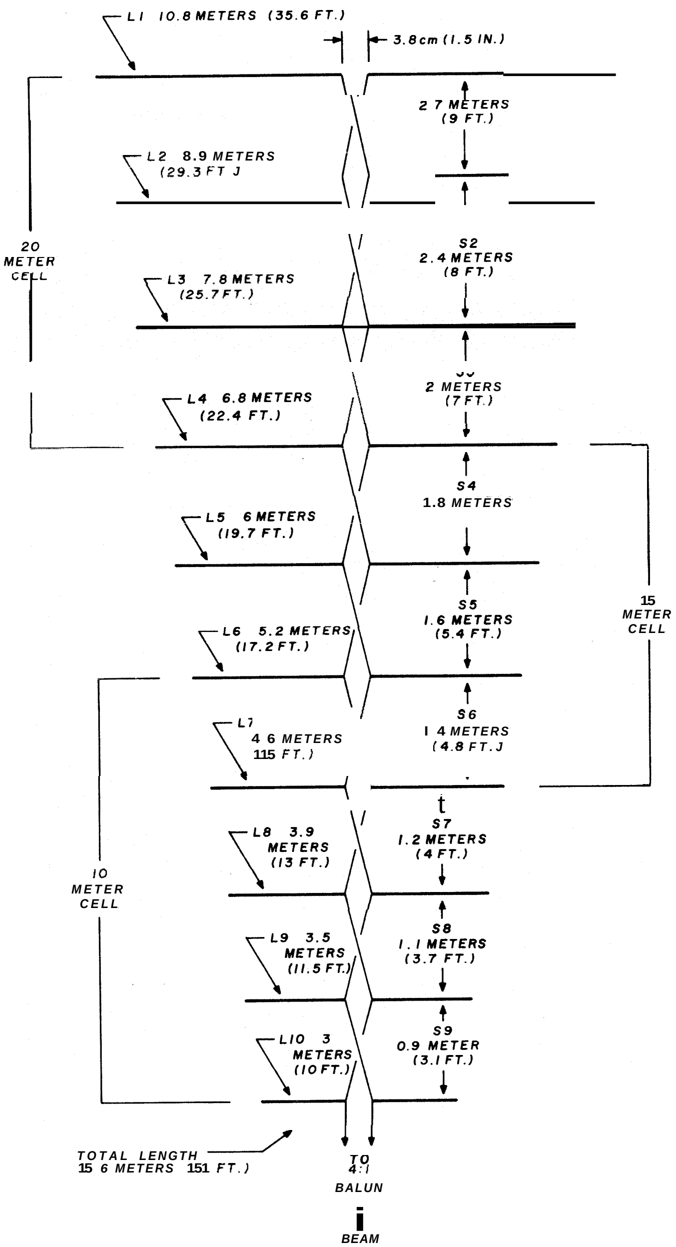

fig. 1. Ten-element 14-29-MHz log periodic antenna for 10, 15, and 20 meters. Designed to τ = 0.875 and σ = 0.13. Gain = 6.7 dBd. Note how the three "cells" for 10, 15, and 20 meters overlap — no waste of wire.

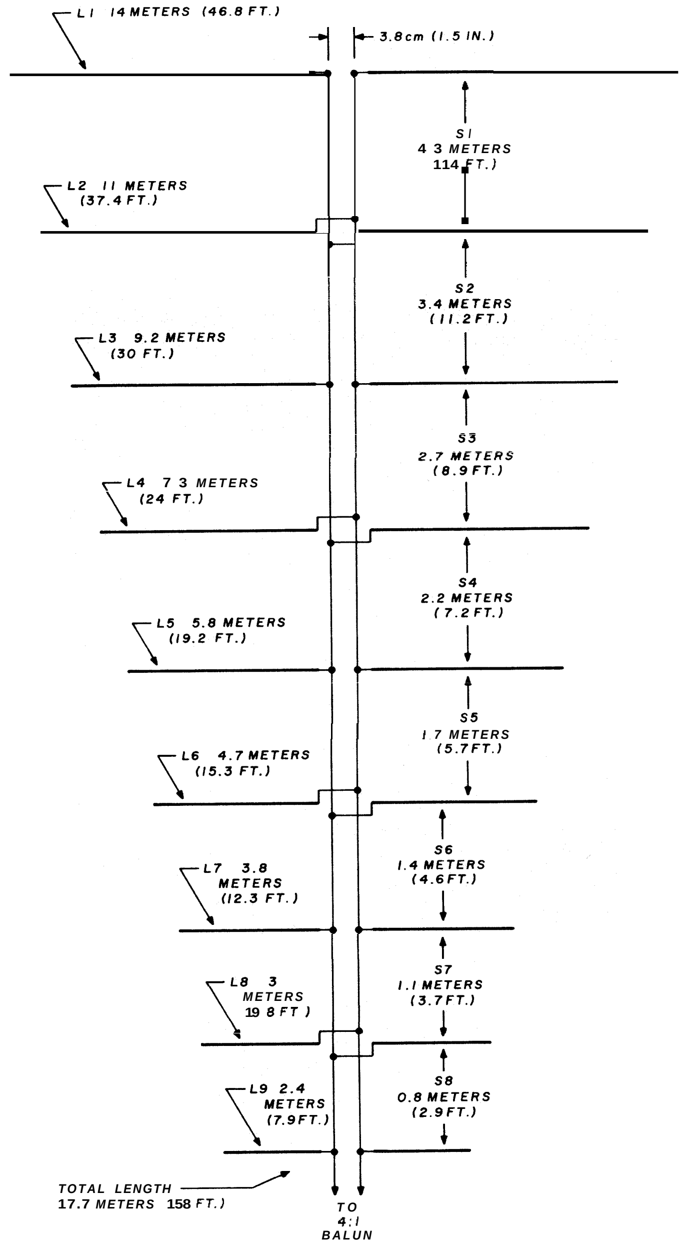

fig. 2. Nine-element LP designed for 10-30 MHz, using τ = 0.80 and σ = 0.142. Gain = 5.9 dBd. This will cover future Amateur bands as proposed in the WARC conference in Geneva.

bands: 10.1, 18.1, and 25.25 MHz at the forthcoming WARC Geneva conference. An LP designed to cover the entire 10-30-MHz spectrum will then be usable on six bands: 10.1, 14, 18.1, 21, 25.25 and 28 MHz.

As Paul states it would be difficult, and quite a mechanical challenge, to design a practical six-band rotatable Yagi similar to the present triband Yagis.

Fig. 2 shows dimensions for a nine-element (B = 3) 10-30 MHz LP designed to τ = 0.8 and σ = 0.142, which gives an array length of 17.7 meters (58 feet) and a 5.9-dBd gain. Although this is a moderate gain, it's probably about as good as some of the present tribanders and about as good as can be expected from an LP with such a short boom length. A big advantage is that gain and SWR are relatively constant over the entire range of 10-30 MHz. A further advantage is that it covers our present 14-, 21-, and 28-MHz bands plus the proposed 10.1-, 18.1-, and 25.5-MHz bands.

references

1. Paul A. Scholz, W6PYK, "Another Approach to Log-Periodic Antenna Design," *ham radio,* December, 1979, page **34.**
2. George E. Smith, W4AEO, "Log Periodic Beam for 15 and 20 Meters," *ham radio,* May, 1974, page 6.

ham radio

communications audio processor for reception

This article describes a system aimed at the special needs of Amateur Radio CW and voice signal reception. It's called the Comm Audio processor (CAP) and uses a combination of special voice-frequency filters, binaural synthesis, Tone-Tag modulation, and controlled antiphasic noise.

design considerations

Much attention was given to aspects of human engineering, signal environment, and operation in the CAP design. For example, most ham receivers provide plenty of band spread and a high-quality, calibrated tuning control, so selecting frequency in an add-on unit would be redundant. Accordingly, the CAP has been designed so that once everything is set up for a given voice or CW mode, most attention and operational control remains within your basic receiver. Aside from human engineering aspects, our very complex signal environment – both manmade and natural – contributed to the system design. In this respect the FCC's frequency allocation system heavily influenced all of the hardware because, through their regulation, there are at least three major radio signal environments in the high frequency spectrum: First, clear-channel frequencies are available for many industrial, public-broadcast, and governmental activities. Second, there are channelized frequency blocks, such as CB and many commercial and government allocations. Finally there is our type. We hams are required to fit our signals into frequency *bands* but aren't regimented into *channels* (so a few freedoms do remain!).

Because we have freedom of choice on any frequency within an allocated band, we're able to get five to ten times more useful communications functions in the ham phone bands than can be obtained in FCC channelized allocations of equal bandwidth, and up to fifty times or more in the CW bands. Certainly this signal density is such that QRM is one of the Q codes most popular; but the pressure has resulted, and will continue to result, in transmitter, receiver, antenna, and control design improvements. Clearly, our signal environment demands that we do not transmit on or receive unnecessary frequencies.

In addition to the human engineering and signal

environmental aspects, there is the belief on my part that most hams, particularly DXers, are interested in enhancing weak-signal reception when there are strong signals nearby. The key to this is retaining an awareness of the noise floor and maintaining linearity in the signal path throughout an entire receiving system. A strong signal will suppress a weak signal either through the use of agc or by allowing compression of any stage. CAP was designed in consideration of these objectives.

In the system shown in **fig. 1**,* design effort has been directed toward an audio system for the Amateur brand of voice and CW communications. The unit contains the following key features, none of which are found to a refined degree in manufactured receivers:

1. Voice-shaped filter – thirteen poles to improve signal-to-noise ratio and reduce QRM by rejecting a broad spectrum between the first and second voice formants.

2. Binaural synthesis for both voice and CW.

3. Tone-Tag – the system that distinctively modulates any signal tuned to 750 ± 50 Hz, the binaural crossover frequency.

4. Nine-pole, 100-Hz bandwidth, stagger-tuned filter for super-steep skirts with a passive prefilter.

5. A continuously adjustable pink (soft) noise source – the how and why of this is covered in detail.

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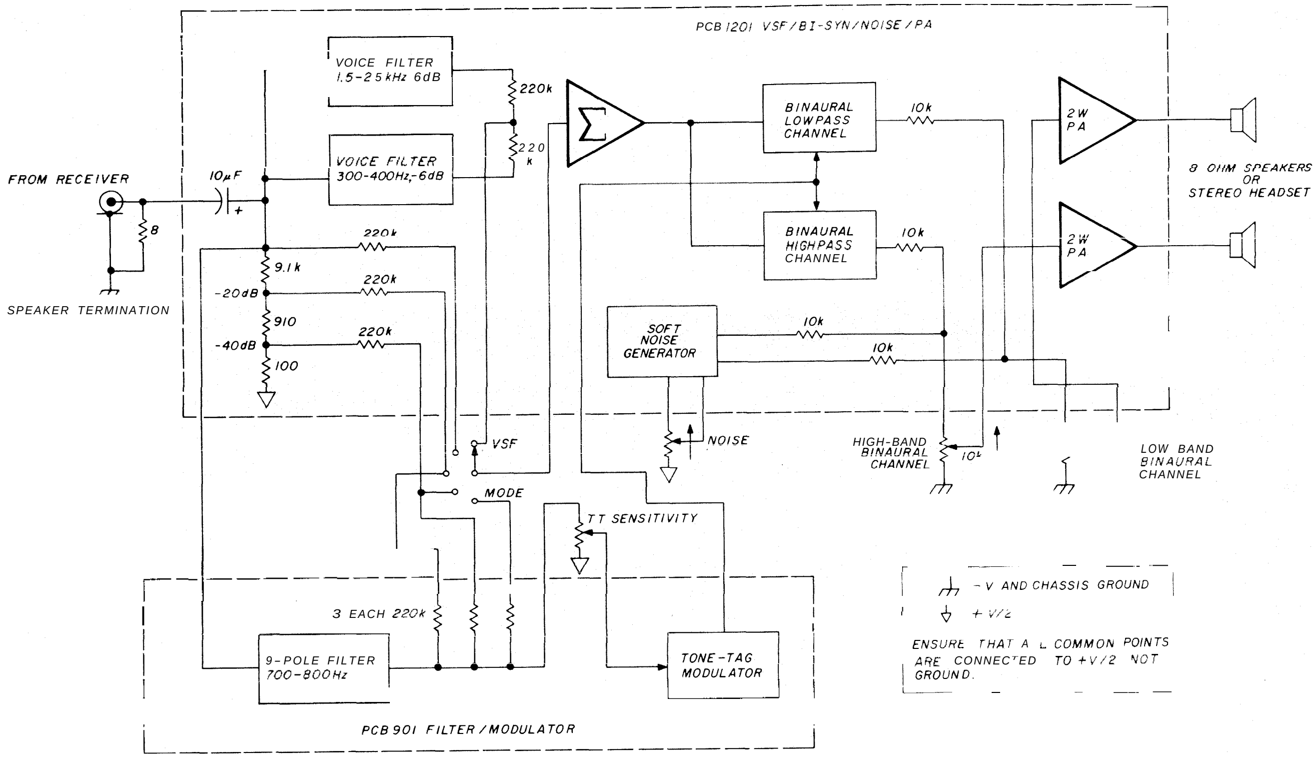


fig. 1. Comm Audio Processor.

Over the years many voice-manipulating systems have been proposed and developed, all aimed at making more effective use of the actual spectral distribution in the human voice. Most recently Dr. Richard W. Harris and J. F. Cleveland developed an excellent technique.[1] In this system, the voice spectrum is transformed following your transmitter's microphone output, as shown in **fig. 2.** The second and third voice formants (contained in the dashed area) are folded by a 3.1-kHz oscillator, mixer, and filters, which results in a band between about 300 and 1600 Hz for transmission.

An inverse system, again using a 3.1-kHz oscillator, mixer, and filters at the received audio end, unfolds (returns) second and third formants to their proper position (dashed). The amount of bandwidth actually involved is not reduced but merely rearranged.

Although frequency tolerances will have to be tightened, this system can nearly double the effective spectrum now used for voice-only, channelized communications. Since it can be applied to services such as CB, police, fire, and many other industrial, commercial, and governmental functions, frequency companding of this type could reduce the pressures to gobble up the ham bands.

*A drilled and etched PC-board set is available for $9.95 postpaid. Send a self-addressed, stamped envelope for a complete product listing to Hildreth Engineering, P.O. Box 6003, Sunnyvale, California 94088. Phone (408) 245-3279.

The Comm Audio Processor design is based on the reality that those working in ham frequency bands up to 28 MHz are involved in nonchannelized communications, which is also heavily influenced by skip conditions.

We play billiards with the ionosphere, which reduces the effectivity of a frequency compandor. If you can't tell, because of skip, where other stations may be, how do you know where to place your bundles of voice energy?

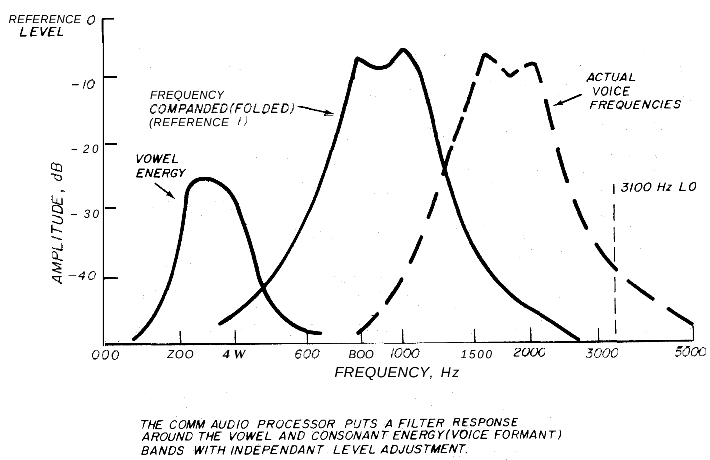


fig. 2. Folded voice-spectrum on approximated energy distribution. The Comm Audio Processor puts a filter response around the vowel and consonant energy (voice formant) bands with independent level adjustment.

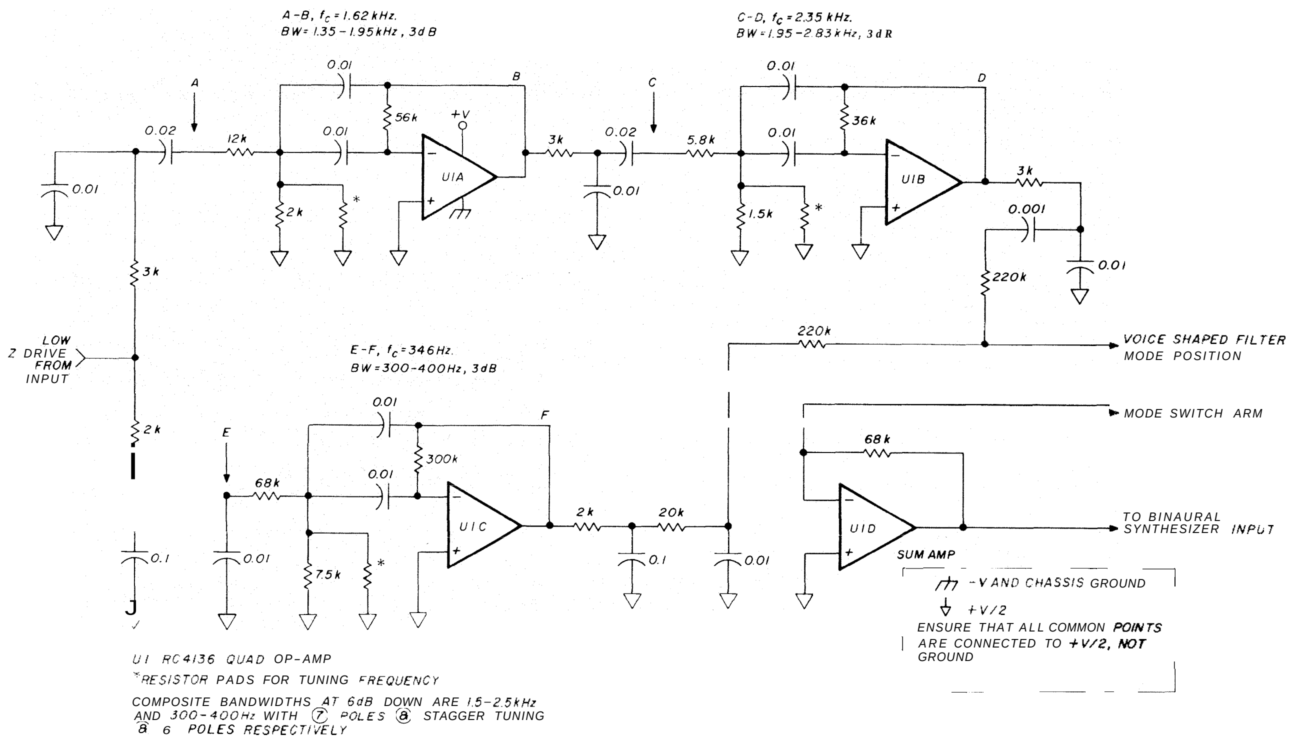


fig. 3. Voice-shaped filter and sum amplifier. Composite bandwidths at 6-dB down are 1.5-2.5 kHz and 300-400 Hz with seven poles and stagger tuning and six poles respectively.

Fig. 3 shows CAP's voice-shaped filter (VSF). In this case, it's not mandatory to roll transmitted voice energy off sharply above 2.5 kHz. It is desirable, however, because 2.5 kHz of bandwidth (or, more accurately, response from 1.5 to 2.5 kHz) will supply the basic needs for voice communications; the human voice does emit unnecessary energy to 10 kHz or so. Even though this energy can be filtered at the receiving end by someone listening to *you,* this energy will appear as lower frequencies – thus unfilterable – for other stations up to 8 or 9 kHz away. The design shown in **fig.** 3 can also be used in your microphone circuit. It would make other hams in our crowded bands very happy.

binaural synthesizer

To synthesize a binaural sound environment, the audio output passband from any receiver is divided (with reasonably sharp filters) into two parts. Frequencies below 750 Hz are fed to one speaker and frequencies above 750 Hz are fed to another. The speakers are located as you would place them for stereo listening. Speaker locations and their resulting stereo amplitude potentials are shown in **fig. 4.** A maximum differential of 7 dB (only slightly more than one S unit of signal strength) holds through most of a typical voice communications bandwidth from about 500 Hz to 3 kHz.[2] Below 500 Hz the human stereo potential diminishes; below 300 Hz it's gone.

Since the 7-dB differential occurs because of the "sound shadow" presented by your head, you may leap to the thought of a huge improvement by using stereo headsets. But wait. It won't happen. You can get a better amplitude differential with headsets, but that's not the whole story. Our brain processes a sound wave, or many, in terms of much more than

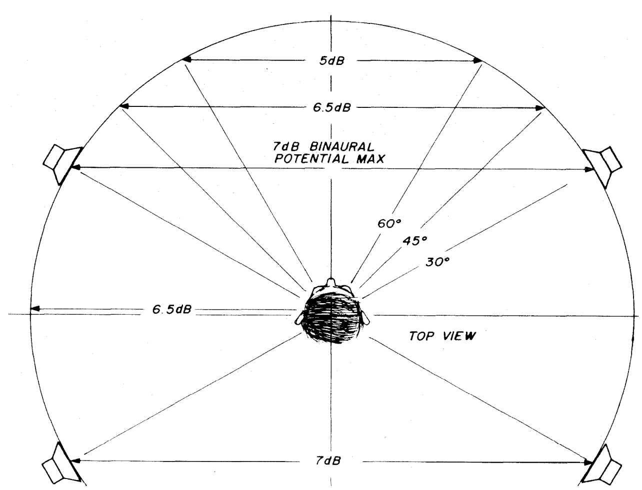


fig. 4. Locating speakers for best binaural amplitude separation. The maximum effective binaural differential based on intensity is obtained when speakers are located forward of, or behind, an imaginary line through your ears by about 30 degrees.

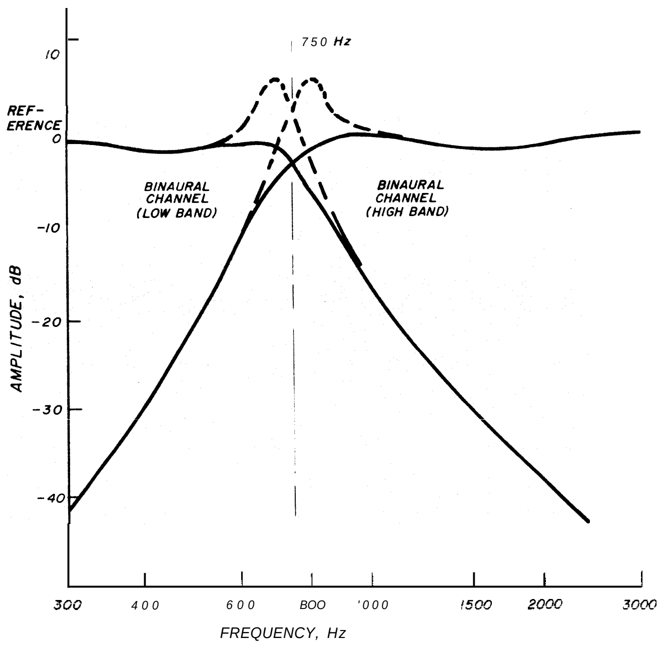


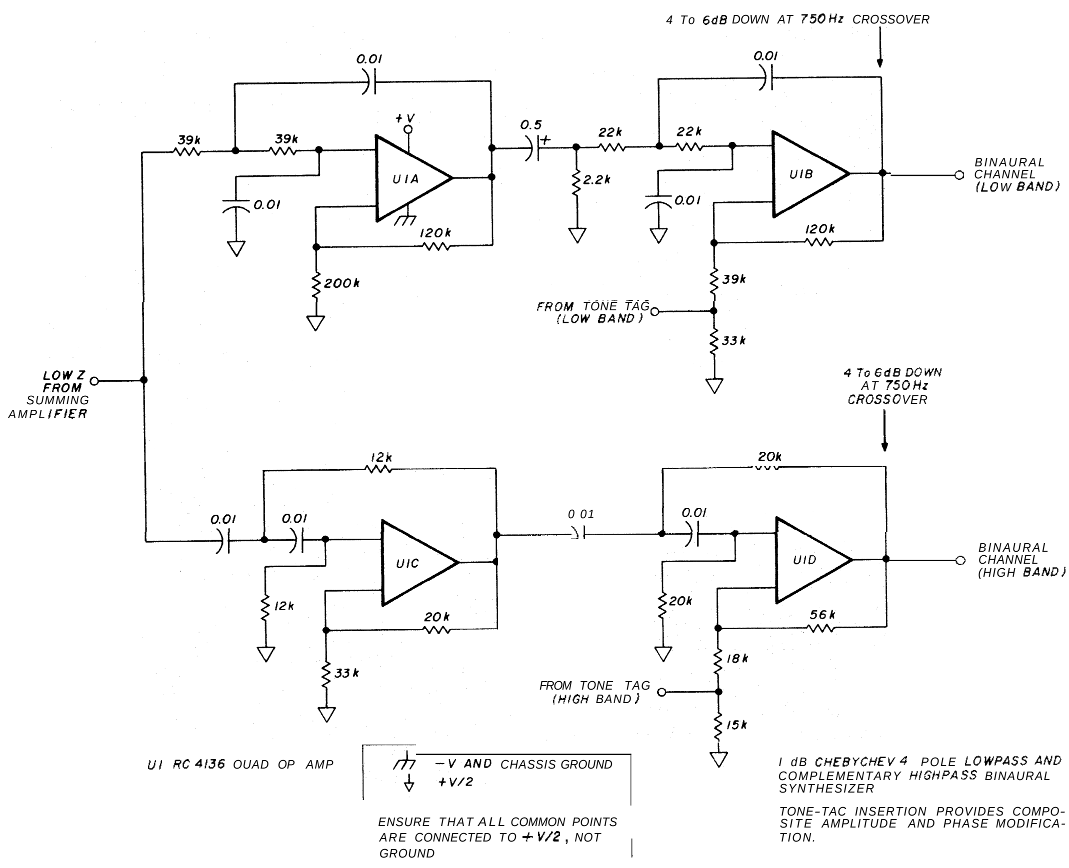
fig. 5 (above). Binaural response. Solid lines show approximate I-dB Chebychev response. Tone-Tag changes response to dashed lines 100 times/second.

fig. 6 (below). I-dB Chebychev four-pole lowpass and complementary highpass binaural synthesizer. Tone-Tag insertion provides composite amplitude and phase modulation.

just relative amplitude. There are also time-of-arrival and phase variation phenomena. Headsets wipe these out in our system. When using speakers with the binaural synthesis method used here, sounds originate to the right or left in space, just as in nature. However, when a CW beat note is tuned to the crossover frequency of 750 Hz, the brain gets equal information in terms of amplitude and time-of-arrival from both left and right azimuths, resulting in the impression of "surround sound." Headsets work well, of course, but you just won't get the improvement you may expect.

In the CAP, I-dB Chebychev four-pole lowpass and complementary highpass filters are used, with a frequency crossover (equal energy from both sides) at 750 Hz. The voltage control voltage source (VCVS) active filter form enables the insertion of Tone-Tag modulation. **Fig. 5** shows a representative response (solid lines). Tone-Tag modulation alternates the response' between the solid and dashed lines, but only as driven by a signal tuned to 750 ± 50 Hz through a 9-pole filter and at a nominal 100-Hz rate. **Fig. 6** is a complete schematic.

In the first article describing a binaural synthesizer for CW receptions a simple cascade of 2-pole filters was used. To improve the crossover slope without adding more poles and to improve Tone-Tag inser-



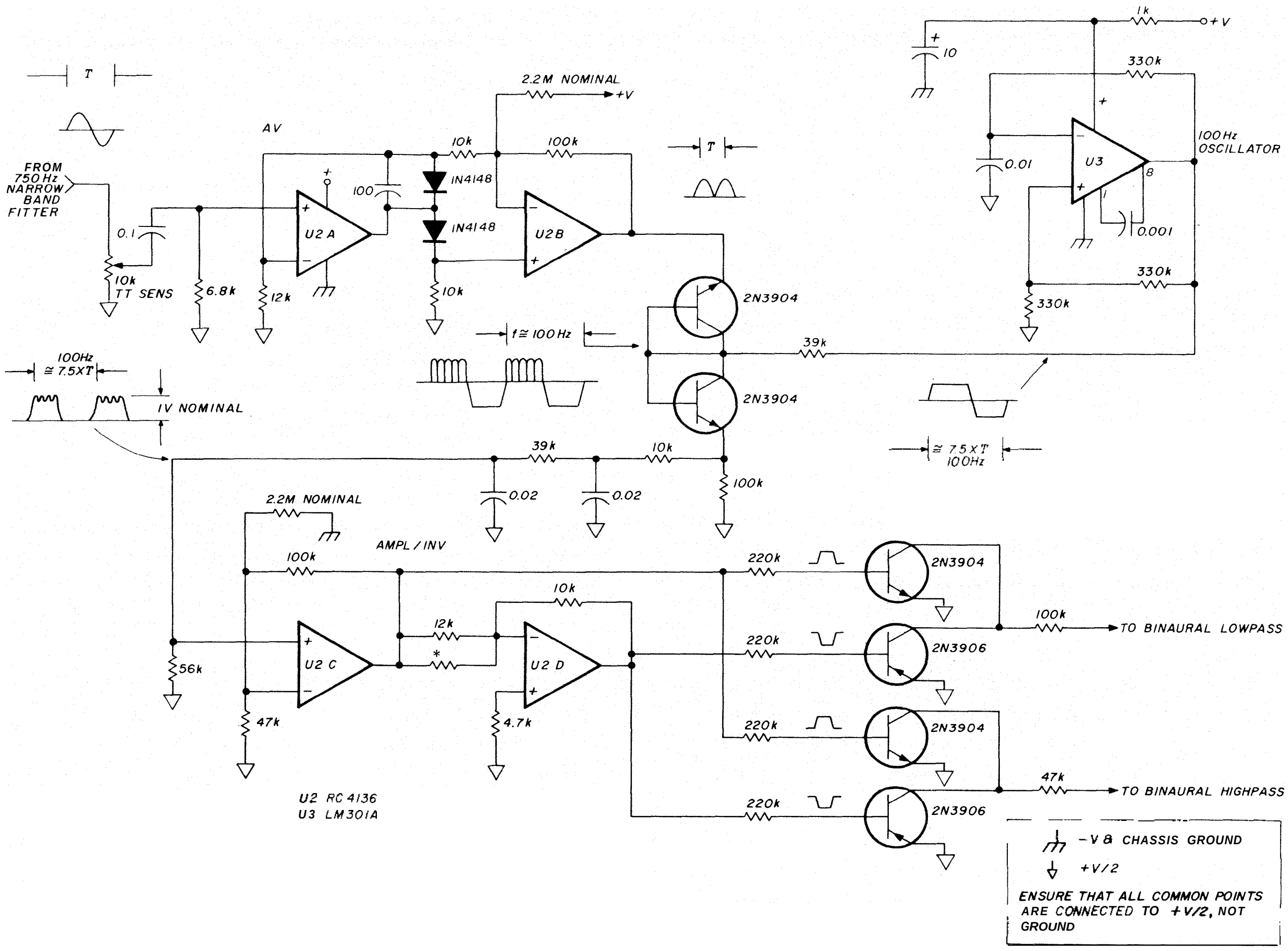


fig. 7. Tone-Tag modulator. Resistor marked with an asterisk at U2D – input is adjusted for minimum tone output from the binaural lowpass when at least 100 mW of 750 Hz signal is at the input above and the signal input line of the binaural section is grounded to signal ground (V/2). Resistor nominal value is 57k.

tion, the design has been changed to the Chebychev type. (Burr-Brown[4] supplies an excellent set of tables for Butterworth, Bessel, and Chebychev designs.) To improve antiphasic response enabled by the binaural system, the low- and high-pass systems are adjusted to be about 4-6 dB down at the 750-Hz crossover frequency.

Tone-Tag

The Tone-Tag system first used (model 1100) was based on simplicity.[5] Now, however, refinement has set in. Our model 1500 changed the modulator from diodes to a balanced bipolar technique to improve weak-signal performance in a noisy background. And CAP adds still further, although minor, improvement. **Fig. 7** shows the heart of our current Tone-Tag modulator design.

The absolute value (AV) circuit – also called a precision rectifier – transforms any incoming signal to a 750 ±50 Hz rectified, unfiltered positive output with a gain of ten. When no signal is present, positive excursions of a nominal 100-Hz oscillator are clamped to a nominal +0.5 volt by a diode-connected transistor working against the low-impedance zero output of the AV op-amp. A second diode-connected transistor, turned around, subtracts most of the +0.5 volt signal and rejects the negative half cycles of the 100-Hz oscillator signal. When an input signal appears, positive excursions appear at the AV output, and the diode-connected transistor releases its clamp on the 100-Hz oscillator, which results in the waveforms shown.

Since more current normally flows in the first diode than in the second, the second diode voltage doesn't quite offset that of the first. The resistor from – of U2B to +V provides a small compensating offset at the U2B output, resulting in a near zero voltage at U2C input.

When no 750-Hz signal is present, U2C provides a nominal gain and a small offset, which sets the modulator transistors to near conduction. U2D inverts U2C output to provide a balanced drive to the com-

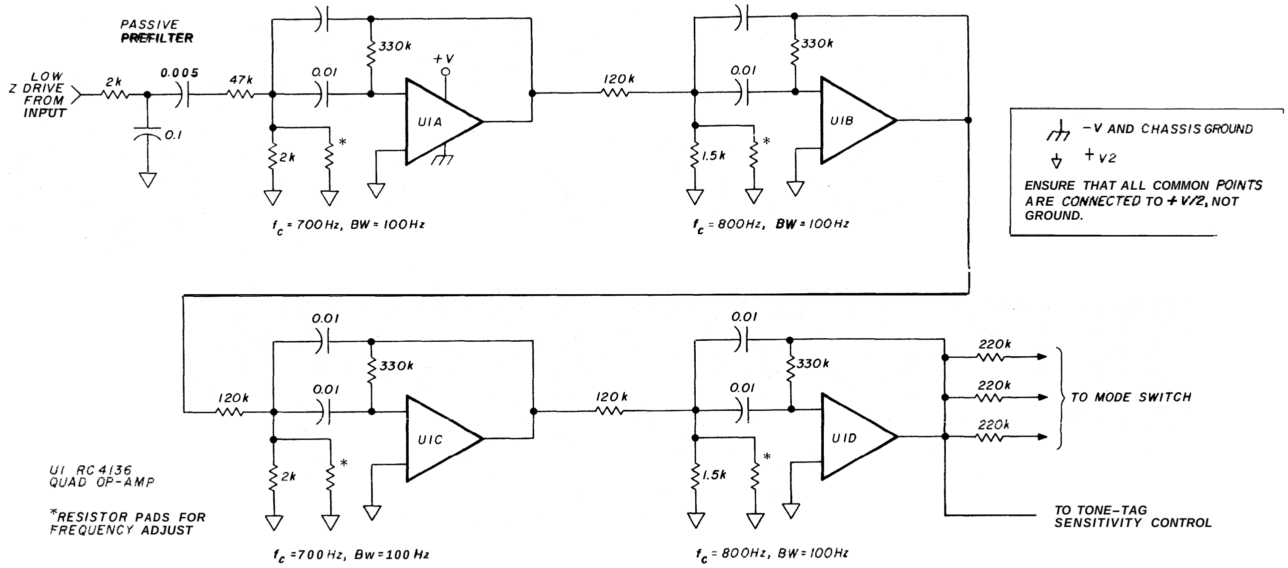


fig. 8. Nine-pole, 100-Hz bandwidth stagger-tuned, 750-Hz filter. Passive prefilter provides pole at 750 Hz with phase-corrective element.

plementary modulator transistor switches. This action is precisely set with the padding resistor marked with an asterisk in fig. 7. Resistor-buffered switch outputs are fed to the binaural synthesizer (fig. 6).

9-pole narrowband filter

A narrowband filter centered at 750 Hz, the binaural crossover frequency, is required to drive the Tone-Tag AV input circuit. Since a filter of this type is popular in its own right, a very steep-skirted, stagger-tuned device was used. A passive prefilter is used in addition to quality RC4136 op-amps to mitigate the effects of transient intermodulation distortion that may be induced by impulse noise or transients (key clicks). The design also minimizes small-signal compression by strong nearby signals acting on the first, and most susceptible, filter stage. The composite is the 9-pole filter shown in fig. 8.

good noise – bad noise

White noise or pink noise (shaped) is often added to communications circuits to mask distracting forms of low-level interference. In most of its applications, this form of noise is soft and fluffy, even pleasant. Pink noise generators are even found in such diverse applications as sleep aids and pain suppressors for some dental operations. When properly used, this is "good" noise.

Then there's that other kind, the abrasive, or "bad," noise that ranges between 10 and 40 dB thick depending on frequency, time of day or year, and your location6 (fig. 9). This noise, available at your antenna terminals, is the sum of many sources. Added to nature's atmospheric electrical disturbances, manmade electrical perturbations mix in to generate the din. Electrical circuits being made or broken create impulse-like radiation that chirps through the spectrum. Auto ignition, power leak . . . the list can go on and on.

Thermal noise – and, to a large extent, galactic noise – can be thought of as a statistical average of a very large number of low-energy electrical perturbations. At the same time, manmade and atmospheric noise comes from a much smaller number of relatively high-energy events. When heard, these noise types influence our receivers and our nervous systems very differently. The sum of the two noise

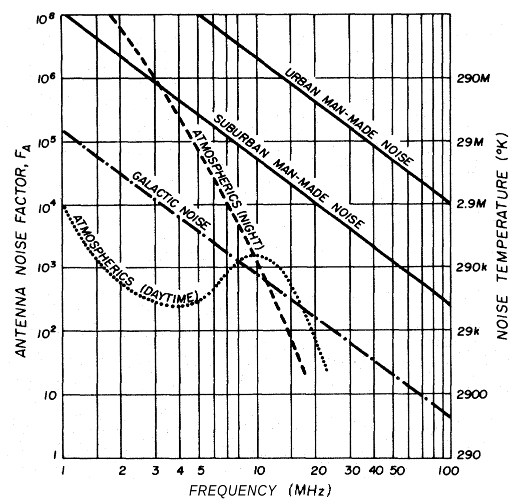


fig. 9. Receiver sensitivity is limited by the external available noise power, which varies with frequency. For a quiet, rural location, galactic noise is the limiting factor down to about 18 MHz, and atmospheric noise dominates below 18 MHz.

types could be described as a foam mattress (thermal noise) perforated by a family of random spikes (impulse noise and signals).

In practice, something approximating "good" noise may be heard by simply removing the antenna from your receiver and listening with an audio bandwidth of 2-3 kHz or more. The other stuff is what you hear when your antenna is reconnected and you tune with high gain to a spot where no signals are present (it's assumed your receiver is not noise-figure limited). Now, with antenna-received noise present, connect your receiver output to one of those ubiquitous multipole active audio filters. With the rf gain "up," you'll hear that tinkling roar that drives so many of us to look for a better solution, for, even though the little devices do a good job in some respects, they can't filter those noise components that sound like a signal itself. In addition, sporadic, impulse-like noise tweeks the filter to produce sounds approximating what you get if you rapidly strum the strings of a violin while damping them with the other hand. It doesn't ring long, to be sure, but the sounds produced are much too like the code structures in a weak CW signal.

listening experiments

In the Tone-Tag system first published[5], I made an attempt to get most of the benefits of a multipole narrow-band filter without listening "through" it. However to use Tone-Tag under high-sensitivity levels, coincidental to feeding the binaural synthesizer signal input port with the output of a narrow-band filter (not the original intent), I found that the filter tinkling sound along with the desired signal was modulated – thus boosted – by Tone-Tag. Although the level was not too high, it was objectionable.

On a monaural basis I decided to try adding soft noise to mask those elements of noise that appear at a narrow-band filter output with the desired signal. The resulting signal sounded reasonably smooth; but weak signals suffered slightly under the noise load. The next step was to add the soft noise in-phase to binaural audio channels while the signal plus residual filter noise irritants were fed to the two channels in phase opposition, or approximately so.

It's well known that copying a signal in noise with a binaural antiphasic noise combination has a 15-20 dB advantage over the monaural case,[7] but the question was this: If a weak signal can be clearly heard when added to white noise in this way, will the signal's filtered residual noise components be masked by the white noise? Results were much better than expected!

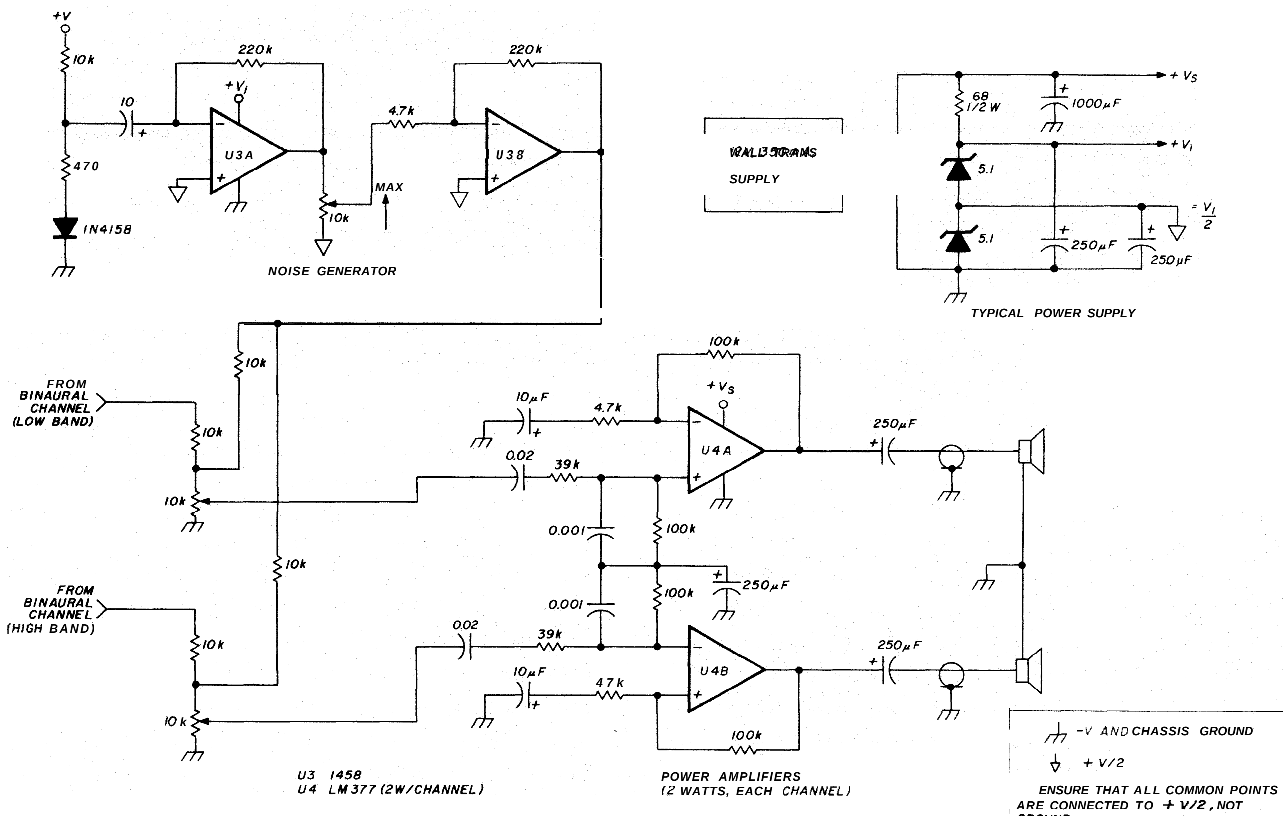


fig. 10. Noise source and audio power amplifiers.

I've spent many hours listening to compare the readability of a weak signal with or without soft noise added. Clearly it's more pleasing to listen to a signal plus soft noise, and that was expected. But readability, in many cases, was *improved,* which was not expected to the degree I found. Actually, I'd hoped to minimize signal readability degradation with the addition of antiphasic binaural noise. It now appears that when a weak signal and abrasive (impulse) noise pass through a narrowband filter, noise elements are transformed into sounds that are so much like the signal's coded structure that they compete with it. The result is a less readable signal. But with noise added in the right way and amount. readability is actually improved.

In no case during these listening tests could I detect signals with the white noise added that could not be detected without the added noise. As expected there was no basic improvement in signal-to-noise ratio. Sut a signal could often be copied with the added noise where it could be heard but not copied when the noise was removed. To date I've found no case where the addition of soft noise has made a signal less readable than without it. Clearly, of course, added noise has usually been at a level just necessary to mask the undesirable noise elements coming out of a narrowband filter. Happily, added noise benefits the critical application of the Tone-Tag referred to above as well!

Since the existing binaural synthesizer already supplies the desired antiphasic* condition, summing in-phase noise at the power-amplifier input adds very little to the cost in the CAP design — and one more control knob. **Fig. 10** shows the noise generator and how it's added at the junction of the binaural synthesizer and power amplifiers. Although most emphasis has been on the judicious addition of soft noise in CW reception, the feature can also be used to some advantage in the reception of voice signals.

operating with CAP

The main mode control switch enables the selection of the shaped-voice filter, a nominal 2.5-kHz flat bandwidth for either voice or CW or three CW filter positions. Any selected mode is fed into the binaural synthesizer filters. **Fig. 11** illustrates. Independent power output controls are provided for each binaural synthesizer channel, which allows compensation for different speaker or headset efficiencies or for special effects. The two remaining controls include Tone-Tag sensitivity and amplitude control for the soft-noise source.

When the voice-shaped filter is selected, tuning your SSB receiver will be easier — more like tuning an a-m signal than when the usual 2.5-kHz flat band-pass filter is used. When the frequency band from about 400-1500 Hz is deeply rejected, a slight-to-moderately mistuned station will not produce much output energy in this band, thus it more quickly disappears as you tune. In addition, under critical conditions often found on 14 MHz, for example, the binaural low-band from 300-400 Hz may be reduced to or near zero with the independent channel output control. Under this condition, the 300-400 Hz low-band segment is still available at approximately 30 dB down in the high-band binaural channel along with its normal 1.5-2.5 kHz response. The noise bandwidth in this super-sharp condition is only about 1 kHz. Clearly recognizable voice is available although it will be nearly devoid of character. Bringing ***up*** the binaural low-band control will progressively enable individual signal recognition. Some signal conditions allow boosting the low and reducing highs for 100-Hz bandwidth. One click on the mode control and you have the prevalent 2.5-kHz nominal, flat-voice bandwidth, but in binaural.

The 2.5-kHz binaural position is also useful for CW in general listening. When Tone-Tag is used with this bandwidth, the modulation effect is mild because of competition with a relatively wide noise bandwidth. But it's still adequate to provide excellent selectivity through the significant tone quality difference relative to other signals in the bandwidth. In addition, of course, signals not in the modulated pass-band appear on the right or left specifically, while the tagged signal is heard in "surround sound."

On the next click of the mode control, everything in the 2.5-kHz bandwidth — ***except*** the 100-Hz band centered at 750 Hz — drops 20 dB. All basic conditions remain the same as above, but the Tone-Tag modulation is now more prevalent, and nontagged signals and noise drop a little more than 3 S units. If this isn't enough, the next position moves the floor down 40 dB — nearly 7 S units.

Now, if you're listening to a signal at about S-5 or so and a signal of about the same strength appears at around 400 or 1100 Hz, for example, you are alerted to the fact that the signal is very strong — nearly 16 dB over S-9 and probably suppressing your S-5 signal at his keying rate (if the signal has key clicks, it will be placing energy in a wide band as well). This alerts you to roll back your i-f (often called rf) gain control to avoid probable compression in your receiver's final down-converter (product detector), assuming your first mixer is not also being overloaded. If overloading is present in the first mixer, insertion of some attenuation between it and your antenna is indicated as well. Actually, however, so much attention has been given to the first mixer over the last

*The term antiphasic describes the case where a signal is fed out of phase to your two ears and noise is supplied in phase. The inverse is also true. The usual case (as in monaural reception) is called homosphasic.

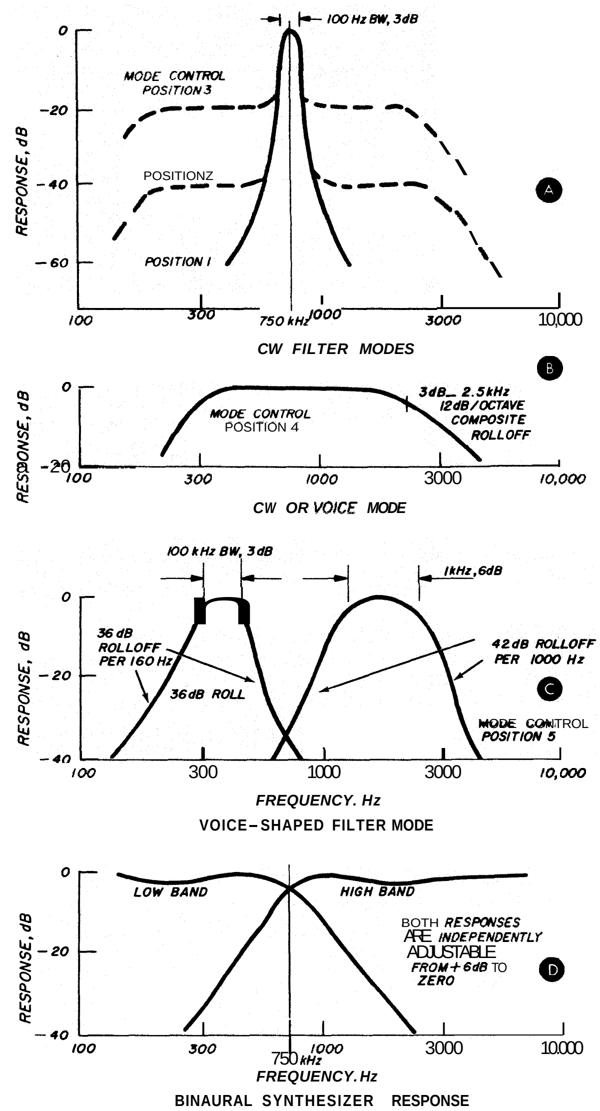


fig. 11. Frequency response. Responses A through C are fed through the response of D to become a binaural function.

two decades that the problem has shifted to the following converter(s) in our current receivers. Corrective action varies somewhat in many receiver designs, but the same general action is taken regardless of whether you are using agc or not.

The last mode select position supplies the basic 9-pole narrowband filter output through the center of the binaural band with no added binaural floor. To make good use of this position usually requires that your receiver have a super-good dynamic range, at least through its product detector.

Tone-Tag may be used or not as you choose in any of the CW modes. In general use, its sensitivity control is advanced to a point of a few degrees above where a given signal is modulated when tuned to the 750-Hz frequency. If no signal is present, increase Tone-Tag sensitivity until background noise is modulated, then back down until the modulated noise is just barely noticed. The best level is also dependent on your receiver audio-output level. Therefore it's possible, after a little experimentation, to generally set the Tone-Tag control on CAP then adjust your receiver gain to the tag level for any signal being received. Tone-Tag's modulator switching uses a nonabrupt design to avoid critical operation and to mitigate the effect of fade generally present on DX signals. In most cases Tone-Tag will make solid those ghostly multi-hop DX signals.

Finally there's the noise control. At the beginning you may nearly wear out this pot. Since the idea of adding noise seems so contradictory, you may constantly test the effect by repeatedly running the noise level up and down when listening to a weak signal. In general, however, a good starting point is found by increasing the soft noise to a point where it's just barely noticeable when the mode control is in the general 2.5-kHz position. The control is left there for all modes. You'll note that the added soft noise power is fixed relative to your receiver gain setting. Through this feature you may vary the ratio of added soft noise to relatively abrasive antenna-derived noise by receiver gain variations.

The binaural function provides a spatial sound environment in all modes without the need for adjustment. In addition, however, with proper physical arrangement and some practice, it can also serve as a tuning aid. For example, in using CAP with a Kenwood TS820S the low-band speaker or headset is placed on the right and the high-band device on the left. In this way, if a signal is heard on the left you know that tuning to the right (clockwise) will move the signal in that direction spatially. This is true with the TS820S because its conversion scheme results in an audio tone that moves from high to low as you turn the knob clockwise. Other receivers will be like this or exactly the reverse. The same may be true when going from band to band (SSB filtering for both voice and CW is assumed). The same tuning directional benefits are also present for SSB voice, although the action is more subtle.

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ham radio

digital techniques: gate arrays for control

Digital-circuit control sometimes requires a specific set of states. Phase-locked-loop (PLL) frequency-control input is such a condition; the requirement is to translate decimal control to some form of binary output. The example here is control of a "channelized" 10-meter transceiver using a National Semiconductor MM55106 synthesizer package.[1]

Fig. 1 shows the analog frequency control with PLL frequency input switching. Tuning is in 5-kHz increments over the band; no shift between transmit and receive. The receiver i-f circuit is assumed to have a 9-MHz crystal filter. This could possibly be a reworked CB set; the 55106 PLL is designed for such applications.

Both VCO outputs are mixed with one crystal-oscillator output to yield frequencies below the 3-MHz maximum of the 55106. Schmitt trigger NAND gates select the mixed VCO outputs, and filter levels are assumed high enough to cross the Schmitt thresholds.

division control

The PLL allows 5- or 10-kHz increments (hardwired for 5 kHz), with range controlled by programmable divider inputs in binary. Decimal division is obtained by dividing the mixed frequency by five. The range is 152-492 transmit; 168-508 receive, with each converted to binary and applied to the PLL divider input control.

Manual control is through four BCD-coded thumbwheel switches. The 10- and 100-kHz increment switches are standard. The 5-kHz switch is mechanically locked so that it will switch between only 0 and 5; the 0 position may be either 4 or 6 with a decal cover. The 8/9-MHz switch is locked mechanically to those two positions. The 20-MHz position may be assumed, or a dummy switch may be used for appearance.

The MM55106 has nine binary divider control lines. P0 is the least-significant bit, P8 the most significant bit. The LSB controls 5-kHz increments, P1 controls 10-kHz increments; P2 controls 20-kHz increments, and so on. The task is now to translate four decimal digits to one large binary word, including the frequency difference between transmit and receive.

Fig. 2 shows the switch interface for either TTL or CMOS. The switch is conventional BCD with a single common, so that the inverters change a low ON (grounded) output to a high state. Pull-up resistors

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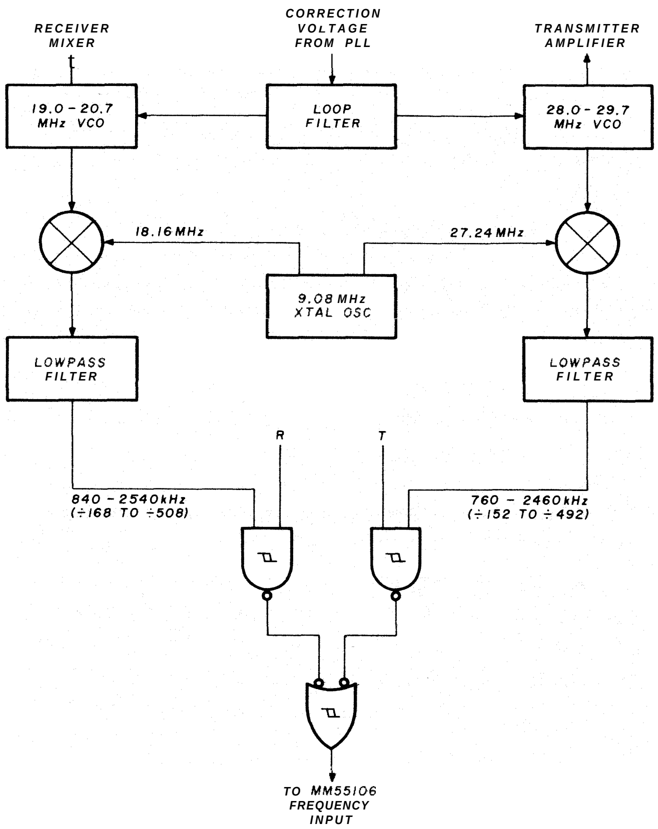


fig. 1. Channelized 10-meter transceiver analog frequency control.

are required, and inverted BCD data is taken from inverter inputs. The 5-kHz and 8/9-MHz switches use only the A switch contact.

translation through addition

TTL packages are available that convert BCD to binary (TI SN74184), but an offset must be provided because the lowest frequency is not zero. The scheme here is to use binary addition for translation with standard packages in TTL or CMOS.

Fig. 3 shows a truth table and schematic for a full adder. A full adder has three inputs. A half adder has only two, using gates G1 and G2 only. Both adders provide a sum and carry out.

Note the Exclusive-OR gates (not mentioned since part one of this series).[2] The truth table shows that inputs A and B are exclusive relative to the sum, and the sum is exclusive with carry in. Two Exclusive-ORs in cascade provide a full sum. Carry out is always generated with both A and B high. The full adder has a carry out with AB or $\bar{A}B$ and a carry in.

Single-bit addition is quite simple. A sum occurs with odd numbers of high inputs. Carry out occurs with two or three inputs high. Note that carry in can be treated as just another input. Four-bit binary adder packages are available. A single carry in to the LSB and a single carry out from the MSB is external, with other carries internal. Several packages have high-speed, "carry look ahead" connections for fast arithmetic but aren't required here.

Table 1 shows the required binary states from each selector for addition translation. An interim bit nomenclature is used with A as LSB, H as MSB. The 5-kHz K line is neglected for the moment.

translation in detail

The 10-kHz column of **table 1** is the same as the conventional BCD switch output in **fig.** 2. The 100-kHz column looks more complex, but note that 100 kHz is the next binary state up from 90 kHz: 1010. Two-hundred kHz is binary 10100, or the same as 100-kHz-state with a left-shift. Shifting a binary number to the left is the same as multiplying it by two.

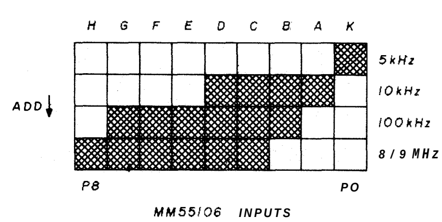
This multiplication with a left-shift allows you to set up an addition of four 100-kHz selector lines to create the six-bit states shown. Some scratchpad work with binary addition will show how the states have been achieved.

The megahertz selector has only two positions. To simplify circuitry, the 8 has been assumed 0 and the 9 assumed 1. A selection of 28.000 would appear to be

table 1. Binary states and addition of selectors.

	10 kHz				100 kHz							5 kHz
position	D	C	B	A	G	F	E	D	C	B	A	K
0	0	0	0	0	0	0	0	0	0	0	0	0
1	0	0	0	1	0	0	0	1	0	1	0	
2	0	0	1	0	0	0	1	0	1	0	0	
3	0	0	1	1	0	0	1	1	1	1	0	
4	0	1	0	0	0	1	0	1	0	0	0	0
5	0	1	0	1	0	1	1	0	0	1	0	1
6	0	1	1	0	0	1	1	1	1	0	0	0
7	0	1	1	1	1	0	0	0	1	1	0	
8	1	0	0	0	1	0	1	0	0	0	0	
9	1	0	0	1	1	0	1	1	0	1	0	

	1 MHz								
position	H	G	F	E	D	C	B	A	
28 receive	0	1	0	1	0	1	0	0	(84)
29 receive	1	0	1	1	1	0	0	0	(184)
28 transmit	0	1	0	0	1	1	0	0	(76)
29 transmit	1	0	1	1	0	0	0	0	(176)



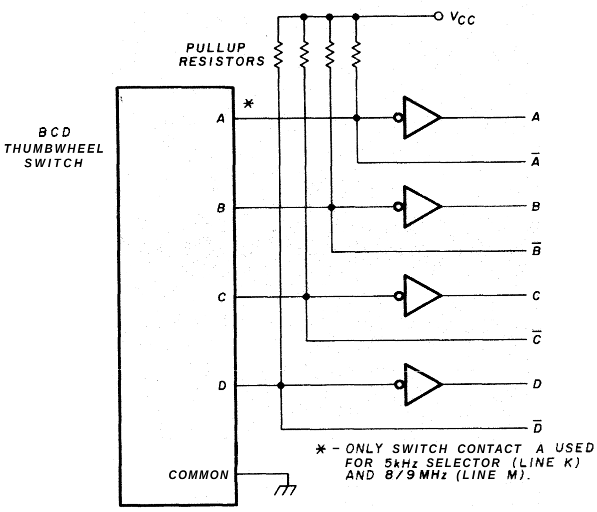


fig. 2. BCD thumbwheel switch interface.

all zeroes, but this won't work. The minimum frequency selected must start with 152 for transmit, 168 for receive; shifting up one megahertz requires adding 200 to each selector. All four-state combinations can be provided with simple gating.

The box with shaded areas in **table 1** shows the bits from each control that will be added for the final division control state. (Note that bits K and A do not add with any others and go directly to the PLL synthesizer).

If you've been paying attention to bit weights (part 5 of this series), you'll notice that values are only half of that required. Including the 5-kHz K line as the LSB will shift everything left once, multiplying by two, and achieving the correct division number.

hardware

Fig. 4 shows the adder grouping and megahertz gating to create the final binary-control state. In this case, megahertz gating is quite simple. Examining **table 1** you'll notice that bit F is the same as M (9-position), while bits C, G, and $\overline{H}$ are the same as $\overline{M}$ (8-position). Bit D is exclusive between M and T (high in transmit). Bit E is created by performing a NAND operation on $\overline{M}$ and T – it will be low only at 28 transmit; high otherwise. All gates except Exclusive-ORs are NANDs.

A half adder sums C_M and S2 from U2. This scheme avoids using an extra four-bit adder. Another package savings is to use the input arrangement for U1 and neglect its carry out. No combination of U1 inputs will result in a carry out; examining all possible states from **table 1** will bear this out.

No adder is used for $\overline{H_M}$ and U3 carry out. Keeping selector controls within band limits holds the maximum divider number to decimal 507 or binary 111111011 for 29.695 MHz receive. No combination of inputs to the adders will yield simultaneous $\overline{H_M}$ and U3 carry, so an OR operation is performed on these two for the P8 MSB. If both were high, it would imply a carry into a tenth bit, which the 55106 cannot handle.

Gates G15 through G20 provide a low state for all frequencies within the band. Sidebands are assumed to be 5 kHz maximum. An interface circuit must disable the transmitter when G20 is high. A high occurs at either 28.000 or 29.7 MHz and higher.

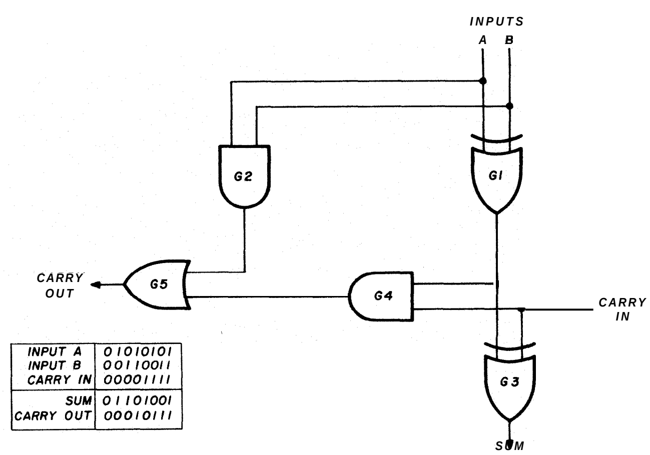
G15 and its inverter performs an AND operation on the zero positions of both the 10- and 100-kHz selectors. G16 expands this action by performing an AND operation on G15 with $\overline{K}$ and $\overline{M}$ to complete a 28.000-MHz gate. Both could be a single 12-input NAND in TTL.

The 100-kHz selector-7 position has an AND performed on it by G17. It does not require a D bit, because BCD states have D as a don't-care for positions 4 through 7. Similarly, gating for positions 8 and 9 could ignore the B and C bits. G18 ORs G17 and $\overline{D_{100}}$ to provide an output high at 700 kHz or higher. This outlet is ANDed with M (9 MHz) for the final high-frequency limit.

design rules

Design boils down to binary-state examination of all inputs *versus* outputs, keeping different package-function capabilities in mind. Intermediate states should be considered, as in the case of ignoring the U1 carry out. Different arrangements might be an advantage. **Table** 2 provides a list of devices with equivalent functions.

As an example, four adders instead of three would eliminate G12, G13, G14, and two inverters. The single remaining Exclusive-OR, G10, could be replaced



INPUT A	01010101
INPUT B	00110011
CARRY IN	00001111
SUM	01101001
CARRY OUT	00010111

fig. 3. Full adder truth table and schematic.

table 2. Device equivalent-function part numbers.

	TTL	CMOS
hex inverter	7404	4069
quad exclusive-OR	7486	4070
quad 2-input NAND	7400	4011
triple 3-input NAND	7410	4023
dual 4-input NAND	7420	4012
8-input NAND	7430	4068
4-bit adder	7483	4008
quad 2-input NAND Schmidt trigger	74132	4093
BCD-to-decimal decoder	7442	MM74C42*

*National Semiconductor part. 4028 CMOS decoder has active-high outputs while 7442 has active-low outputs.

by two NANDs. Package count is the same, but some spares are available.

Another alternative takes advantage of the offset difference of 16 between transmit and receive. This is the fifth highest binary bit. Four adders are used with direct M and $\overline{M}$ inputs plus an extra D bit input from $\overline{T}$ to add the difference in receive; G10 through G14 are deleted. This case is possible only because all binary states C through H, for 28 and 29 transmit, are opposite. Only the transmit states are used, because adding 16 creates receive states.

Certain cases may have the added control lines in a state higher than desired. Addition is still possible by using only those bits desired and the following expression in decimal numbers:

$$P + D - S = A$$

where P = maximum binary control input plus one
D = desired number for control input
S = control switch number
A = offset number to be added to all others

An example has S = 400 with desired D to be 170 and 9 binary lines. Nine binary bits yield 511 ($2^9 - 1$), so P is 512. The offset to be added is 282 decimal, or 100011010 binary. All carries beyond the 9th bit would be ignored.

A 6- or 2-meter control needs four megahertz positions. Only switch bits A and B are needed; see table 1 for 0-3 and 4-7 positions. A BCD-to-decimal decoder with active-low outputs could be used with its C bit input as the transmit/receive line. The decoder's eight outputs could then be ORed with NANDs for the necessary states. This package was originally de-

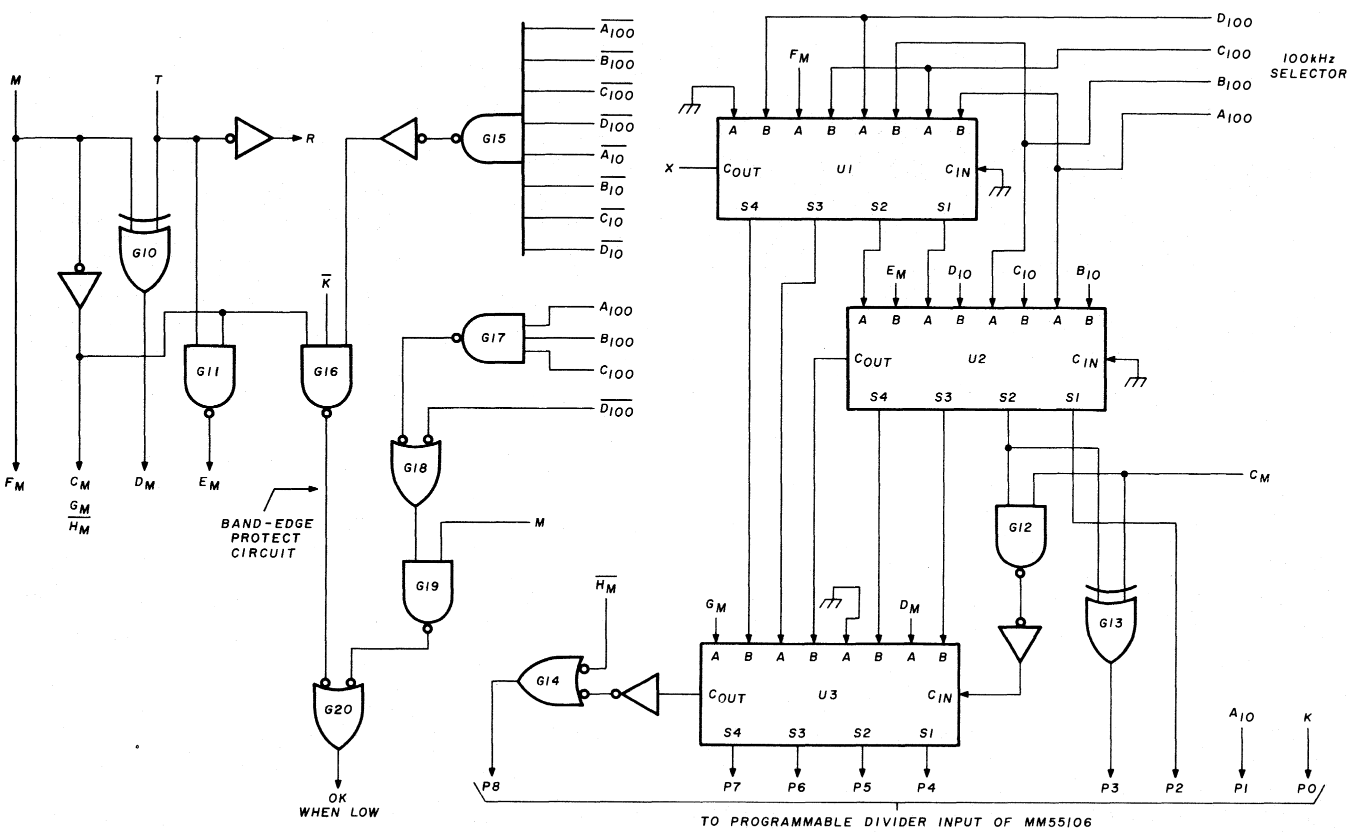


fig. 4. Translation gating and adders for MM55106

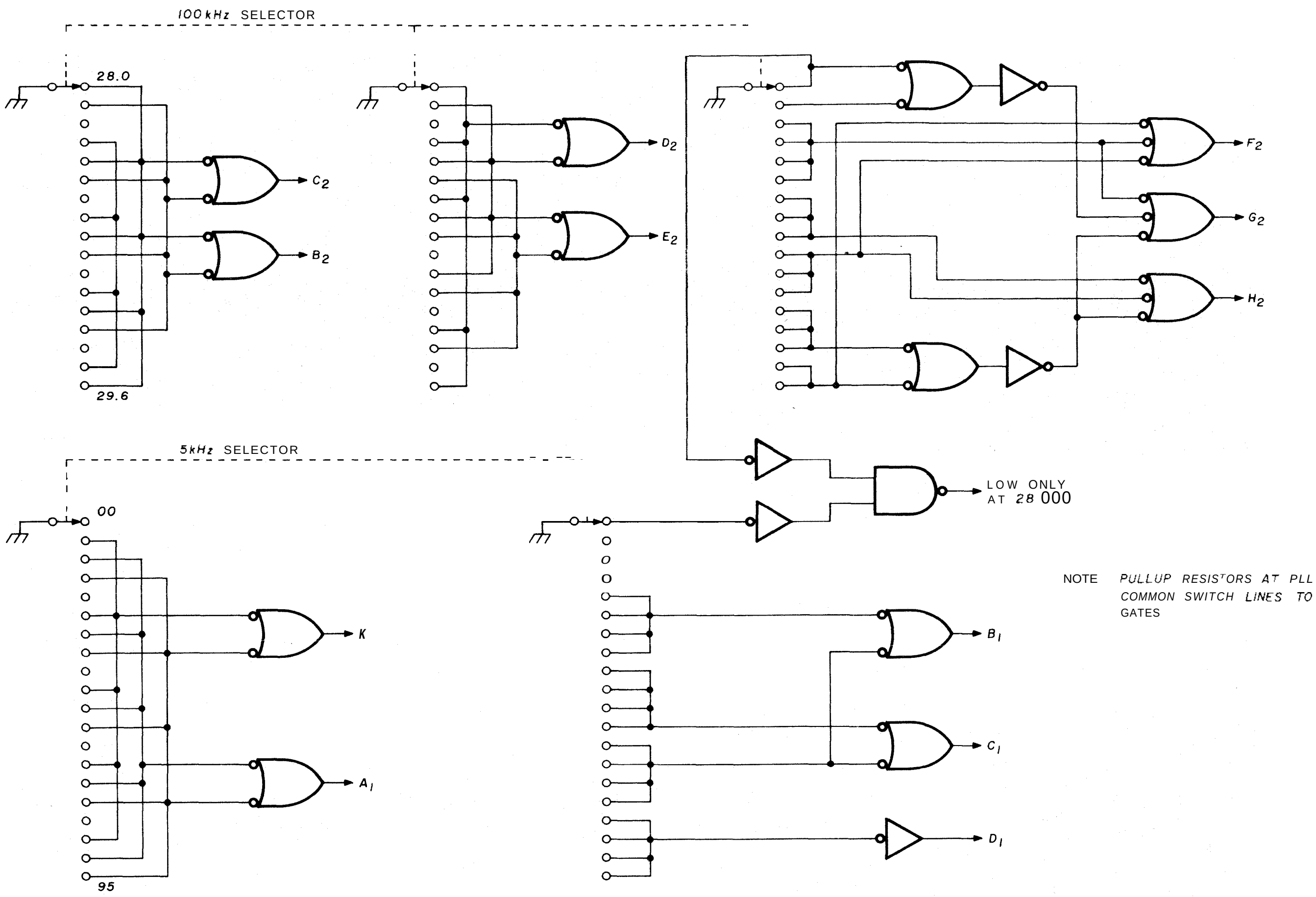


fig. 5. Dual rotary switch circuit equivalent of BCD thumbwheel selectors.

signed as a decimal indicator decoder but will work very well in this application. You might think a decimal switch would work best here. This is true for manual control, but automatic tuning from an up/down counter array needs some form of decoding.

another manual control

Two rotary switches can be used in the 10-meter rig. One can be seventeen positions for 100-kHz increments of 28.0 to 29.6 MHz, while the other can be twenty positions for 5-kHz increments. Indicator dials with adjacent edges can display a continuous number at any position.

Cam-actuated leaf or microswitches are the least troublesome. PC wafer-contact types are good but expensive. An alternative is a pair of 15-degree index rotary switches with stator contacts wired as in **fig. 5.**

The dual-rotary circuit uses NAND-gate ORing to produce a high from grounded inputs. Pull-up resistors are required for all gate inputs. The 5-kHz selector produces BCD states in A_1 through D_1 plus K for the LSB. The 100-kHz selector generates transmit states of decimal 76 to 236 (152 to 472 when left-shifted) in B_2 through H_2.

ORed switch contacts feed two four-bit adders with overlap occurring only in bits B, C, and D. An E bit input would come from $\overline{T}$, as in the other circuit, for the difference of 16 between transmit and receive. The second adder inputs for bits F, G, and H would be grounded. Band-edge protection is needed only at 28.000 MHz, because the high band-end is automatically controlled through the detent stop.

Many different ways are available to make a complex control input. Regardless of function, all designs must consider the required output, desired input control, and various possibilities of control and package functions. Binary-state examination is not only necessary, but will also provide a good basis for troubleshooting.

references

1. *CMOS DATABOOK*, National Semiconductor Corporation, 1977, pages 4-22.
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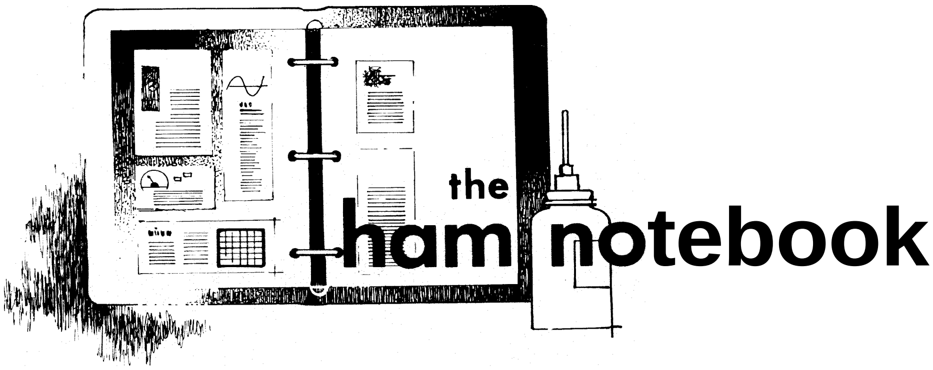
ham radio



reducing interference in the TS-820/TS-820S

The TS-820 modifies the audio response when switching from SSB to CW; the CW response is peaked at about 1000 Hz. This response is often useful in SSB reception as a way of reducing interference.

The peaked response is obtained by grounding lead CWG of the audio board (pin 4 of AF1). The ground should be one of the audio grounds; MEG or pin 12 of IF4 is the one normally used.

Those with the TS-820 can use the DISPLAY HOLD switch for this purpose. For the TS-820S, the unused contact pair on this switch could be used; this would require turning on the HOLD control whenever the modified response is desired for SSB, which is probably satisfactory. Alternatively, an external switch can be used.

R. P. Haviland, W4MB

improved CW agc for the Ten-Tec Omni-D

Shortly after receiving a new Ten-Tec Omni-D transceiver, I noticed that the agc time constant remained unchanged whether operating ssb or CW.

Upon examination of the i-f agc circuit, it became apparent that adding a resistor in parallel with C22, on the i-f agc circuit board, would decrease the recovery time. A value between 100k and 1 meg will work fine depending on your CW agc requirements. I used a 220k resistor.

Switching this resistor in, in the CW mode only, was accomplished by adding another wafer to the mode switch. The two wafers presently mounted on this switch are both double-pole six-position, with only four positions used. There's plenty of room on the mode switch shaft for the additional wafer, but you'll have to use 6-mm (114-inch) longer M3 (4-40) round-head screws to mount it. I purchased the additional wafer, which is also double-pole six-position, directly from Ten-Tec, although there may be other sources I'm not aware of.

I added the resistor from one of the stringers of the new wafer to ground on the front panel. I then soldered a piece of hookup wire between the high side of C22 on the i-f agc board and the CW position on the new wafer. Be sure to use the same switch on this wafer; there are two, unless you purchased the single-pole type.

According to Ten-Tec, this modification will not void your warranty.

Don McDougall, W6OA

salvaging water-damaged coax cable

The effectiveness of the shield braid on coaxial cables depends on the wires remaining bright and shiny so that the individual conductors remain in constant contact with each other throughout the entire feedline length. When coaxial line becomes contaminated, through damage to the outer plastic jacket, improper sealing of line terminations — or, even worse — no sealing at all, water will enter the line and can eventually penetrate the full length through capillary attraction, the braid acting like a metal sponge.

Such contamination renders the line unserviceable for rf applications, as the moisture soon corrodes the individual braid wires, destroying the surface continuity necessary for effective shielding. Characteristic line impedance changes, as the center dielectric becomes a combination of water, plastic, and corrosion. Rf losses increase drastically.

If contamination isn't extensive, it may be possible to cut off the contaminated portion to the point where the braid is again bright and shiny. This should be followed by at least a loss test: feeding a known power through the line and measuring the power delivered at the load. The results can then be compared with published specifications for loss at a specific frequency. However, replacement of the line and careful attention to sealing should be considered. A simple line-loss test ignores many important factors.

Although no longer serviceable for rf, water-damaged coaxial line, especially heavier lines such as RG-8/U or RG-213/U, may find other applications, especially for low-voltage, high-current dc.

Military specification requirements for RG-8/U and RG-8A/U lines originally required that the center conductor be composed of seven strands of 0.03-inch (0.76-mm) bare copper. This provided a twisted conductor of 0.09-inch (2.3-mm), approximately equal to no. 12 (2.1-mm) solid copper in current-carrying characteristics. Braid specifications required 192 strands of no. 33 (0.16-mm) bare copper wire, the combined circular mill area of which is only slightly smaller than a no. 10 (2.6-mm) solid copper wire.

Paralleling the center conductor and braid produces a single conductor with a CMA of 15270 cm, larger than a no. 9 conductor and only slightly smaller than no. 8. RG-213IU cable used in this manner will have a CMA equivalent almost identical to no. 8 copper wire.

Current-carrying capability of such cable will depend on many factors (such as ventilation), but it should carry at least 60-70 amps continuously and much higher currents for shorter periods of time. This makes salvaged cable useful for heavy-current cable in mobile equipment or for use in grounding applications.

For such applications soldering will be essential but will be complicated by the corrosion. Usually the corrosion may be removed by dipping the exposed braid and center conductors in a chemical cleaner such as those for cleaning silverware and copper kitchen utensils. After dipping, wash the cleaned copper in water and it will solder easily.

Add four heavy battery clamps to short lengths of salvaged coax and you can homebrew battery jumper cables for light-duty service.

Many currently manufactured coax cables have much fewer wires in the braid, so before using salvaged coax check the braid closely. It may be necessary to derate its current-carrying capability if it doesn't provide approximately 95 per cent coverage.

Robert Wheaton, W5XW

measuring air pressure across transmitting tubes

As Bill Orr notes (June, 1979, ham radio) inadequate cooling of air- cooled transmitting tubes may be dangerous for their health. After long deliberation I hit upon a convenient way to measure this air pressure.

Remove the innards of a single-hole-mount BNC connector (UG-1094) and mount it at some convenient spot in the pressurized compartment. To measure air pressure, slip the end of a short length of 9.5-mm (318-inch)ID clear plastic tubing over the connector. Insert the other end into a glass of water to the point where air bubbles stop. Measure the length of tube in the water: this is the air drop. When finished, close off the opening with a male BNC cap, which can be removed any time.

By the way, a convenient way to bring air to a chassis is through the flexible tubing, 64 mm (2-1/2 inches) in diameter, used on shop-type vacuum cleaners. The tubing and end fittings, suitable for mounting on plenum chambers and chassis, are available as replacement parts. Shop-Vac 2-112-inch (64-mm) diameter flexible hose in 6-foot (1.8-meter) lengths is part number 1722, and their 2-1/2-inch (64-mm) flange ferrule fittings, which are easy to mount with standard hardware, are part number 1714. A Y joint is available to direct air from one source to two chassis.

One of the surprises in measuring air pressure was that the popular muffin fan develops an insignificant amount of air flow for the 4X250B.

Guy **Black, W4PSJ**

modifications to the Wilson Mark II and IV

The Wilson Mark II and IV 2-meter handheld transceivers are fine units; they have excellent battery life and a very good receiver. For those with the early version, there have been some factory changesfor which modification kits can be obtained from Wilson. The kits are for the HI/LOW power switch located on the bottom of the unit and a battery-condition LED that blinks as batteries get weak. Both are good features.

plug-in crystals

One drawback of these transceivers is that you must solder the crystals into the PC board. This presents problems in emergency situations and when traveling. It would be much better to open the case and quickly exchange at least a few channels. Crystals with a six-position switch make this unit easier to use in the dark or in an emergency – compared with those that use no crystals and take forever to dial up that needed frequency. In the dark it's just plain impossible!

When investigating this quick-change in my Mark IV, I located a supply of some subminiature crystal jacks. These work very nicely. It takes longer to explain how to install them than to do the job. I installed three channels; most of the hams in the area installed all six channels.

installation

Open the unit and remove the crystals from the channels for which you wish to have plug-in capability. Enlarge the existing holes with a no. 48 1.95-mm (0.076-inch) drill. These new holes will be slightly larger than needed, but we want a margin of freedom between hole and jack. Drill from the track side of the board to be sure not to disturb the copper track from the laminate.

The best way to solder the modification jacks is to use a crystal with the jacks mounted on the pins. Insert the assembly into the enlarged holes. Solder the crystal assembly and proceed with the next assembly.

After installation some units seemed to require a little pressure when closing the case. This is because of the varying thickness of the rear case/battery compartment. In these units, I affixed a piece of white paper and a carbon to the rear case with the carbon toward the paper, so that when the case is closed the crystal jack ends will press onto the carbon and paper, marking the paper where there is contact. With the dots on the paper as a guide, use a 3-mm (1/8 inch) drill and remove a very *slight* amount of plastic from that point. Not much is required. This should allow you to reassemble the handheld. I've a supply of subminiature crystal jacks at $0.15 each.

Don Smith, W9EPT/AAM9EPT

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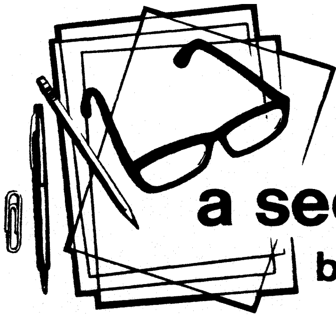


a second look

by Jim Fisk

The 1979 World Administrative Radio Conference (WARC) concluded just as the January issue was going to press, so I was unable to comment editorially on our good fortune in Geneva. It's no secret that I've been cautiously optimistic about WARC for some time, and as I predicted almost a year ago, Amateur Radio remains in possession of the same high-frequency spectrum as we've had for the past several decades; three new HF bands are our bonus for the 1980s. The international Amateur Radio community went to WARC very well prepared, thanks to long years of behind-the-scenes work by the International Amateur Radio Union (IARU), the various national societies, and dozens of Amateur Radio volunteers – these are the same people who remained optimistic about WARC throughout the long years of preparation. The Wizard of Woe, who expected WARC to be a complete disaster and forecast the end of Amateur Radio in the 1980s, was last seen skipping off to the Land of Oz with one foot in his mouth.

I first became aware of plans to pursue new ham bands at 10, 18, and 24 MHz more than five years ago, but quite frankly I considered the possibility of any new high-frequency spectrum as more pipe dream than reality. When W1RU announced the new bands at the Dayton Hamfest in 1974, I think most amateurs thought the pressure for new bands was a ploy to maintain the *status quo;* you know the old game: ask for more than you want so you can keep what you already got. Apparently the movers and doers of Amateur Radio who started preparing for WARC in the mid 1970s thought otherwise.

A. Prose Walker, W4BW, Chief of the FCC's Amateur and Citizens Division in those years was instrumental in arranging for the propagation studies that were conducted to show how long-distance Amateur communications links would be greatly enhanced and made more reliable with new allocations at 10, 18, and 24 MHz; Prose deserves a great deal of praise – and thanks – for getting the WARC preparations underway at such an early date. Without his foresight and experience I don't think Amateur Radio would have survived WARC in such good shape. Amateur Radio is also indebted to the dozens of dedicated volunteers who served on the FCC's Amateur Radio Advisory Committees, often at great personal expense, and to the ARRL and IARU staffers who convinced other national societies to get their WARC preparations underway early and then coordinated those activities.

What about those new bands? What are their propagation characteristics? How soon can we expect to have them available for Amateur Radio use? To answer the last question first, it will be quite awhile before you begin hearing ham signals on these bands; probably 1982 for 10 MHz, and not before 1985 for 18 and 24 MHz. The long delay for the top two bands is because Amateur Radio stations will not be permitted to take possession of these *exclusive* frequencies until all existing fixed services have moved to new assignments; the transition will occur between July, 1984, and July, 1989, so it will be nearly five years at the earliest, and significantly longer if any of the present users feel like dragging their feet.

It may not be obvious, but the frequencies of the three new ham bands were very carefully chosen – each band is very near the geometric mean of the two existing, adjacent bands. This is optimum band placement for maximum propagation enhancement, so it should be possible to maintain long-distance radio communications for many more hours each day than is possible with our present allocations.

Practically all of the new rigs which use phase-locked loops already cover the new bands (although they are not presently programmed to transmit there); some will require simple modifications, but you can be sure the manufacturers are already working on them – and will have mod kits available by the time the new bands are opened. And when the new frequencies do become available, most of the equipment makers will be marketing ham gear with even more features than they offer now! Thanks to WARC, the future for Amateur Radio is bright, and the decade of the 1980s promises to be exciting – both to Amateur Radio and to the technology which will become available to us.

Jim Fisk, W1HR
editor-in-chief

comments

compact loop antenna

Dear **HR:**

W6TC's article on the compact loop antenna in October *ham radio* essentially describes the antenna I have used for several years. I improvised the antenna out of desperation because my station was over pure sand, and grounded verticals simply did not work. Contrary to the title of his article, I was searching, not for a DX antenna, but a general-purpose antenna for all bands that would work reasonably well in an area with extremely poor ground conductivity. In addition, I wanted an antenna that would cover the entire 75/80-meter band. The antenna has satisfied all these requirements.

The antenna books give the resonant impedance of a full-wave loop as about 105 ohms; measurements confirmed this figure. To cover the 75180-meter band, the loop was cut for about 3950 kHz (248 feet). A 17 μH roller coil in series with the ungrounded side at the feed point resonated the antenna at 3550 kHz. The impedance had dropped to about 25 ohms. Fed with RG-81U coax, the antenna has an SWR never more than 2:1.

The original antenna was in a pentagon configuration. The feed point was near ground, with the roller coil in a weather-proof metal box. The two ends of the loop sloped upwards in a sort of "inclined V" configuration to the first pair of masts, one on either side of the lot. With a maximum height of 20 feet, this antenna was definitely *not* a DX antenna. It was a high-angle radiator, as expected, but on 75 meters it outperformed a dipole at the same height up to distances of 400 to 500 miles. At increasing distance, even an inefficient vertical antenna did better.

In the CW portion of the 40-meter band a series inductance of 9 μH is required for resonance. This results in rather high rf voltages on the antenna side of the coil, and an SWR of about 2.5:1. Numerous contacts with 1-watt QRP CW on 40 meters, in the middle of the day, at distances of several hundred miles and with consistent reports of S6-S9, indicate excellent performance.

At my present station, the antenna has been erected in a delta-loop configuration with about the same results. Feeders have been dispensed with (except for 10 feet of coax from transmitter to tuner). Romex wire carries the loop ends through the ceiling of shack up to the peak of the roof, and through the gable. (Since I'm in a remote area, I can get away with this. However, I can run a TV set on rabbit ears in the shack – except on 40 meters.)

W.S. Skeen, **W6WR**
Hornbrook, California

W3VT memory keyer

Dear HR:

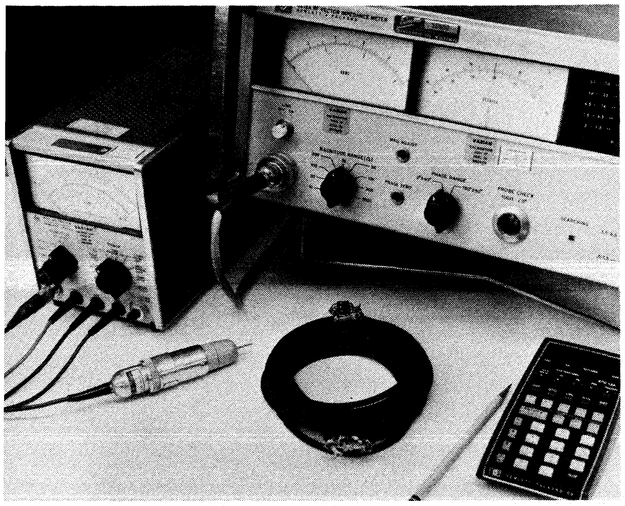
A builder of the deluxe memory keyer featured in the April issue has written, pointing out that under some conditions a slight click occurs in his unit during manual sending each time the keyer clock control line goes high. This click was also present in the original version, but was barely perceptible in the monitor and did not seem objectionable; perhaps in some other units it might be severe enough to be a problem. Following is an explanation and the solution.

The click occurs during manual sending when sending from memory has been stopped at a memory location with a high data bit; this can happen when the memory is stopped by the stop button or interrupted by manual sending. The memory output control U13B is held high by the low on the keyer control line at U13B pin 10, inhibiting the output even though the other three inputs (pins 9, 12, and 13) are high. When the control line goes high, U3B inverts the control line signal and applies the low to U13B pin 9, which continues to inhibit memory output. There is a momentary gap in the inhibition process, however, due to propagation delay in U3B.

The cure is simple and requires only slight surgery on the memory board.

First, remove the keyer clock control lead from the designated eyelet on the board. From this eyelet, connect a 0.047-μF (or 0.1-μF) capacitor to the nearby ground foil. Next, remove the jumper which connects the clock control line to pin 9 of **U4,** and replace the jumper with a 150-ohm resistor. Connect a new jumper from the bottom end of this resistor (near

(Continued on page 74)



a new class of coaxial-line transformers

A review of transmission-line transformers and balun theory, including problems with magnetic cores — Part 1 of a two-part series

Most coaxial-fed antennas require a balun for optimum performance; many also require a matching transformer. Typically, these baluns and transformers are made with magnetic core material such as ferrite. These devices are subject to arcing and linearity problems. Simple baluns and a new class of rf transformers which are not subject to these problems use only coaxial line in their construction – and they are easy to make.

To match the low impedance of two closely spaced dipoles, I needed a broadband 4:1 transformer for some experiments on a low-band phased array. My first inclination was to go to the handbooks to design a ferrite-core transmission-line transformer. However, I decided that there had to be a better way. I had just read Doug DeMaw's article, "The Whys and Hows of Bifilar Filament Chokes" in *QST*.[1] He expressed concern about saturation of the core material and corona from the windings to the core when operated at high power. My search began for a way to make broadband transformers without magnetic core material.

background

In conventional low-frequency transformers, closely coupled primary and secondary windings of the appropriate turns ratio are used. At radio frequencies, because of inevitable leakage reactance, narrowband tuned transformers are generally used for impedance transformation. Quarter-wave matching transformers may be used but are also narrowband; they are a quarter wavelength at only one frequency.

The availability of solid-state rf power devices with their capability for broadband performance created the need for broadband interstage matching, thus causing rapid development of transmission-line transformers.[2,3,4] Development of ferrite materials also expanded rapidly during this period.

The low-frequency response of transmission-line transformers is limited by winding inductance, and the high-frequency response is limited by resonances from stray capacitance; therefore, ferrite material is used to extend the low-frequency limit of small transformers by increasing inductance. Thus broadband transmission-line transformers and baluns using ferrite cores have come into wide use in solid-state rf circuitry and Amateur antenna systems. While these cores are very useful, they have some disadvantages.

By George Badger, W6TC, 341 La Mesa Drive, Portola Valley, California 94025

This article shows how to build and design broadband rf transformers and baluns without magnetic cores.

problems with magnetic cores

Amateurs build or buy highly linear SSB equipment and effective lowpass filters to avoid TVI. We then subject our clean, harmonic-free signals to the uncertainties of ferrite-core transformers or baluns in our antenna systems. The cores in these devices are subject to saturation and, therefore, nonlinearity. High permeability ferrite cores are also susceptible to permanent damage at flux densities of a few hundred gauss.[5] Tune up your linear into the wrong antenna just once and the damage is done.

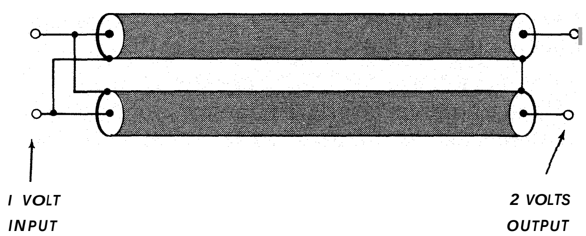


fig. 1. Broadband transmission-line transformers are made of two or more transmission lines connected in parallel at one end and in series at the other. One volt applied to two coax lines in parallel at the input results in 1 volt across *each* of the lines at the output. If these two lines are connected in series at the output as shown, the output will be 2 volts. In this way a 1:4 impedance stepup is achieved. Sufficient impedance must be provided over the length of the outside conductors to prevent the connections at one end from shorting the other end.

Magnetic materials such as ferrite, powdered iron, and specialty steel tapes have added greatly to the performance of components available to circuit designers. However, these materials should not be used in high-power circuits or antenna systems unless they are adequately characterized regarding power-handling capability and saturation effects. This is necessary so that interaction of the material with your system can be thoroughly understood. Put another way, sufficient core material must be used to keep the flux density well below the saturation level. Data on harmonic distortion measurements, taken at high power on a popular commercial ferrite core balun, are presented in part 2 of this article.

Ferrite baluns and transformers are usually wound with copper wire coated with thin enamel insulation. Pairs of wires are placed close together or twisted to make transmission lines, which are wound tightly onto the core. The conductors must be close, because the surge impedance of the wire pairs must be correctly related to the impedances to be matched. For this reason, thin insulation is often used. Inadequate insulation may result in arcing between wires or to the core when operated at high power.[1,2] Those who have blown a balun in the heat of competition are all too familiar with these problems.

transmission-line transformers

Basically a transmission-line transformer consists of two or more parallel lengths of transmission line connected in parallel at one terminal and connected in series at the other **(fig. 1)**. For example, if two lengths of coaxial line are connected in parallel at the input and 1 volt is applied, **1** volt appears at the output end of each of the two lines. If the output ends are connected in series so that the two voltages add, the output is 2 volts, thus creating a 1:2 voltage increase (1:4 impedance transformation). **Fig. 1** also can be used to describe a 4:1 impedance reduction; for example, from 50 ohms to 12.5 ohms.

Sufficient rf impedance must be provided between the input and output ends of the transformer of **fig. 1** to prevent the connections at one end of the lines from shorting the other end of the lines. The impedance is usually provided by wrapping the transmission lines around magnetic cores.

the Collins balun

By far the best balun I've ever used is the Collins balun which, to my knowledge, was first described in a book published by the Collins Radio Company entitled *Fundamentals of Single Sideband.*[6] The Collins balun derives its name from this reference. I believe the earliest reference to an Amateur application was in an article by K2HLT in G.E. *Ham Notes* in 1960.[7] The Collins balun is rarely mentioned in Amateur literature, which is surprising in view of its superb performance. However, Bill Orr, W6SAI, describes one in his *Radio Handbook.**

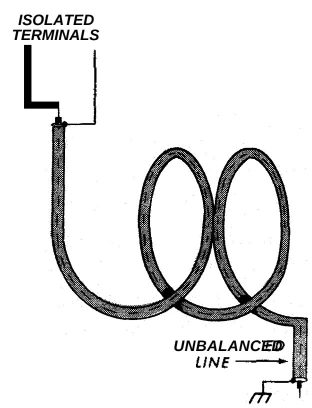


fig. 2. Simple coiled length of coaxial line isolates output terminals from ground.

Perhaps the reason the Collins balun hasn't gained popularity with Amateurs is that it's quite bulky when made with RG-8/U. The balun is extremely simple. No exotic materials are used in its construction; only coaxial cable and insulated wire. I've used these baluns for years with various antennas and never had a failure. One has been on my three-element **10-15** meter quad for eight years with no sign of deterioration. There are only two disadvantages to the Collins design: 1) when made with RG-8/U, the balun is bulky — too large for installation on a clean-design antenna system; and 2) the balun is useful only at **50** ohms. This article shows how to eliminate these disadvantages.

balun theory

Baluns convert energy from unbalanced coaxial line to balanced two-wire line by isolating the two balanced terminals from ground. As in the transmission-line transformer, this is often accomplished by coiling transmission lines around magnetic material so the impedance to ground from both output terminals is high compared with the characteristic impedance of the input coaxial line. By using this technique, shown in **fig.** 2, the two balanced terminals are "floated" with respect to ground by the isolation provided by the coiled-line impedance. However, a simple coiled length of transmission line is often not adequate because it doesn't contribute to the balance of the system.[9] For a balun to make this contribution, the impedance ground from both terminals must be nearly matched.

Accordingly, in the Collins balun, a dummy length of coax is wound as a continuation of the isolating winding, so that the coil consists of the original length of coiled coax of **fig.** 2 plus an equivalent length of dummy line, as shown in **fig.** 3.

The dummy-line center conductor is unused and is left floating, or both ends may be shorted to the outer conductor if desired. The dummy length of tine causes the impedance to ground, from each of the two output terminals, to be nearly equal. The isolation impedance (common-mode impedance) is held higher than the coax-line characteristic impedance over a wide frequency range by the distributed capacitance and inductance of the combined coil. The coil must have sufficient inductance so the impedance, at the lowest operating frequency, is higher than the line surge impedance. As the frequency is increased, the impedance increases through parallel self-resonance, then decreases as the frequency is further increased.

Because the self-resonant circuit consisting of the distributed capacitance and inductance of the combined coil is loaded by the low characteristic impe-

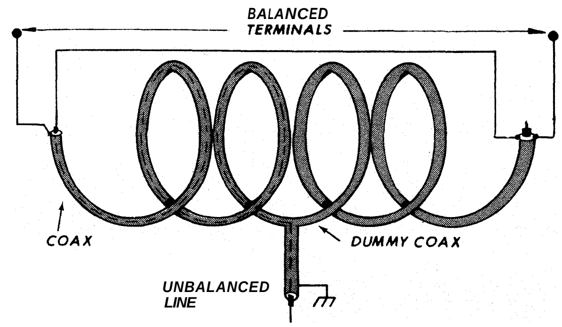


fig. 3. Broadband coil balun is evolved from the coiled coaxial line of fig. 1.

dance of the line, the impedance versus frequency curve is broad. Balun performance therefore is not critical with respect to frequency. Data taken on measurements of the common-mode impedance on a typical Collins balun are presented in Part 2.

The symmetry provided by the dummy line makes balun performance less dependent on common-mode impedance and is therefore often essential in baluns and balanced systems.[9] The isolation, balance, and impedance match of this class of balun are superb over the hf Amateur bands. Specific designs, performance data, and a systematic design procedure are presented in Part 2.

new class of transformers

Faced with the need to match a very low-impedance antenna system, I decided to try to develop a 4:1 transmission-line transformer based on the principles of the well-proven Collins balun. The transformer was successfully developed; in fact, a new class of wideband transformers evolved from this work.

One of the nice things about an avocation — as compared with a vocation — is that you're not on a time schedule. I found that the performance of the 4:1 transformer was so good that the idea of other transformer designs based on the same principles looked interesting. I shelved the phased-array project long enough to enjoy the freedom to explore the possibilities of these transformers. The result was a series of broadband balanced and unbalanced transformer designs that are extremely simple, made entirely of coax, and, most important, don't depend on ferrite or powdered-iron materials.

design concept

Because the Collins balun so successfully isolates the balanced output terminals from the unbalanced coaxial line input, it seemed reasonable that a similar broadly resonant configuration would provide the isolation necessary to the series and parallel lines of **fig. 1.** From previous experience I'd found that it's unnecessary to wind the Collins balun on a cylindrical form as shown in **fig.** 3. It's sufficient to random-wind the turns without a coil form, as shown in the photo. I decided to try winding the two lines of **fig. 1A** into a continuous winding similar to the Collins balun of **fig.** 3.

My first try was with a nine-turn coil random wound on a nominal 25-cm (10-inch) diameter shown in **fig. 4** (Only three turns are shown in the drawing for clarity.) Two 254-cm (100-inch) lengths of 50-ohm line were used. The transformer was tested by inserting a 12.5-ohm low-inductance load at the output and measuring the input impedance with a Hewlett Packard Vector Impedance Bridge. While the transformer made the 12.5-ohm resistor appear to be a 50-ohm resistor over a wide range, the useful range was centered in the broadcast band. As the frequency was increased through 3.5 MHz, the input impedance magnitude increased rapidly, and the impedance phase angle was greater than 30 degrees. However, the useful frequency range was 4:1. Encouraged, I removed half the turns, expecting equivalent results more nearly centered over the ham bands. Results were disappointing. A good match was achieved only over a narrow frequency range.

line impedance

Note that the load at the transformer output of **fig. 4** terminates two coax lines connected in parallel. If the lines have 50 ohms characteristic impedance the load must be 25 ohms for the lines to be properly terminated. At the input the two lines are connected in series, so the input impedance will be 100 ohms. Thus the transformer of **fig. 4** will match a 25-ohm balanced load to a 100-ohm balanced line, but per-

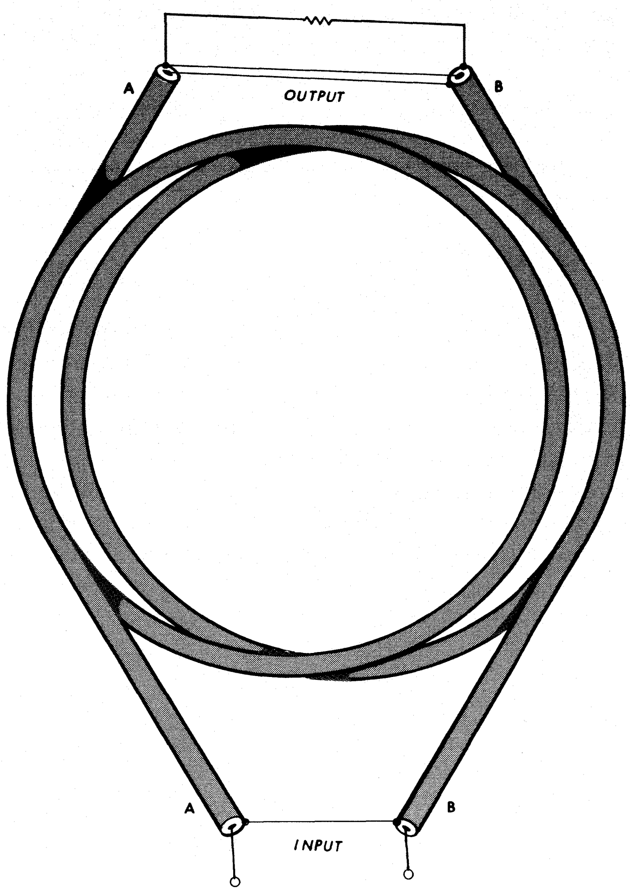


fig. 4. Two lengths of coax (A and B) wound into a multiturn coil similar to the Collins balun of fig. 3. Only three turns are shown for clarity. Note that the input ends of the lines are connected in series and the output ends are connected in parallel. Made with 50-ohm coax, this transformer will match a 100-ohm balanced line to a 25-ohm balanced load.

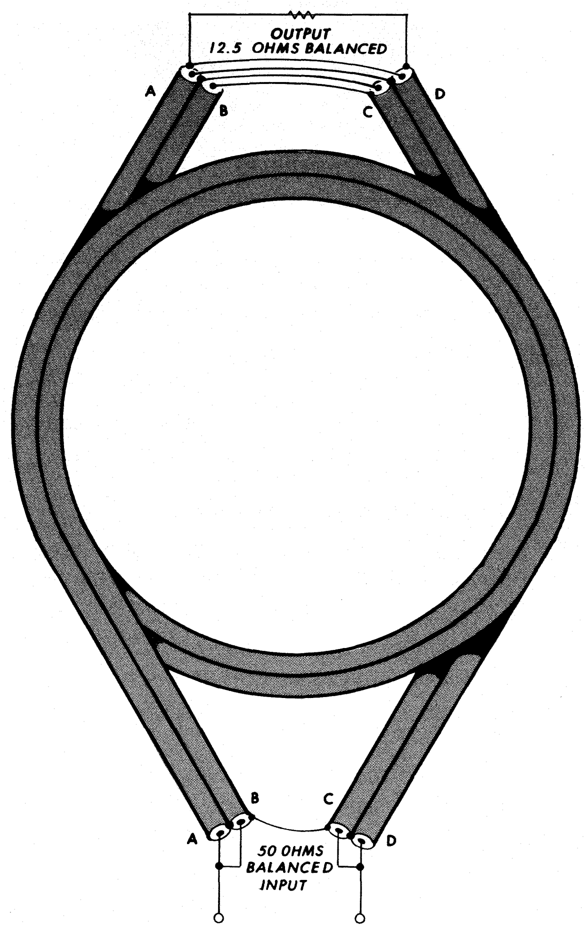


fig. 5. A **4:1** rf transformer (shown in the photo) consisting of two parallel 127-cm (50-inch) lengths of RG-58A/U random wound into a 7 turn, 11.5 cm (4.5 inch) nominal **diameter** coil. Each of the two paralleled **50-ohm** lines **(AB-CD)** is connected in series at the input and in parallel at the output. The excellent performance is shown in table 1.

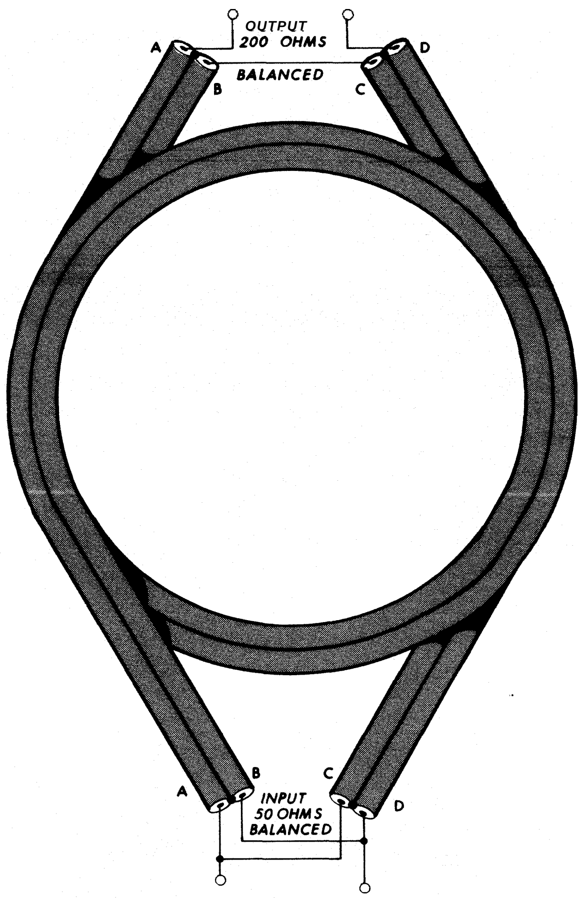


fig. 6. A 50- to 200-ohm balanced-to-balanced transformer made from two 127-cm (50-inch) 100-ohm lines each consisting of two 50-ohm lines in series. The 100-ohm lines are connected in parallel at the input and series at the output. Performance data are shown in table 2.

formance will be poor when trying to match a 12.5-ohm load to a 50-ohm line.

50112.5-ohm transformer

To match a 50-ohm line to a 12.5-ohm load, 25-ohm transmission line must be used so that the 50-ohm source will be correctly terminated by two 25-ohm lines in series. Similarly, the two 25-ohm lines in parallel at the output will be correctly terminated by a 12.5-ohm load. I decided to make 25-ohm lines by connecting pairs of 50-ohm line in parallel. A transformer similar to that in **fig.** 3 was wound with two lengths of transmission line, each made with a pair of 50-ohm RG-58A/U lines in parallel (25-ohm line).

After much cutting, winding, soldering, and data taking, the optimum design of **fig. 5** evolved. The design lookssomewhat more complex than the simple configuration of **fig. 4** because each length of transmission line consists of two 50-ohm lines in parallel; however, in other respects the configuration is exactly the same as that of **fig. 4.** Note that the two 125-cm (50-inch)25-ohm lines are connected in parallel at the output and are properly terminated by the 12.5-ohm load. This transformer design was useful over a wide frequency range. The VSWR is low over the high-frequency Amateur bands. The excellent data are shown in **table 1.**

construction

While it's not necessary to cut the lengths of coax to exactly 127 cm (50 inches), this length wound into a seven-turn, 11.5-cm (4.5-inch) nominal diameter coil is optimum to cover the high-frequency bands. The detail of coil winding is unimportant. You can bind the lines tightly with tape or leave them loose. You can hold the coil in your hands and separate the turns several cm (2 inches) before the magnitude or phase of the match are greatly affected.

However, the length of the short between the coax-line outer conductors at the input end of the balun is critical. Performance is degraded at the high-frequency end of the useful range unless the coaxial outer conductors are soldered together into a common joint as shown in the photo and **fig. 5.** Performance deteriorates if the shorting lead is as long as 2.5 cm (1 inch). The cross-connected leads at the output should be short. In the version shown, the length of the center conductors exposed beyond the outer conductors is about 2 cm (0.75 inch). Construction is shown schematically in **fig. 5** and in more detail in the photo.

measurements

Impedance measurements were made with the Hewlett-Packard rf vector impedance meter model 4815A shown in the photo. The vector impedance meter reads directly in impedance magnitude and phase angle. Data in this article are presented using the abbreviation Z for magnitude of the impedance and θ for the corresponding phase angle. Balance measurements, described in Part 2, were made with the HP rf voltmeter model 4-10C also shown. The terminating resistor is critical to the evaluation of these balun transformers. The VSWR looking into the transformer is, of course, no better than the quality

table 1. Frequency, impedance magnitude. phase angle and calculated VSWR for the balanced-to-balanced 4:1 transformer of fig. 5.

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR
1.8	50	13	1.26
3.5	51	8	1.15
4	51	6	1.11
7	49	3	1.06
14	54	5	1.12
21	55	1	1.10
28	53	– 1	1.06
30	53	– 1	1.06

of the load. Shown in the photo is a bundle of eight parallel-connected, 100-ohm, 114-watt resistors soldered directly to the 50/12.5-ohm broadband transformer for minimum inductance.

performance

Referring to **table 1,** note that the VSWR between 3.5-30 MHz is less than 1.15. Even on 160 meters, the VSWR is only 1.26. Data were recorded only in the ham bands; however, with each configuration, I swept the impedance bridge over the full range looking for spurious resonances. In the designs presented here, none were found within the frequency range of the data shown.

Data are presented in tabular form. Displaying experimental results in this convenient form required calculation of VSWR from the impedance-magnitude and phase-angle data. These calculations are long and tedious, so a Hewlett Packard HP65 programmable calculator was used. N6AIG suggested the idea and wrote the program. The program is very useful and is included in the appendix. The program calculates VSWR based on an impedance of 50 ohms and can be modified easily for any impedance.

501200 ohm transformers

After the 50112.5 ohm transformer was optimized, it seemed reasonable to expect that a 501200 ohm (1:4 impedance stepup) transformer could be made using the same principles. In this case, 100-ohm lines were used so they would match 50 ohms when connected in parallel at the input and would be properly terminated by 200 ohms at the output. The 100-ohm lines were made with two pairs of RG-59AIU cable; each pair was connected in series. The outer conductors of each pair were connected together at both ends. Each pair is 127 cm (50 inches) long. The lengths of coax were wound into a compact package of seven turns at a nominal diameter of 11.5 cm (4.5 inches). The package looks very similar to the 50112.5 ohm transformer shown in the photo. The transformer is shown schematically in **fig. 6.** Performance data recorded for this transformer are listed in **table 2.**

To show the effect of increasing the length of the lines and increasing the number of turns, another similar 501200 ohm transformer was built. Data taken on this transformer are also shown in **table 2.** The modified transformer was optimized for the low bands. It is similar to the configuration of **fig. 6.** However, the coax lines are twice as long, 254 cm (100 inches), and have twice as many turns on the same nominal diameter. Note that the increased length makes the transformer useful on 160 meters and substantially improves VSWR on 80 and 40 meters. The low-band transformer performance is good through 10 MHz; however, the VSWR climbs to 1.37 at 14 MHz.

table 2. Data taken on the 50-ohm balanced to 200-ohm balanced transformer. (A) lists data for the transformer made with 127-cm (50-inch) lines as shown in fig. 6. (B) lists data for another version optimized for the low bands. This transformer has the same configuration as that of fig. 6; however, the two 100-ohm lines are made with pairs of 254-cm (100-inch)lengths of RG-58A/U.

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR
	A. Data for transformer of fig. 6		
3.5	56	20	1.45
4	57	18	1.41
7	52	7	1.14
14	51	1	1.03
21	47	5	1.11
28	46	15	1.32
30	46	16	1.34
	B. Data for iow-band version		
1.8	55	10	1.22
3.5	55	2	1.11
4	55	0	1.10
7	51	– 4	1.08
10	46	– 2	1.09
14	46	– 17	1.37

The common-mode impedance of the transformers described in this article is sufficiently high so that the transformers can be driven from either a balanced or an unbalanced coax transmission line. Two-stage balun transformers with improved isolation are described in Part 2.

I determined the efficiency of the 4:1 transformer of **fig. 5,** by carefully measuring the input complex impedance and the complex impedance of the load, driving it with a signal generator and measuring the input and output rf voltages. I used the HP4815A rf Vector Impedance Meter and the HP410C rf Voltmeter shown in the photo. Input power was determined by calculating the power in the real part of the input impedance; the output power was calculated similarly using the complex load impedance data. Efficiency was determined by dividing the output power by the input power. Measurements and calculations were made for each of the bands, from 160 through 10 meters. As you might expect, efficiency was lowest on 10 meters; however, efficiency was greater than 95 per cent on all bands.

power-handling capability

The 4:1 transformers may be made with pairs of RG-58A/U lines connected in series for 100-ohm surge impedance or connected in parallel for 25-ohm surge impedance. Each length of RG-58AIU must therefore handle only 50 per cent of the power delivered by the transformer. Because RG-58A/U is rated for more than 500 watts at 30 MHz, these RG-58A/U transformers will handle 1000 watts of rf. I verified this by connecting two 12.5-ohm transformers back-to-back into a dummy load. At 1 kW, heating was not discernible at 7 MHz. At 30 MHz, the transformers became warm to the touch after about one minute, key down. Part 2 describes how to build baluns capable of several kilowatts overload into severely mismatched loads.

Part 2 of this article, which will appear in March, 1980, *ham radio*, will describe how to build more useful balun transformers and three specific 1:1 baluns, including one for vhf. These designs are capable of conservatively handling high power into widely varying loads. The impedance match (VSWR) and balance of these new transformers and the baluns will be compared with popular commercially available rod and toroid core balun transformers. In addition, harmonic distortion data taken at 2 kW PEP on a typical commercial ferrite core balun will be included.

Balun performance is usually described when working into a "flat" load. In the real world, baluns must work into widely varying loads as frequency is changed across the band. Measured performance of baluns with varying loads will be included and comparative data presented in Part 2. How to design balun transformers to your needs and how to modify a currently popular balun will also be described.

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appendix

Program for calculating VSWR for a given load impedance, Z_L using the HP-67 calculator.

$$\text{reflection coefficient, } \Gamma = \frac{Z_L/Z_0 - 1}{Z_L/Z_0 + 1}$$

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|}$$

running instructions

step	instruction	input	keys	output
1	(clear memory — enter program)			
2	initialize	—	h RTN R/S	—
3	input magnitude	$\lvert Z_L \rvert$	R/S	—
4	*input phase*	*angle°*	R/S	
5	copy VSWR	—	—	VSWR
6	return to step 3		R/S	

ham radio

Yagi antenna design

experiments confirm computer analysis

Comparison of calculated Yagi performance with NBS measurements confirms validity of theoretical approach

A mathematical model of the Yagi antenna, as described last month,' is required for computer analysis of antenna gain, front-to-back ratio, and operational bandwidth. Once the model has been constructed, a computer program can be written. To avoid endless consultations, I prefer not to supply actual computer programs; for those readers who would like to create their own programs, I will explain the information flow I use in programs for computing element currents and Yagi performance.

writing programs

I first create a labeled input file containing all of the necessary information about the Yagi. The label itself is usually an abbreviated reminder of the particular Yagi. This input file contains a statement of the number of parasites and the number of drivers. Next is given information on each parasite, *i.e.*, X and Y coordinates, length (*LE*) and radius (*RO*), all specified in terms of wavelength at a given central design frequency. Next is given information on each driver, *i.e.*, X and Y coordinates, length (*LE*) and radius (*RO*), driving point voltage and phase.

The element-current program first calls for the input data file label and asks at what frequencies (in terms of the center design frequency) the computations are to be made. With this starting information, the input file is called and read. For each frequency specified, a computation is made to determine the (complex) values of all terms in the *Z* matrix; self-impedances are calculated (from **eqs. 4, 5,** and **6,** reference 1) and mutuals are given by an interpolation routine or power series approximation (using **eqs. 7** and **8**).[1]

Once the Z matrix is complete and the voltage vector noted from the input file, matrix inversion is accomplished; this generates the individual element complex current solutions. The entire result is then written into an output file for later use. I have found it convenient to include in the output file a statement of frequency for each computation and the number of parasites and number of drivers. For each element, I include its X and Y coordinates, its resonant frequency and *Q,* and magnitude and phase of its current. The element-current program continues to compute until all initially specified frequencies are satisifed.

This output file now becomes a useful *input* file to other computer programs which are designed to produce displays of interesting Yagi properties. One such program calculates and plots the pattern both for the H-plane and the E-plane. This is done by reading the new input file, and for a specified sequence of elevation angles (say every 1 or 2 degrees), computing the radiated energy flux density. This is easily done from the known element positions and the known element complex currents; the elevation angle and positions geometrically determine the effective phase delay of each element from the reference origin. The vector sum of all contributions is then proportional to radiated field strength, the square of which is proportional to radiated energy flux density. This calculation provides values which are referenced to a single element carrying a unit current; this can be referred to an isotropic source by correcting for the gain of a single element (2.15 dBi).

This (energy flux density) pattern is calculated for each frequency (read from the new input file) and

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can be easily plotted by one of the plot routines. I believe it is most useful if the pattern is specified in dB relative to an isotropic source,* which I label dBi; such values will be shown throughout this series. In this pattern program I have also found it convenient to compute and display the driver's driving point impedance; the largest energy flux value found in the entire elevation angle search (usually in 1 degree intervals), which I label as MAX H-GAIN (dBi), and the angle at which it occurs, labeled H-ANGLE; and the front-to-back ratio (ratio of energy flux at H-ANGLE to energy flux at 180 degrees minus H-ANGLE) labeled FRONT/BACK GAIN (dB).

Another program of great utility is one which reads

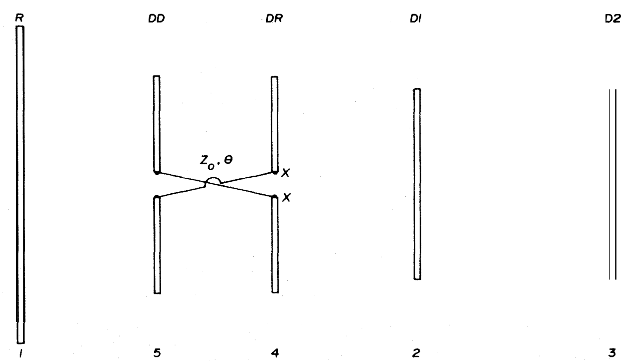


fig. 1. Layout of a Yagi with a broadband feed system of the type manufactured by KLM. The main driver (DR) and dependent driver (DD) are connected through a transposed transmission line with characteristic impedance Z_o and phase angle θ. Performance of this type of Yagi can also be analyzed by computer, as discussed in the text.

the new input file (element current file) and computes and displays a number of useful Yagi properties. These are maximum gain (dBi) and angle at which it occurs, reverse gain (dBi) at the same reverse elevation angle, front-to-back ratio, and the driving point impedance of the first driver. Since this program does not spend time displaying the complete patterns, it is fast and therefore capable of rapidly running through a large number of different situations.

other antenna types

Programs can be written that will perform equivalent calculations for different antenna systems. First of all, a somewhat different element current program is needed for a conventional Yagi if the driver *current* is specified rather than its voltage (using **eqs. 10** and **11** from reference 1). You can also write a different program to handle a broadband drive system as used in the KLM antennas, but to do this requires a little manipulation. In this type of antenna the drive system is not just a single element but actually consists of a voltage-fed main driver whose input terminals also feed a dependent driver in parallel through a crossed transmission line with characteristic impedance Z_0 and phase delay angle Θ. With conventional transmission line equations you can relate voltages and currents at both ends of this transmission line.

As an illustration, a 5-element KLM-style beam can be simulated as shown in **fig. 1.** The parasites are elements 1, 2, and 3; the main or master driver is 4, and the dependent or slave driver is 5 (connected through the transposed transmission line). Start with the five linear equations in matrix form:

$$\begin{vmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} & Z_{15} \\ Z_{21} & Z_{22} & Z_{23} & Z_{24} & Z_{25} \\ Z_{31} & Z_{32} & Z_{33} & Z_{34} & Z_{35} \\ Z_{41} & Z_{42} & Z_{43} & Z_{44} & Z_{45} \\ Z_{51} & Z_{52} & Z_{53} & Z_{54} & Z_{55} \end{vmatrix} \times \begin{vmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \\ I_5 \end{vmatrix} = \begin{vmatrix} 0 \\ 0 \\ 0 \\ V_4 \\ V_5 \end{vmatrix} \qquad \textbf{(1)}$$

V_4 (the main driver voltage) is given, but V_5 (dependent driver voltage) must be determined from the transmission line equations which relate V_5 and I_5 to V_4 and transmission line current at terminals XX. For a transposed lossless transmission line:

$$V_5 = -V_4 \sec\Theta - jZ_0 I_5 \tan\Theta \qquad \textbf{(2)}$$

so that the final five linear equations become in matrix form:

$$\begin{vmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} & Z_{15} \\ Z_{21} & Z_{22} & Z_{23} & Z_{24} & Z_{25} \\ Z_{31} & Z_{32} & Z_{33} & z_{34} & Z_{35} \\ Z_{41} & Z_{42} & Z_{43} & Z_{44} & Z_{45} \\ Z_{51} & Z_{52} & Z_{53} & Z_{54} & Z_{55}+jZ_0 \tan\Theta \end{vmatrix} + \begin{vmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \\ I_5 \end{vmatrix} = \begin{vmatrix} 0 \\ 0 \\ 0 \\ V_4 \\ -V_4 \sec\Theta \end{vmatrix} \qquad \textbf{(3)}$$

Note that the input file for this antenna must contain information about the number of parasites, the number of independent or master drivers, and the number of dependent or slave drivers; the dependent or slave driver must contain a statement as to which is its master driver, and must provide Z_0 and Θ for the transposed transmission line. The Z matrix and V vector must be modified as shown, then the program can proceed as before.

validation

With the tools for computing a wide spectrum of antenna characteristics at hand, it is crucial to ask for some experimental validation. I have already commented on the inaccuracies inherent in physical modeling (although such inaccuracies are believed to be of little consequence), and have discussed approximations which have been used in various computations. These approximations are expected to be most serious for those antenna properties

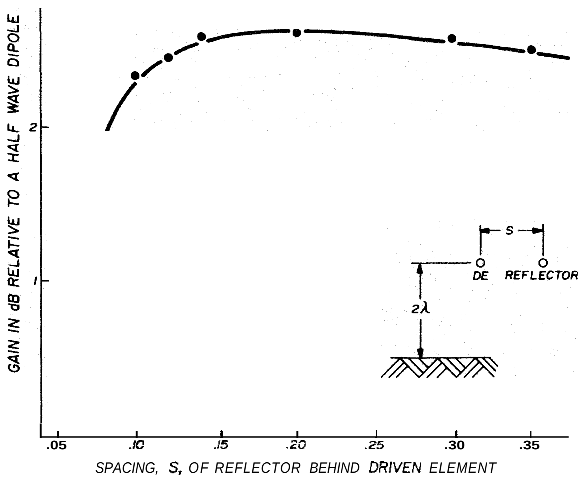


fig. 2. Gain in dB of a driven element and reflector for different spacings between elements (*from* NBS *Report 688*).

which depend critically on precise element currents, such as F/B ratio. On the other hand, it should be possible to calculate other properties such as gain or directivity with reasonable accuracy (comparable to the accuracy of the currents themselves). The calculations are expected to produce superior results for short antennas (few elements) and for frequency regions relatively close to design center frequency.

Experimental information suitable for validation is frustratingly difficult to obtain! While a great amount of experimental results have been published, it is extremely difficult to find examples of *accurate* measurements made under conditions where external factors have been properly considered or eliminated.

NBS Yagi experiments

For example, let's examine the experimental results reported by Peter Viezbicke in *NBS Technical Note 688*.[3] This publication provides a rich range of gain and pattern measurements for a wide variety of Yagi configurations. In most cases, all relevant dimensions of the Yagi design are given, so this publication is a fine vehicle to test the validity of theoretical computational methodology; shortly, I will attempt to make such comparisons. At the same time, *Technical Note* 688 is a frustrating report; it contains several cases of inconsistent information and many times there is lack of supporting technical information. Careful documentation and editing at the time of publication would have made this report of a really excellent series of measurements much more valuable.

I must first comment on the NBS experimental approach. Viezbicke states that *all* tests on the (receiving) test antennas were carried out using a nonconducting Plexiglas boom mounted three wavelengths (3λ) above ground (readers of the NBS paper will note that the very first figure shows 2λ). The (transmitting) generator antenna was mounted 320 meters (1050 feet) away and at a height above ground to illuminate the receiving antenna at "grazing angles." I interpret this to mean that the transmitter is also at a height over ground of 3λ. The nature of the intervening ground, which is highly relevant, is never mentioned! A reference half-wavelength dipole is mounted about 5λ to the side of the antenna under test (also at 3λ above ground). All tests were made using a frequency of "400 MHz" but the frequency precision is never mentioned, and this may be quite important, as shall be seen.

If my interpretation of the experimental setup is correct, received field strength at the receiving site will be the vectorial sum of the field from the direct ray and the field from the ray reflected from the ground at a point midway between transmitter and receiver. At these grazing angles, the reflectivity of the ground is probably near unity and for horizontally polarized radiation ground reflection will give a phase change of 180 degrees. Thus, the two rays interfere and nearly cancel each other, making the received energy nearly zero! It can be seen instantly that the actual received energy is *extremely* sensitive to the nature of ground midway between transmitter and receiver, *i.e.*, its true reflection coefficient, its height, and its degree of flatness (which influences its focus-

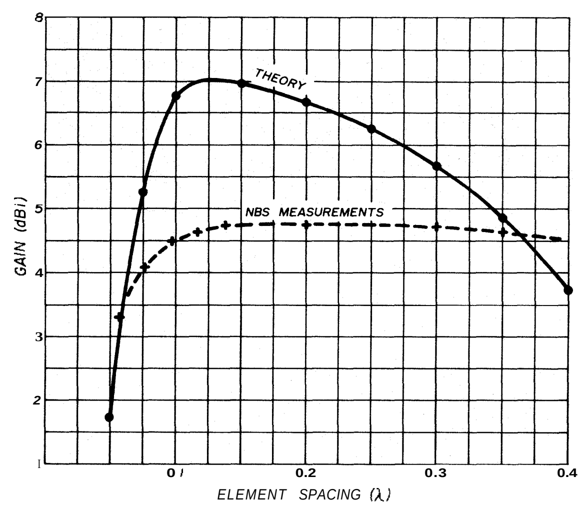


fig. 3. Comparison of computed gain with published NBS measurements (see fig. 2) for the same 2-element Yagi shows very poor agreement. As discussed in the text, however, the disagreement was apparently caused by measurement errors; when the NBS E and H plane patterns for the same antenna ($s = 0.2$) are summed and properly averaged, the isotropic gain is calculated to be 6.71 dBi (compared with 6.70 dBi in the THEORY curve in the graph).

ing properties). Moreover, since the reference antenna "sees" a different ground patch midway to the transmitter, its inherent sensitivity can be different from that of the antenna under test. For example, if the ground patch seen by the reference antenna is only 1 foot (30 cm) higher than the ground patch seen by the test antenna, a systematic error of about 2 dB will occur! It is clear, therefore, that the NBS experimental setup invites systematic errors.

Yagi gain

Let me turn now to the experimental results reported by Viezbicke. I will start with the measured gain of a 2-element (dipole and reflector) Yagi as a function of spacing S. This is shown in the NBS paper as Fig. 1 and for convenience is reproduced here as **fig.** 2. Unfortunately, Viezbicke failed to state an essential dimension — the actual length of the reflector itself; I assume that it is 0.482λ (which was usually used for other Yagis). On this assumption I have calculated a theoretical result shown in **fig.** 3. The theoretical gain is expressed in dBi (referenced to an isotrope); a direct comparison can be made by adding 2.15 dB to the experimental NBS findings.

The comparison in **fig.** 3 is totally unsatisfactory; not only are the shapes of the plots different, the peak gains (at say $S = 0.2$) are very different! Viezbicke gets 4.77 dBi, whereas I calculate 6.70 dBi. In an attempt to understand this conflict, look at Fig. 12 in the NBS publication which illustrates the measured pattern of the same 2-element Yagi ($S = 0.2$) in both E and H planes. I have numerically summed and properly averaged the (power) response over the total 4π solid angle and have found the directivity, or isotropic gain, to be 6.71 dBi! Thus, the measured pattern exhibits gain which agrees with my theoretical result (see **fig. 4**). I can only conclude that some unexpected system error was present for the NBS series of measurements shown in their Fig. 1.

table 1. Optimized lengths of parasitic elements for Yagi antennas of six different lengths (reflector spaced 0.2λ behind the driven element, element diameter 0.0085λ).

	Length of Yagi in Wavelengths					
	0.4	0.8	1.2	2.2	3.2	4.2
Length of Reflector, λ	0.482	0.482	0.482	0.482	0.482	0.475
Director Length, λ						
1st	0.442	0.428	0.428	0.432	0.428	0.424
2nd		0.424	0.420	0.415	0.420	0.424
3rd		0.428	0.420	0.407	0.407	0.420
4th			0.428	0.398	0.398	0.407
5th				0.390	0.394	0.403
6th				0.390	0.390	0.398
7th				0.390	0.386	0.394
8th				0.390	0.386	0.390
9th				0.398	0.386	0.390
10th				0.407	0.386	0.390
11th					0.386	0.390
12th					0.386	0.390
13th					0.386	0.390
14th					0.386	
15th					0.386	
Spacing between directors, in λ	0.200	0.200	0.250	0.200	0.200	0.308
Gain relative to half-wave dipole, dB	7.1	9.2	10.2	12.25	13.4	14.2
Isotropic gain, dBi	9.25	11.35	12.35	14.40	15.55	16.35

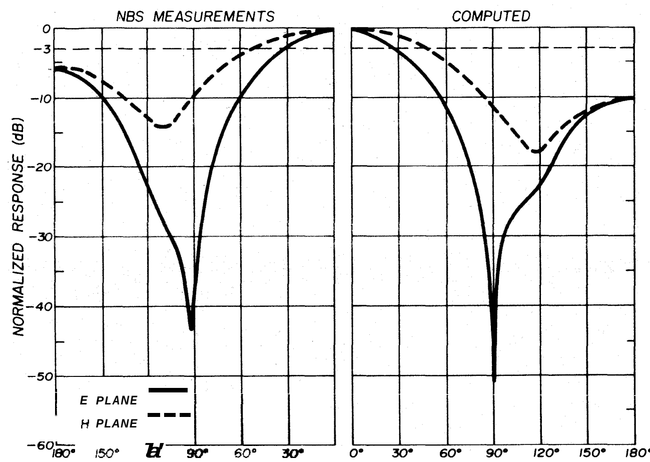


fig. 4. Comparison of the E and H plane radiation patterns for a 2-element Yagi with 0.2λ spacing as measured by the National Bureau of Standards, left, and predicted by theory, right, shows good agreement.

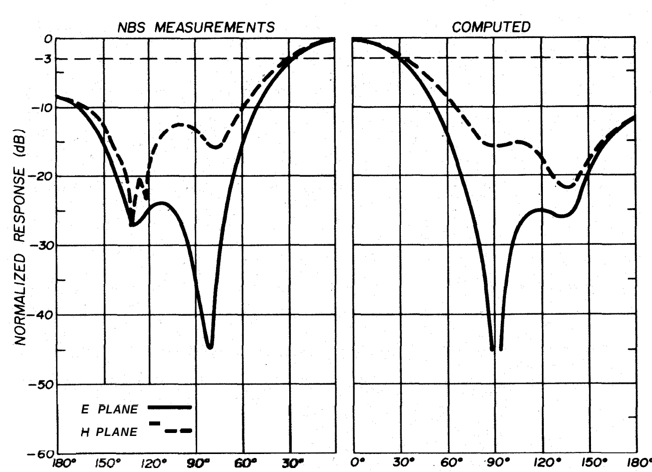


fig. 5. Comparison of the E and H plane radiation patterns for a 3-element, 0.4λ long Yagi, measured by NBS, left, and predicted by theory, right.

In the NBS paper a set of six specific Yagi designs is shown ranging in overall length from 0.4 to 4.2 wavelengths. If you refer directly to the NBS publication you will note that the first director for the 0.4λ 3-element design is shown in NBS Fig. 9 to be 0.442λ long rather than 0.424λ, as shown in NBS Table 1. With this correction, the specifications for these six NBS Yagi designs are listed here in **table 1.**

I have calculated the theoretical gains for these six Yagi designs and also their theoretical patterns for comparison with those shown in the Viezbicke paper (NBS Figs. 14 through 19). For convenience in making comparisons, the NBS patterns and my newly calcuiated theoretical patterns are shown side by side

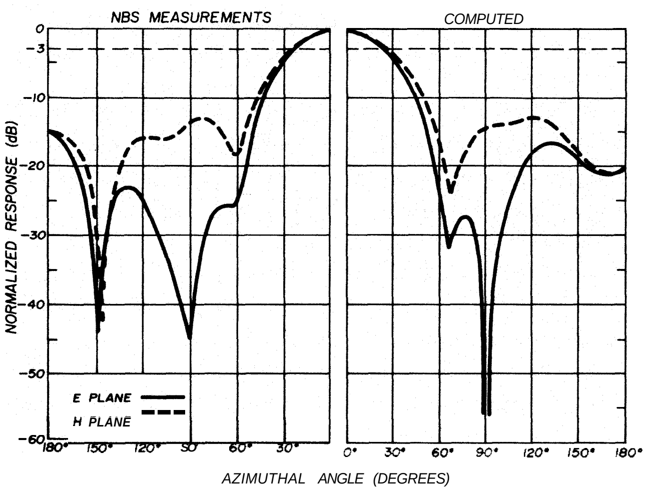


fig. 6. Comparison of the E and H plane radiation patterns for a 5-element, 0.8λ long Yagi, measured by NBS, left, and predicted by theory, right.

in **figs. 5** through **10.** It is apparent that for all cases there is a striking similarity, not only in the qualitative details of lobe structure, but in most quantitative aspects! Careful scrutiny of the experimental results show some variances between the two halves; theoretically, of course, the two halves are totally symmetrical. Agreement of this kind is gratifying and demonstrates that the experimental patterns were made with great care and also that the theoretical calculations seem to give valid answers. This is especially comforting since the 4.2λ Yagi design is very long and contains many parasites. This is precisely the situation where theoretical approximations (cylindrical element resonant lengths, mutual, and self-impedances) are most sensitive.

The gain of each of these Yagi designs can be obtained in several ways, and it is illuminating to compare all methods. **Table 2** shows a comparison of three experimental methods, all derived from the

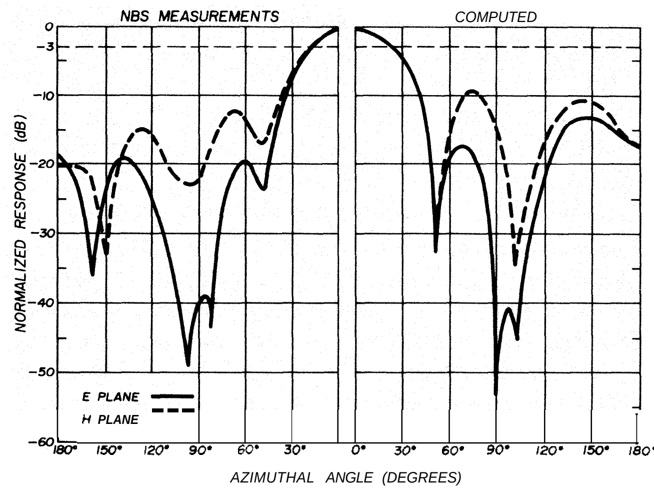


fig. 7. The E and H plane radiation patterns of a 6-element Yagi with a length of 1.2λ – measured [left) vs theory (right).

NBS data, and my theoretical calculations. Column 1 shows the NBS measured gain referenced to a half-wavelength dipole, but corrected to give isotropic gain (dBi). Column 2 shows the gain calculated from the measured half-power main beam angles by the usual formula, $dBi = 10 \log (41253/\Theta_H \Theta_E)$, where the H and E half-power angles are measured in degrees.[4] The third column is derived entirely from the experimental NBS patterns; in each case I have calculated the directivity by appropriately summing all 10 degree intervals. This pattern averaging, if carefully done, should yield a reasonably reliable result, free from any systematic error (due to ground reflections, for example). The last column is the result of my theoretical calculations made on each design.

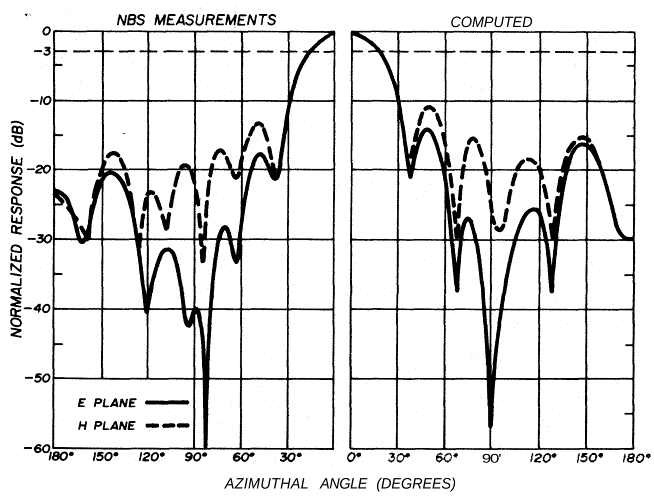


fig. 8. The E and H plane radiation patterns for a 12-element Yagi with a length of 2.2λ – measured (left) vs theory (right).

In table 2, note that column 1 (NBS measurements) and column 4 (theory) are in agreement within Viezbicke's stated estimated accuracy of about 0.5 dB with the exception of the value for the 2-element Yagi (discussed earlier). Column 2 is derived from measured half-power angles; it is quite difficult to determine such angles with any great precision and, moreover, the method is nothing more than a crude approximation at best. Column 3 (derived from NBS experimental patterns) is in remarkable agreement with theory in all cases, especially considering that the summation in 10 degree intervals is really too coarse and that values in these 10-degree intervals were only "eyeballed" from the published NBS patterns.

table 2. Gain of six different NBS-designed Yagi antennas, in dBi, as determined by four different methods.

NBS Yagi type	NBS measurements	calculated from half-power beamwidth	pattern integration	computer derived
2 **element** (0.2λ)	4.77	7.50	6.71	6.70
3 **element** (0.4λ)	9.25	10.02	9.62	9.16
5 **element** (0.8λ)	11.35	11.86	11.41	10.73
6 **element** (1.2λ)	12.35	13.90	12.64	11.80
12 **element** (2.2λ)	14.40	15.28	14.28	14.04
17 **element** (3.2λ)	15.55	16.63	15.47	15.20
15 **element** (4.2λ)	16.35	17.38	16.22	15.71

table 3. Measured and calculated gain in dBi of Yagi antennas with average director lengths.

NBS Yagi type	director length	NBS measured gain	computed gain
5 **element** (0.8λ)	0.4260λ	11.27	10.68
6 **element** (1.2λ)	0.4240λ	12.24	11.71
12 **element** (2.2λ)	0.4017λ	13.92	13.62
17 **element** (3.2λ)	0.3946λ	14.83	14.68
15 **element** (4.2λ)	0.4008λ	15.55	15.15

effect of director length

Fig. 7 in the NBS report shows an interesting experimental result: The selected designs are superior in gain to simplified Yagis (all directors of equal length) for booms longer than about one wavelength. Although Viezbicke did not give the lengths of the directors used for the simplified Yagis, I have made calculations using directors with lengths that are the *average* of those used in each of the five longer Yagis. His measurements of gain and my theoretical calculations are shown in table 3. Again, these results are in satisfactory agreement.

Fig. 11 shows a graph of my theoretical results for

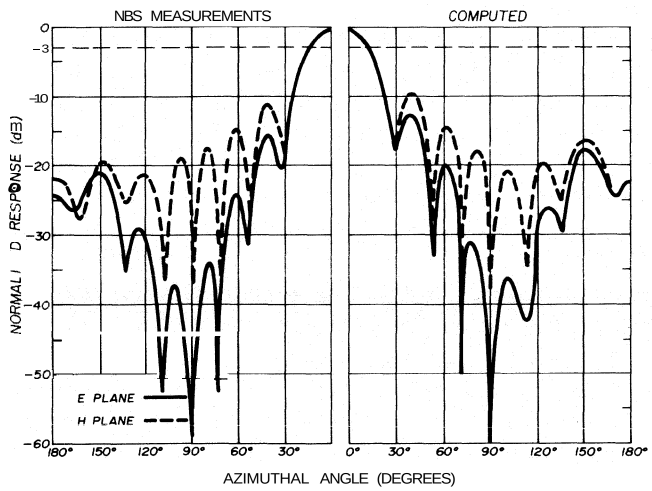


fig. 9. E **and** H **plane radiation patterns of a 17-element, 3.2λ long Yagi — measured (left) vs theory (right).**

both tables **2** and 3, together with the published NBS experimental points. Note that the theoretical results support the idea that while the simplified (equal director length) Yagi is just as good as a more sophisticated design for booms shorter than one wavelength, a slight gain improvement in the gain of long Yagis is apparently possible by using directors of different lengths.

Viezbicke's Fig. 9, reproduced here as **fig. 12,** shows the results of an interesting experiment in which gain for a given Yagi design was measured using a series of different director lengths. Gain curves were produced for each of a wide range of element diameters. These measurements allow an interesting test of the theory of the reactance of cylindrical elements (see eqs. **4,** 5, and ***6*** in my previous article).[1] In principle, the peak of each gain curve

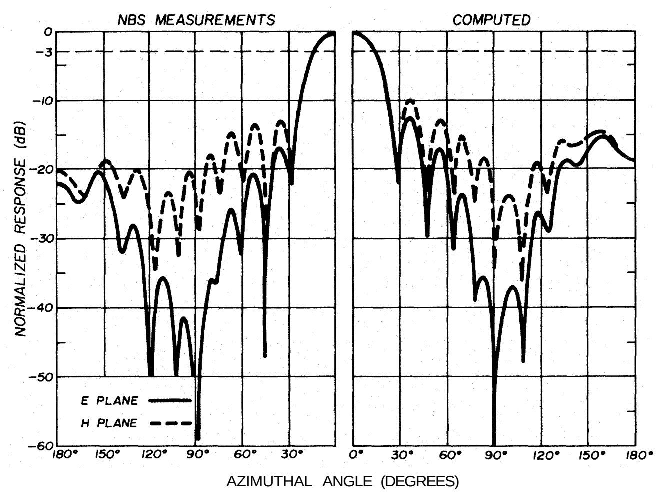


fig. 10. E **and** H **plane radiation patterns of a 15-element, 4.2λ long Yagi — measured (left) vs theory (right).**

should correspond to a single element reactance; this reactance value should, of course, be the same for all peaks. The reactance of the directors can be calculated from **eqs. 4, 5,** and **6** using the lengths of an element which gives the peak gain for each element diameter. From **fig. 12,** I have estimated, as carefully as possible, the lengths of elements which produce peak gain; **table 4** shows the results.

The standard deviation actually corresponds to an element length variation of only 0.3 per cent; it is likely that such an error could easily occur in estimating the position of the peak gain or even in physically constructing the elements used in the experiment. Thus, this experiment seems to give strong support for the theory of cylinder length resonances.

table 4. Element lengths (estimated from fig. 12) which produce maximum gain, their resonant length *LER*, and the reactance *X* in ohms. Average reactance is -77.22 ohms: standard deviation is 2.24 ohms.

element diameter	length inches	(cm)	LER	*X* (ohms)
0.08 cm	13.163	(33.4)	0.4816λ	−81.07
0.16 cm	13.033	(33.1)	0.4791λ	−75.63
0.32 cm	12.767	(32.4)	0.4759λ	−75.88
0.64 cm	12.358	(31.4)	0.4714λ	−78.57
0.95 cm	12.093	(30.7)	0.4680λ	−78.28
1.27 cm	11.860	(30.1)	0.4650λ	77.93
1.59 cm	11.674	(29.7)	0.4622λ	−76.68
2.54 cm	11.191	(28.4)	0.4548λ	−73.70

gain variations

Perhaps one of the most interesting experimental results of the NBS work is the strange "oscillating" gain characteristic shown in Viezbicke's Figs. 4, 5, and 6, reproduced here for convenience in **figs. 13, 14,** and **15.** I have attempted to calculate a number of these cases and although the oscillating gain phenomenon does show up, the detailed agreement between experiment and theory is not impressive. However, I have found that the exact behavior of these long and heavily (director) loaded arrays is critically dependent on frequency and/or exact director lengths. Accordingly, I have run calculations for several frequencies around the central frequency (normalized to unity). The results are shown in **fig. 16,** where the theoretical calculated points for each frequency are connected by line segments for clarity.

Superimposed on the plots of **fig. 16** are Viezbicke's experimental points, which should correspond to the theoretical calculations. The theoretical results generally show that the accurately computed points do not really lie on a smooth curve; this is the reason connecting lines are used. Theory *does* give the oscillating gain phenomenon, but it also shows

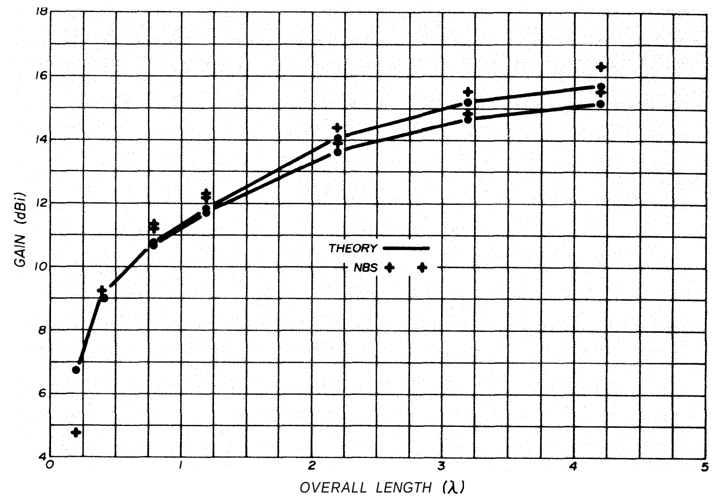


fig. 11. **Graphical comparison of the data from tables** 2 **and 3 showing theoretical (computed) gain and measured gain figures published by NBS. Computed curves support NBS *Report 688* which shows slight gain increase is possible with different-length directors.**

that the details depend sharply on the *exact* transmitter frequency. Pretty good agreement can be obtained between experiment and theory if it is assumed that slightly different frequencies (±1 to 3 per cent) were used from experiment to experiment. It is perhaps significant that the most likely frequency does not appear to be systematically high or systematically low. It is unfortunate that Viezbicke neither specified frequency accuracy, nor published the exact frequency for each experiment.

The theoretical results are interesting from three points of view. First, for a given frequency the calculated points do not appear to lie on a smooth curve; it is likely that the gain varies somewhat with side- and back-lobe structure. The calculated *F/B* ratio varies greatly from point to point and it would seem reasonable that there would be some reaction on forward

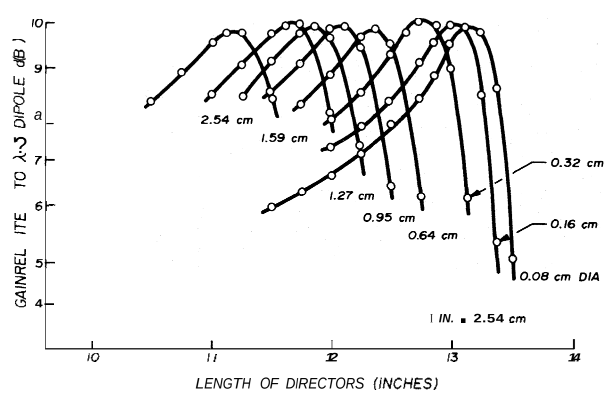


fig. 12. Measured gain vs director length at 400 MHz for a Yagi 1.25λ long, using three directors of different length and diameter, spaced 0.35λ.

gain. Secondly, for long Yagis the results vary significantly with the exact frequency used! Third, detailed calculations show that beyond the end of the first "oscillating" gain cycle, Yagi gain is not maximum in the plane of the Yagi; instead, the front beam appears to develop a central dimple causing maximum gain to occur at an appreciable elevation angle. (The plots in **fig. 16,** however, show only the gain at zero elevation angle because this is what was presumably done in the NBS experiments).

Note that all three of these behavioral aspects of long Yagis derive from theoretical computations; they were not even suspected by the NBS experimenters! This illustrates the point that an extensive variety of Yagi configurations can be explored *much*

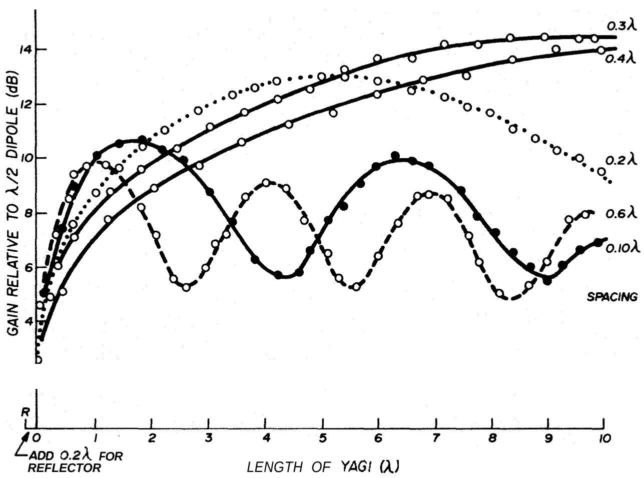


fig. 13. Yagi gain as a function of length (number of directors) for different constant spacings between directors of length equal to 0.382λ (Fig. 4, *NBS Report 688*).

faster theoretically, and with much more relative accuracy, than can be accomplished experimentally. As a result, new concepts and understanding are emerging.

summary

Let me summarize the overall comparison of the experimental results of the NBS group with the theoretical results. The comparisons show total agreement on Yagi *gain* for an astonishing range of models with the single exception of the direct measurements on a 2-element beam. A gain figure derived from the measurement of the *pattern* of the 2-element bearn, however, *does* agree with theory; it is my belief that some unanticipated error was made by the NBS group in their measurements.

The comparisons also show excellent agreement for Yagi *patterns,* again over a wide range of models. This should be a very sensitive test of the accuracy with which mutual and self-impedances are repre-

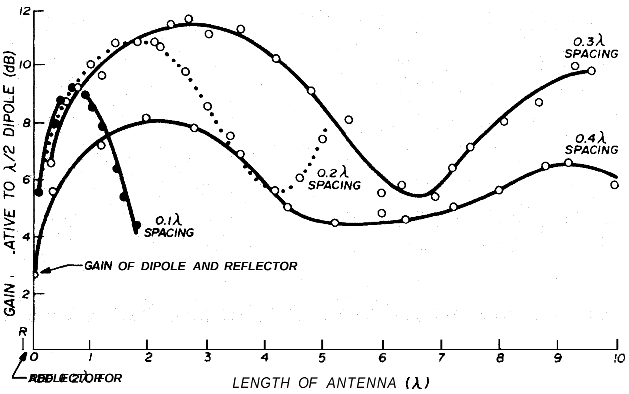


fig. 14. Yagi gain as a function of length (number of directors) for different constant spacings between directors of length equal to 0.411λ (Fig. 5, *NBS Report 688*).

sented, as well as the resonant frequency calculations of cylindrical elements. This latter point is also strongly supported by the consistent values of parasitic element reactance for elements whose diameters vary over a range of greater than 30 to one.

Finally, the strange "oscillating" gain phenomenon observed by Viezbicke can be reproduced theoretically. Agreement is only qualitative if the NBS group used an *accurate* 400.0 MHz frequency in their tests; however, the comparisons suggest that they may have actually used a nominal "400 MHz," with frequencies within a range of ±3 per cent. In this event, the comparisons indicate a potentially remarkable agreement in quantitative aspects as well.

Since these comparisons include very long Yagi models (up to 6λ long) and correspondingly large numbers of parasitic elements (up to 40), it seems certain that the computational methodology I have outlined should be generally trustworthy. This should be especially true for shorter antennas (with fewer elements) with which I will be primarily concerned throughout this series of articles. Especially noteworthy is the fact that all computed results contain

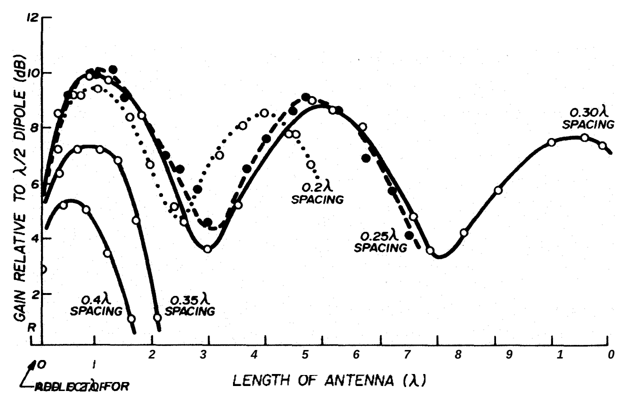
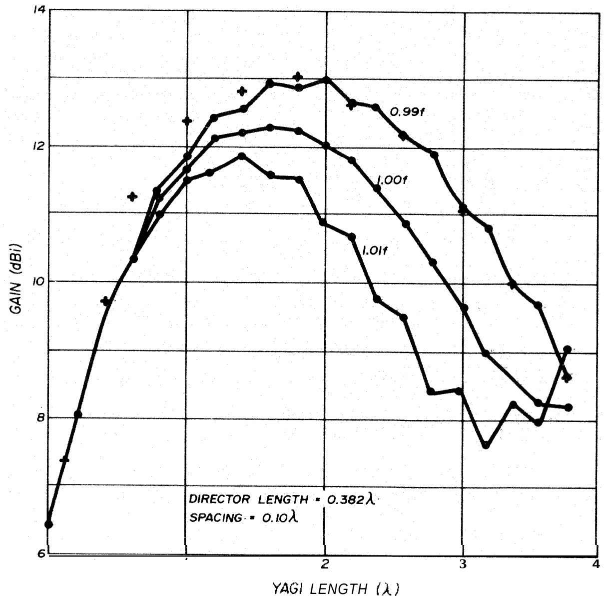
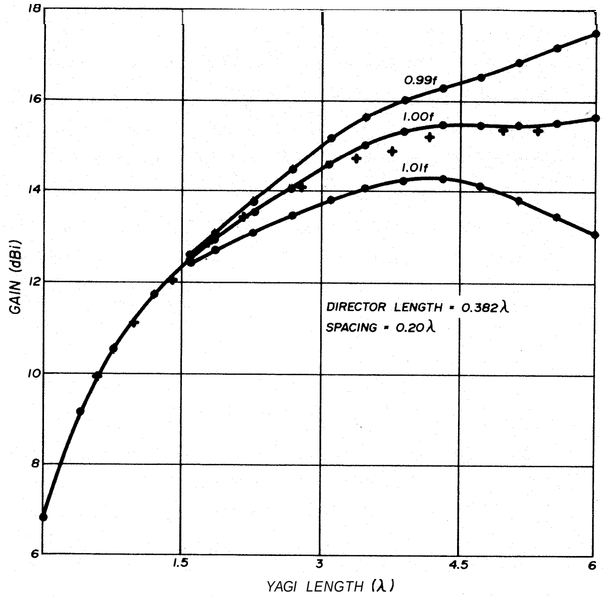
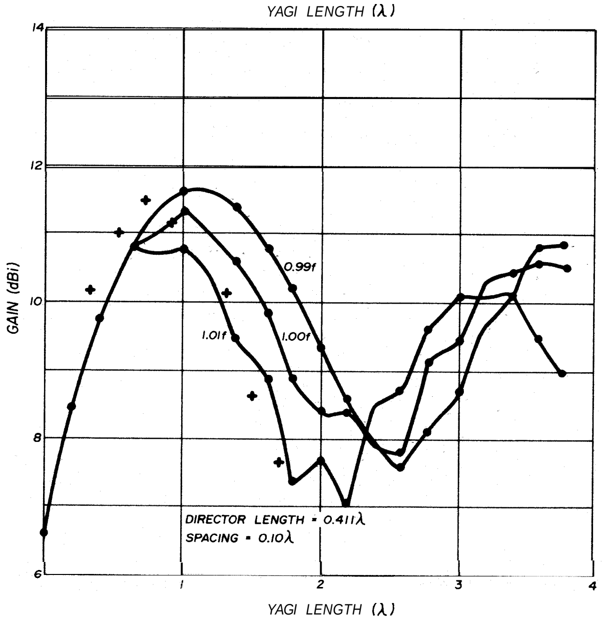
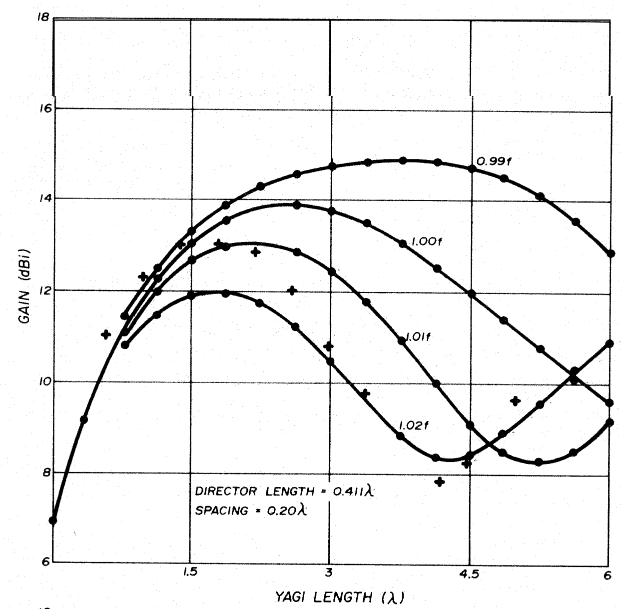


fig. 15. Yagi gain as a function of length (number of directors) for different constant spacings between directors of length equal to 0.424λ (Fig. 6, *NBS Report 688*).









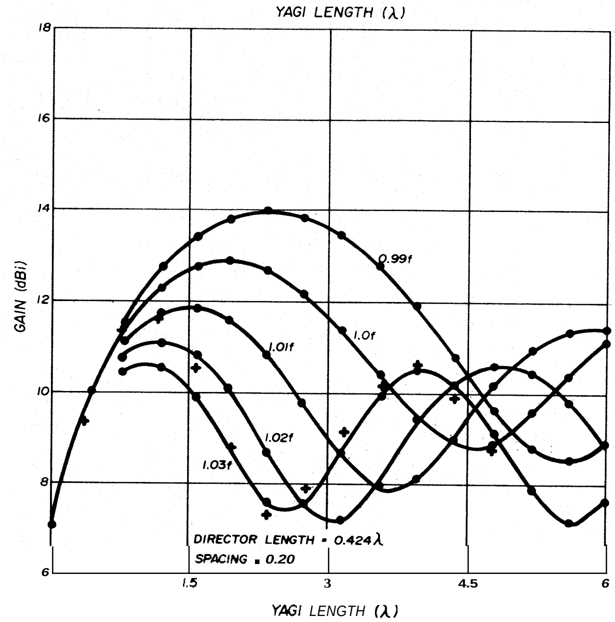


fig. 16. Graphs of Yagi gain vs overall antenna length for various director spacing and director lengths at small frequency differences; frequency designations on each plot have been normalized to the center design frequency. Points which are not interconnected are experimental data published in NBS *Report 688*.

no adjustable constants or parameters; they are all derived from basic physical principles using adequately accurate mathematical approximations.

references

1. James Lawson, "Calculating Yagi Performance," *ham radio,* January, 1980, page 22.
2. James Lawson, "Antenna Gain and Directivity Over Ground," *ham radio,* August, 1979, page 12.
3. Peter Viezbicke, "Yagi Antenna Design," NBS *Technical Note 688, U.S.* Department of Commerce. Washington, DC, December. 1976.
4. John Kraus, *Antennas,* McGraw-Hill, New York, 1950, page 25.

ham radio

high-performance broadband balun

Theory and design of broadband 1:1 baluns, with construction details for an improved model of the well-known W1JR balun

Sometimes equipment that looks the simplest to build is really the trickiest to make work, and vice versa. Take, for instance, a balun. What sounds simpler than wrapping a few turns of transmission line around a toroid and getting a balun? Or do you? My experience is that you may, but then again you may not.

Over the last few years many articles have appeared on the 1:1 balun, but absolutely no information — to the best of my knowledge — on why they work, how to design one, or, equally important, how to test one.

Almost 100 years ago, the eminent British physicist, Lord Kelvin, said, in effect, "If you can express what you are speaking of in numbers, you understand it. If you cannot express it in numbers, your knowledge is very meager and unsatisfactory." From this standpoint, our knowledge of 1:1 baluns is "very meager and unsatisfactory" indeed!

In this article I give a brief description of the historical development of the balun, explain how a 1:1 balun works, include a procedure for designing a balun, show some pitfalls that are all too easy to fall into, and finally present some test results on a balun I have built.

brief history

The concept of a broadband balun made by winding a transmission line into an inductance originated in 1944 by G. Guanella[1] of the Brown-Boveri Company in Switzerland. In 1959, Ruthroff[2] of Bell Laboratories wound the transmission line onto a ferrite toroid and obtained bandwidth ratios as high as 20,000:1. His paper appears to be the takeoff point for the present balun designs. In the Amateur literature, Dick Turin,[3] Jerry Sevick,[4] Bill Orr,[5,6] and Joe Reisert[7] have all presented articles on broadband baluns.

theory of operation

Consider a balanced transmission line of characteristic impedance $Z_{ch} = R_L$, wound into an inductance, and feeding a balanced load R_L, as shown in **fig. 1 (A).** The normal transmission line currents, i_1 and i_2, are out of phase; their magnetic fields cancel, and the inductive reactance is essentially zero or, at least, very low compared with the transmission-line characteristic impedance. As far as these currents are concerned, the balun appears only as an added length of transmission line, of characteristic impedance Z_{ch}, tied to the end of the usual unbalanced transmission line whose characteristic impedance is also Z_{ch}. If the load is balanced with respect to ground, the voltage at the junction point G will be zero with respect to ground; the current i_g will also be zero.

That this is true can be seen from the following argument: Assume that the voltage at G is not zero; the unbalance can be represented by a voltage generator, which causes a current i_g to flow, as shown in **fig. 1 (B).** This current divides into two in-phase currents, $i_g = i_3 + i_4$, which flow through the two windings of the balun. Since these currents are in phase, their magnetic fields will add instead of cancelling, and the balun winding will appear as an inductance. If the inductive reactance of the balun winding is sufficiently large compared with $R_L/2$, the in-phase currents that cause the unbalance will be essentially zero compared with the normal transmission line currents, i_1 and i_2.

Thus, if the balun is working properly, it will provide a high impedance to in-phase components and a low impedance to out-of-phase components, effectively isolating the balanced and unbalanced sides of the balun.

In theory, then, the problem of designing a balun boils down to simply winding an inductance to provide a high impedance over the required frequency

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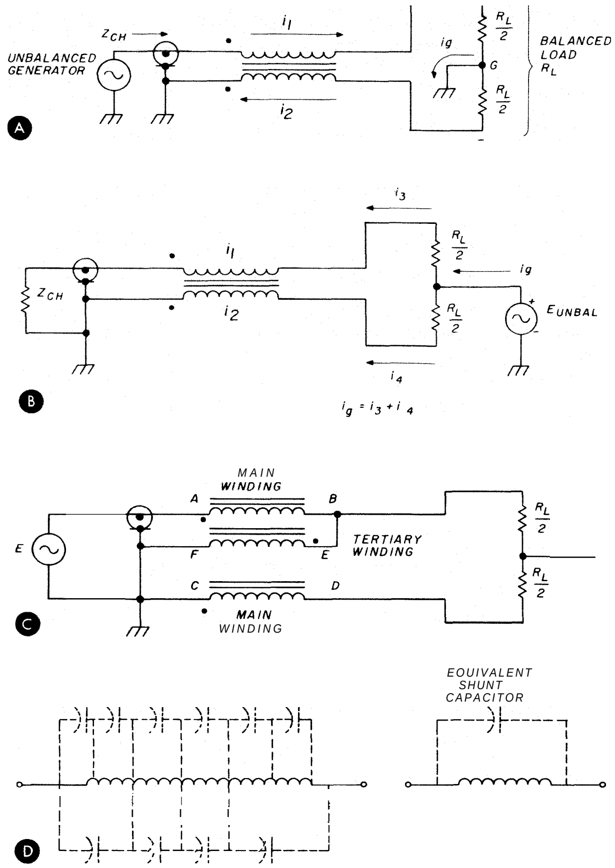

fig. 1 (A): Basic 1:1 balun schematic. Circuit transforms an unbalanced transmission line of impedance Z_{ch} to a balanced load R_L, where $R_l = Z_{ch}$. (B): Schematic of a 1:1 balun showing an unbalanced condition represented by an unbalanced generator, E_{UNBAL}. (C): A 1:1 balun with a tertiary winding, E-F. (D): The left-hand drawing shows an inductor with a few of the possible infinitesimal stray capacitors. These can be equated to one shunt capacitor, as shown in the right-hand drawing.

range. Unfortunately, life is not always simple; neither are baluns, for although their theory is simple, their actua! design is quite involved. First, designing an inductance to provide a high impedance over a wide frequency range can be a substantial problem. Second, for the winding to be a high impedance over any frequency range, the current required to magnetize the core must flow without disrupting the desired signal current and the voltage relationship existing in the balun or load. In a 1:1 balun, this is accomplished by adding a tertiary winding, as shown in **fig. 1 (C)**. This tertiary winding, in turn, must be designed and constructed so that it does not create any additional problems of its own. But, as we'll see later, although the tertiary winding will considerably complicate matters, it is absolutely essential if acceptable balance and good low-frequency response are to be obtained. I'll now discuss these problems in detail as they are related to balun design.

inductance

In designing an inductor to be used over a wide bandwidth, the basic problem is capacitance — stray capacitance. Schematically, an inductor is drawn as a coil of wire; note the solid line in **fig. 1 (D)**. Physically, however, stray capacitance is present as suggested by the dotted lines. Actually, the capacitance is distributed; *i.*e., it exists between every infinitesimally small length of coil and every other infinitesimally small length of coil; it *is not* discrete, as implied in **fig. 1 (D)**. The effectiveness of these incremental capacitors depends on how much energy each one stores. The energy stored by a capacitor is equal to $energy = \frac{1}{2} CE^2$, where C is the capacitance in farads and E is the voltage across the capacitor terminals; therefore, it is important to minimize the capacitance across those parts of an inductance where the voltage is the largest; *i.e.*, at the ends of the coil. Unhappily, this is not always possible!

Since it's very difficult to consider all the incremental capacitors in analyzing the behavior of an inductance, it has become standard practice to hypothesize a single capacitor connected across the entire coil storing the same amount of energy as all of the incremental capacitors and to call this the stray capacitance of the coil.

As the frequency across the inductance is increased, a frequency will be reached where the inductance and stray capacitance will become parallel resonant; at this frequency, the inductor will appear as a very high resistance. As the frequency is further increased, the reactance decreases to zero. At the series-resonant frequency, the coil impedance becomes very low. For a balun, the series-resonant region must be avoided, as all isolation between the balanced and unbalanced sides of the balun is lost. My experience has been that it's necessary to limit operation to about one octave below (one-half) the series-resonant frequency, or else the balun phase and amplitude balance balun will be upset.

Probably the most readable and useful discussion of wideband inductor design in either the professional or Amateur literature was written by Vernon Chambers[8] over 25 years ago. Chambers' article refers to the design of rf chokes, but the basic problems are similar to baluns. More recently, Doug DeMaw[9] has written a very helpful article on toroid inductor design.

As implied previously, one of the most effective ways of reducing the distributed capacitance is to separate the ends of the winding as much as possible. This brings us to the concept of the super toroid.

super toroid

The super toroid was developed in the 1970s by

T.A.O. Gross,[10] primarily as a means of reducing stray pickup from external fields, but was introduced to the Amateur community only recently by Reisert.[7] With the super toroid, one-half the circumference of the toroid is wound in the usual manner. The winding is then taken across the diameter of the core and the last half wound in the opposite direction, as shown in **fig.** 2. The advantage of this type of winding for balun application is that the ends of the winding, where the voltage is the highest, are at opposite sides of the core where the capacitance is minimum.

Every inductor I've wound using the super toroid technique has had bandwidth characteristics superior to the same core wound with an equal number of turns in the usual manner! Why not use the super toroid concept, then, in the balun? The problem is the tertiary winding.

tertiary winding

As we know, a balun functions by providing a high impedance, usually inductive, between the balanced and unbalanced sides. To obtain this inductance, it's necessary that a path be provided for magnetizing current. Ruthroff[2] states that with the balanced load disconnected, a dc path must exist between the unbalanced input side and ground. The way of providing this is to use a tertiary winding as shown in **fig. 1(C)**.

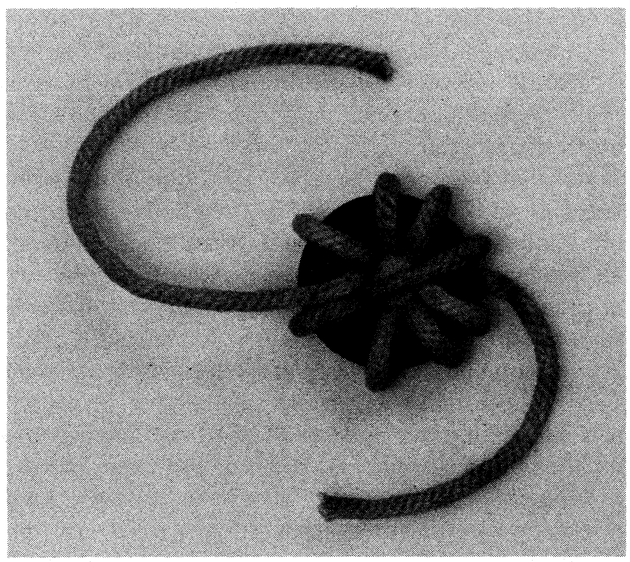

fig. 2. A toroidal core wound with a super toroid winding. This core is wound with clothes line to make it show up better in the photograph.

This circuit is easier to visualize if redrawn in the form of an autotransformer as shown in **fig.** 3. Voltage levels with respect to ground are also shown, and it is seen that the effect is that of a 1:1 autotransformer with the voltages on the balanced side low-

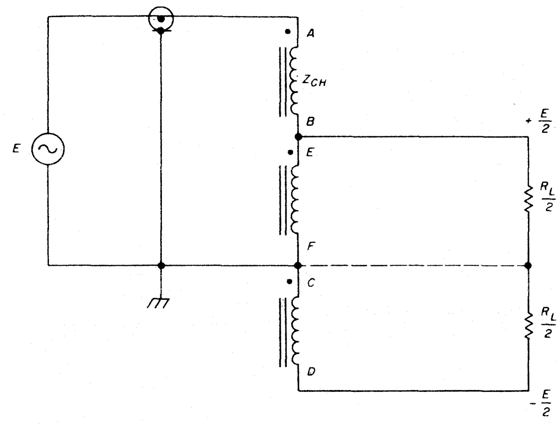

fig. 3. A balun with a tertiary winding drawn in the form of an autotransformer. The voltage across each winding is $\frac{E}{2}$ volts.

ered by $\frac{E}{2}$ volts, compared with the unbalanced input voltage. The total voltage across either the input of the balun or the balanced output, however, is still the same, E volts.

The path for the magnetizing current is also evident: from the unbalanced generator through windings **A-B** and then E-F to ground. The important point to note is that the magnetizing current does not pass through either half of the load impedance. If the balun is to couple a balanced generator to an unbalanced load, and the unbalanced side is open circuited, the magnetizing current will pass through the tertiary winding, E-F, and the tower half of the transmission line, **C-D,** back to the balanced generator without passing through the load impedance.

practical considerations

To show the practical need for tertiary winding, I'll discuss the effects that the lack of a tertiary winding will have on balun performance. Consider the 1:1 balun with no tertiary winding, as shown in **fig. 4.** Because of the instrumentation considerations, it's easier to use a balanced load, consisting of two equal-value resistors, each of value $\frac{R_L}{2}$, and to make measurements at the unbalanced side rather than vice versa, which would require a balanced impedance bridge. The center of these two resistors may be grounded, as in **fig. 4 (B),** or ungrounded, as in **fig. 4 (A).** If the balun is working properly, there will be no difference whether the resistors' center is grounded or not. Any differences can give an important clue as too the type of problem being encountered. In all cases, I'll assume the total load impedance, R_L, is equal to the characteristic impedance, Z_{ch}, of the transmission line.

First, let's assume the center tap of the load is not grounded, as in **fig. 4 (A)**. The resistive component

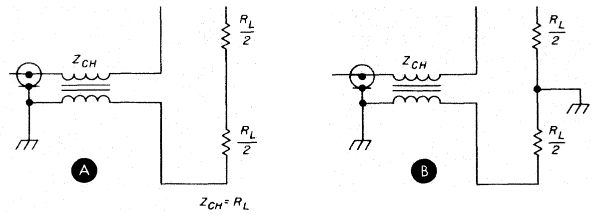

fig. 4. Test circuit for a 1:1 balun *without* a tertiary winding, with and without the load impedance centertap grounded. $Z_{ch} = R_L$ **in both cases.**

of impedance looking into the unbalanced end will be R_L ohms regardless of the frequency. In fact, it will be R_L ohms if measured with a dc ohmmeter.

Thus, you can be lulled into a false sense of security if you depend or: impedance measurements alone. This happened to me. I ran down to the low frequency limit of the signal generator, and the impedance remained on 50 ohms! But let's look at what happens to the balance in the output voltage in both magnitude and phase. This latter point does not appear to have been considered by most writers on baluns.

phase shift

Ideally the voltages between point B and ground and D and ground (fig. **3**) should be equal in magnitude and opposite in phase. Each of these voltages should also be equal to one-half the voltage between **A** and ground (the input voltage). As the frequency is lowered below that at which balun action is effective, the voltage at B will approach that at **A,** and the voltage at D will approach zero. The total voltage across R_L is still equal to the input voltage but not balanced with respect to ground.

The need for magnetizing current can be verified by making voltage and phase measurements across the two halves of the balanced load. The phase and

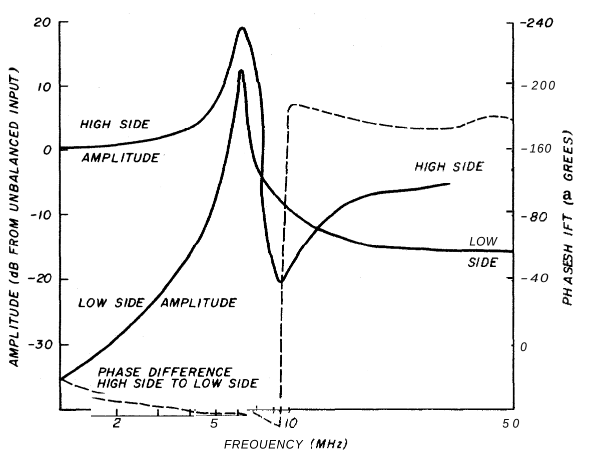

fig. 5. No tertiary winding. Phase and amplitude of balanced output voltages — load centertap not grounded, as in *fig. 4(A).*

magnitude of the output voltage as a function of frequency are shown in fig. **5** for the case where the load resistance centertap isn't grounded. The amplitude measurements are shown in dB with respect to the input voltage; the voltage across each half of the load resistor should be one-half, or 6 dB down, from the input voltage. The phase measurements are the phase difference between the balanced output voltages; these, of course, should be 180 degrees apart. These measurements are plotted in fig. **5.**

After looking at fig. 5, you're probably thinking, "No well-behaved balun would act like that. There must be something wrong with his test setup." My answer is that the balun is *not* well behaved and there's *nothing* wrong with the test setup.

Let's look at the same balun, measured at the same time, with the same equipment, but with the

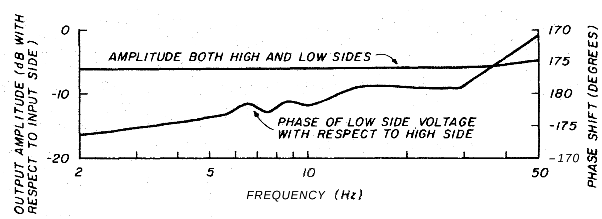

fig. 6. Same as in *fig. 5* but with the load centertap grounded to provide a path for the magnetizing current. This balun is usable between 3 and 36 MHz.

load-resistor centertap grounded. The magnetizing current now has somewhere to flow. The amplitude and phase responses shown in fig. **6** are very close to what one would expect. The balun appears to be usable over 3-35 MHz, just by grounding the load impedance centertap! This, I believe, demonstrates the necessity of providing a path for the magnetizing current.

Before leaving the subject, look at the input impedance of the balun with and without the load centertap grounded, as shown in fig. **7.** There is very little difference between the two impedance curves, so we can conclude that impedance measurements by themselves are not necessarily a good measure of a balun's actual performance.

So, a path for the magnetizing current is absolutely necessary, but grounding the centertap may not be practical in all cases and it causes an unbalance even when it is possible. This result leads us to look at another way of providing a path for the magnetizing current, namely, a tertiary winding.

coupling between windings

As previously noted, the tertiary winding is connected between the high side of the balanced output and ground as shown by E-F in figs. **1(C)** or 3. You'll

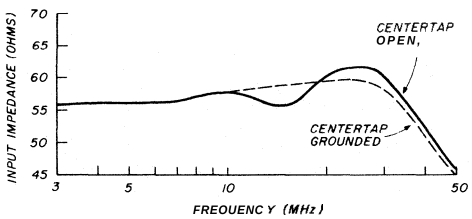

fig. 7. Input impedance of 50-ohm balun without tertiary winding. Load impedance = 55.7 ohms balanced.

observe that it must support a voltage of $\frac{E}{2}$ *volts,* the same as each of the main windings. In fact, if accurate amplitude and phase balance are to be maintained at the balanced terminals, the voltage across all three windings must be the same at all frequencies. This implies that the coupling between all three windings must be very tight; indeed, the tightness of coupling is one factor that determines the balun bandwidth, a need that doesn't seem to have been pointed out before.

The degree of coupling between the main windings is not a problem as these windings are part of a transmission line. The coupling between these windings and the tertiary winding, however, is crucial. I'll discuss how to accomplish tight coupling when I describe construction details.

The requirements imposed on the tertiary winding are basically the same as those on the main winding: first, the tertiary must support a voltage of $\frac{E}{2}$; second, the series-resonant frequency should be at least one octave above the highest operating frequency; and third, the tertiary winding must have the same number of turns as the main winding and be very tightly coupled to it.

impedance levels

There seems to be a difference of opinion about the inductive reactance required on a balun winding. The usual number given suggests that the reactance at the lowest operating frequency should be at least ten times the characteristic impedance of the balun. This might be a good number to use if the entire input voltage, E, were across the balun winding. Actually, only one-half the input voltage is across any one winding so that the reactance-to-characteristic-impedance ratio can be reduced to 5. In his article, DeMaw[9] suggests a value of 4; this ratio will give a maximum VSWR of 1.64:1, which is better than most Amateurs achieve anyway.

design procedure

With the preceding background, a design procedure can now be worked out as follows:

1. Number of turns and core size. The first problem is to select a core size and determine the number of turns. These two choices are interrelated. If good high-frequency performance — above 30 MHz — is required, small physical size is important. Where good low-frequency performance is important and/or high power will be used, a large size is indicated. If all three characteristics are desired, obviously compromises must be made.

Based on the experience of Reisert,[7] I started with an Indiana General F-568-1 core of Q-1 material. Next, I wound a test coil on the core; it's not necessary to use actual transmission line for this test winding provided the winding has the same form as that of the final winding. I used no. 16 (1.3-mm) enamel wire with the turns spread out to occupy the same space as the final transmission line (RG-141/U coax in this case).

Next, I measured the coil impedance over the frequency range of interest, 3.5-30 MHz. The reactance should be at least four to five times the characteristic impedance over the entire frequency range desired.

It's a good idea to extend the impedance measurements to at least one octave above and below the wanted frequencies to ensure that the impedance is well behaved. You should find that the reactance peaks up at a given frequency, then falls off, and eventually goes negative. The peak is the parallel-resonant frequency of the winding. The frequency at which the reactance is zero is the series-resonant frequency; this frequency must be at least twice the highest frequency at which the balun will be used.

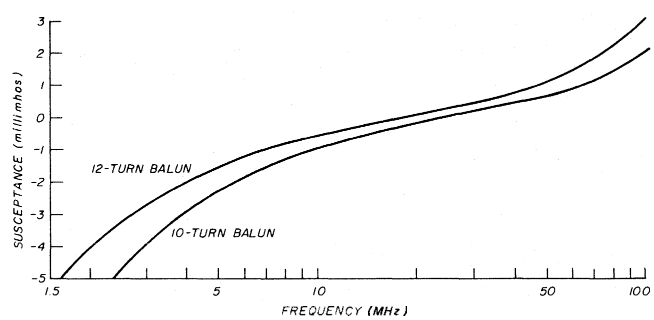

fig. 8. Susceptance of a ten- and twelve-turn coil on a F-568-1 core of Q1 material.

Generally speaking, it will be found that, for a given core and winding form, the ratio of highest-to-lowest usable frequency will be approximately constant. Changing the number of turns will just slide this ratio up or down on the frequency scale.

Measurements made on a 10- and 12-turn balun are shown in **fig. 8.** I've plotted susceptance

$(B = \frac{1}{X})$ since a high reactance or low susceptance is desired; the curve would be off the paper over most of the frequency range for any reasonable reactance scale. If it's required that the reactance be 250 ohms (five times the characteristic impedance), the required susceptance range is between – 4 millisiemens and 4 millisiemens.* This places the low-frequency limit at about 2 MHz for the twelve-turn winding and 3 MHz for the ten-turn.

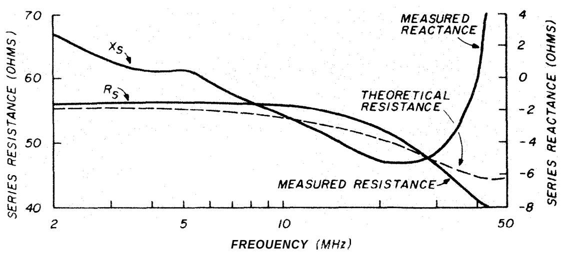

fig. 9. The measured resistive and reactive components of input impedance of a 12-turn balun with a tertiary winding. The load impedance was 55.7 ohms. The theoretical resistive component is shown by the dotted line.

The real (resistive) component of impedance, if measured on a series equivalent basis, should be no more than one-tenth, and preferably one-twentieth, the load impedance. If this component is measured on a parallel basis, it should be at least ten to twenty times the load impedance.

If you can't get enough turns on the core, you'll have to use a larger core or a higher permeability material. I found that twelve turns would cover 2 MHz through 30 MHz. Comparing my frequency range with Reisert's,[7] I found that Reisert required the reactance of the winding to be at least ten times the characteristic impedance of the line, while I allowed the reactance to drop to five times the load impedance.

We can now turn our attention to the transmission line.

2. Transmission line. It doesn't seem to be generally appreciated, but the characteristic impedance of a balun transmission line *must* match the load impedance. This is not because of VSWR considerations but to obtain a flat frequency response at the high-frequency end of the balun. This implies two requirements which are seldom true in the Amateur case; First, the antenna impedance must be accurately known, whereas it is usually estimated. Second, even if you do know your antenna impedance, how do you design a transmission line suitable for balun use with that impedance? Since the design of balun transmission lines would take an article itself, I'll not consider it further.

*Millisiemens are the SI units of conductance and susceptance replacing the millimho; the abbreviation is mS.

To see the effect of a mismatch, let's jump ahead and look at the measured input impedance of the finished balun, which is shown in fig. **9.** Here, the 50-ohm coax was terminated in a 55.7-ohm balanced load. By Amateur standards, this is a very close match, yet the input impedance sags by more than 10 per cent at 30 MHz. Why?

To find out, let's go back to the basics. The length of coax used to wind the balun measured a quarter-wavelength long at 45.33 MHz. At this frequency, then, the input impedance for a 55.7-ohm termination should be 44.88 ohms caused by the quarter-wave impedance inversion effect. For lower frequencies, the theoretical input impedance, as calculated from the transmission-line equation, is shown by the dotted line; in **fig. 9** note that the measured and theoretical are within 1 or 2 ohms to beyond 30 MHz. I could therefore have dramatically flattened the input impedance curve by securing a better impedance match at the output end.

If a balun designer were to look at the drooping input impedance curve, without realizing its true cause, he would think he had a bad design. On the contrary, there's nothing wrong with his balun that a better match wouldn't cure.

A mismatch has another detrimental effect – namely, it adds reactance. Assuming the mismatched load is a pure resistance and the mismatch is small, which I have, the maximum reactance will be generated when the balun is about one-eighth wavelength long, or at about 22.6 MHz in this case. Notice the dip in reactance around that frequency.

This also points out the fact that, where a mismatch may occur, the balun winding should be as electrically short as possible. If the balun described were to be used on a triband beam for, say, 10, 15, and 20 meters, at least one turn, and possibly two, should be removed from the balun.

The rise in reactance at the low-frequency end is caused by the shunting effect of the tertiary winding. If it's assumed that the series equivalent reactance must be less than one-tenth the load impedance, the limiting frequency for this effect is less than 1.5 MHz.

Since this balun will be used in a 50-ohm system, a 50-ohm line will be used: RG-141/U, which is coaxial cable. I'm not particularly happy about using coaxial cable for both electrical and mechanical reasons, but I haven't been able to construct a 50-ohm balanced line that I consider satisfactory; so coax it is.

construction

Before winding the toroid, let's consider the tertiary winding. I said earlier that the tertiary winding should have the same number of turns as the main winding and must be coupled very tightly to it. This latter requirement means that the tertiary must be wound as close as physically possible to the main winding.

To ensure this, I removed the fabric covering from the RG-141/U but left the outer Teflon layer of insulation. I placed a length of no. 18 (1-mm) enamel wire against the Teflon and covered it with heat-shrink tubing. When the tubing had shrunk to size, I had a compact assembly that could be wound as a single ***unit*** on the Indiana General F-568-1 toroid core.*

The proof of a balun is in how well the output voltages are balanced with respect to ground in both phase and amplitude. **Fig. 10** shows these measurements. The amplitude deviation is from equal amplitude (zero dB), and the phase is the actual phase of the low side of the balun with respect to the high side. This value, of course, should be 180 degrees.

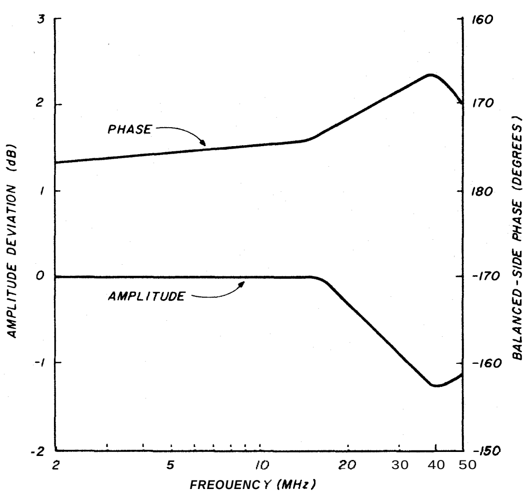

fig. 10. Amplitude and phase measurements at the balanced terminals.

These measurements were made with a Hewlett-Packard model 8405A Vector Voltmeter. Notice that both deviations peak at about 40 MHz. This could be a manifestation of the impedance mismatch and the quarter-wavelength frequency. If so, it points out again the importance of an accurate impedance match between balun and load impedances. Even considering the mismatch effect, the balance isn't too bad. It's within 1 dB and 10 degrees to 30 MHz.

conclusion

I've attempted to explain how a balun works and to clarify some of the problems you must be careful about when designing and building a balun. I've emphasized the importance of 1) tight coupling between the main and tertiary windings of a balun and 2) good impedance matching between the balun and load. I feel strongly about this to the extent that I believe balun manufacturers should include impedance information in their specifications.

I've also given one method of building a balun to minimize the adverse effects of these problems. I don't pretend that the balun design I've presented is the only way around the physical limitations of a balun; it's simply my best so far.

I hope my discussion of balun problems inspires others to experiment and to produce a better balun. With ingenuity, testing and a bit of luck, the balun of the future can indeed be better than ever.

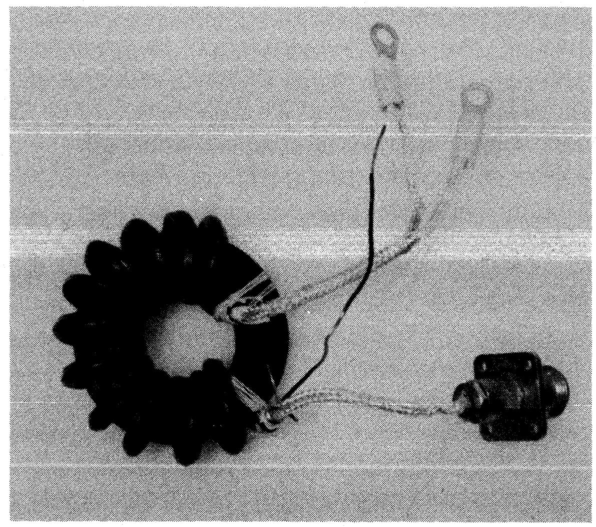

fig. 11. The finished balun.

*A complete kit of parts for the high-performance balun is available from Radiokit, Box 429, Hollis, New Hampshire 03049.

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ham radio

More **Details?** CHECK – OFF Page **94**

third-generation Touch-Tone decoder

Another approach to the *Touch-Tone* decoder for your repeater or remote-base station

If you own a repeater or a remote-base station, you've probably been faced with the problem of coming up with a good *Touch-Tone** decoder. Now you can throw away your toroids and NE567s. Also, you can throw away the falsing problems that go with these devices and still not throw away half a month's salary in the process.

I'm talking about the Mostek MK5102N-5 decoder chip.[1] This device accepts the audible *Touch-Tones* and converts them to either row-column information or BCD four-bit binary information. The chip also provides a valid strobe for logic interfaces. The only external parts required are a 3.579545-MHz TV colorburst crystal and a band separation filter-limiter circuit.

To prevent falsing, the decoder requires 40 milliseconds of valid tone before the strobe and outputs generate output information. When used in the circuit described here, an input level variation above 20 dB presents no falsing problems. Differences in level between the high and low tones (twist) of over 6 dB don't affect performance.

"Touch-Tone **is the registered trademark of the American Telephone and Telegraph Company.**

features

The circuit provides the following features for use in a repeater or remote-base application:

1) Sixteen outputs for use in control circuits

2) Requires five tones of proper frequency and order within five seconds of the first digit before a control output is generated

3) Provides *, #, 1, and 0 for use in autopatch applications (optional)

4) All ten digits are available for use in a regeneration autopatch system (optional)

5) Strobe and BCD information are available for use in external circuits

6) Control outputs (command outputs) and phonepatch outputs can be either high-true or low-true

7) Requires a single +12 Vdc supply

8) Cost is approximately $100 for all features

Touch-Tone decoding has been covered in numerous articles. Therefore, I won't rehash all the basic information. Reference 1 gives a complete description on how the MK-5102-N5 works. For our purposes let's just say that we put high group and low group in and get BCD information out.

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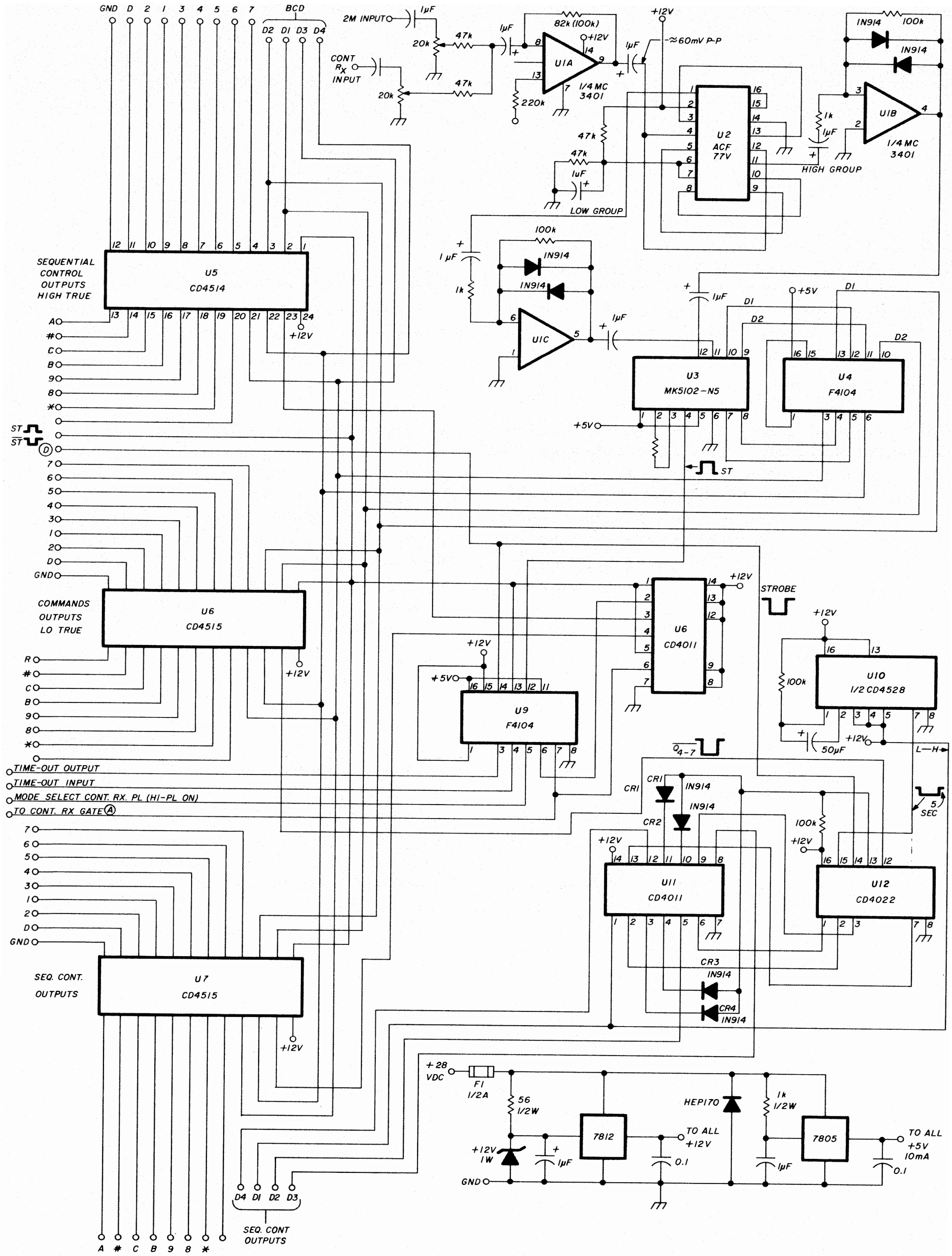

fig. 1. Improved *Touch-Tone* decoder schematic using the Mostek MK5102N-5 decoder chip. A kit is available for the complete decoder or optional circuits (see text).

circuit description

The *Touch-Tone* decoder schematic is shown in **fig. 1.** Let's start at the audio input. IC-1A is one-fourth of a MC3401 op-amp used as a buffer amplifier. The MC3401 is a special op-amp constructed to run on single-ended power supplies. You'll note that provisions are made for two inputs on the decoder (RX audio and control RX audio). These inputs should be low level (approximately 20-30 mV p-p). If your receiver audio level is much higher than this, you may have to change the input-pot value or the value of the feedback resistor in the op-amp (betweenpins 8 and 9).

The gain and input pot should be adjusted for a level of approximately 100 mV p-p (22 mV rms) when a zone corresponding to the digit **5** is being received (measured at IC-2 pin 4). Take care in setting up this level. If excessive levels, such as 1 volt p-p, are shown at this point, the decoder won't function properly. If both inputs are used, provisions should be made for gating the two signals so that only one signal is present at a time. This can be done by using a COR or PL logic level from the control receiver to gate a CD4066 analog switch.

When a signal (or PL) is received on the control receiver, the repeater input is gated off and the control input is turned on. When in the repeater-input mode (no control signal), the opposite occurs. If only one input is needed, delete the additional pot, resistor, and coupling capacitor.

separation filter-limiter

I mentioned earlier that the MK5102 needs a band-separation filter and limiter circuit. The ACF7711 (IC-2) performs band separation. This could be done by cascaded bandpass amplifiers using op-amps as in reference 1. However, the ACF7711 has the advantages of size and temperature stability.

To obtain the stability of the ACF7711 in a discrete op-amp circuit, high-precision capacitors and resistors are required, which aren't cheap. The two 47k resistors and the 1-μF capacitor on IC-2 pins 6 and 7 allow the filter to function on a single-ended supply. Don't omit the 1-μF capacitor at this point.

Speaking of capacitors, let's make a point at this time. All the 1-μF capacitors should be of the low-leakage type. The two outputs (high and low group) of the separation filter feed into limiters (IC-1B and IC-1C). These amplifiers feed square-wave signals to the decoder. (For a more detailed description of the band separation filter and limiter, refer to the data sheetsfor the ACF-7711 and the Mostek MK-5102-N5, which are available from references 2 and 3 respectively.)

Incidentally, a pin-for-pin equivalent of the ACF-7711 is made by Data Signal Corporation of Watertown, Massachusetts. However, I was informed by their sales manager, Mr. Clarence L. Walker, Jr. (after two letters and three phone calls to reach him) that: "We have plenty of business without dealing with a bunch of cheap hams who always want something for nothing." So be it. I hope you bear this in mind if you plan to do business with this company.

decoder-translator

The limited signals are fed into the high- and low-group inputs of Mostek decoder IC-3. The output format pin (pin 6) is grounded so that the outputs are in the form of BCD-coded information. Note that the MK-5102-N5 is a 5-Vdc CMOS device. However, the control logic used in our repeater is high level CMOS (12 Vdc). This was done to obtain a higher noise immunity level on the circuits. This function is performed by IC-4 and IC-9.

IC-4 translates BCD information. IC-9 translates the strobe to high level CMOS. The F4104 IC provides both noninverted and inverted outputs. In the strobe case, both outputs are used. The BCD information from IC-4 is fed to IC-5, IC-6, and IC-7. The BCD information is also fed into the board edge connector for use in other auxiliary equipment (such as LED number display, auto-dialer, *Touch-Tone* regenerator, or additional control circuits).

The CD4514 and CD4515 (IC-5, 6 and 7) are BCD-to-16 output decoders. The CD4514 and CD4515 are identical except for the output states. The CD4514 has true outputs high, while the CD4515 outputs are low. Other than this, the two are pin-for-pin identical. What this means is that IC-6 could be replaced with a CD4514 if you needed command outputs that were high instead of low. The same is true for IC-7. The device used in IC-5 must be a CD4514 (unlessyou don't want things to work).

mode-select logic

After running through the translator, the strobe is fed to IC-8 pins 1 and 5. Pins 1 and 5 are inputs to two sections of the NAND gates. The other inputs (pins 2 and 6) of these two NAND gates go to IC-9 pins 6 and 7. These two outputs of IC-9 are the non-inverted and inverted outputs of the input on pin 5. IC-9 pin 5 is what I call the "mode-select input" for the decoders.

The state of the mode select (high or low) determines whether the sequential control outputs or the phone patch outputs are enabled. To explain this, let's go through the logic when the mode select input is high. This condition corresponds to the control mode when the sequential control decoder is enabled.

A high on IC-9 pin 5 produces a high on pin 6. This action causes a high on IC-8 pin 2. When a valid Touch-Tone digit is decoded, the strobe is high. This action places a high on IC-8 pin 1.

When both inputs (pins 1 and 2) of IC-8 are high, the NAND output is low (pin 3). This output is connected to IC-5 pin 23 (the enable pin for the CD4514). When this pin is low, the inputs are enabled. When high, the outputs all go to a low state. It's necessary to return all the outputs to a low state for the sequential decoder to function correctly. During this time, the phone-patch outputs are disabled because IC-7 pin 23 is held high. A high on IC-9 pin 5 places a low (inverted output) on IC-8 pin 6. Consequently, even though pin 5 is strobed high, the output remains high. This high state keeps IC-7 disabled.

When the mode-select input is low a similar process occurs, except that IC-7 is enabled and IC-5 is disabled. IC-6 outputs are disabled by the sequential decoder. (This will be explained later.) The mode select input can be controlled by a COR or PL output on the control receiver. If your application doesn't require the phone patch output, the mode select pin should be wired to +12 volts. The phone-patch decoder (IC-7) can be deleted from the board. If you want the phone patch outputs but not the command outputs, ground the mode-select input. If this is done ICs 5, 6, 10, 11, and 12 can be omitted.

sequential decoder

The heart of the sequential decoder is a CD4022 (IC-12) Johnson Counter, the electronic equivalent of a stepping relay. For the counter to advance, the two clock inputs must be in the correct states and the master reset must be low. Each time the inputs are in the correct conditions (a code match) the counter advances one position. When the counter advances for the fourth time, a carry output is generated. This output is called Q4-7. Q4-7 will be low during the fourth through seventh advances of the counter.

To explain, let's run through a typical code sequence. The decoder is set up so that the first four digits are a common address for command functions. The fifth digit determines the actual command output from IC-6. From the time that the first digit is sent, the next four digits must be sent within approximately five seconds. The reason for this is that when someone starts sending random digits on a pad, he will have a hard time hitting the correct digits within a five-second period.

Say that our address is 4-1-3-7. The outputs from the IC-5 are strapped to the D1 through D4 inputs (see **fig. 1**). One way to do this is to use a wire wrap edge connector on the PC board. The correct code is wire wrapped from outputs to inputs. In our case, the output pins for 4, 1, 3, and 7 would be strapped to D1, D2, D3, and D4 respectively. The four digit inputs go to one of the inputs of each section of the quad NAND gate, IC-11. The remaining input of the NAND gates is connected to IC-12, pins 2, 1, 3, and 7, the Johnson counter. These four pins correspond to 0_0, 0_1, 0_2, and 0_3 of the counters. These are the first four positions in the counter outputs.

The digit inputs, IC-11, diode CR1 through CR4, and the counter outputs form a sequential coincidence circuit. Each time a correct digit is in the proper position of the sequence, the counter will advance one position on the outputs.

In our example the first digit is 4. A "4" places a high on IC-11 pin 1. The counter is in the reset position, so output 0_0 (pin 2) of IC-12 is high. These two highs cause a low on the NAND-gate output (IC-11 pin 3). This signal is coupled through CR1 to the counter clock input. The other counter clock input is connected to the inverted strobe output of translator chip IC-9.

These two simultaneous inputs cause the counter to advance if the master reset pin (IC-12 pin 15) is low. The master reset is connected to the IC-10 $\overline{Q}$, a CD4528 monostable multivibrator. IC-10 is triggered by the high input from D1 (IC-10 pin 4). The resistor and capacitor values on IC-10 pins 1 and 2 multivibrator determine the time constant (the length of time that the MR pin [pin 151 on the Johnson Counter stays low).

The values shown in **fig. 1** give you about five seconds to enter the remaining three address digits (D2-4) and the command digit. When the coincidence circuit has received digits 4-1-3-7-9, a low is placed on IC-12 pin 12 (Q4-7 output). This low turns on the enable input of IC-6 (pin 23). With the chip enabled, the BCD information on IC-6 input is converted to one of the sixteen corresponding output digits. These outputs are the command outputs, which drive the logic functions in your remote base or repeater.

The CD4515 produces low true outputs. As previously stated, a CD4515 can be used as IC-6 if you need high outputs to drive your logic. The command outputs will remain latched until the timer resets the master reset on the counter. These outputs will then return to all low for the CD4514 or all high for the CD4515. If you need latched command functions, the outputs could be used to drive a CD4043 or CD4044 quad R-S latch. If you need to drive relays, a CD4049 or CD4050 could be used as a buffer for the relays. If you need to drive outside information (amp or other equipment), the BCD-coded information and strobe are brought out to the edge connector pins.

power supply

If a 12-volt battery is used, the resistor and zener diode should be used. (Omit the 7812 IC. It must have at least 14 Vdc to work). If a higher voltage is used, the 7812 regulator can be used (omit the zener). Select the value of the 560-ohm resistor for your particular input voltage. The value of the 1k resistor on the 7805 regulator should be selected so that approximately 7.5 volts is present at the regulator input. A 5-volt diode could be used. However, I prefer the regulator IC because it filters out a lot of the garbage present on the supply line. The HEP170 diode is for reverse-polarity protection. The fuse can be either on the board or mounted externally. If externally mounted, install a jumper on the PC board where F1 is shown.

a kit is available

To assist those who have trouble finding parts, I've prepared a kit for the decoder. Please send checks or money order to Reliable 2-Way Radio, 513 W. 10th St., Casa Grande, Arizona 85222. Here's a list of the various options and prices:

complete kit — all parts and PC board	$140
assembled and tested	add $165
kit — phone only (no sequential decoding)	$130
complete kit less PC board	$115
PC board only	$ 25

Prices include sockets for all ICs (Molex pins for ACF7711). The price doesn't include edge connector for board. The edge connector is a Masterite, part no. 000206-1159. The connector is available for $6 from the above address.

If you have questions or comments about the circuit, send me a letter with an SASE for a reply. I'll be glad to answer your questions if I can. Good luck on your remote base or repeater.

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ham radio

More Details? CHECK—OFF Page 94

how to modify surplus cavity filters for operation on 144 MHz

Simple conversion of surplus 417/GRC filters for top performance on 2 meters

While browsing through an electronics surplus catalog recently I came across a rather obscure item that immediately caught my attention: a small photograph of what appeared to be a dual-section resonant cavity assembly described as a "Bandpass Filter for 417/GRC Receivers." The published operating frequencies are listed in **table 1.** Since I had been playing around with homebuilt cavities for some time and was putting one to good use at my 2-meter base station, I thought it would be worthwhile to look into the surplus filters.

I ordered the F-194/U bandpass filter that covers the 2-meter Amateur band* and was very pleased to find that it was indeed a dual-section tunable cavity resonator, beautifully built both electrically and mechanically, probably at considerable government expense. The unit was neatly calibrated, rugged, and gold plated with low-loss Teflon insulation.

I ordered several more of the same model for experimental use but was told that that particular unit was sold out. It then occurred to me that perhaps the lower frequency units could be converted for use on the 144 MHz 2-meter band. I placed an order for the only low-frequency models that were available: F-239/U (58.5-67 MHz), F-192/U (100-121 MHz) and F-193/U (121-142 MHz). I was soon pleased to find that I was able to convert all three models for operation in the 2-meter Amateur band. The purpose of this article is to make more Amateurs aware of this unique surplus item and to outline the modification procedure for 2-meter operation.

bandpass filters

Cavity type bandpass filters have been around a long time and have been referred to by many names, such as resonant re-entrant cavity, coaxial bandpass filter, coaxial tank filter, coaxial TVI filter, tuned cavity filter, stripline filter, trough line filter, etc. Such cavities have been built in many different sizes and shapes; filter construction projects in the Amateur magazines have been based on common household items from beer cans to coffee tins, metal chassis, rectangular project boxes, and paint cans. The theory of operation of a cavity resonator can be described briefly as such: A cavity is an enclosure or partial enclosure of any size and shape having conducting walls or surfaces that can support oscillating electromagnetic fields within it and possesses certain

*$12.95 from Fair Radio Sales, Post Office Box 1105, Lima, Ohio 45802.

By **William Tucker, W4FXE, 1965** South Ocean Drive, 15-G, Hallandaie, Florida 33009

resonant frequencies when excited by electrical oscillations.

Most Amateurs know that a quarter-wavelength of coaxial cable, shorted at one end, is equivalent to a parallel resonant circuit; the resonant cavity is similar and can be seen as a wide-diameter quarter-wavelength coaxial line shorted at one end, using air as the dielectric. These filters have been used for many years as the very high Q tank circuits in vhf and uhf transmitters and as the tuned rf circuits of receivers.

The radio-frequency current is maximum at the shorted end of the coaxial line or cavity, so a great deal of care must be exercised to ensure a good low-resistance rf contact. Because of skin-effect, copper, silver, or even gold is used to provide high conductivity.

The open end of the coaxial cavity exhibits a very high impedance and, therefore, high voltage. If construction requires insulation at the open end, care

table 1. Resonant frequencies of the surplus tunable bandpass filters for 417/GRC receivers.

	MHz		MHz
F-238/U	50.0-58.5	F-196/U	184-205
F-239/U	58.0-67.0	F-197/U	205-226
F-240/U	67.0-76.0	F-199/U	224-254
F-241/U	75.0-84.0	F-200/U	254-284
F-242/U	84.0-92.5	F-201/U	284-314
F-192/U	100 - 121	F-202/U	314-344
F-193/U	121 - 142	F-203/U	344-374
F-194/U	142 - 163	F-204/U	374-404
F-195/U	163 - 184	F-236/U	550-600

must be used to select very low-loss materials so the very high Q will not be degraded.

Optimum performance can be expected from a coaxial resonant cavity when the electrical length of the cavity and its inner conductor is a full quarter-wavelength long; the frequency of the cavity can be changed by varying the length of the inner conductor. In those cases where a full quarter-wavelength cavity is impractical, shorter lengths can be used with capacitance loading. A coaxial line or cavity shorter than a quarter-wavelength is electrically equivalent to an inductive reactance and requires the addition of capacitive reactance to equalize and achieve resonance; this is accomplished by adding a low-loss capacitor across the open end of the cavity. A variable capacitor provides a convenient method of tuning the cavity to the desired frequency. Generally speaking, the least amount of added capacitance results in highest Q and maximum efficiency (fig. 1).

surplus filters

The physical size of bandpass filter assemblies

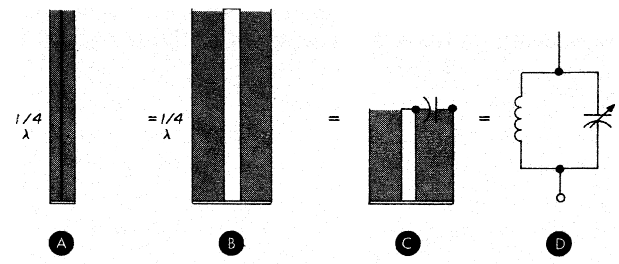

fig. 1. Quarter-wavelength coaxial cable, shorted at one end (A), is equivalent to shortened, capacitance-loaded cavity (C). Performance is similar to lumped L-C tuned circuit (D).

used in the 417/GRC receivers is the same, although the length of the cavities vary according to frequency coverage. The maximum length of the actual cavities in these assemblies is 9 118 inches (23.2 cm). That length is fine where it approaches a quarter-wavelength at the higher frequency ranges, but is far short of optimum at lower frequencies. Though the full 9 1/8 inches (23.2 cm) of cavity space is available, the designers did not take full advantage of the space. Shorter lengths were used on all cavities starting with the F-193/U (121-142 MHz). The F-193/U cavity, for example, is 7 314 inches (19.7 cm) long, and the F-194/U (142-163 MHz) is only 6 314 inches (17.2 cm) in length. Both the F-239/U and the F-192/U use the full 9 118 inches (23.2 cm). For conversion to 144-148 MHz, it is best to select one of the lower frequency units to take full advantage of the additional length. (Note that a full quarter-wavelength on two meters is approximately 20 inches [51 cm]). The reduction in length below the optimum quarter-wavelength lowers the Q of the cavity, but it is still higher than the much lower Q afforded by conventional lumped L-C circuits at these frequencies.

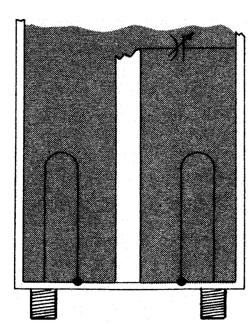

fig. 2. Placement of loops for input/output coupling to the cavity. For light loading (highest Q), the loops must be small and spaced well away from the center conductor.

coupling

Rf energy is usually coupled into and out of this type of resonant cavity with pickup loops placed diametrically opposite each other in the electromagnetic field that exists in the shorted high-current end of the cavity (fig. 2). The loops are similar and allow the cavity to be used bilaterally. The size of the pickup loop, its position, and its proximity to the center conductor are the factors that determine the degree of coupling to the cavity. Large loops and close proximity to the center conductor provide close coupling; small loops spaced away from the center conductor result in loose coupling. Variable coupling in large commercial cavities is provided by rotatable pickup loops.

Close coupling reduces the selectivity and lowers insertion loss; conversely, loose coupling increases both selectivity and insertion loss. Two or more cavities can be cascaded for wider bandpass with steeper seiectivity skirts, or to provide increased seiectivity,

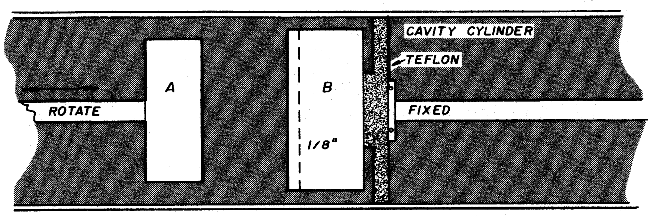

fig. 3. Original configuration of the F-193/U filter which was designed to cover 121-142 MHz; carefully remove 1/8 inch (3 mm) of material from part B to convert to two meters.

depending upon the degree of input/output coupling used in each cavity.

filter applications

Bandpass filters of the type discussed here are most often used to minimize or eliminate intermod and desense interference which originates from sources outside the desired band. On 2 meters such interference may originate in the 150-170 MHz commercial band; considerable interference also originates in the local fm and television broadcast bands and in the 120-136 MHz aircraft band. In most cases the cavity bandpass filter is placed in series with the transmission line to the antenna and attenuates all signals which fall outside its sharp passband.

As an example, I am located not far from two powerful broadcast stations, an fm station on 91.3 MHz and a Channel 2 television transmitter (video on 55.25 MHz). The two signals get into the front end of my receiver through the antenna, mix, and produce a broad, garbled fm signal centered on 146.55 MHz. Insertion of a bandpass cavity in the transmission line completely removes the offending signals and allows simplex communications on 146.55 MHz with the weakest signals.

Most Amateurs use transceivers on 2 meters so the cavity is in the transmission line during transmission as well as reception. This is a bonus because the cavity provides many benefits when used in the

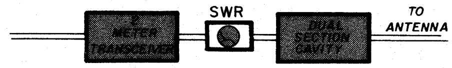

fig. 4. Setup for adjusting the converted dual-section cavity of 144-MHz; tune for minimum VSWR at the desired operating frequency.

transmitted output. Spurious out-of-band emissions, including harmonics, are greatly attenuated, thus minimizing possible illegal interference with other services; this also reduces the possibility of TVI.

surplus filter conversion

Conversion of the 417/GRC bandpass fiiters does not require the addition of any parts. The F-193/U unit, which was originally manufactured to cover 121-142 MHz, requires only a simple modification to tune it up on the Amateur 2-meter band. Remove the six hex nuts from the rear of the assembly and carefully slide out the stationary portion of the cavities and its housing. Referring to **fig.** 3, part **B** is the fixed portion of the variable concentric capacitor which provides a fixed lumped capacitance because of its proximity to the cavity wall; variable capacitance is introduced by the movable capacitor section, part **A, fig.** 3, which is controlled by the front panel knob. The conversion is made by simply removing no more than 1/8 inch (3 mm) from part B (shown by the dotted line) with a hacksaw, grinder, or hand file; remove all burrs, reassemble the cavity, and the conversion is complete except for testing.

A dipmeter may be used effectively to check the

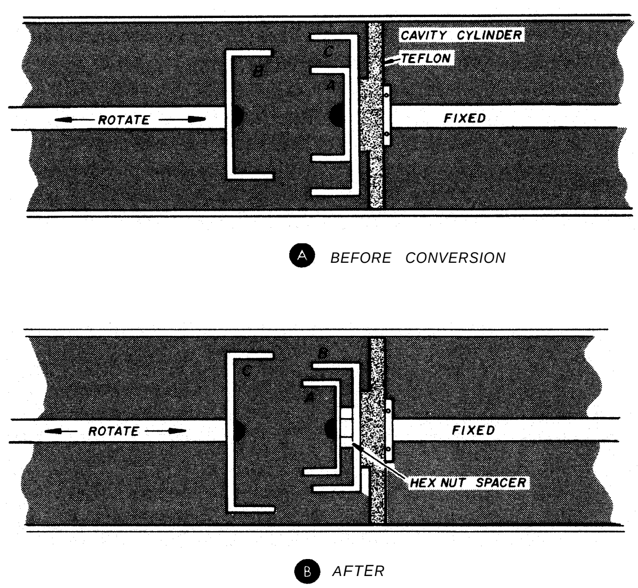

fig. 5. Conversion of the F-192/U filter (100-121 MHz) to two meters. Original arrangement before conversion is shown at (A); (B) shows the modification with a hex nut which moves the resonant frequency to the 144-MHz band.

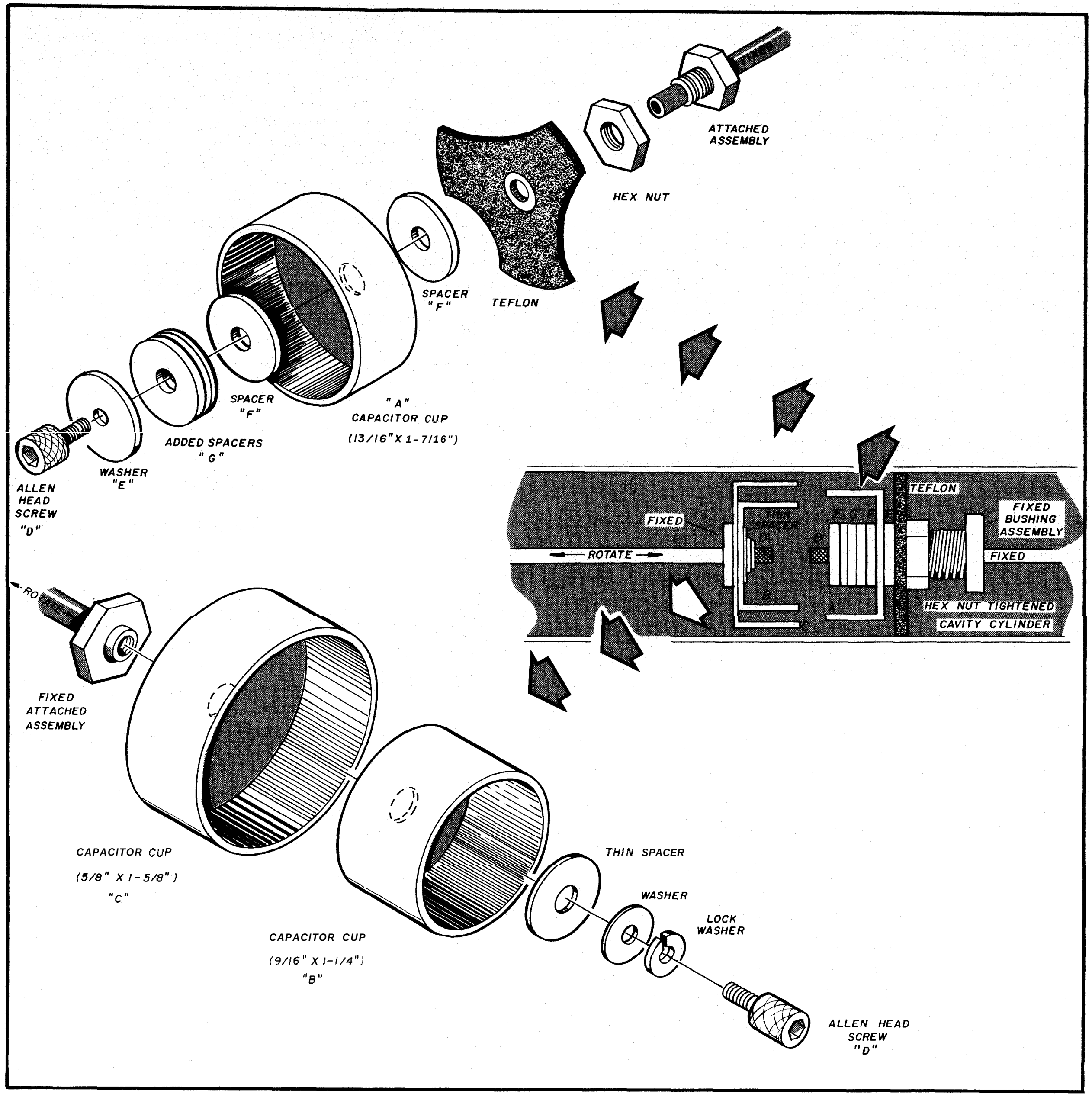

fig. 6. Modification of the F-239/U filter (58.5-67.0 MHz) for operation on the 144-MHz Amateur band [see text).

frequency range. Use a very small pickup loop connected to either port of the cavity (about 1/2 inch [12 mml or less) and couple it lightly to the dipmeter. With the front dial turned fully clockwise, a sharp dip should occur around 130 MHz; fully counterclockwise, the dip should occur at about 155 MHz. Note that the dips are very sharp in this lightly loaded condition and can be easily missed. Check both input and output in the same way. The dial can be recalibrated in any manner you wish.

If a dipmeter is not available, a receiver can be used to find out if you are in the 2-meter range. Tune in a rather weak signal and as you tune the cavity to resonance at that frequency, the signal should pick up nicely.

After you have determined that the conversion is successful, you can install the unit permanently in the transmission line as shown in **fig. 4.** Adjust the cavities for lowest VSWR. When transmit frequency is moved appreciably, about 500 kHz, you may have to retune the cavity for lowest VSWR. I suggest you keep the VSWR meter and the cavities as close as possible to the transceiver.

The F-192/U filter was originally designed to cover

100-121 MHz. Follow the same procedure as previously when taking the assembly apart. The conversion simply involves rearranging the configuration as shown in **fig. 5,** before **A,** and after B. The hex nut used as a spacer is borrowed from the rear of the assembly. Make sure the Teflon insulator is reassembled with the extrusion part as shown because this provides a lowest leakage path at the high impedance point. When testing this unit, note that the capacitor will short out at the full clockwise position. The frequency range should be about 135-175 MHz.

The F-239/U filter was designed to cover 58.5-67 MHz and requires more extensive modification. Before taking the assembly apart, remove the three screws visible on the outside surface of the cavity. Carefully slide out the fixed portion of the assembly and remove all the fixed and variable capacitor sections from the center conductor and the rotor section. Select the parts needed as shown in **fig. 6** and reassemble. You will have quite a few pieces of hardware left over for your junk box. The only additional parts you will need are two or three thin washers that can be used as spacers (part **G, fig. 6);** as shown, the modified cavities cover 120-170 MHz. If more selectivity is desired, the pickup loops can be shortened by any convenient method.

In addition to using the dual cavity assemblies in the conventional manner as bandpass filters, they can also be used as series-resonant traps by shunting them across the line as shown in **fig. 7;** in this application the filter is used as a wave trap. If you wish, you may separate the two cavity sections by disconnecting the connecting link between them; you will then have two separate series-resonant traps as shown in **fig. 8.** You can also separate the two cavities for use as two individual bandpass filters by removing the connecting link and installing two connector sockets. Because of the limited space, I suggest you use BNC connectors.

summary

When you become familiar with the action of these fine cavity assemblies and have digested the many good articles that have appeared in the various handbooks and magazines, you will find many other uses for them around your hamshack. You could even build a duplexer for a repeater using these units as the foundation.

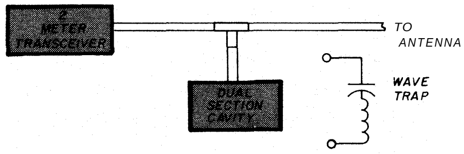

fig. 7. Use of the dual section cavity as a tuned wave trap.

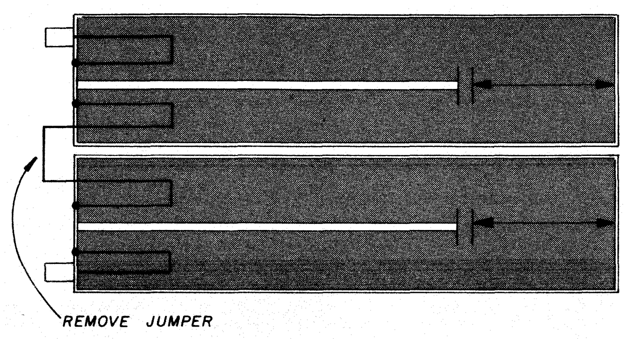

fig. 8. Separation of the two cavities for use as two tuned, single-section wave traps.

I have inciuded a iist of magazine articles from the past ten years for the convenience of those readers who wish to delve further into the subject of cavity filters.

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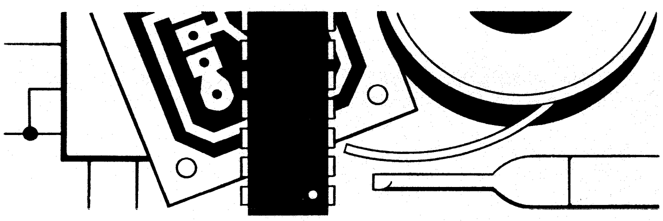
ham radio

the weekender



Bench Power Supply

A good quality regulated bench supply is probably one of the most useful devices you can have in the workshop or ham shack. Whether you're just getting started in electronics and looking for a first project or are an experienced builder or experimenter in need of an additional bench supply, here's a supply that will fill your needs. This versatile power supply can be built in just a few hours, uses readily available components, and the finished product is a neat, professional looking package. You'll be proud to say, "I built it."

description

Quick and easy construction is a feature of this power supply, which is accomplished with a single PC board. Such construction keeps mechanical and sheet-metal work to a minimum. Most components are on the PC board. Full voltage and current metering are provided. The output is isolated from the case to allow its use as a positive or negative supply.

Output current is 3 amperes over the range of 1.5-15 volts, with short circuit and overload protection provided by the IC voltage regulator U1. Output ripple is at a low level. The power supply is physically small. The basic design isn't limited to bench-type power supply applications, because the PC board lends itself to installation in fixed equipment. No changes in the PC board pattern are required for use as a fixed supply.

the circuit

The bench supply schematic is shown in fig. **1.** Component count is very low. The basic design consists of a full-wave power supply followed by a three-terminal IC regulator. Power transformer T1 is an 18-volt 4-ampere unit. It isolates the power supply from the ac line while furnishing 18 volts to the bridge rectifier. T1 primary is fused, and the transformer frame is held at ground potentiai through the three-wire ac line cord. The on-off switch is connected to the output-level control for convenience and safety. The bridge rectifier is a single-unit device rated at 4 amperes 100 PIV. The pulsating dc output from the rectifier is filtered by a pair of 1500-μF, 35-Vdc radial-lead electrolytic capacitors.

The filtered dc is applied to the input of the LM-350K IC voltage regulator. The output level is controlled by a divider network formed by R1 and R2. Load current is displayed by M1, a 3-ampere dc meter. M2, a 15-volt dc meter, displays output level. The power-supply output is obtained from five-way binding posts. The output jacks, meters, and level control and switch are the only components not mounted on the PC board.

construction

PC board construction was chosen for the power supply for simplicity, ease of construction, and repeatability. Heavy glass-reinforced board should be used because the power transformer is fairly heavy and we want to avoid cracked and broken land patterns as the board ages and is subjected to stress. The etching pattern is shown in fig. **2.**

After etching and drilling, the on-board components can be mounted and soldered as in fig. 3. Pay particular attention to the polarity of the capacitors and the bridge rectifier. Coat regulator U1 with heat-sinking compound on the bottom surface where it mates with the heat sink. Fasten these items to the PC board as a unit with M3.5 (6-32) hardware. The regulator is not insulated from the heat sink.

This completes component mounting and soldering. Check the PC board against fig. 3 to ascertain that all polarized components are properly oriented.

The ac cord and front-panel interconnect wiring, also shown in fig. 3, can be installed next. The interconnecting wiring should be no. 16 (1.6 mm) stranded to minimize voltage drop and provide flexibility. Solder these wires to the PC board and leave them about 25 cm (10 inches) long for connection to the

By **Ken Powell, WB6AFT, 6949** Lenwood Way, San Jose, California 95120

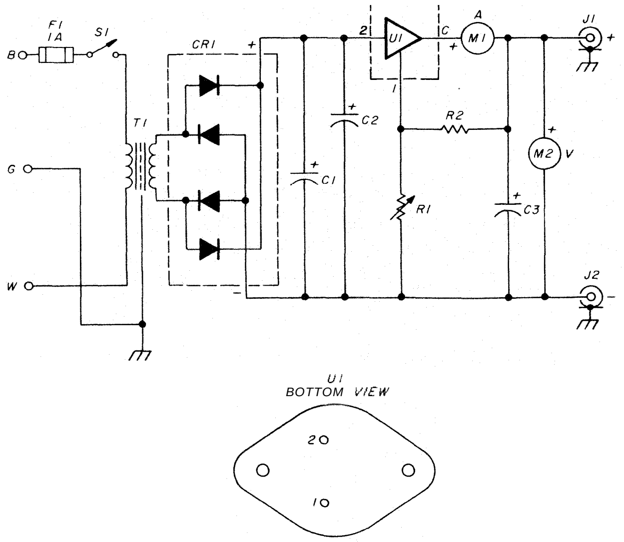


fig. 1. Schematic of the bench power supply. Basic design consists of a full-wave power supply followed by a three-terminal IC regulator.

Parts list for fig. 1:

C1, C2	1500-μF, 35-Vdc	BA 18A1506-5
C3	2.2-μF, 35-Vdc	RS 272-1407
CR1	4 amp, 100 PIV	RS 276-1171
F1	1-amp fuse	RS 270-1283
FH1	fuse holder	RS 270-739
HS1	heatsink	BA 12A2229-9
IC1	LM-350K	CSC LM-350K
J1, J2	binding posts	RS 274-662
M1	3-amp meter	CSC D1-918
M2	15-volt meter	CSC D1-920
R1	5K	RS 271-1714
R2	220,112-W	RS 271-015
T1	18-volt, 4-amp	RS 273-1514
S1	switch, spst	RS 271-1740
CASE	14.7 x 22.5 x 14.4 cm (5-7/8 x 9 x 4-718 inches)	RS 270-281

BA: Burstein-Applebee, 3199 Mercier St., Kansas City, Missouri 64111

CSC: Circuit Specialist Co., Box 3047, Scottsdale, Arizona 85257

RS: Radio Shack, local stores

Note: A kit is available which includes the following parts:
1 etched and drilled PC board
1 regulator, LM-350K (IC$_1$)
1 heatsink (HS$_1$)
2 capacitors, 1500 μf/35 Vdc (C$_1$ and C$_2$)
1 resistor, 220 ohms (R$_2$)
1 capacitor, 2.2 μf/35 Vdc (C$_3$)

The cost of this kit is $21.00 plus $1.50 postage and handling. Ask for kit #PS-25-3 from J. Oswald, 1436 Gerhardt Avenue, San Jose, California 95125.

front panel components. They can be trimmed to length as the front panel is installed.

Drill and punch the front panel using caution to avoid damaging the smooth finish on the panel. If a chassis punch isn't available for making the meter holes, a coping saw will work fine in the soft aluminum. When you're satisfied that all the front panel components fit well, label the on-off level control with dry transfer labels. Apply a coat of clear acrylic to the panel to protect the lettering. The front panel components can be mounted after the panel is dry.

At this time the PC board should be mounted to the cabinet base. Use standoffs to elevate the board above the base. Cut a small notch into the lower edge of the cabinet back and install a strain-relief or grommet for the ac line cord. Lay the front panel down in front of the PC board and complete the interconnecting wiring as in **fig.** 3. This wiring can be laced or spot-tied for a neat appearance (photo). The cabinet can now be assembled. Your project should look pretty much like the photo of the completed prototype. If a cabinet is used other than that listed in the parts list, you may have to modify the construction procedure a bit, but the end result should be the same.

test and applications

After assembly, the front-panel meters should be adjusted for mechanical zero. There are no other adjustments, so the little supply is now ready for the smoke test. Apply ac power and advance the output level control clockwise. The voltmeter should track nicely from about 1.5 to slightly over 15 volts. Output ripple can be checked on a scope if you have one, but, if not, don't be overly concerned about it. The two prototype units I built were smooth and ripple free.

The little supply is now ready to go to work. The voltage and current range will let you power up quite a variety of equipment and circuits. I've used the

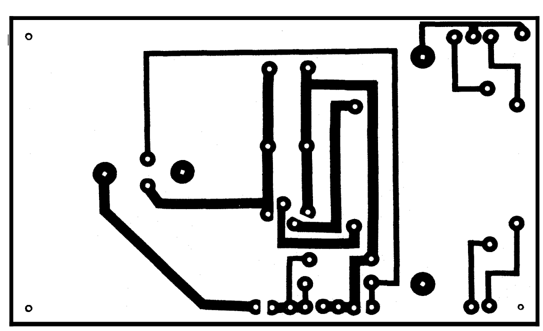


fig. 2. **Full-size** printed-circuit layout; component placement is shown in fig. 3.

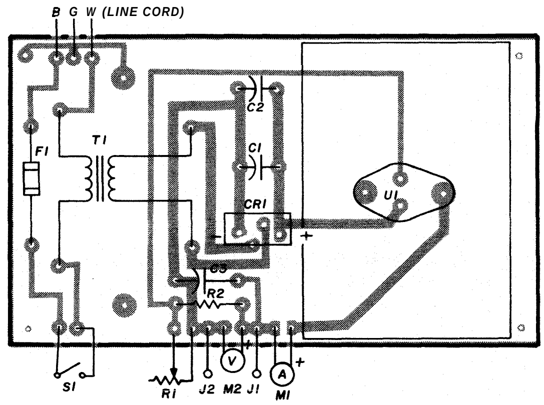


fig. 3. Component side of the circuit board; photographs show placement of the power transformer, heatsink, and other larger components.

supply on TTL projects, op amps, vhf rigs, and a number of auto radio and tape player combinations.

The properties designed into the IC regulator take most of the worry out of the unit. If you apply a shorted piece of gear to the output, or connect a rig that draws more current than the supply can furnish, the regulator will shut down and no damage will occur. If the unit gets too hot, it will shut down. When normal conditions are restored, the supply will bounce back and go to work again. It's a tough little performer and just the ticket for a guy like me, who's apt to make an error now and then.

conclusion

There are many variations that could be made in the supply to suit individual requirements. Components aren't critical, although I don't think I'd go to lower values in any case. Changes such as using different meters with shunts and multipliers would be no problem, and a single meter with shunts, multipliers, and switching could be used. For fixed applications the level control could be replaced with a fixed resistor and the meters eliminated. Layout and mechanical details aren't critical, but the regulator IC should be kept physically close to the filter capacitors.

The cabinet should be well ventilated. If the supply is powered up outside the case, keep in mind that 117 Vac is on the PC board and the interconnecting wiring! The low voltages usually associated with solid-state equipment tend to encourage carelessness.

The supply was a very enjoyable project in that the assembly time was minimal (about three hours) and no sophisticated equipment was required for building, de-bugging, or adjustment. I don't think anyone should have any apprehension about trying this one for a weekend project.

ham radio

More Details? CHECK—OFF Page 94

voltage-controlled resistance for Wien Bridge oscillators

A simple improvement that uses a photocell as a voltage-controlled resistor

Wien bridge oscillators[1,2] require some sort of voltage-controlled resistance in the feedback network to maintain unity loop gain as the frequency is varied over a wide range. In practice, part of the oscillator output voltage is used to control the variable resistance element so that the output voltage remains constant, or nearly so. The voltage-controlled resistance must not respond to the output-waveform instantaneousvalue, but only to its average amplitude. Thus, the voltage-controlled resistance device must have a long time constant compared with the lowest-frequency period.

I've experimented with these circuits over the years, using incandescent lamps, diodes, and thermistors.3 These methods all have shortcomings of one kind or another, and I think the idea described here works better than any of those others.

homebrew device

Shown in **fig. 1** is a simple voltage-controlled resistor that can be homebrewed with readily available parts. Such devices may be available already assembled but perhaps not as available as the two parts needed to build your own. The controlled resistor is a cadmium sulfide (CdS) photocell, obtainable from Radio Shack for 99 cents (catalog number 276-116). Its resistance changes from several megohms in darkness to about 100 ohms in bright light. It's most sensitive to yellow or green light. A green light-emitting diode (LED) causes the CdS resistance to vary inversely to the dc control voltage applied to the LED circuit.

To prevent ambient light from interfering with operation, the LED-CdS assembly should be sealed as shown in **fig.** 2. The cover may be almost any type of scrap material, such as a square of copper-clad circuit board. Drill a hole in the cover for the LED to stick through into the photocell, then epoxy the parts together and spray thoroughly with black paint. Be sure to mark the anode lead of the LED before epoxy is applied.

circuit details

Fig. 3 shows the complete oscillator. U1 is a dual jfet op amp, TL072CP, made by Texas Instruments. It gives excellent performance in this circuit. The oscillator signal at pin 7 of U1A is buffered and detected to provide a dc control voltage for the LED. The

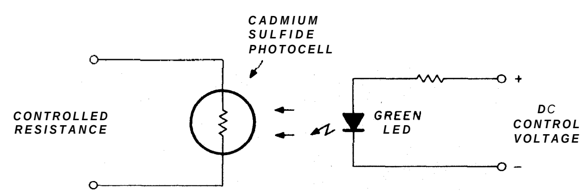


fig. 1. Homebrew voltage-controlled resistor.

1N914 is required because the emitter-base breakdown rating of the 2N2222 is only 5 volts. The 10-ohm resistor in series with the filter capacitor is required for stability. The CdS photocell is connected into the oscillator circuit in the same way a thermistor would be used. Other information on Wien bridge oscillators may be found in the references cited.

operation

The 10-k trimpot is adjusted for maximum undistorted output in the normal manner. If the trimmer capacitors on the tuning capacitor are properly adjusted, output amplitude should be flat within about 0.1 dB between 30 Hz and 200 kHz. If frequencies lower than 30 Hz are requried, the 500-μF filter capacitor must be increased in value to prevent distortion.

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

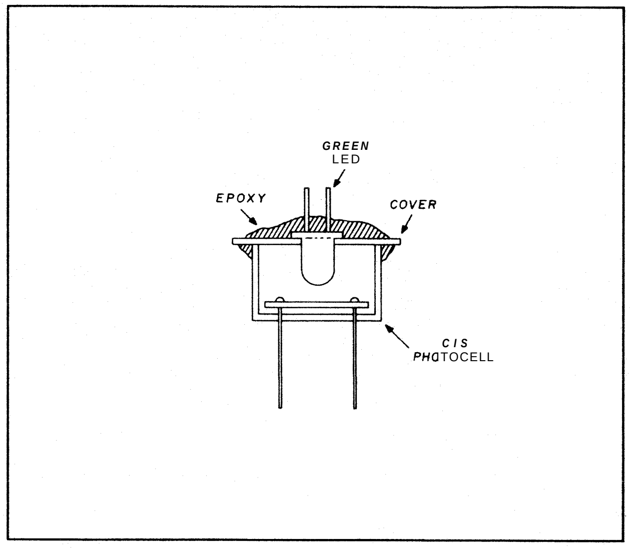
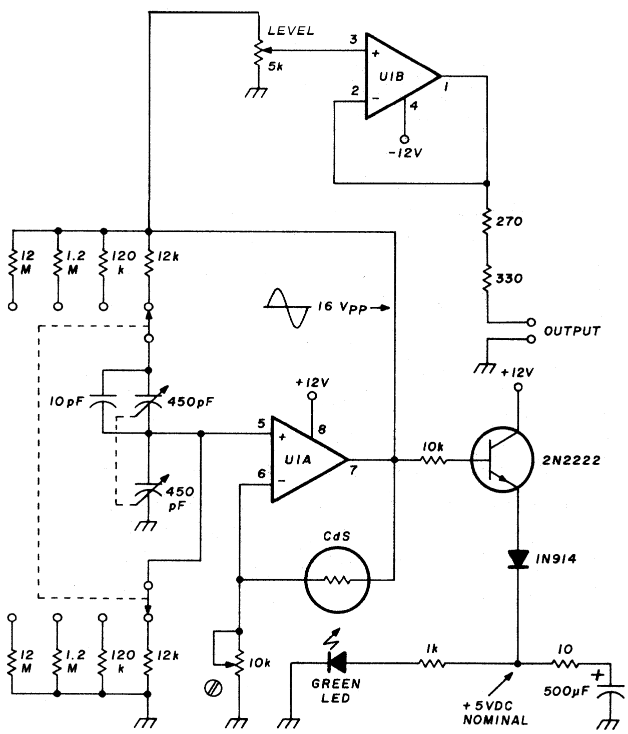


fig. 2. Cutaway view of hornebrew voltage-controlled resistor.



U1 jfet **op amp,** TLO72CP. **For availability contact T.I. Supply Co., 6000** Denton **Drive, Dallas, Texas 75235.**

fig. 3. Wien bridge oscillator with improved gain control circuit.

references

1. H. Olson, W6GXN, "Resistance-Capacitance Oscillators," *ham radio,* July, 1972, pages 18-25.
2 H. Olson, W6GXN, "A Transistor Wien Bridge Oscillator," 73 April, 1967, pages 36-39.
3. C. Hall, WA5SNZ, "New Op Amp Challenges the 741," *ham* radio, January, 1978, pages 76-78.

ham radio

More Details? CHECK — OFF Page 94

a carrier-operated relay for your Heath HW-2036

Simple mods
you can make
to enjoy your
a-mlfm radio
and your 2-meter rig
at the same time

If **you have problems** doing more than two things at once, then this modification is for you. Try it if you own a HW-2036 2-meter rig and have trouble trying to decide between listening to an interesting conversation on the rig and soothing background music from your a-mlfm radio while traveling down the road.

I've installed a COR (carrier operated relay) in the 2036 to switch the a-mlfm radio on and off when the 2036 is receiving a signal. The a-mlfm set goes off when a signal is received on the 2036, which results in only one radio being on at a time — a real blessing for eliminating the hectic atmosphere in an automobile when both radios are running simultaneously and you are in rush-hour traffic.

operation

The circuit is simple and I'm surprised that Heath hasn't made it an available option on the 2036. The basic idea is to turn off the a-mlfm radio during the transmit and receive modes of the 2036.

The benefit of turning off the a-mlfm radio during the transmit is twofold. First, there's no possibility of transmitting your soothing background music over the 2-meter rig, which is not looked upon favorably by the FCC, to say the least. The second benefit is that listening to the a-mlfm radio is automated while going down the road.

circuit

The diodes (fig. 1) are used solely for isolation to keep this circuit from affecting anything going on at the connection points in the 2036. The RC network on the relay gives a delay of 1-2 seconds to keep the a-mlfm radio off during short repeater dropouts or when working simplex.

By **Gary L. Long, WD5HYQ**, Route 5, Box 267, Muskogee, Oklahoma 74401

construction

I located all the parts beside the lever switches on the 0/5-kHz switch side of the HW-2036. A little Super Glue will hold everything in place for wiring. The transistor isn't critical; just be sure it can handle enough source current for the relay.

0/5-kHz – COR bypass. If you live in a signal area with very few repeaters that require the 0/5-kHz switch, you can make it a dual function switch (0/5-kHz – COR bypass). This beats installing another switch in the rig and messing up the looks of the 2036. This addition is useful when the a-m/fm radio is broadcasting something you don't want interrupted while listening to it. All that's necessary is to remove the wire going to SW6 from pin X on the synthesizer board and reconnect it through an isolation diode to the switch as shown in **fig. 1.**

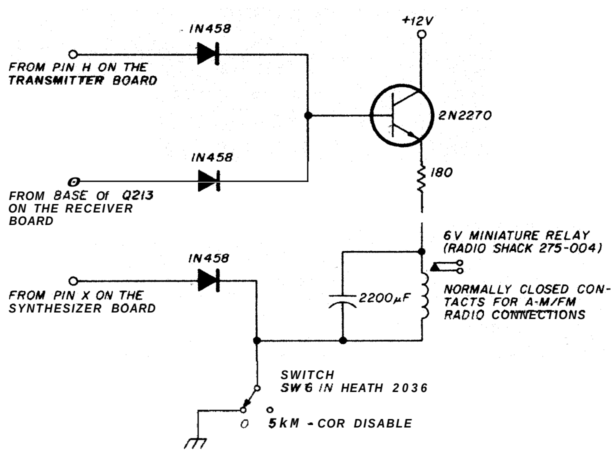


fig. 1. Circuit for turning off your a-m/fm radio during transmit-receive modes of the Heath HW-2036 2-meter transceiver. Circuit features a 015-kHz – COR bypass arrangement.

A-m/fm connections. All that's left to do is to connect a pair of wires to the relay contacts and feed them out through the same slot in the back of the radio where the power cable comes out. The easiest way to connect these wires is to pull the a-m/fm radio fuse and connect these wires right at the fuse block. Don't forget to install an in-line fuse on one of the wires before you turn on the power.

one last item

I had been plagued with a birdie problem, and while I had the rig apart I decided to go hunting for birdies. This critter appeared on 146.85 MHz, 146.52 MHz, and several other places. My solution was to put a mica capacitor across the base-collector junction of Q409 on the synthesizer board. The value appears to be critical. The best response was obtained in my rig by using about 820 pF.

ham radio

plasma-diode experiments

Plasma diode detector using a neon glow lamp — today's answer to the slotted-line standing-wave detector

The use of the plasma in glow lamps and specially designed discharge tubes for microwave detection is not new.[1-4] In most cases, means are provided for focusing the microwave energy on a selected point of the discharge, with the detector output appearing as a variable component of the lamp current.

The plasmas operate at quite high power levels as far as biasing is concerned (20 mA and 250 volts). This type power supply is out of line for the experimenter working with transistors and ICs.

The drawback of the conventional kind of plasma detector is its high noise. Relying on Maxwell's displacement current, research workers have, however, constructed glow-lamp microwave detectors that are free from the annoying lamp-current noise. In certain cases they are even more sensitive than the old standby, the crystal diode. These new detectors operate in the gigahertz region up to 100 GHz (3-mm wavelength).

starvation-current mode

A recent research and development project at Sercolab revealed that worthwhile detector sensitivity can be achieved in a "starvation current" glow-lamp mode, with the demand on supply power reduced to about 0.1 per cent. This kind of operation brings the glow-lamp microwave detector within every experimenter's reach, since the entire power supply consists of a few 9-volt transistor batteries. With capacitively coupled microwave energy, the glow-lamp detector combines an acceptable signal-to-noise ratio with a broad frequency range, a wide dynamic range, and the feature of being almost indestructible – something that can't be claimed for its competitor, the crystal diode.

By Dr. Harry E. Stockman, Sercolab, P.O. Box 78, Arlington, Massachusetts 02174

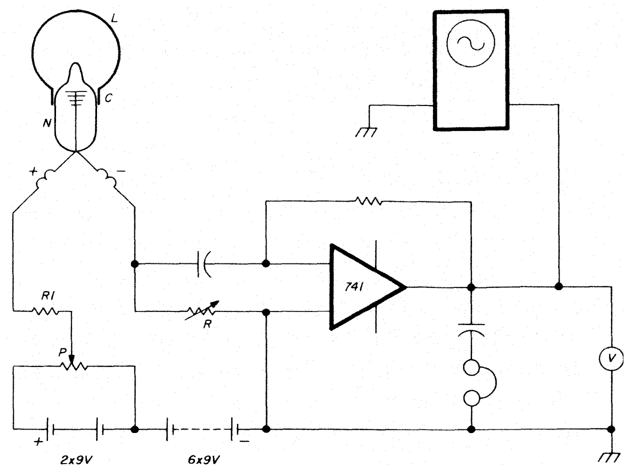


fig. 1. Experimental circuit for evaluating plasma diodes in the "starvation-current" mode. LC is the loop antenna; N is a neon lamp. Other components are discussed in the text.

Its characteristic of being able to take a heavy overload without showing any defect in operation makes the glow lamp a handy microwave detector. Furthermore, it's inexpensive. A brand-new neon lamp costs only 25 cents. (Radio Shack's NE 2 and NE 2H are good examples.)

Today, triode and tetrode glow lamps, such as the TRJ 250, are being used successfully as microwave detectors, providing the desired coupling internally.* Alternatively, an external capacitor plate can be attached to a diode, cemented to the glass envelope. In the detector described below, a diode lamp with protruding terminal wires is used with an external rf-electrode arrangement that forms a loop antenna. A dipole antenna with capacitor plates provides an alternative.

test setup

Fig. 1 shows a simple test rig. In the starvation mode, only the tip of the cathode electrode glows. The capacitor plates, C, of the loop antenna, L, are loosely coupled to the discharge. (We are looking into the glow lamp, N, in the plane of the two electrodes.) Each lamp terminal wire is formed into two solenoid turns around any temporary 3-mm (1/8-inch) core, thus providing an rf choke.

Potentiometer P is a small 25k trimpot. Variable resistor R is 250k. The values are not critical. R1 is a 5k safety resistor. A 741 IC is the detector – output amplifier; but two 741s in cascade are better, meaning that we may use a 1458 IC.

The output indicators are a scope, an electronic voltmeter, and earphones. Eight 9-volt batteries are used, held together with a rubber band. The lamp-supply current amounts to only 0.1 mA, so the life of the major part of the battery is almost the shelf life.

*The TRJ 250 tube and other glow lamps suitable for microwave detection are obtainable from General Instrument Signalite Division, 1933 Heck Avenue, Neptune, New Jersey 07753.

the antenna

The experimenter can readily make up an assortment of loop antennas by clipping metal strips from a coffee can, or better, from a neatly polished, somewhat thicker, copper sheet. The width of the strips may be 6 mm (1/4 inch).

The quickest way to obtain a properly tuned system is to try out different-size loops for a given wavelength. The inductance can be reduced somewhat by flattening the loop. The loop position relative to the glow lamp is shown in fig. 1. This is only one of several possible positions. Actually, the best results were obtained with the loop folded 90 degrees out of the paper (fig. 1), so that its axis is parallel with the direction of the electrode system. The loop is then moved up, down, and lengthwise for optimum coupling to the plasma. For the loop orientation with respect to the transmitter, the rules are about the same as for everyday shortwave work.

test equipment and adjustments

Two low-power laboratory oscillators were available for the tests – one 1.2 GHz (25-cm wavelength) and a 2.4 GHz 112.5-cm wavelength). Each was amplitude modulated with a 1000-Hz tone.

In adjusting the ***receiver,*** the P and R controls are set in a combination that does not promote CR relaxation oscillations, which are common in neon-lamp hookups. The adjustment of the controls is such that maximum detector audio output is provided.

Once the proper control setting has been found, the detector will remain stable. If, after a long time of operation, the lamp goes out, potentiometer P voltage is temporarily increased to make the lamp fire. Then it is decreased to its previous value. Most lamps fire in the 60-70 volt range. The extinguishing voltage is then some 10 volts or less.

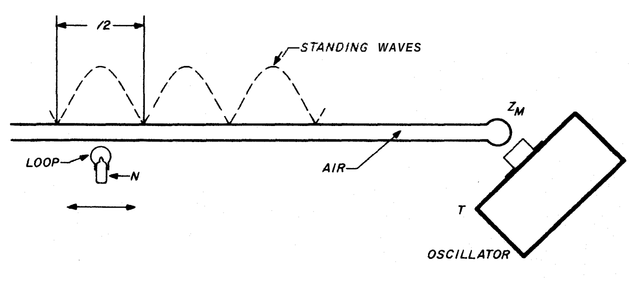


fig. 2. Plasma detector of fig. 1 implemented for plotting standing waves on a transmission line. Loop must be moved along the line at a constant distance from the line for consistent data.

uses

Among practical uses of this detector is the determination of the oscillator wavelength. The glow-lamp detector is placed at the side of the transmitter with just enough coupling for the modulation tone to be heard in the earphones. A meter stick is then placed vertically on the bench, and a reflector is moved up and down along the stick. Readings are taken at every sharp null on the meter or in the earphones. A suitable reflector may be made from heavy aluminum foil cemented to the back of a writing-pad cardboard.

A bigger reflector is better. One with a quarter-meter side is a deluxe article. The average distance between the nulls is one-half the wavelength. Really, doing it this way we're too close to the transmitter, and if the experiment is repeated in the horizontal plane, with a larger distance between transmitter and receiver, better accuracy results.

In another experiment, we may rig up a two-wire line in air with a wire distance of 13 mm (1/2 inch), using one end for coupling and the other end either open or closed. Another arrangement is to put a bit of TV downlead on a wooden table, similarly arranged at the ends.

Fig. 2 shows how the excited end of the line is formed into a loop, coupled through the mutual impedance, Z_M, to the oscillator. With the line removed, the transmitter is tuned for minimum direct pickup by the receiver. Then, with the line in position, Z_M is assigned the compromise value that gives a clear standing-wave pattern with only sufficient rf power for good deflections on the voltmeter.

The detector is then moved along the line and the readings are tabulated, whereupon the standing-wave pattern can be plotted. Loop L must be moved at a constant distance from the line.

The two-wire air line gives sharp nulls, and the wavelength obtained agrees quite well with that measured in the preceding reflector experiment. The TV cable nulls are not equally sharp, and the measured wavelength is smaller than the true wavelength because the insulation dielectric constant is larger than unity.

references

1. G. Burroughs and A. Bronwell, "High-Sensitivity Gas Tube Detector," Tele-Tech, vol. II, August, 1952, pages 62-63.
2. N.S. Kopeika and N.H. Farhat, "Video Detection of Millimeter Waves with Glow-Discharge Tubes," *IEEE* Transactions on Electron Devices, vol. ED-22, No. 8, August, 1975, pages 534-548.
3. N.S. Kopeika et al, "Commercial Glow-Discharge Tubes as Detectors of X-Band Radiation," IEEE Transactions on Microwave Theory and Technology, vol. MTT-23, October, 1975, pages 843-846.
4. N.S. Kopeika, "A New Type of 10-GHz Receiver," 73, September, 1978, pages 222-223.

ham radio

90-degree phase-shift network offers 2:1 bandwidth

Differential 90-degree phase-shift network built with coaxial line sections provides wide operating bandwidth

This brief **note** describes an easy-to-build network capable of providing 90-degree differential phase shift between two outputs. Over a 2:1 bandwidth the VSWR maximum is less than 1.2:1 and the deviation from quadrature phase is ±2 degrees. This network should find application in antenna feed configurations and other instances where extremely broadband differential phase shift is not required.

network description

The phase-shift network is $\lambda/2$ long at band center and is shunted at midpoint with an $\lambda/8$ shorted stub and an $\lambda/8$ open stub. For a 50-ohm system the characteristic impedance of the $\lambda/2$ line and both stubs is 31 ohms. A sketch of the circuit is shown in fig. **1**. For clarity, balanced transmission is shown although actual realization most likely will be coaxial line. The other half of the differential phase shift network is 314 λ length of 50-ohm transmission line. The complete network is shown in fig. **2.**

The network in fig. **1** has a dispersive phase characteristic, which means that the phase shift through the network does not vary linearly with frequency. An equivalent circuit of a dispersive network is shown in fig. 3. The phase through the network is retarded below band center and is advanced above band center. At band center the parallel circuit is resonant and has no effect. In this manner, a constant 90-degree phase differential is maintained between the dispersive network and the 314 λ transmission line for an octave bandwidth. Optimum impedance match for the network is achieved by using 31-ohm impedance line.

Transmission line of 31-ohm impedance is not available unless specially made; however, a 30-ohm line can be realized by paralleling lengths of ***50-*** and 75-ohm line. Fig. **4** suggests one method you may use to build the circuit. **Fig. 5** is a computed plot of VSWR versus frequency for the dispersive network made from 30-ohm transmission line.

conclusions

This note describes a simple 90-degree phase-shift network, which is easy to build from readily available materials and adequate for 2:1 bandwidths. In instances where only 2:1 bandwidths are required in

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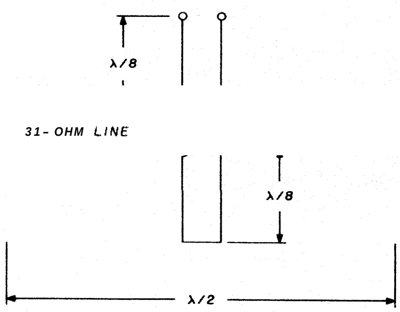
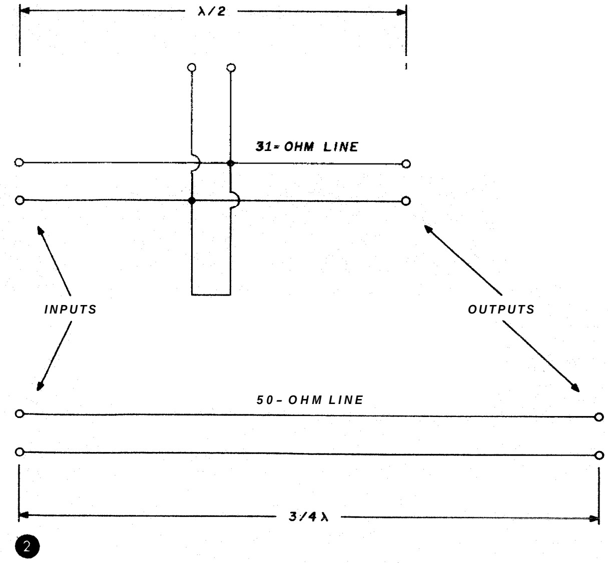
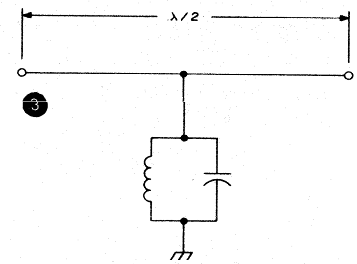
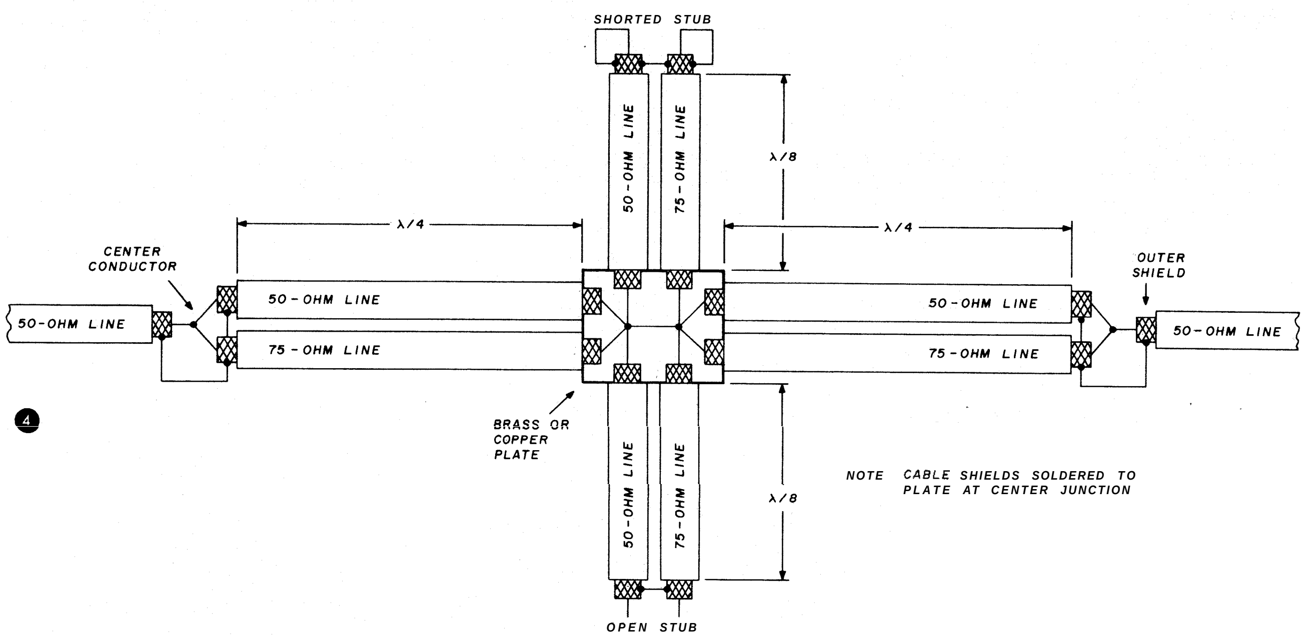








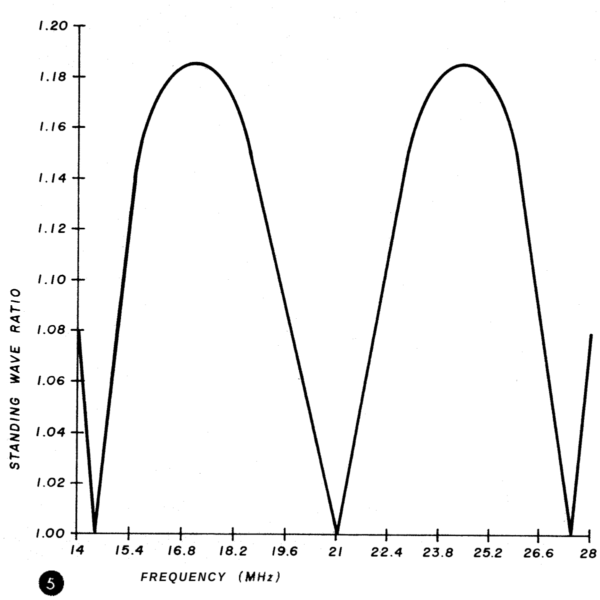


fig: 1. Phase-shift section

fig. 2. Complete phase-shift network.

fig. 3. Equivalent circuit of phase-shift network.

fig. 4. Construction details for phase-shift section.

fig. 5. VSWR of dispersive phase-shift section.

the rf range, this network is simpler than lattice networks, which require baluns for use with coaxial line. Other transmission-line-type phase shifters, such as those described by Schiffman,[1] require coupled stripline or coaxial line sections and are applicable primarily for microwave frequencies.

reference

1. B.M. Schiffrnan, "A New Class of Broad-Band Microwave 90 degree Phase Shifers," *Transactions IEEE MTT,* April, 1958, pages 232-237.

ham radio

comments

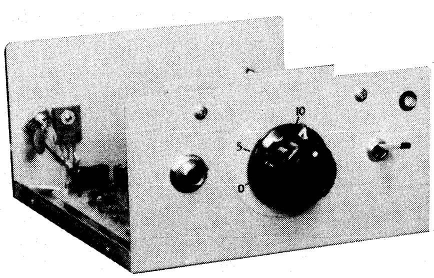
(Continued from page 6)

U4) to the top end of diode CR5 near the lower right hand corner of U13. Reconnect the clock control lead to either end of this new jumper. Finally, cut away a short segment of the foil to open the path between U13 pin 10 and CR5.

The effect of these changes is to delay the buildup at U13B pin 10 of the high on the clock control line until U3B has time to take U13B pin 9 low. The resistor and capacitor values are not too critical, and if this doesn't cure the click completely, try a slightly higher resistor value (up to about 270 ohms).

Robert C. Cheek, W3VT
Tokyo, Japan

split-band speech processing



Dear HR:

This note is prompted by Wes Stewart's article in *ham radio* for September, 1979, which I liked very much for his understanding of the subject and the unmasking of some of its secrets. However, there are a couple of points which can stand clarification; they have been alluded to in the original article describing the process (Fisk, *ham radio* June, 1976) but not discussed in detail.

The first thing I noticed in Stewart's design is the measured distortion, which is about twice as high as that of the *Vomax* unit by Maximilian Associates. The detrimental effect of distortion on the processing gain has received considerable attention in both my own *(ham radio* February, 1971) and Fisk's article; Stewart's design will probably show 2 dB less talk power gain than the commercial device, for which 10 dB is claimed. Explanations of the causes of distortion are a little tedious, but not really difficult.

Stewart's band splitting filters have a *Q* of 1.84; in contrast, the *Vomax* design value is 2.5, with the same passband ripple and slight reduction in overall bandwidth. I shall try to demonstrate the importance of the higher *Q* value. For a hard-clipped signal near the center of a band, the third harmonic distortion in that band due to finite filtering is 5 per cent for *Vomax,* 7 per cent for Stewart's design. The difference is not very significant. For frequencies above the filter peak, the numbers are essentially the same.

Unfortunately, the situation is very much worse for signals below the peak frequency of a particular band. Imagine a signal near the peak of one band, but below the peak of the next higher band by a factor of 1.73 ($\sqrt{3}$). If this signal is distorted (clipped), there will be no filtering as the third harmonic falls above the band peak by the same factor. If there is to be no distortion, then the maximum clipping must be held to whatever response is provided by a tuned circuit which is off-resonance by a factor of 1.73; this is 9.5 dB and 6.8 dB respectively for the *Vomax* and Stewart designs. Fortunately, this signal component amplitude is down by the same amount when combined with the lower band output, so that some distortion, i.e., clipping, is permissible.

For a total distortion limit of 10 per cent, this number is 2.5 dB, giving a total of 12 dB for *Vomax* and probably about 1 dB for a total of 8 dB for Stewart's circuit. Conversely, for the same amount of clipping, the lower *Q* design will produce higher distortion, approximately proportional to the square of the *Q* ratio. Obviously the higher the Q, the lower the distortion; this can be obtained within our design restraints, however, only by a smaller total bandwidth, higher passband ripple, or both. This distortion of signals below the peak frequency of a band increases rapidly with amplitude above a specific level, so that an effective agc of the preamplifier is an absolute necessity. For the perfectionist, I would suggest a design incorporating six or more bands permitting the use of higher Q filters for reduced distortion.

If I have one criticism of Stewart's article, it is his lack of concern for the input amplifier. The low-cost SL1626 IC is a loose-tolerance device with output level variations between samples of more than 8 dB. This necessitates individual calibration of the clipping level control for each assembly. Even worse is the effect of its output distortion, stated as 2 per cent typical with no maximum specified, on the performance of the processor.

Any distortion produced ahead of the processing circuitry will be magnified; i.e., the processor acts as a distortion enhancer. Imagine a signal near the center of the lowest band; if it contains second and third harmonic distortion, these individual components will appear near the centers or peaks of the second and third bands. With 14 dB clipping in the lowest band, the distortion components will be amplified by the same amount, or five times! Therefore, 2 per cent distortion in the input amplifier will show up as 10 per cent added to the processing distortion in the output combiner. While developing the final design for the *Vomax* unit, I investigated a number of available integrated circuits for use as input amplifiers with agc. I was unable to find one with acceptably low distortion (0.5 per cent), so resorted to the circuit implementation with discreet components.

To summarize, split-band processing, while an acknowledged valuable and effective technique, poses a number of design problems, which require careful attention to detail.

Walter Schreuer
Maximilian Associates
Swampscott, Massachusetts

I have studied Mr. Schreuer's comments, and with the exception of a couple of minor points, I find myself in agreement with them. Our major difference seems to be our different design goals. Mr. Schreuer's goal was a high quality, high performance, high cost commercial product; my goal was the design of a unit possessing similar characteristics that could be built by the average ham who could not or would not spend nearly $200 for the Vomax.

As mentioned in the opening paragraph of my article, the price paid for increased performance is increased complexity and cost. Even Mr. Schreuer agrees that to approach perfection the design would require six or more bands. Obviously, we must stop somewhere. Mr. Schreuer and I seem to disagree only on where to stop.

Higher Q filters can offer lower distortion; however, I feel that higher precision components should then be used in the active filters and summing amplifier. While I personally have access to precision parts, I felt the average home builder would have trouble finding them and designed accordingly. Complete design data was given in the text and appendices for those wishing to try design improvements and the experimenter is encouraged to do so.

I must add that Mr. Schreuer took the liberty of comparing the results of using higher Q using my center frequencies. If we compare results using a Q of 2.5 and the center frequencies used in the Vomax, *the off-resonance attenuation computes to −8.4 dB, not the −9.5 dB claimed by Mr. Schreuer; a small point to be sure but perhaps worth mentioning.*

Regarding the SL1626, the tolerance on output level is only 6 dB, not the over 8 dB claimed by Mr. Schreuer. Except in a production environment, this is not a big problem. Mr. Schreuer raises a good point regarding the distortion characteristics. Perhaps he could share his expert knowledge in this area with those who wish to go to the increased complexity of the discreet circuit.

In summary, I feel the design I presented in the September issue fills a need for an easily reproduced, inexpensive speech processor that delivers more performance enhancement for dollars spent than anything other than perhaps antenna improvements.

Wes Stewart, N7WS
Tucson, Arizona

talking clock

Dear HR:

The diagram of the Talking Clock addition in October, 1979, *ham radio* has an error: the leads connecting to pins 10 and 11 of the 74157 should be interchanged.

The Talking Readout described in the June, 1979, issue was described as it was interfaced to the Spectronics DD-1, which I understand is no longer in production. However, I have successfully interfaced the unit with the Kenwood DG-1, intended for use with the Kenwood 520S. Since the article was published, the multiplexer portion of the circuit has been made available on a separate board, which can be inserted into the DG-1, reducing the interconnection to about nine leads rather than thirty, as multiplexing is done *in situ*. The BCD data is accessed as it enters the six 7454s. Since this is latched or stored data, no gating is required, so pin 4 of the 7454 of the talking digital readout control board may be tied to +5 volts.

Similarly, the Yaesu ZD-1 external readout for the FT101-Z should be readily interfaced; BCD data for each digit exist at pins 5, 10, 2, 12, sf the six 74LS196s. Gating should be required but is available at 001, which drives the LED cathodes.

Interfacing to other transceivers may be complicated by the increasing use of seven-segment counter outputs from counters. When I have located a suitable seven-segment to BCD decoder, such as the 74C915, a redesign will be undertaken.

I would be happy to answer any correspondence regarding either article from readers who include an SASE. Inquiries related to other transceivers should include a detailed diagram of the digital readout portion.

Ray Brandt, N9KV
Janesville, Wisconsin

Dear HR:

I would like to express my appreciation for your series of articles on "Digital Techniques"; it expanded my knowledge of digital devices considerably. I hope that this series is continued, and I would like to see these articles expanded into the more complex integrated circuits used in microprocessors. Understanding the MCS6522 and MCS6530 would be typical.

I subscribe to *ham radio* not for the straight Amateur Radio articles but for articles on things like the phase-lock loop, power supplies, counters, operational amplifiers, and other technical subjects.

Keep up the good work.

Norbert K. Fox
Guttenberg, Iowa

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MARCH 1980

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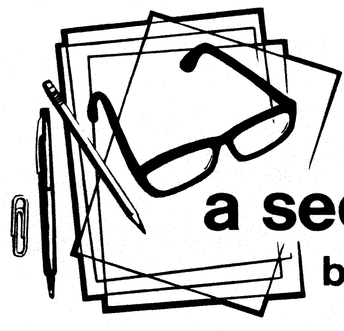


a second look

by Jim Fisk

Just as we were about to go to press with this issue, the FCC adopted rules permitting the use of ASCII (American Standard Code for Information Interchange) on the Amateur bands; at press time the FCC had not yet set a date when Amateur ASCII transmissions would be permitted, but early March was suggested, so it's quite likely that you will hear some ASCII signals on the ham bands by the time you receive this issue of the magazine. If you don't have a terminal unit, however, you probably won't be able to tell any difference from the Baudot code used until now for Amateur RTTY! ASCII is more versatile than Baudot, and is capable of handling 2½ times more individual characters (or unified commands), so it's much more powerful; it will be especially popular with Amateurs who are also computer hobbyists and wish to link their computers and exchange programs over the air.

For those readers who are not familiar with ASCII, it is a standard code used extensively in digital data transmission, and is commonly used when computers "talk" to each other. RTTY and computer enthusiasts have been trying for some time to obtain ASCII authorization for use not only in conventional communications, but also for exchanging digital data, computer control of repeaters and remote stations, and even for exchanging digitized voice and video on the high-frequency bands.

The new FCC rules permit the use of ASCII transmissions (carrier-shift keying, F1) on the same frequencies presently authorized for RTTY between 3500 kHz and 21.25 MHz at a maximum rate of 300 baud. On the RTTY frequencies between 28 and 220 MHz transmission rates up to 1200 baud may be used with carrier-shift keying (F1), AFSK (F2), or amplitude tone-modulated keying (A2); above 420 MHz these modes may be used with rates up to 19.6 kilobaud.

There is also considerable interest in ***packet*** radio, a digital mode used since 1978 by Canadian Amateurs on the 220-MHz band. Packet radio is basically a time-shared use of the same frequency channel that results in a tremendous savings of spectrum. Look at it this way: it might take you 20 seconds to type a line on your terminal — and only a few thousandths of a second for the computer to process that line and enter it into storage. If the line is not transmitted character by character as you typed it but is sent as a short, high-speed burst at the end of each line, you have packet radio.

At the keying rates permitted on 80 through 15 meters, the typical transmission time for a one-line packet would be something less than 2 seconds, a 10:1 savings over the time required to enter the line into the terminal in the first place. Therefore, up to 10 different stations could communicate on the same packet channel; the significantly higher keying rates permitted at vhf would permit 100 stations or more on the same frequency! You'll be hearing a lot more about packet radio in the months ahead, but in the meantime you may want to read VE2BEN's excellent "Introduction to Packet Radio" which appeared in the June, 1979, issue of *ham radio.*

Jim Fisk, W1HR
editor-in-chief

comments

thought that it was difficult for Chinese to read, and shows what would be expected when there may be as many as thirty or so strokes in one character. Obviously, that requires rather good facsimile definition or it might not be possible to read.

E. H. Conklin, K6KA
La Canada, California

Hellschreiber

Dear HR:

The December issue of *ham radio* describes the Hellschreiber typing keyboard machine, and the article mentioned that the shortest pulses are 8.16 ms, producing a speed of 122.5 baud and a minimum bandwidth of 61 Hz.

Unfortunately for the system, the abrupt rise and fall times involved are quite broad. A similar system that has been on 14140 kHz from the Hsinhua News Agency in Peking for years (but now possibly removed at the request of the Intruder Match), is more like 3 or 4 kHz wide at a distance 6241 miles; it was hard to live with. There continue to be other signals here in the mornings on 3577, 3595, and 3845 kHz just before the 80-meter band closes to China.

The system I heard during WW2 was used by German fighter aircraft in interception, and directed from ground by this equipment. That was not so bad because vhf was used, and at some distance in frequency from other communications circuits. I think that it is a mistake to encourage the use of this system of 14 emission, which is *not* authorized by FCC Regulations, Section 97.61, except on frequencies of 51.1 MHz and higher.

My tape of the Hsinhua transmissions was printed by G5XB, who thought that it was difficult for Chinese to read, and shows what would be expected when there may be as many as thirty or so strokes in one character. Obviously, that requires rather good facsimile definition or it might not be possible to read.

Dear HR:

The December issue of your magazine arrived this morning and as usual I sat down to skim it – saving the serious stuff for later.

I was surprised at the Hellschreiber article. You see, I have one of these machines, sitting above the rafters of the shack, waiting until someone came up with the other one.

That last remark is deliberate: It relates to the time of WWII when I was working with the Signal Service Section of the Signal Corps in Liege, Belgium. We were located in rear of the 15th Army; they were sending back captured German equipment to our depot, and we had no orders how to process it. We were very busy reworking our own equipment. I was acting as senior officer in charge of salvage and incoming equipment.

Among the items coming in on the rail cars was this type of equipment. I intercepted three of these Hellschreibers and shipped two home complete and one in parts less the case. Luckily, the case size just fit the maximum package size that could be shipped home. I was also depot security officer, and as such knew what could be shipped and what couldn't. Numerous articles were shipped at my personal expense to the Signal Corps Laboratories at Fort Monmouth, where I had spent several weeks in 1942.

Thanks to the poor work of the Army mail, and over-emphasis on what could, and could not, be received within a country still at war, only one of the Hellschreibers arrived at my father's home address. The other and the parts were not received, but I did receive some papers telling me that it was illegal to ship this stuff (our orders stated that souvenirs could be shipped only if certain papers were put on the outside of the package).

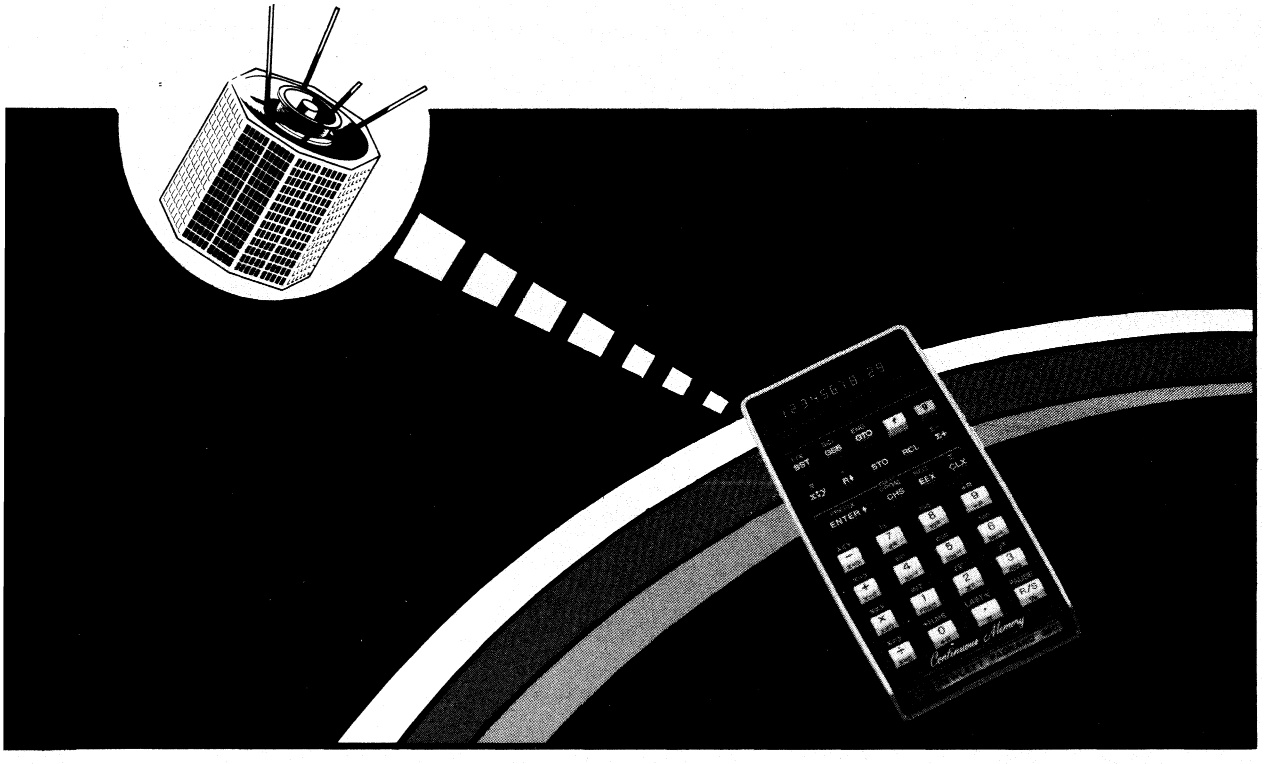
To say I was disgusted and angry is to put it mildly, but having only one Hellschreiber and no spare parts, I simply put it away until I could find use for it. I arrived home in December, 1945, and with a new wife and setting up a home and finding work, it was forgotten for many months.

Therefore, there is at least another of these machines in the United States. It will be marked inside with my call W6DKZ. The parts were not identified, as I thought no one would be interested in them. I am still looking for the missing machine, and would like to get in contact with anyone who might be saving it, as I intended it for a museum. If it turns up, I'll try out the two between some friends here in Santa Clara Valley.

The Hellschreiber machines are all that the writer says they are, although I did not know they would work well through QRM. They were made to work on wire lines as simplex or duplex, with isolating coils, and since they employ a tone and amplifier, they don't interfere with speech on the lines.

Henry B. Plant, W6DKZ
San Jose. California

Apparently the German Wehrmacht was not alone in their use of the Hellschreiber system during World War II. Ed King, WA8PFB, of Louisburg, West Virginia, reports that he has a U.S. Signal Corps BC-918B which has a similar ink pad and worm gear mechanism for "writing" on paper tape, but a photo-cell is used for the input, Ed's BC-918B has a 20-pin plug so it's part of a larger system, but Ed has been unable to locate the matching unit, or even to find a technical manual. Does anyone have any more technical details on this equipment or any ideas how it was used, or know where there might be a technical manual? ***W1HR***



satellite computer

for under $150

The Hewlett-Packard HP-29C is put to work for OSCAR enthusiasts in computing orbital geometry

The availability of card-reading calculators, such as the Hewlett-Packard 67 and SR-52, has made it possible for satellite users to track **OSCAR** without so much as using a pencil and paper. Fine programs for both machines have been written and published,[1] but there's one pitfall associated with the two systems – cost. For those who bought one of the $400 gems for purposes other than satellite tracking, keying in the appropriate **OSCAR** program is certainly worth the effort. However, it's not likely that an Amateur Radio operator will spend that much money on a calculator for just satellite tracking, since an entire ground station needn't cost more than a few hundred dollars.

the calculator

A more realistically priced programmable calculator that can serve the **OSCAR** community as well as the geosynchronous-satellite enthusiasts is the Hewlett-Packard Model HP-29C. Although the HP-29C doesn't have the luxury of a card reader, it does have continuous-memory capability, which, in my opinion, matches and in some instances outweighs the virtues of those card-readers that are dumbfounded when power is interrupted. The HP-29C may be purchased at college bookstores for as little as $135 – a modest sum compared with the price of automatic keyers that send CQ and little else! Even if purchased only for **OSCAR** purposes, the **HP-29C's** price is worth considering when you realize its full potential as an SWR calculator, dipole-length computer, or grocery-bill adder. Truly it's a good investment.

the program

Developing an **OSCAR** program that fits within the 98-step memory limit of the HP-29C demanded a

By Richard Cleis, WB6POU, 438 South Havenside Avenue, Newbury Park, California 91320

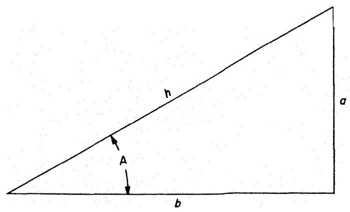


fig. 1. Pertinent to the program is the coordinate-conversion feature of the HP-29C calculator. When solving a right triangle, given h and A, the P-R key yields a and *b*. Conversely, given a and *b*, the R-P key yields h and A.

process far removed from the popular memory-gobbling Napier's Rule formulas of expensive calculators. My method requires no discrete trigonometric functions, but depends on the HP-29C's ability to perform rectangular-to-polar coordinate conversions and the reverse operation. The program does the following:

1. Calculates azimuth and elevation to satellite

2. Calculates slant range

3. Runs in real time

4. Displays real time

5. Can be set backward or forward any number of minutes

6. Can be set backward or forward any number of orbits

7. Can be set to start at any time

8. Works for any ground-station location

9. Uses only 11 seconds per calculation

10. Calculator can be turned off without losing track of OSCAR's position

11. Program can be used to find azimuth, elevation, and slant range to a prescribed synchronous satellite

The polar-rectangular (P-R) conversion key of the HP-29C is valuable because it yields the length of two sides of a right triangle when given the hypotenuse and the angle included in the hypotenuse and one side. The rectangular-polar (R-P) key does the operation in reverse (see fig. **1**). Nearly every aspect of the program uses these two functions.

finding the rectangular coordinates of OSCAR

One portion of the program determines the movement of the satellite by incrementing the number of degrees that the satellite travels in its approximate circular orbit. Assume OSCAR travels arc distance T degrees from an arbitrary point as shown in fig. **2.** After the program recalls *T* and the radius r, of the orbit, values *s* and dare calculated with the P-R function. Recalling the inclination of the satellite, and performing a P-R conversion in conjunction with s, yields z and v. The next portion of the program finds angle A and distance e through an R-P operation on v and d.

During the period that OSCAR takes to travel arc distance *T*, the earth will have rotated beneath the satellite. The x and y axes have thus rotated counterclockwise, as shown by the dotted line. The angle of rotation, R, is added to angle A, obtaining the system shown in fig. 3. Performing a P-R conversion on e and the new angle, A + R yields y and x, while *z* remains as calculated earlier.

What's been accomplished during the previous calculations is the assignment of the satellite to a three-dimensional rectangular coordinate system. When the position of the tracking station is assigned to its rectangular coordinates on the same system as that of the satellite, the relative rectangular position of the satellite is determined by subtracting the corresponding x, y, and z values (fig. **4**).

finding azimuth, elevation, and slant range

After the subtraction operations, the new x, y, and *z* values are used to set up the final coordinate system (fig. **5**). Length *g* and angle *B* are calculated by converting z and x to their polar equivalents. The complement of the tracking station's latitude is then added to *B*, since we want to know the azimuth and elevation according to the earth's surface at the station rather than with reference to the *x*/*y* plane parallel to the equator. Next, a P-R conversion is performed on *g* and *B* + (90-latitude) to obtain t and *u*.

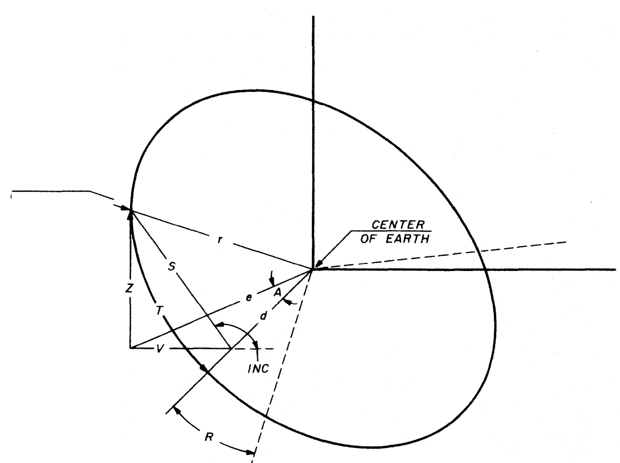


fig. 2. Geometry for determining satellite movement in its orbit. Values s and dare calculated with the calculator P-R function.

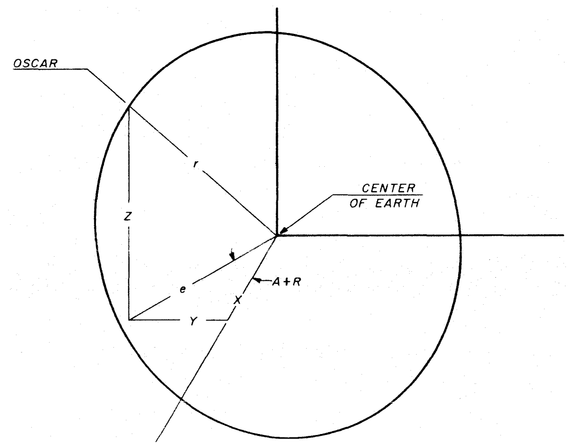


fig. 3. Accounting for earth rotation after the satellite has traveled arc distance *T*.

Azimuth is derived by an R-P operation on the *u* and the y value calculated earlier. Since distance v is also found during the previous step, elevation and slant range can be derived by performing an R-P conversion on t and v.

preparing the program

The first time the calculator is to be used for tracking a satellite, the program must be keyed in. Twelve data registers have to be filled. Since the HP-29C remembers when it's turned off, only half of the memories must be entered again so long as the station location is not altered. Registers 1 and 2 will change daily because they're the reference orbits' equatorial times and crossings respectively.

Four other registers (.1, .2, .3, and. 3) must be changed when you want to track a satellite other than the one for which the calculator is set up. In other words, if OSCAR 8's progression, inclination angle, orbit radius, and arc rate are stored in the data registers, they must be changed to represent OSCAR 7's parameters.

tracking a satellite

Suppose the calculator is set for tracking OSCAR 7 and you want to work it on December 20. The first step is to look up the *reference node data* for that date. Turn on your HP-29C, punch in the time in H.MS format. Store the data in register 1. Next key in the longitude. Store it in register 2.

You must now guess the number of orbits that will pass before acquisition will occur. If you think three orbits will suffice, 3 should be entered and **GSB 1** depressed (see program, **fig. 6**). The calculator will then display the time at which the satellite crosses the equator after three passes beyond the reference node.

Press the **x-y** key to display the crossing longitude relative to the station. If the crossing is 15 degrees east of the station, the displayed result is – **15,** since the difference between station longitude and satellite longitude decreases on successive orbits.

If the node is 15 degrees west of the station, **+15** will be displayed, because successive nodes are further away from the station.

You must guess with the **GSB-1** function until you find an appropriate node that will produce an orbit likely to pass through the station window. Nodes between – 45 and + 25 are likely candidates for stations with a latitude near that of Los Angeles.

If you want to begin tracking the satellite from a nodal point, wait until the calculated crossing time arrives, then push **GSB** 2. The HP-29C does about 10 seconds of number crunching, then displays the time for about a second, displays the slant range for a second, displays the azimuth for a second, displays the elevation for a second, then repeats the cycle until 1 minute is used up. During each minute, the time is incremented, so that the calculator is tracking in real time.

The ascending nodes of OSCAR 7 are often too far south for North-American stations to see, and they may occur during the middle of the window for southern Florida stations. Also, the descending passes aren't visible until the satellite travels up the back side of the earth — a 40-minute wait.

Getting to the point, few operators will desire to begin tracking the satellite from the time it crosses the equator. The program is equipped with solutions to this problem.

notes on calculator subroutines

Two subroutines permit the track to be started at

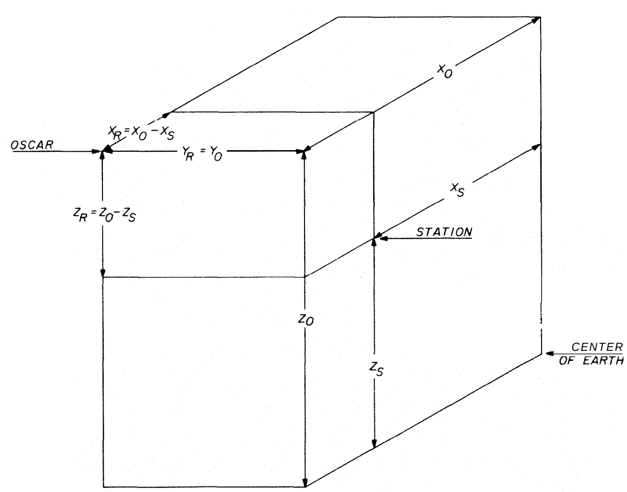


fig. 4. The relative rectangular position of the satellite is determined by subtracting the corresponding x, *y*, and *z* values.

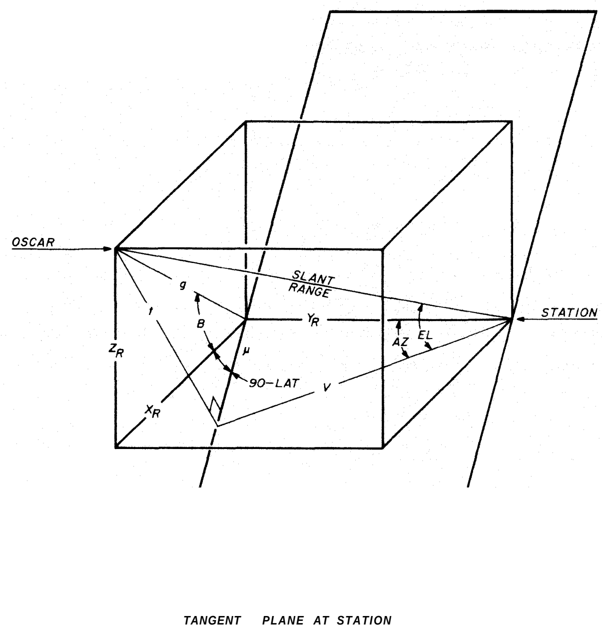


fig. 5. Geometry for determining final coordinates of the satellite (azimuth, elevation, and slant range).

any time and enable the time of acquisition to be easily found. Keying in any number of minutes and pressing **GSB** 3 will advance the time to the prescribed value. If a negative number is used the time will be back-tracked. This subroutine is similar to **GSB 2;** so after **GSB** 3 is called, the time will be altered, then the program will run in real time as if **GSB 2** had been used.

The other time-altering subroutine is **GSB 4.** By entering any time in H.MS format, then pressing **GSB 4,** the program will begin tracking at the prescribed time.

To enhance your search for the acquisition time, the satellite elevation will be displayed as a negative number if it's below the horizon. If you make a guess that produces an elevation of, say, −**45** degrees, then you can immediately assume that many minutes must be added. If an elevation of −3 degrees is discovered, then acquisition is only a few minutes away.

To realize the usefulness of the features offered in this program, it must be stressed that the program is time based. The only register that changes is the time register, so the calculator may be turned off at ***any time*** during ***any*** subroutine without losing track of the satellite's position. If the calculator is turned off, turning it on and pressing **GSB 2** will start it tracking at the time it was last calculating. The other subroutines will also function as if the calculator were never turned off.

It's also important to note that the subroutine **GSB 1** is independent of the number of times that any other subroutines are used. In other words, whatever

User Instructions

Program Title HP-29 OSCAR PROGRAM
Name RICHARD A. CLEIS Date MARCH, '80
Address
City State Zip Code

STEP	INSTRUCTIONS	INPUT DATA/UNITS	KEYS		OUTPUT DATA/UNITS
	INITIALIZING PROGRAM				
1	Keyin program				
2	360	Degrees	STO	7	
3	1/60	Hrs/Min	STO	6	
	ENTERING TRACKING STATION LOCATION				
1	Longitude of tracking station	Deg. W	STO	4	
2	90-Latitude of tracking station	Deg. N	STO	5	
3	Latitude of tracking station	Deg. N.	ENT		
4	Radius of earth: 3964	Miles	f	R	X_{ts}
5			STO	.4	
6			x/y		Z_{ts}
7			STO	.5	
	ENTERING SATELLITE DATA				
1	-60^x increment/period in minutes	Deg/Hr	STO	.1	
2	Inclination of orbit	Degrees	STO	.2	
3	Radius of orbit	Miles	STO	.3	
4	Arc rate of satellite: 21600/period in minutes	Deg/Hr	STO	3	
	ENTERING DAILY OSCAR DATA				
1	Time of equatorial crossing	H. MS	STO	1	
2	Longitude of equatorial crossing	Deg. W	STO	2	
	FINDING DAY'S EQUATORIAL CROSSINGS				
1	Key in number of orbits to skip after reference		GSB	1	Xing time
2	Longitude with respect to station yielded here:		x/y		Xing long.
	CHANGING MINUTES: Keyin ± # of minutes		GSB	3	New data
	SETTING NEW TIME: Keyin desired time		GSB	4	New data
	STARTING TRACK WITHOUT CHANGES		GSB	2	
	FINDING SYNCHRONOUS SATELLITE DATA				
1	0		ENT		
2	Station longitude-satellite longitude	Deg. W	ENT		
3	Radius of synchronous orbit	Miles	GSB	5	Sat.data

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	g LBL 1	15 13 01	
	RCL 7	24 07	
	RCL 3	24 03	
	/	71	
	X	61	
	RCL 1	24 01	
	g H	15 72	
	+	51	
	STO .0	23.0	
010	GSB 8	12 08	
	CHS	32	
	RCL .0	24.0	
	f H.MS	14 72	
	g RTN	15 12	
	g LBL 8	15 13 08	
	3	03	
	7	07	
	STO 0	23 00	
	RCL .1	24.1	
020	RCL .0	24.0	
	RCL .1	24 01	
	g H	15 72	
	−	41	
	X	61	
	STO 8	23 08	
	f LAST x	14 73	
	RCL 3	24 03	
	X	61	
	RCL .3	24.3	
030	f R	14 44	
	STO 9	23 09	
	ROLL	22	
	RCL .2	24.2	
	x/y	21	
	f R	14 44	
	RCL 9	24 09	
	g P	15 44	
	x/y	21	
	RCL 8	24 08	
040	+	51	
	RCL 4	24 04	
	+	51	
	RCL 2	24 02	
	−	41	
	g RTN	15 12	
	g LBL 2	15 13 02	
	GSB 8	12 08	
	x/y	21	
	g LBL 5	15 13 05	
050	f R	14 44	
	x/y	21	
	STO 9	23 09	
	ROLL	22	
	RCL .4	24.4	
	−	41	
	x/y	21	

STEP	KEY ENTRY	KEY CODE	COMMENTS
	RCL .5	24.5	
	−	41	
	x/y	21	
060	g P	15 44	
	x/y	21	
	RCL 5	24 05	
	+	51	
	x/y	21	
	f R	14 44	
	CHS	32	
	RCL 9	24 09	
	x/y	21	
	g P	15 44	
070	x/y	21	
	g x> 0?	15 51	
	GTO 9	13 09	
	RCL 7	24 07	
	+	51	
	g LBL 9	15 13 09	
	STO 9	23 09	
	ROLL	22	
	g P	15 44	
	RCL .0	24.0	
080	f H.MS	14 72	
	RCL 9	24 09	
	g LBL 7	15 13 07	
	ROLL	22	
	f PAUSE	14 74	
	g DSZ	15 23	
	GTO 7	13 07	
	1	01	
	g LBL 3	15 13 03	
	RCL 6	24 06	
090	X	61	
	STO + .0	23 51 .0	
	GTO 2	13 02	
	g LBL 4	15 13 04	
	g H	15 72	
	STO .0	23 .0	
	GTO 2	13 02	
100			
110			

REGISTERS

0 used	1 Time Eq X	2 Long. Eq. X	3 Arc d / Hour	4 Long QTH	5 Lat. QTH	6 1/60	7 360	8 used	9 used
S0 Time	S1 60 Inc. / period	S2 Inclin.	S3 Radius Orbit	S4 x_{QTH}	S5 y_{QTH}	S6	S7	S8	S9
A	B	C	D	E	I				

fig. 6. User instructions for the satellite-tracking program using the HP-29C calculator.

the calculator is doing (on or off), if a number, n, is keyed in and **GSB 1** is depressed, the time and longitude of the nth node before or after the reference node will be calculated. Analogous independence rules can be applied to the other subroutines.

program versatility

Suppose the satellite's fourth orbit after the reference node has just been tracked, and the time is 0300. You want to work the following pass, so you can choose one of three alternatives:

1. Enter **5 then GSB 1**
2. Enter **100** then **GSB** 3
3. Enter **4.40** then **GSB 4**

The first choice calculates the node of the next orbit. From that point you may start looking for the acquisition time with the other subroutines. The second choice adds 100 minutes to the time; an orbit lasts about 100 minutes, so the program is positioned somewhere near the window of the following orbit. From inside the window, it's easy to find the acquisition time. The third choice is almost exactly the same as the second, except you must mentally add the 1-hour, 40-minute orbit period to the time (0300).

synchronous satellites

Finding azimuth, elevation, and slant range of a synchronous satellite requires entering a zero, the difference between the satellite and station longitudes, then the radius of a synchronous orbit. After pressing **GSB 5,** the calculator will display the time of the last satellite orbit (meaningless for this purpose), then it will display slant range, elevation, and azimuth in that order.

credibility

The methods used in this program should work for any near-circular orbit and any ground-station location. I've successfully used the program for OSCAR 7 and OSCAR 8 on ascending and descending passes. Although I've never worked with the Russian satellites, I feel confident that the program will work as well for them.

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ham radio

More Details? CHECK–OFF Page 94

new class of coaxial-line transformers

Coreless 4:1 and 1:1 balun transformers are described with a systematic design procedure for making your own — Part 2 of a two-part series

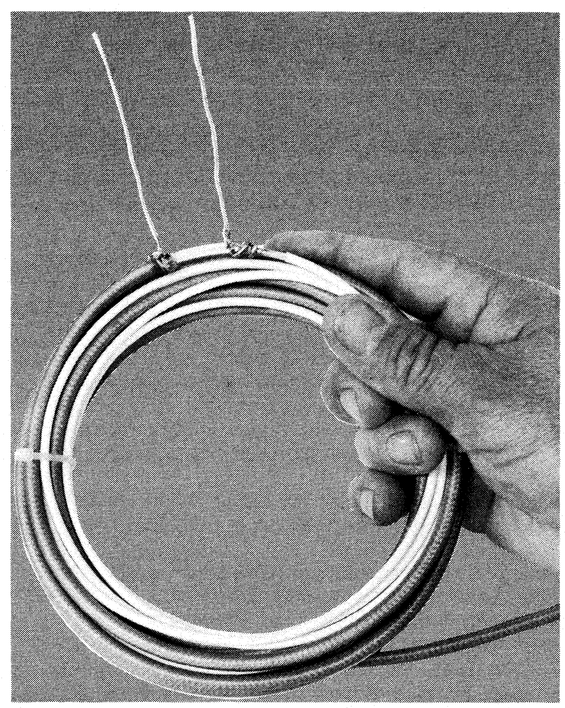
Part 1 of this article reviewed the theory of transmission-line transformers and baluns, as well as problems with magnetic cores such as arcing, distortion, and harmonics. A simple balun that doesn't depend on magnetic materials was described. A new class of coaxial transmission-line transformers based on the same principles as the coreless balun was introduced. Also described were two specific transformer designs with experimental performance data.

In this article I will describe additional 4:1 and 1:1 balun transformers, including one for vhf. Impedance, VSWR, and balance data on these specific designs and on commercially available balun transformers are compared. I have included data on baluns working into various loads, with information on how to build and modify balun transformers. A systematic design procedure, evolved during the development of these transformers, is summarized.

how to make coreless baluns

While the balanced-to-balanced 4:1 transformers described in Part 1 are interesting, more useful configurations are 50-ohm unbalanced to 12.5-ohm balanced, and 50-ohm unbalanced to 200-ohm balanced, balun transformers. These were made in two stages using coreless baluns together with the balanced-to-balanced coreless transformers of figs. 5 and 6 described in Part 1.

The first step in making these balun transformers was to arrive at an optimum 50-ohm 1:1 balun design. I tried many lengths of coax and many configurations before choosing the design shown in fig. 1. A length of RG141/U* Teflon coaxial transmission line longer than 127 cm (50 inches) was used. A dummy length of line 127 cm (50 inches) long was soldered to the outer conductor of the Teflon coax 127 cm (50 inches) from the end as shown in fig. 1.



How to add a simple compensating winding to the W1JR balun to provide superior balance. Thanks to W6ZO for building the balun and suggesting the easy modification. Low reactance adjustable load shown is connected with 1.27-cm (0.5-inch) copper strap for match and balance measurements.

*Available from Radiokit, Box 429, Hollis, New Hampshire 03049. RG-142B/U (Belden 83242-100) may also be used.

By George Badger, W6TC, 341 La Mesa Drive, Portola Valley, California 94025

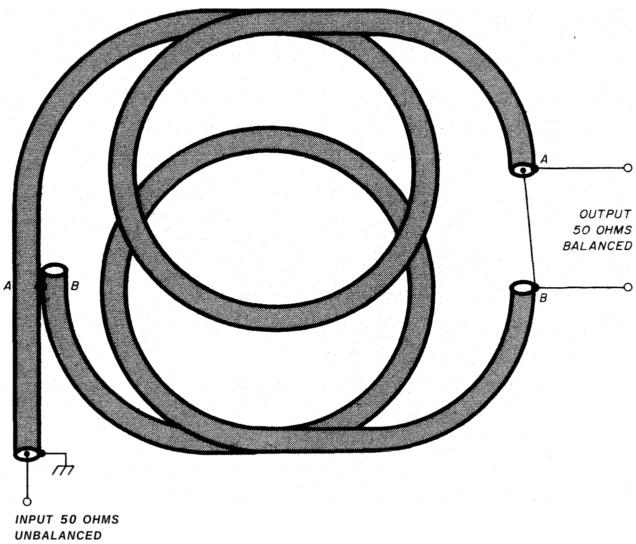


fig. 1. 50-ohm unbalanced to 50-ohm balanced coaxial balun. From the common point (system ground) to the output terminals, coax line A and compensating line B are each 127 cm (50 inches). The lines were wound into a seven-turn random-wound coil of 11.5 cm (4.5 inch) nominal diameter. For clarity, only three turns are shown. Performance data are shown in table 1.

The resulting 254 cm (100 inches) of line was then random wound into a nominal 11.5-cm (4.5-inch) diameter seven-turn coil. **Fig. 1**, for clarity, shows only three turns of coaxial line. The dummy length of line used for the compensating winding was made with RG-58A/U. The advantage of using small coax is that the balun is compact and results in a convenient configuration for mounting on beam antennas. Performance is shown in table 1. Note the excellent balance data. Balance was determined by measuring the rf voltage with respect to the common point (ground) at each of the output terminals when terminated with a floating matched load. The difference between the readings taken at each frequency was divided by the sum of the readings expressed as a percentage . The rf voltage was measured with an HP model 410C rf voltmeter.

Rather than use coax for the compensating winding, to save money and space I decided to try a length of hookup wire. I tried some surplus no. 12 (2.1-mm) Teflon insulated wire. Hookup wire instead of coax for the compensating winding results in an excellent design. The balun is shown in the photo. A balun made this way was compared with one made entirely of coax. The two designs used for this comparison were optimized for the low bands. Data taken on these designs are shown in table 2. This table compares the use of Teflon coated no. 12 (2.1-mm) wire with coax for the compensating winding. VSWR performance of the Teflon wire version was at least equal to that of the all-coax balun, and the balance was actually better.

The balun design optimized for the 80 through 10 meter bands (table **1**) was made with 127-cm (50-inch) lines. The balun designs optimized for the 160 through 20 meter bands (table **2**) were made with 254-cm (100-inch) lines.

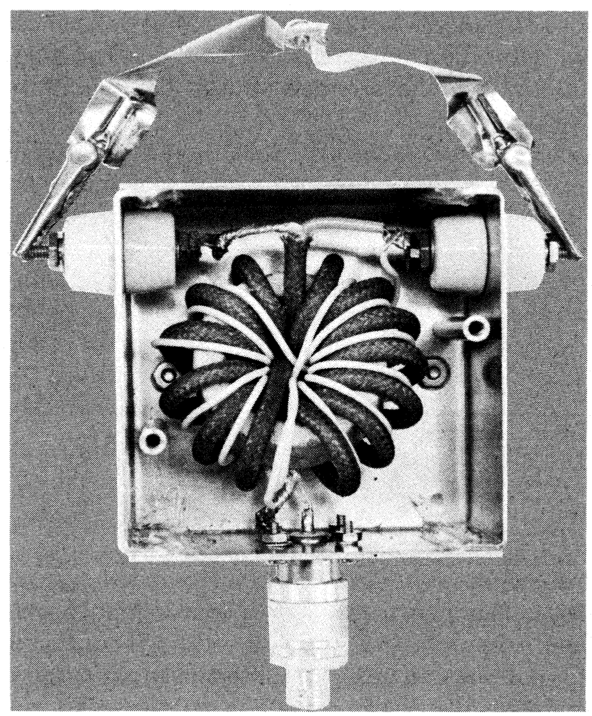
vhf balun

A vhf version of the coreless 1:1 balun is shown in **fig. 2.** The balun has a nominal diameter of 5.8 cm (2.25 inches). The length of coax from the output to the common point is 45.7 cm (18 inches). An equal length of coax line is used for the dummy. The lengths of RG-58A/U were wound into a five-turn coil. Table 3 shows the performance of this balun.

two-stage balun transformers

After the 50 to 12.5 ohm and 50 to 200 ohm balanced-to-balanced transformers (Part 1) and the 50-ohm unbalanced to 50-ohm balanced balun (see **fig. 1**) were optimized, I combined them into two-stage 50 to 12.5 ohm and 50 to 200 ohm unbalanced-to-balanced configurations. These two-stage transformers are shown in **figs.** 3 and **4.** Tables **4** and **5** show performance data.

The first stage converts from 50-ohm unbalanced



Compact broadband 1:1 balun. The only materials used are a short length of RG-141/U and insulated hookup wire. The balun provides excellent match, balance, and several kW reserve power-handling capability.

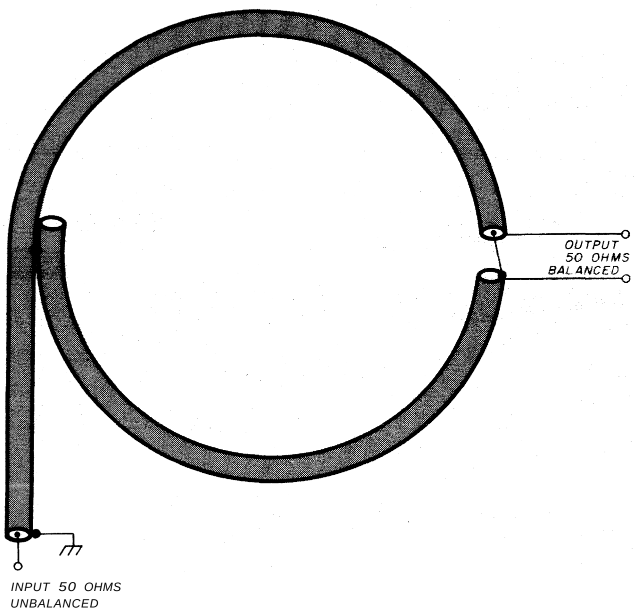


fig. 2. Vhf version of the coreless balun. Length of each line from the common point to the output terminals is 45.7 cm (18 inches). The lengths of RG-58A/U were wound into a five-turn coil of about 5.8 cm (2.25 inches) diameter. For clarity, only one turn is shown. The center conductor of the compensating coax line winding may be left floating or shorted to the outer conductor at both ends. Performance data are shown in table 3.

to 50-ohm balanced; the second stage converts from 50-ohm balanced to 12.5-ohm or 200-ohm balanced loads. Note that the bandwidth of these two-stage balun transformers is somewhat less than that of the individual stages.

When the two stages are coiled together into one compact bundle of coax, the way in which connections are made between the two stages is important. Note the lead crossover between the first and second stage **(fig. 4).** Performance was significantly better when the leads between stages were cross-connected because of the magnitude and direction of rf current flow over the coaxial-line outer conductors. The leads between the two stages must be short. The length of grounding wire, **AB,** was not critical.

a 50112.5-ohm unbalanced-to-unbalanced transformer

A transformer particularly useful for matching low-impedance unbalanced loads, such as a mobile whip or short ground plane antenna, is shown in **fig. 5.** Note that this configuration differs from the designs described earlier because the line lengths aren't random wound into a common coil and, therefore, aren't coupled together. Because of the unbalance-to-unbalance connection, both ends of the outer conductors of line **CD** are grounded. Thus, line **CD,** if coiled with and therefore coupled to coil **AB,** would act like a shorted turn, reducing the common-mode impedance of coil **AB.** Both ends of line **CD** are at the same potential so no isolation impedance is required. Thus the line may be positioned in any convenient way that doesn't couple to coil **AB.** The line is shown folded in the drawing to minimize coupling. The line may be twisted and taped to the incoming 50-ohm line. Line **CD** must be the same length as line **AB** so that the two rf paths are equal, thus preserving the phase relationship. Lines **AB** and **CD** are each 127 cm (50 inches) long and are made of two paralleled lengths of RG-58A/U.

Performance data on the 50/12.5-ohm unbalance/unbalance transformer is shown in **table 6.** The VSWR data show the harmful effect of the shorted turn when **CD** is coiled and coupled to **AB.** The VSWR curve could be centered to improve the match at the low end by adding length to lines **AB** and **CD** by the design techniques described later.

efficiency and power

The power-handling capability and efficiency of these new transformers made with RG-58/U coaxial cable were analyzed in Part 1. The 4:1 baluns shown in **figs. 5** and **6** of Part 1 and **fig. 5** of Part 2 can handle 1 kW at 30 MHz. This is twice the rating of RG-

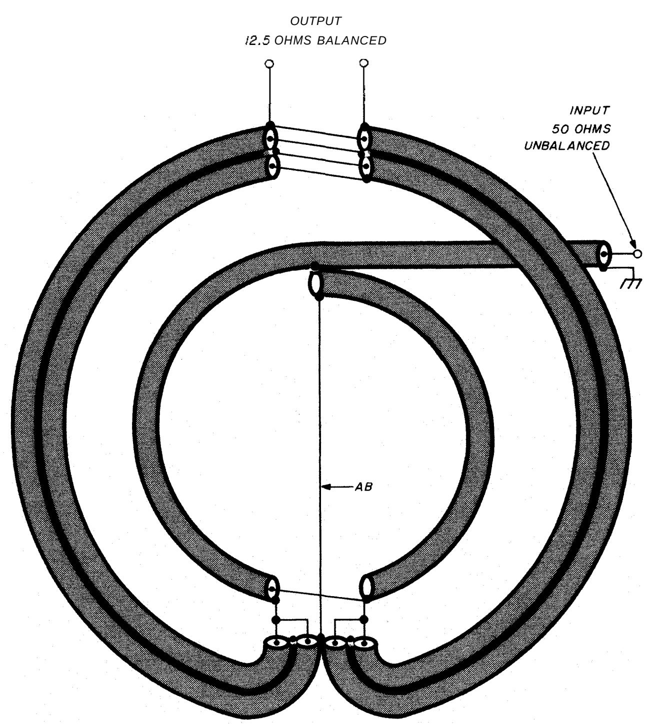


fig. 3. This 50-ohm unbalanced to 12.5-ohm balanced balun transformer is a two-stage design combining the transformer of fig. 5 (Part 1) with the balun of fig. 1. Performance data are shown in table 4.

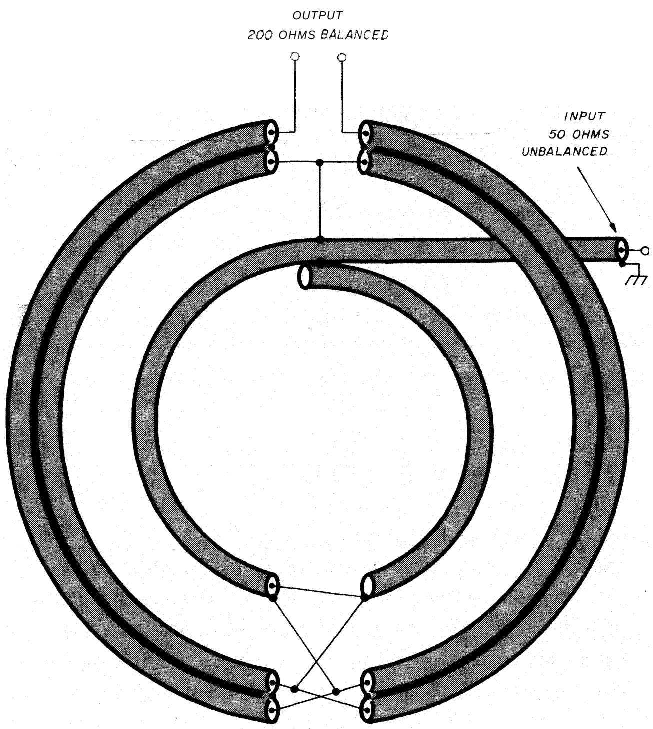


fig. 4. This 50-ohm unbalanced to 200-ohm balanced balun transformer is a two-stage design combining the transformer of fig. 6 (Part 1) with the balun of fig. 1. Performance data including balance are shown in table 5.

58A/U cable because the line pairs are connected in series or parallel.

In the case of the coreless 1:1 baluns, all of the power is transmitted through a single coax. Therefore, for high-power applications, the 1:1 baluns must be made with RG-8/U or RG-141/U transmission line. RG-141/U is the same size as RG-58A/U but it's about four times as expensive. The dielectric used in this coax is Teflon; therefore, the baluns can handle about 5 kW at 30 MHz. Using RG-141lU or RG-142BlU results in a rugged balun of reasonable size. From my experience, these compact baluns made with Teflon coax are virtually indestructible in Amateur use.

Efficiency of the Teflon coax balun shown in the photo was tested by the method described in Part 1.

table 1. Performance of the **50-ohm** balun shown in fig. 1. Balance expressed as a percentage is shown. This design was optimized for 80 through 10 meters.

F_0 (MHz)	Z (ohms)	θ degrees	VSWR	balance (per cent)
3.5	48	16	1.33	2.8
4.0	49	14	1.28	2.1
7.0	50	10	1.19	1.3
14.0	50	8	1.15	2.5
21.0	51	8	1.15	4.2
28.0	52	9	1.18	1.3
30.0	53	9	1.18	1.3

Efficiency was better than 95 per cent over the useful bandwidth shown in **table** 2.

comparison with commercial products

Just how good are these balun transformers regarding match and balance? The best way to answer this question is to compare them with popular, commercially available products. The devices described here were compared with a commercial ferrite rod core 1:1 balun and a commercial toroid-wound 1:4 balun transformer. Performance comparisons are summarized in **tables 7** and **8. Table 7** shows the comparison between the commercial 1:1 ferrite-core balun and the corless balun of **fig. 1. Table 8** compares the performance of the commercial 1:4 toroid balun transformer with the two-stage 501200 ohm balun transformer shown in **fig. 4.** On the average, the VSWR and balance are

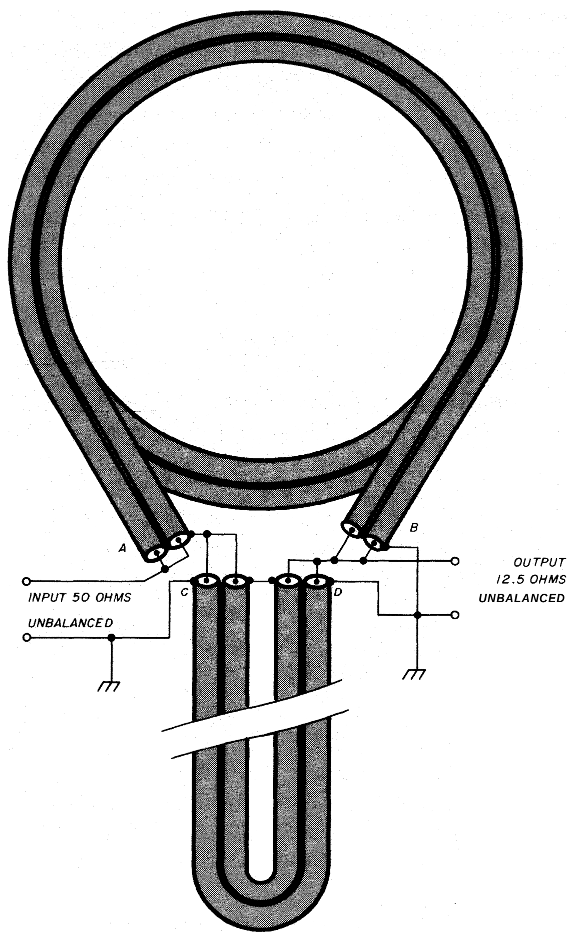


fig. 5. This **50/12.5-ohm** unbalanced-to-unbalanced transformer consists of two 127-cm **(50-inch)** parallel pairs of **RG-58A/U** coaxial cable connected in series at the input and in parallel at the output. Line CD must be the same length as line AB and should not be coupled to AB. Data comparing the coupled and uncoupled cases are shown in table 6.

table 2. Comparison of two baluns optimized for the low bands. Balun A was made entirely of coax as shown in fig. 1 but with 254-cm (100-inch) lines. Balun B is identical except for the dummy compensation line, which was made with an equivalent length of insulated no. 12 (2.1-mm) wire. These baluns are optimized for the low bands, so performance is good on 160 meters and poor an 10 meters.

	balun A				balun B			
F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)
1.8	53	10	1.20	3.8	51	6	1.11	1.90
2.0	53	9	1 18	4.7	51	5	1.09	1.40
3.5	53	4	1.10	4.5	51	2	1.04	.68
4.0	53	3	1.08	4.5	51	1	1.03	.67
7.0	53	0	1.06	5.1	49	1	1.05	2.00
14.0	53	1	1.02	6.1	46	5	1.13	4.00

better for the devices described here than for the commercial balun transformers tested. I evaluated only two commercial balun products, which were selected at random.

balun performance with varying loads

All the test data were taken with terminations for which the balun transformers were designed. In the real world, balun transformers are connected to antennas. Antennas are rarely ideally matched; as operating frequency is changed across the band, both resistive and reactive components of the antenna impedance change. It is therefore important to understand the influence of the balun when terminated with other than the characteristic impedance of the line and balun.

I tested the coreless balun of **fig. 1** and a commercial 1:1 balun at **3.5** and 14 MHz with loads varying from 16 to 150 ohms. These measurements are summarized in table **9.** Impedance magnitude, phase angle, and calculated VSWR of the loads are listed. Measurements taken through the balun of **fig. 1** and the commerical balun are also recorded. Both resistive and reacitve components of the impedance looking through the baluns varied widely from the data taken on the loads alone. In general, however, the resulting VSWR was not significantly altered.

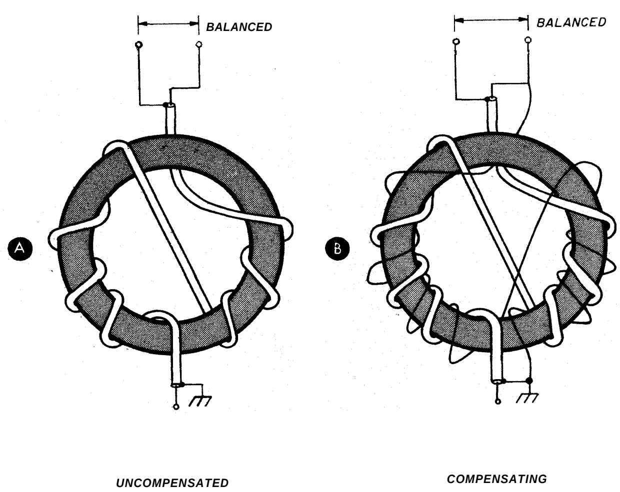


fig. 6. Compensating winding added to W1JR balun* for improved performance. A length of no. 16 (1.3-mm) Teflon insulated wire equal to the coax line, wound and connected as shown, improves balance and VSWR of the uncompensated balun.

table 3. Performance of the vhf balun shown in fig. 2.

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR
21	60	10	1.29
28	60	5	1.22
30	60	15	1.22
50	53	−1	1.06
56	52	−1	1.05
70	48	3	1.07
80	49	6	1.11
90	50	8	1.15
100	54	12	1.25

balun rf distortion measurements

Saturation effects in magnetic-core materials in balun transformers may contribute to nonlinearity and cause generation of harmonics with attendant TVI problems. However, to my knowledge, this problem has not been addressed in the literature and no measurements have been made to lend experimental validity to these concerns. For this reason a popular commercially available rod magnetic core 1:1 balun was measured for nonlinearity at a power level of 2 kW PEP.

The two-tone test method[1] offers a convenient means for measuring harmonic distortion. It's the method commonly used for determining the linearity of power tubes and solid-state devices. If two rf signals are linearly combined and are equal in amplitude, the resultant envelope varies periodically from zero to maximum. When a two-tone rf signal is passed through a nonlinear device, many new signals are produced, including harmonics and products resulting from harmonics and the original signals. Products that fall near the original signals in frequency are known as odd-order products (3rd, 5th, 7th, 9th, 11th). The measurement of the amplitude of these products with respect to the amplitude of one of the original signals is an excellent method for evaluating the harmonic distortion products generated by a nonlinear device.

The two-tone method was used to measure the harmonic distortion contribution of the commercial ferrite balun. In this experiment, the two rf signal sources were 2000 Hz apart at 2.001 and 2.003 MHz. The signals were combined and amplified to 2 kW PEP and fed through the balun to the load. The distortion products were measured with a modified HP310A Wave Analyzer. Power output at the 50-ohm load was measured with an HP3400A rms Volt-Meter. **Table 10** summarizes the results of the measurements. Note the 3rd-order distortion product increased from 43 to 39 dB below one of the two original signals, a 4-dB deterioration. Under the set of power-amplifier operating conditions chosen, the 5th- and 7th-order products decreased, and the 9th-

table 4. Performance characteristics of the two-stage **50/12.5-ohm** transformer of fig. 3 consisting of the balun of fig. 1 combined with **the 4:1** transformer of fig. 5 (Part **1**).

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)
3.5	53	22	1.49	3.5
4.0	53	20	1.44	2.7
7.0	56	9	1.21	2.1
14.0	55	−1	1.10	3.3
21.0	47	−1	1.07	0.0
28.0	45	10	1.23	4.3
30.0	47	12	1.25	6.5

order distortion product again increased.

It's clear from these measurements that you can't assume that a magnetic-core device, such as a ferrite core balun, is perfectly linear at all power levels. Unless flux density is held below the saturation threshold for the core material used, magnetic-core baluns and transformers can affect the linearity of your equipment and may cause TVI through the generation of harmonics.

W1JR balun improvement

Joe Reisert, W1JR, made an excellent contribution to the state of the art in his article, "Simple and Efficient Broadband Balun," in the September, 1978, issue of *ham radio*.[2] An improvement in the balance of the WIJR balun can be made by the very simple addition of a length of insulated hookup wire wound on the toroid as a continuation of the coax winding.

*Also proposed by K4KJ and discussed in the February, 1980, issue of *ham radio*, page 28.

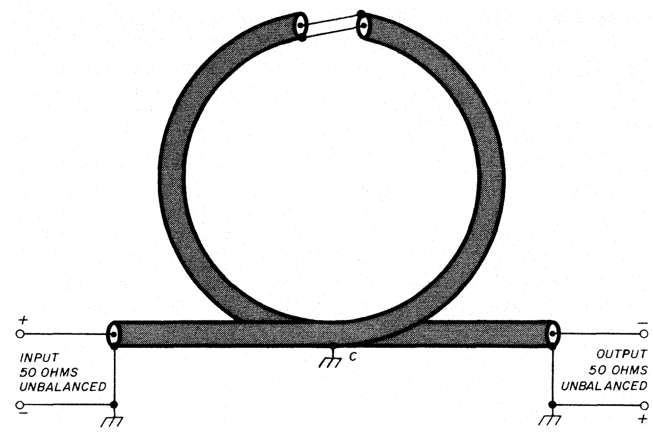


fig. 7. Phase inverter based on the same principles as the **coreless** balun. This useful coaxial line component changes the phase of an rf signal applied at the **50-ohm** input terminal by 180°; the phase reversal is produced by the cross connections between the two coaxial lines at A-B. Connections at **A-B** are isolated from ground by the self-resonance of the coiled coax lines. Construction. dimensions, and connections are the same as the **coreless** balun shown in fig. 1. From the common point C (system ground) to the output terminals, coax lines A and B are each 127 cm (**50** inches) long. The lines are wound into a seven-turn random wound **coil of** 11.5 cm (**4½** inches) nominal diameter. For clarity, only one turn is shown. Performance data is shown in table 12.

See **fig. 6** and the photo. The length of the compensating winding must, of course, be equal to the coax length. This modification was made at the suggestion of Ray Rinaudo, W6ZO.* It's based on the principles described in Part 1, showing how the length of coiled coaxial line of **fig. 2** (Part 1) is evolved into the compensated balun of **fig.** 3 (Part 1).

Data showing VSWR and inherent balance of the compensated and uncompensated baluns are shown in **table 11..** The balance measurement was made by terminating the balun with 50 ohms, driving at the frequencies shown, and measuring the voltage with respect to ground (enclosure) at each of the output terminals. Note the very significant variations in balance shown for the uncompensated balun, compared with the reasonably good inherent balance shown in the right-hand column. Balance is defined

table 5. Performance characteristics of the two-stage **50/200-ohm** transformer of fig. 4 consisting of the balun of fig. 1 combined with **the 4:1** transformer of fig. **6 (Part** 1).

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)
3.5	60	25	1.63	1.3
4.0	60	25	1.63	0.6
7.0	60	3	1.21	0.6
14.0	48	0	1.04	0.6
21.0	51	10	1.19	0.0
28.0	60	2	1.20	3.3
30.0	60	−1	1.20	3.3

table 6. Performance of 50112.5-ohm unbalanced-to-unbalanced transformer of fig. 5. The two right-hand columns compare VSWR of coupled and uncoupled configurations as explained in the text. VSWR on the 80-meter band can be improved by increasing the length of the coax lines as explained in the design procedure.

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR (uncoupled)	VSWR (coupled)
3.5	49	20	1.4	1.6
4.0	50	19	1.4	1.6
7.0	54	14	1.3	1.5
14.0	58	11	1.3	1.5
21.0	61	6	1.2	1.6
28.0	55	4	1.1	1.4
30.0	53	4	1.1	1.4

as the difference between the rf voltage readings at each of the two output terminals to ground (enclosure) divided by the sum of the two readings, expressed as a percentage.

Fig. 7 shows how to build a useful component for reversing the phase of an rf signal in a coaxial line. This phase inverter is useful for coaxial-fed W8JK antennas and other close-spaced phased arrays. The phase reversal takes place at the cross connection of two coax lines at terminals **A** and B. Terminals **A** and B are isolated from ground by the self-resonance principles described last month **(fig. 3)** and in **fig. 12.** Terminals **A** and B are not shorted by the grounded common connection between the coax outer conductors at C because of the high impedance over the outer conductors of the coiled coax lines. Of course, 180-degree phase shift can be accomplished in coax with a half-wavelength line; phase shift by this method, however, depends on frequency. The simple device shown in **fig. 7** inverts phase by 180 degrees independent of frequency; it inverts rf phase by exactly 180 degrees, with respect to equivalent length of coaxial cable, over a very broad band of frequencies. Measured broadband VSWR performance of the phase inverter is shown in **table 12.**

Phase inverters optimized for other frequency ranges may be designed according to the systematic design procedure for balun transformers detailed at the conclusion of this article.

summary of results

Of the various coreless rf devices made during the project, eleven are described in Parts 1 and 2 of this article. For convenience, they are summarized in Table 13, which correlates the construction of each device with measured performance data.

The transformers described in this article were, for the most part, designed with a combination of intuition and practical experience with coax baluns. However, as the project evolved, I gathered information that can be organized into a systematic design procedure. For example, N6AIG suggested a method of analysis starting with diagramming all of the possible ways to connect the ends of two or more coaxial cables.

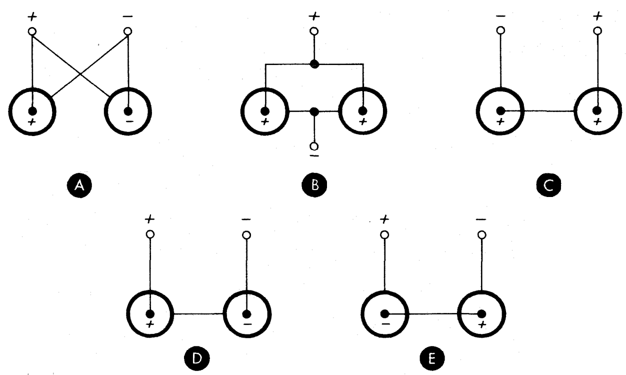


fig. 8. Various connections for pairs of coaxial lines. Polarity is arbitrarily assigned to the terminals, and the resulting polarity of the center conductors with respect to the outer conductors is indicated.

The diagram of **fig. 8** shows most of the connections possible with a pair of lines, and **fig. 9** shows some of the combinations for four lines. Similar diagrams can, of course, be drawn for any number of lines. Polarities are then assigned to the network terminals.

Next, assign polarities to each of the coax center conductors with respect to the outer conductors. Ex-

table 7. Performance of the 50-ohm coreless balun shown in fig. 1 compared with a commercial ferrite-core balun. VSWR was calculated from the impedance magnitude and phase data and is referred to 50 ohms.

	coreless balun				ferrite-core commercial balun			
F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)
3.5	48	16	1.33	2.8	49	11	1.21	11.8
4.0	49	14	1.28	2.1	49	9	1.17	12.0
7.0	50	10	1.19	1.3	50	9	1.17	11.6
14.0	50	8	1.15	2.5	55	11	1.24	7.9
21.0	51	8	1.15	4.2	63	12	1.37	1.4
28.0	52	9	1.18	1.3	72	5	1.46	3.9
30.0	53	9	1.18	1.3	75	8	1.54	1.6

table 8. Performance characteristics of a popular commercial 1:4 toroid balun transformer compared with the two-stage 50/200-ohm balun transformer of fig. 4.

	coreless balun transformer				commercial toroid balun transformer			
F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)	Z (ohms)	θ (degrees)	VSWR	balance (per cent)
3.5	60	25	1.63	1.3	53	6	1.12	1.8
7.0	60	3	1.21	0.6	53	8	1.16	2.5
14.0	48	0	1.04	0.6	54	16	1.34	12.0
21.0	51	10	1.19	0.0	57	27	1.66	18.0
30.0	60	– 1	1.20	3.3	69	44	2.53	21.0

amples of these assignments are shown in **figs.** 8 and **9.**

Now make a table similar to **table 14.** This table will be an aid in analyzing each of the possible end-connection combinations. Construct the table by choosing the input and output connections you want for your application, taking into account balance/unbalance and impedance. Show these connections in columns 1 and 2.

Next, determine whether there is a polarity match. The input connection must be compatible with the output connection. This is determined by inspecting the polarities assigned to the inner conductors. For example, **A** cannot be matched with B in **fig.** 8 because **A** has one + and one – center conductor polarity, whereas B has two + polarities. **A** and D are compatible because both have + and – polarities. Indicate whether there is a polarity match in column 3.

Whether the connection is balanced or unbalanced can be determined by inspection; this information is entered in columns 4 and 5. For example, **A** is balanced and B is unbalanced. An unbalanced connection can be converted to a balanced connection by the addition of one or more compensating lines. For example, unbalanced connection F or **fig.** 8 may be converted to a balanced configuration by connecting the outside conductor of a dummy length of coax to the positive terminal. Wind the coax as a continuation of the line connected to the negative terminal.

Part 1 explained how the compensating winding creates balanced terminals by showing how the isolated terminals of **fig.** 2 (Part 1) evolve into the balanced terminals of **fig.** 3 (Part 1).

Input and output impedances of transformers made with 50-ohm lines are shown in **table 14,** columns 6 and 7. For example, for connections **A-D** (third line in **table 14,** the input impedance is 25 ohms, because two 50-ohm lines are connected in parallel at the input. The output impedance is 100 ohms, because the two 50-ohm lines connected in series at the output are properly terminated with 100 ohms.

The transformation ratio (column 8) is simply determined from columns 6 and 7. If the transformer in this case had been made with 75-ohm line, the input impedance would be 37.5 ohms, and the output

table 9. This table compares the performance of the corless balun of fig. 1 with that of a typical 1:1 commercial ferrite core balun with varying loads. VSWR is calculated with respect to 50 ohms from the impedance magnitude and phase-angle data.

	load			coreless balun			commercial ferrite balun		
F_0 (MHz)	R (ohms)	0 (degrees)	VSWR	R (ohms)	0 (degrees)	VSWR	R (ohms)	0 (degrees)	VSWR
3.5	16.0	7	3.1	22.0	34	2.9	17	25	3.2
3.5	20.0	4	2.5	25.0	28	2.4	22	20	2.5
3.5	25.0	3	2.0	28.5	25	2.1	26	18	2.1
3.5	33.0	2	1.5	37.0	21	1.6	35	15	1.6
3.5	50.0	1	1.0	54.0	20	1.3	51	13	1.3
3.5	75.0	0	1.5	77.0	20	1.8	75	14	1.6
3.5	100.0	0	2.0	98.0	25	2.0	97	25	2.3
3.5	125.0	– 1	2.5	127.0	24	2.9	122	18	2.6
3.5	150.0	– 1	3.0	140.0	31	3.4	144	20	3.1
14.0	16.5	22	3.3	53.0	60	3.7	33	60	4.1
14.0	21.0	20	2.6	54.0	48	2.6	36	51	3.0
14.0	25.0	18	2.2	54.0	34	1.9	38	46	2.6
14.0	32.0	14	1.7	57.0	30	1.8	43	35	2.0
14.0	51.0	7	1.1	56.0	2	1.1	61	15	1.4
14.0	75.0	5	1.5	62.0	– 20	1.5	80	0	1.6
14.0	100.0	3	2.0	62.0	– 36	2.0	100	– 13	2.1
14.0	125.0	1	2.5	65.0	– 46	2.6	117	– 21	2.6
14.0	150.0	1	3.0	66.0	– 50	2.9	126	– 26	2.9

table 10. Summary of the distortion contribution of a typical commercial ferrite core balun at 2 **kW** PEP. The linearity of a high-power linear amplifier was measured with and without the balun connected between the amplifier and the load.

	odd order products				
	3rd	5th	7th	9th	**11th**
distortion products, amplifier without balun (dB)	43	43	52	63	60
distortion products, amplifier with balun (dB)	39	48	56	60	60

impedance would be 150 ohms. Incidentally, the transformer of this example **(A-D, table** 14) is the same configuration as that shown in **fig.** 4 of Part 1 except that input and output connections are reversed. All the possible connections are not listed in **table** 14, as indicated by the dashed lines. Completing the list of possible combinations is left as a challenge to the reader.

The next step is to determine how the lines are to be coiled and to determine the coupling between the coils. Draw a schematic diagram similar to **fig.** 1 of Part 1. A sample drawing of the example **A-D** in **table** 14 is shown in **fig.** 10. For analysis, assign an arbitrary input rf voltage of 100 volts. In this case, the input is balanced, so a balanced input voltage of ±50 volts is assigned. A total of 100 volts is applied to each line, so the output voltage is 200 volts (± 100 volts with respect to ground). Note that 50 volts appears along the outside conductor across the length of each of the lines.

Determine the magnitude and polarity of the voltage along the outer conductors by tracing the applied voltage. Current will tend to flow over the outside of the outer conductors in the direction shown by the arrows. Sufficient impedance must be provided to prevent shorting the applied voltage. Another example of this technique for analysis is shown in the schematic of the unbalanced-to-unbalanced 50/200-ohm transformer of **fig.** 11.

The required common-mode impedance is provided by coiling the coaxial lines. The lines of **fig.** 10 may be coiled together (closely coupled), because the voltage across the two lines is the same. The direction of current flow dictates that the lines must be coiled together as shown in **fig. 4** of Part 1. If the lines are coiled in opposite directions so that the current flow, as indicated by the arrows, is in the same direction, the coils will have positive mutual coupling and maximum common-mode impedance (input/output isolation). This is the reason the lines are coiled together into a continuous winding as shown, for example, in **fig.** 4 of Part 1.

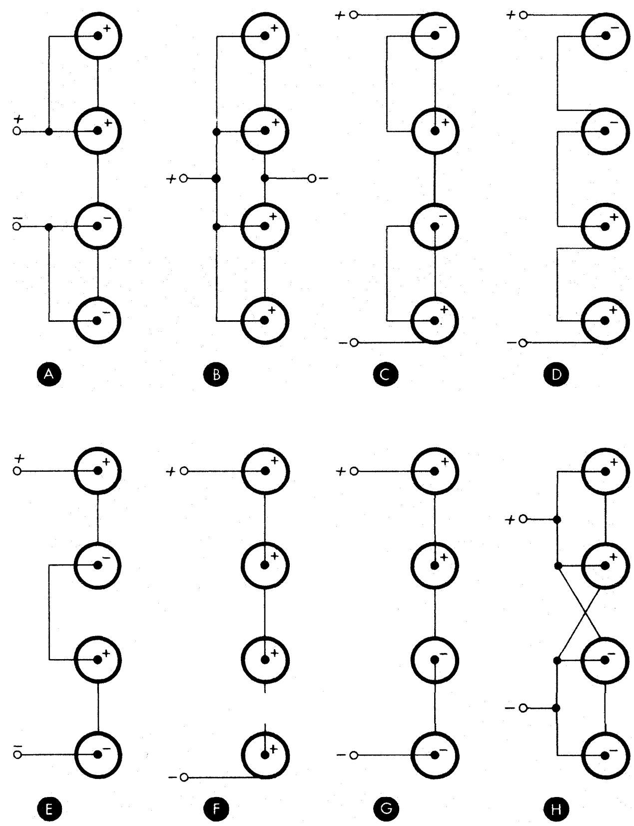


fig. 9. Some of the many possible connections for four coaxial lines. Polarity of the network terminals as well as the polarity of the coax center conductors is shown.

table 11. Comparison of WIJR **Balun**[2] with and without compensating winding. Additional winding of insulated hookup wire on the toroid shown in the photograph and the drawing (fig. **6B**) substantially improves balance as shown below. Because of lead length, load **VSWR** is high at 14 MHz and above. See load data below.

load				WIJR **balun,**[2] uncompensated				**W1JR balun,**[2] **compensated**			
F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR	Z (ohms)	θ (degrees)	**VSWR**	balance (per cent)	Z (ohms)	0 (degrees)	**VSWR**	balance (per cent)
1.8	49	1	1.03	48	1	1.05	99.0	49	7	1.13	1.5
3.5	49	2	1.04	49	4	1.08	93.0	50	6	1.11	1.5
4.0	49	3	1.06	49	4	1.08	92.0	50	7	1.13	1.5
7.0	49	5	1.09	51	8	1.15	71.0	52	8	1.16	0.0
14.0	49	10	1.19	63	11	1.35	3.7	61	8	1.28	0.0
21.0	50	15	1.30	76	4	1.35	38.0	68	0	1.36	7.1
28.0	52	19	1.41	79	−12	1.66	47.0	51	−11	1.21	15.0
30.0	52	21	1.46	67	−17	1.53	48.0	36	−1	1.39	39.0

Depending on the choice of the many input/output configurations partially listed in figs. *8* and **9,** the coaxial lines may or may not have current flow similar to that in example **A-D.** In some configurations, the direction of current flow on two lines is the same, in which case the lines must be wound together in the same direction. In other configurations, the voltage drops are not the same, so the coils should not be tightly coupled. The configuration shown in fig. **11** is an example. In a number of unbalanced-to-unbalanced configurations, the voltage drop in one or more of the lines is zero (zero current flow). These lines should not be coupled with lines having voltage drop, because lines with zero voltage drop act like shorted turns. The unbalanced-to-unbalanced transformer of fig. 5 is an example of this.

The outer conductor of coax **CD** (fig. **5**) has the same potential at both ends. It should not, therefore, be coupled to the coiled length **AB.** The effect of trying to couple incompatible coils together is shown in the two right-hand columns of data in table **6.**

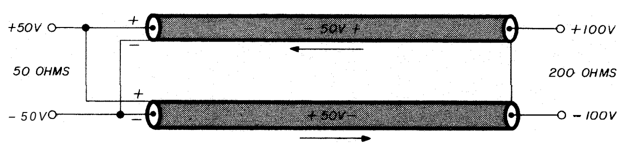


fig. 10. Schematic of the example A-D table 13 and fig. 4 of Part 1. A balanced input voltage of 100 volts (±50 volts) is assigned for analysis. 50 volts appears across the outer conductor along the length of each line, causing current to flow in the direction of the arrows. Enough impedance must be provided along the outside of the lines to prevent shorting the applied voltage.

Fig. **9** shows some of the many possible input and output connections for a group of four coax lines. Terminal and coax center-conductor polarities are indicated in the same format as in fig. *8.* A number of the four-line connections are shown in table **15** to serve as examples. Transformers consisting of any number of lines may be analyzed in this way. The highest transformation ratio depends on the number of coax lines. Table **14** shows that two lines can achieve a transformation of four, and table **15** shows that four lines can achieve transformation ratios up to sixteen. The highest transformation ratio available is equal to the square of the number of lines.3 The length of all lines should be the same, to preserve phase relationships.

The final step in the design procedure is to determine the optimum length of coax line and the number of turns in the coil. If the coil has too few turns, performance will be poor at low frequencies; if it has too many turns, performance will be poor at high frequencies. As an example, in the balun described in

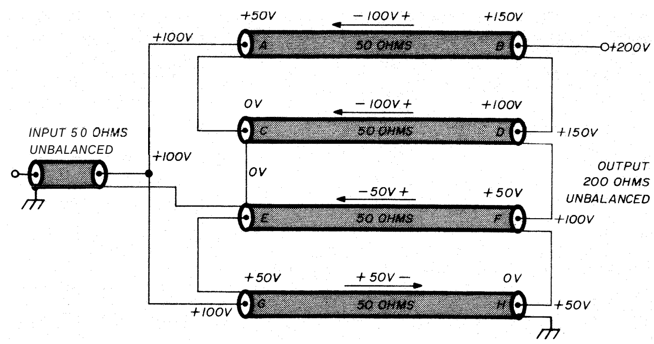


fig. 11. Another example similar to fig. 10 showing how to analyze a transformer configuration for direction and magnitude of voltage drops over the outside conductors of the coax lines. Lines AB and CD should be wound together in parallel and in the same direction for positive mutual coupling, like coiled line AB in fig. *5.* Lines EF and GH should be wound together in opposite directions for the same reason.

fig. **1,** each RG-58A/U line is 127 cm (50 inches) long. The lines were random wound into a 11.5-cm (4.5-inch) nominal diameter coil. VSWR performance over the useful frequency range is shown in table **1**. Note the increase in VSWR at the ends of the frequency range and compare this with fig. **12.**

Fig. **12** shows the impedance across the output terminals of the balun of fig. **1** plotted as a function of frequency. When the balun is open circuited; that is, when the center conductor at the output is disconnected from the dummy coaxial line, the self-impedance of the coax coiled outer conductor can be measured. The vector impedance meter was connected from point **A** to point **B** (fig. **12**), and the impedance magnitude was measured between 1 and 70 MHz. Note that the impedance from **A** to **B** is always greater than 50 ohms, the line surge impedance, over the useful frequency range of the balun (table **1**).

When designing a balun or transformer to your needs, make the coil self-resonant frequency approximately equal to the average of the upper and lower frequency limits of the band of interest. For example, if you want to design a transformer to cover 3.5-30 MHz, the open-circuited coil self-resonant frequency should be about 16 MHz. If you want your balun/transformer to be optimum for 160, 80, and 40 me-

table 12. Broadband VSWR performance of the 50-ohm to 50-ohm coax phase inverter shown in fig. 7.

F_0 (MHz)	Z (ohms)	θ (degrees)	VSWR
3.5	54	16	1.34
4.0	54	14	1.30
7.0	57	4	1.16
14.0	53	– 3	1.08
21.0	48	– 2	1.06
28.0	46	4	1.11
30.0	46	7	1.16

Table 13. Tabular summary of the balun transformers built and measured by W6TC.

input/output impedance ohms	ratio	input/output balance	bandwidth MHz	construction	measured data
50/12.5	4:1	balancedlbalanced	1.8-30	fig. 5, part 1	table 1, part 1
50/200	1:4	balancedlbalanced	3.5-30	fig. 6, part 1	table 2A, part 1
50/200	1:4	balanced1balanced	1.8-14	fig. 6, part 1	table 2B, part 1
50/50	1:1	unbalanced1balanced	3.5-30	fig. 1, part 2	table 1, part 2
50/50	1:1	unbalancedlbalanced	1.8-14	fig. 1, part 2	table 2A, part 2
50/50	1:1	unbalanced! balanced	1.8-14	fig. 1, part 2	table 2B, part 2
50/50	1:1	unbalancedlbalanced	21-100	fig. 2, part2	table 3, part 2
50/12.5	4:1	unbalancedlbalanced	3.5-30	fig. 3, part 2	table 4, part 2
50/200	1:4	unbalancedlbalanced	3.5-30	fig. 4, part2	table 5, part 2
50/12.5	4:1	unbalancedlunbalanced	7.0-30	fig. 5, part 2	table 6, part 2
50/50	1:1 (inverted)	unbalancedlunbalanced	3.5-30	fig. 7, part 2	table 12, part 2

ters, make the resonant frequency about **4.5** MHz. For two or three adjacent band designs, the frequency isn't critical; ±10 per cent accuracy is sufficient.

I found that a grid-dip meter can be used to make this measurement. Make certain that the balun/transformer is open circuited at the output, as described above. The coil diameter should be 15-20 times the coax line diameter. Remember that these baluns and transformers make use of a broadly resonant circuit consisting of the distributed inductance and capacitance of the coax line outside surfaces. Accordingly, if the device is mounted on a metal plate or in a metal enclosure, the resonant frequency should be determined while in place. I tried extending the low-frequency limit of one of these transformers by adding lumped capacitance to the coiled coax at the output. (For example, between **A** and B in **fig. 12.)** This artificial loading does, in fact, lower the frequency of minimum VSWR; however, bandwidth is severly reduced.

design procedure summary

In summary, a systematic design procedure for coreless transformers and baluns is as follows:

1. Select input and output coaxial line connections (see examples in **figs. 8** and **9**).

2. Assign polarities.

3. Make certain the polarities of the input and output connections match (column 3, tables **14** and **15).**

4. Check input and output balance by inspection (columns **4** and **5).**

5. Check input and output impedance by inspection (column 6 and 7).

6. Determine transformation ratio (column 8)

7. Draw a schematic diagram similar to that in **figs. 10** and **11** for analysis.

8. Assign an input voltage, such as 100 volts.

9. Determine the polarity and magnitude of voltage across the length of the lines.

10. Determine the direction of current flow over the outer conductors of the lines.

11. Determine which lines must be coiled, if they can be coiled together, and the sense of the mutual coupling.

12. Select the length of line, coil diameter, and number of turns using a grid-dip meter to resonate the coil near the average between the upper and lower frequencies of the band of interest.

13. Make certain that the **rf** paths through all coax lines are equal, to preserve phase.

table 14. This table is a partial list of the coax connections shown in fig. 8. It is used as an aid in analyzing the connections.

column 1 connection (in)	2 connection (out)	3 polarity (match)	4 balance (in)	5 balance (out)	6 impedance in (ohms)	7 impedance out (ohms)	8 transformation ratio (in/out)
A	B	no					
A	C	no					
A	D	yes	balanced	balanced	25	100	1:4
B	C	yes	unbalanced	unbalanced	25	100	1:4
B	D	no					
—	—	—					
—	—	—					
—	—	—					

table 15. This table shows some of the connections for four coax lines shown in fig. 9. The impedances listed in columns 6 and 7 assume the use of 50-ohm lines.

column 1 connection (in)	2 connection (out)	3 polarity (match)	4 balance (in)	5 balance (out)	6 impedance (in) (ohms)	7 impedance (out) (ohms)	8 transformation (ratio) (in/out)
A	B	no					
B	C	no					
A	D	yes	balanced	balanced	50	200.0	1:4
F	B	yes	unbalanced	unbalanced	200	12.5	16:1
E	C	yes	balanced	balanced	200	50.0	4:1
B	C	no					

advantages of coreless balun transformers

The advantages of this new class of broadband coaxial line transformers over magnetic-core transformers are as follows:

1. They are inexpensive.

2. They are linear; there are no materials in the system that can saturate.

3. They use readily available materials: only coax and hookup wire.

4. They are lightweight and compact.

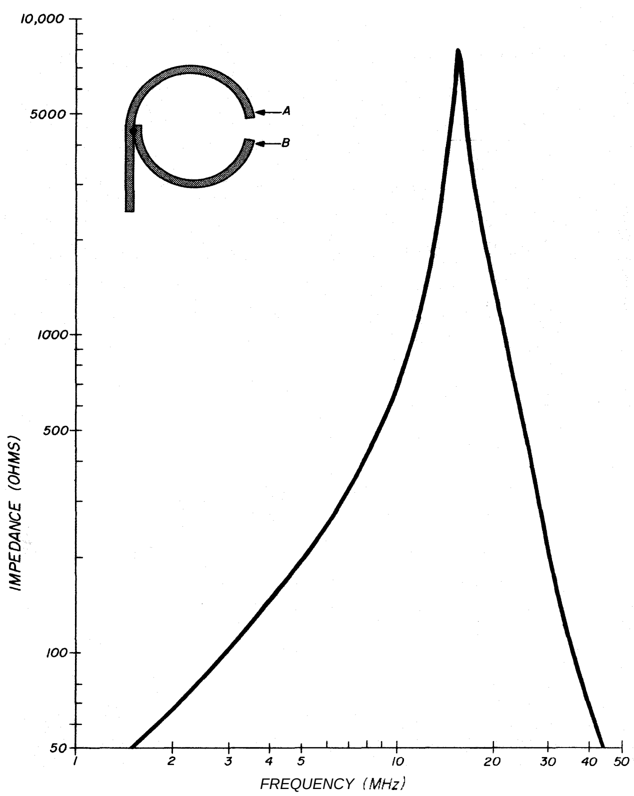


fig. 12. Impedance from point A to point B of open-circuited balun. Impedance at AB is greater than 50 ohms, the line surge impedance over the useful frequency range of the balun.

5. They are weatherproof because of the materiais; an enclosure is not required.

6. They have low VSWR.

7. They are inherently balanced.

8. They have high power-handling capability limited only by the coaxial line chosen. When made with Teflon coax, they are virtually indestructible in Amateur service.

9. There are no closely spaced or tightly twisted enameled wires and no ferrite or powdered iron core materials that can result in arcing.

conclusion

The purpose of this article is to show how high-performance balun transformers can be built free of the disadvantages of magnetic core materials. I hope I've presented enough data on this new class of devices for you to be able to reproduce one or more of the designs described here, or to design one of your own to meet your requirements. My goal has been to provide enough information for others to be able to reproduce these useful balun transformers, even though they may not have access to fine instruments such as the Hewlett Packard vector impedance bridge, rf voltmeter, or programmable calculator.

acknowledgment

I am indebted to the EIMAC gang (the laboratory staff at EIMAC) and the staff at CTC for counsel, constant encouragement, and after-hours use of their laboratory facilities.

Note: The HP-67 program for calculating VSWR that should have appeared in the appendix of part 1 of W6TC's article (February, 1980, *ham radio*) can be found on page 70 of this issue.

references

1. *Care and Feeding of Power Grid Tubes,* EIMAC Division of Varian, San Carlos, California, 1967.
2. Joe Reisert, W1JR, "Simple and Efficient Broadband Balun," *ham radio,* Septermber, 1978, page 12.
3. Ruthroff, "Some Broad-Band Transformers," *Proceedings of the IRE,* August, 1959, pages 1337-1342.

ham radio

considerations regarding microphones and simple speech processing

A look at simple homemade microphones and speech processors

This article describes a microphone stand that can be built easily and that's much more convenient to use than the typical commercial unit. Also described are simple preamplifier and clipper circuits that can be added to a phone station between microphone and transmitter.

improved desk microphone

In spite of all the equipment manufactured for sale to Amateurs, many desirable items still can't be readily purchased. Many times these items are simple to build, and many times the item needed is a simplified version of what's commercially available. In any case, it's seldom that the scratch-built item isn't a big cost saver.

In the case that prompted this article, the audio gain in my low-band rig was marginal. Close talking in a moderate voice into a standard crystal microphone was required for full SSB output. This condition may not be unusual based on my own experience and that of others I've talked to. Additionally, standard microphone stands have always left a lot to be desired, to my way of thinking. First, they're seldom adjustable in height; second, they must be placed off to the side if you want to take notes or fill in your log while talking. One of my friends claims the best he can do is get his nose up to the bottom of the microphone; in my case, I have to bend over to speak into a microphone mounted on a typical commercial stand. The solution to this problem is a boom-type microphone stand.

The mechanical end of this kind of project is wide open with respect to cost and complexity. If you have the shop equipment, the boom stand can be a major project for tools such as a lathe and drill press. It's largely a matter of the materials and tools you have and your personal taste."

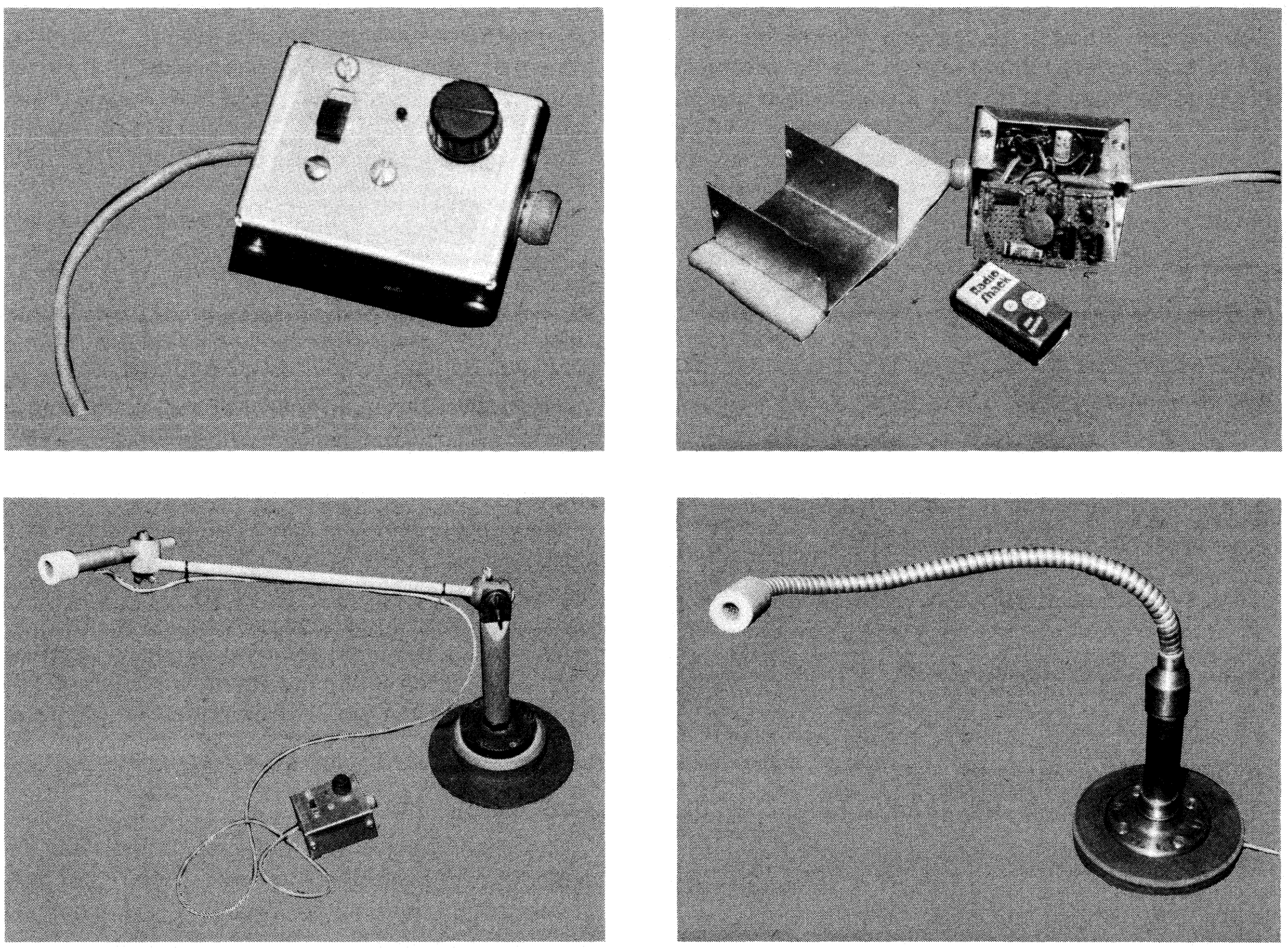
My original stand was made from junk-box parts and some pieces of birch dowel. It looks a bit "Tinker-Toyish," but it serves the purpose very well. The second design, which is shown in the diagram, requires no unusual tools and works better than the original.

building the microphone stand

The base is made from two or more layers of 6.5-mm (¼-inch) tempered Masonite (fig. 1). The base should be at least 153 mm (6 inches) in diameter and may be weighted if a heavy microphone is used. The upright section is a "plumber's delight" made from readily available plumbing fittings. The boom is a piece of Greenfield flexible tubing, which is available at electrical supply houses. This type of tubing is smooth and flexible. To stiffen it, use a piece of aluminum clothes line wire inside of it.

"Metal is recommended for audio and radio-frequency shielding.

By George A. Wilson, Jr. W1OLP, 318 Fisher Street, Walpole, Massachusetts 02081



Top left, panel view of the two stage **preamplifier/clipper** described in the text. The on-off switch is at the left, the LED pilot light is in the center, and the gain control is on the right. The gain setting in the preamplifier controls the amount of clipping, while the gain setting in the transmitter sets the maximum modulation (or audio drive) level. Preamplifier gain will decrease as the battery voltage decreases making it necessary to readjust the gain control. The control in the transmitter should require readjustments only after major tuning changes. Top ***right,*** preamplifier circuit built on a perfboard and contained inside a mini box. The circuit is loose-mounted by wrapping it in foam plastic (under the box cover in the picture). The plastic insulates it from terminals and connections, which are part of the box itself, and provides all the mechanical strength needed. The on-off switch is separate from the gain control so that the preamplifier can be turned on and off without changing the gain setting. Input-output connectors can be varied to suit the needs of any **transmitter/microphone** combination. Bottom left, the first boom microphone built by the author. This type of construction led to a very useful microphone but requires more tools to build than the design in the text. The boom arms and upright section are constructed from birch dowels. This approach doesn't provide the audio and radio-frequency shielding that all-metal construction does. However, it's cost effective, since the parts were all from the **junkbox.** Bottom ***right,*** the microphone design described in the text. It is made of readily available materials easily assembled in the home workshop. If you lack the tools to make a round base, use a square or six-sided base, which will work just as well. This design provides all-metal shielding of the microphone element and that part of the cable inside the stand. The Greenfield flexible tubing is stiffened with an internal piece of aluminum clothesline wire. This approach provides the ultimate in adjustability and convenience.

pilot light circuit

The LED pilot light in the two-stage circuit is of special interest. If the circuit is left on when not using the rig, the battery will last only a few days, depending on how new it is. Having learned the hard way, I realized that a pilot light with minimum power requirement was needed. Several power-saving techniques were rejected before a friend suggested connecting the pilot light in series with the amplifier. Neat! A pilot light that actually causes a slight *de*crease in the current drain! It's not too bright, but it's adequate for the purpose. (If used with the single-stage preamplifier circuit, it may be necessary to add a resistor* to ground after the **LED.** This will cause more current to follow through the LED — at least 2 mA is required for most LEDs.)

Another solution to the power problem is an external source powered from the ac line. If the preamplifier is to be built into the rig there's little problem, even if it is a tube rig. In the latter case, a voltage source, such as the well by-passed cathode bias on an amplifier stage, can be used. The circuit will operate over a range of at least 7-15 volts. External supplies intended for calculators and similar devices can be used, or a miniature regulated supply can be built into the amplifier.

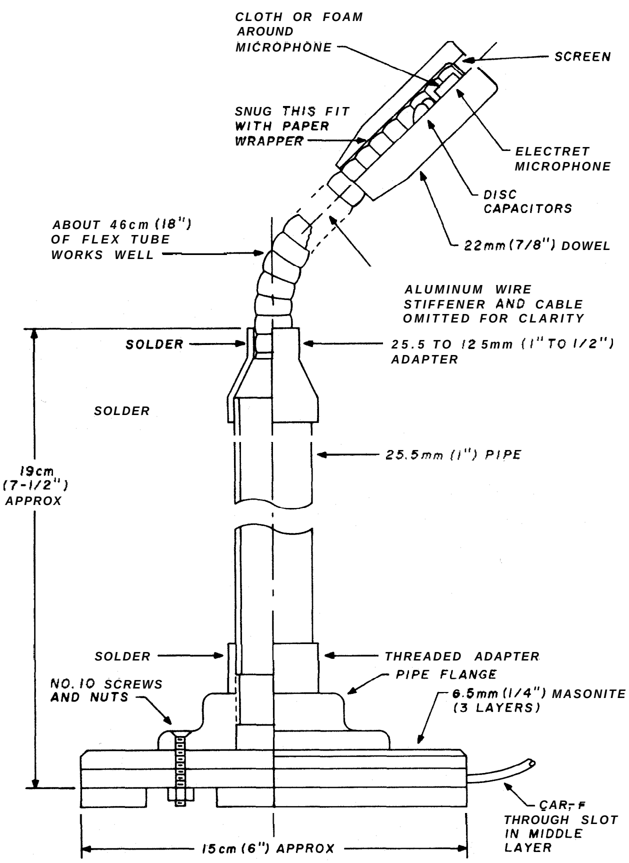


fig. 1. Mechanical layout of the desk microphone described in the text. The construction shown is for guidance only. Most builders will substitute materials on hand and their own techniques of fastening things together. A push-to-talk switch can be mounted on the base if desired.

adapting the Electret cartridge

If an Electret microphone is used, it can be easily adapted to the boom using a short piece of 22-mm (718-inch) birch dowel. Drill the dowel to fit the tubing outside diameter (16 mm, or 5/8 inch) but do not go quite all the way through. Turn the dowel around and drill through from the other end with a 12.5-mm (%-inch) diameter drill. Slip in a piece of screening from the 16-mm (518-inch) end. Sand the dowel, round the edges to suit yourself, and slip the dowel over the end of the Greenfield tubing. Use a piece of 77 x 128 mm (3 x 5 inch) card stock to "snug" the fit if necessary.

Slip the microphone into the end of the Greenfield tubing and make it snug with a layer or two of cloth, sponge rubber, or plastic foam. Make the connections before putting the microphone element into place. Don't neglect a ground connection to tubing at the microphone end!

I've received many on-the-air compliments on the Electret microphone, which compensate for its low output and the power requirement for its internal amplifier.

Other microphone elements can be adapted in a similar manner or with a little ingenuity. For example, the top of a plastic bottle can be cut off and used to house a larger microphone element. If your workmanship isn't the greatest, a "blast shield" can be purchased (Olson Radio has them). A shield of this type will cover the whole microphone housing and make things look quite professional.

plumbing details

The Greenfield tubing and pipe fittings can be soldered together using a propane torch, standard plumbing flux, and solder. Details of the plumbing lash-up are left to you, since the availability of junk-box material may determine your approach. The stand in the photos has a 128-mm (5-inch) length of 25.5-mm (I-inch) copper pipe with a reducing fitting to hold the Greenfield tubing. A 25.5-mm (I-inch) copper pipe threaded adapter was used to mount the pipe flange at the bottom.

All told, including a Radio Shack Electret microphone element, the microphone cost less than $10.00. Most of the cost was in the plumbing fittings. Not a bad price for an extra convenient microphone!

Unlike mikes, most commercially available speech processing equipment is complex and expensive. Many hams need speech processing for two reasons: a) lack of audio gain in the transmitter, and b) a need to limit the audio level to prevent overmodulation and consequent wide bandwidths. In the first case, we're short-changing ourselves by not using the transmitter's full capability. In the second case, we're causing inconvenience to others by splattering outside our allowed bandwidth. In fact, we're operating illegally in this case. Both of these problems can be resolved by building an outboard solid-state speech processor — or buying one.

*Try resistors in the 2-5 kilohm (¼-watt) range. The lower the resistance, the brighter the LED will be.

simple microphone amplifier

If lack of gain and/or power for an Electret microphone is your problem, the single-stage bipolar transistor amplifier shown in **fig.** 2 can be used. The circuit includes a regulated voltage source for an Electret microphone. The zener regulator output is for the microphone's internal amplifier. It reduces the 9-volt battery voltage to about 4 volts, a nominal supply voltage that will satisfy most microphones. Check your microphone and use an appropriate zener if the supply voltage should be higher or lower.

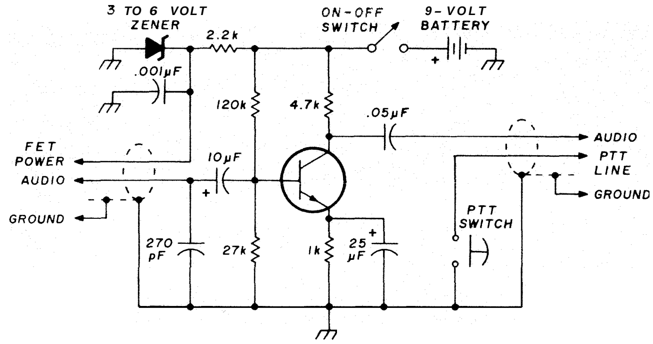
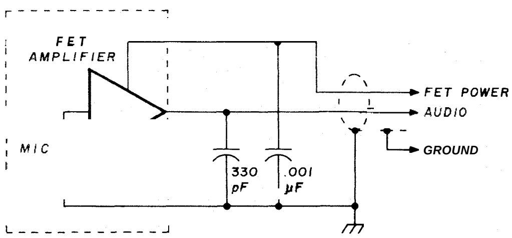


fig. 2. Simple amplifier for use with low-output microphones. A regulated supply for an Electret microphone is included. All resistors % watt. Transistor is **2N2925** or equivalent. Capacitors are disc ceramic except where polarity is marked, these are eiectroiytics. This circuit is not recommended for microphones with impedances higher than 50,000 ohms. The output is intended to match transmitters with high-impedance microphone inputs.



Typical Electret microphone circuit. Output impedances vary but are typically in the order of 1 kilohm. Disc ceramic capacitors tend to short out rf that may be picked up. This type of microphone is inexpensive and provides excellent speech quality. The fet amplifier is built in.

After pricing zeners for the 4-6 volt range, I bought a blister pack full for less than $2.00 and found four usable diodes among them. A test circuit can be made by connecting the diode in series with the 2.2-kilohm resistor specified in **fig. 2,** a voltmeter, and the 9-volt battery. Check the diode in both directions. It will read about 0.6 volt in one direction and the zener voltage (if less than 9 volts) in the other direction. Mark the diode polarity on the diode if it was not marked as received.

The transistor amplifier circuit shown is very tolerant of the transistor used. Most any medium- or high-gain (h_{fe} = 100-300) NPN transistor should do the trick. The 2N2925 is typical of this type and is usually in plentiful supply.

The circuit can be built on a small piece of perfboard. The 9-volt battery and circuit were wrapped in plastic foam to keep them from rattling about and were tucked into the minibox without rigid mounting. The capacitors across the microphone, and at all leads entering or leaving the minibox enclosure, were intended to minimize rf entering the circuit and causing problems.

two-stage speech processor

If you're looking for some gain and would like full audio drive without splattering, the circuit of **fig.** 3 is your answer. This circuit includes two stages of gain, a diode clipper circuit, and an RC filter that reduces

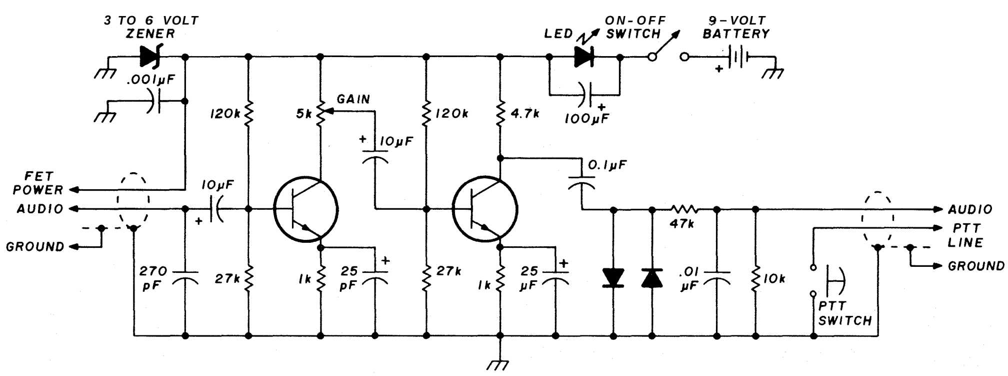


fig. 3. Two-stage **amplifier/clipper** circuit. A regulated supply for an Electret microphone is included. All resistors % watt. Transistors are **2N2925** or equivalent. Capacitors are disc ceramic except where polarity is marked; these are electrolytics. Diodes are **1N4149/1N914.** Input-output impedances are similar to those of the single-stage amplifier of fig. **2.**

the distortion introduced by the clipper circuit. It also includes a voltage source for an Electret microphone and a novel pilot-light circuit (described later). This circuit has enough gain to allow the signal to be clipped or flat-topped by the back-to-back diodes. They set a peak-to-peak audio voltage level of 0.6 volt maximum that can't be exceeded by a loud voice or high gain settings in the preamplifier. Because the diodes fix the output level, the transmitter gain control can be used to set a relatively fixed drive level that won't cause splatter.

The diode clipping distorts audio quality by introducing harmonics of the voice frequencies. These harmonics are reduced by the RC filter that foiiows the diodes. This circuit is far from a cure-all, however, and this clipper-filter circuit must be classified as a rudimentary speech processor.

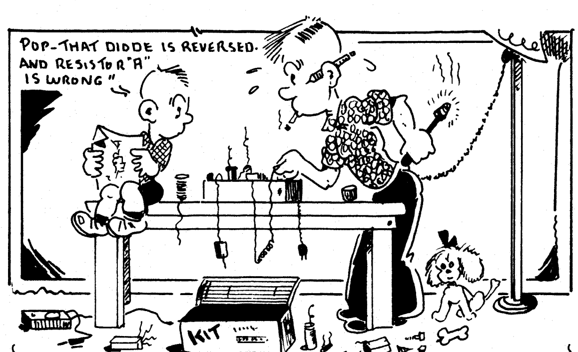
gain control

The potentiometer in the two-stage amplifier is a gain control and will set the amount of gain ahead of the diode clipper. If set high, the gain will be great. The diodes will clip a great deal, and the ambient background noise will modulate the rig with ease. This setting will also introduce more audio distortion. The peak output level is set by the diodes, and the transmitter modulation level is set by the gain control in the transmitter. Once this control is set for any tune-up condition, the transmitter peak drive level will not be exceeded, even under close-talking conditions.

Some experimentation will be necessary. It will be necessary to readjust the transmitter gain control when major changes in transmitter tuning are made. (This adjustment has to be made even without a speech processor.) In my opinion, a simple processor of the sort shown can be very useful but shouldn't be overworked to the point that your listeners complain about distorted voice quality and the dishes clattering in the kitchen.

On-the-air results with these circuits have been gratifying. Their results-to-simplicity ratio is high!

ham radio



logarithmic detector
with a post-injection marker generator

A specialized piece of test equipment that provides accurate frequency markers for crystal-filter alignment

The best way to align a crystal filter is to drive it with a swept signal generator that provides horizontal input to a scope, then detect the filter output and display it vertically. The detector should be logarithmic; that is, ***linear in*** *dB.* And it's helpful if accurate frequency markers are available on the display. This article describes such a box, which I suppose is properly called a "logarithmic detector with post-injection marker generator." That title seems a little long, so I call it the target for my sweeper.

This is a specialized piece of test gear, so many hams wouldn't consider building it; how many times does one align a crystal filter? The fact is, though, there isn't much in this thing in the way of parts, and it's easy to get running. I'd call it a "longweekender" project, and it sure is fun to play with.

detector speed and dynamic range

In any logarithmic detector there's a tradeoff between speed and dynamic range. Take a peek ahead at fig. 5, which shows our box giving a nice account of itself between –60 and –10 dBM. That data alone doesn't guarantee the performance we need because the data in fig. 5 is static; it doesn't show how long the detector takes to settle down when the input level is changed. To see why this is important, consider a typical test situation. We're running our sweeper at 30 Hz (to get a flicker-free display) and our filter is working great. On the display there's a single hill or blip having a width of, say, one-fifth of the screen (our sweep width is five filter bandwidths).

The scope vertical signal has a bandwidth of at least 150 Hz, probably more if our filter response has steep sides; figure it out yourself. This means that, if the displayed response is to have any meaningful relationship to the filter response, the detector must be fast, very fast as logarithmic detectors go. To get this speed we have to give up range; the 50-dB range of fig. 5 is poor in comparison with the 100-dB range available in instruments that can take their time about producing a reading.

performance test

Perhaps the best way to specify the performance of the target is to describe the simple test I apply whenever I use it. Required are a signal generator capable of being 50 per cent modulated at 400 Hz, a calibrated step attenuator, and a dc-coupled scope. (You don't have a dc-coupled scope? Neither do I, but I *do* have an electronic switch that chops up an

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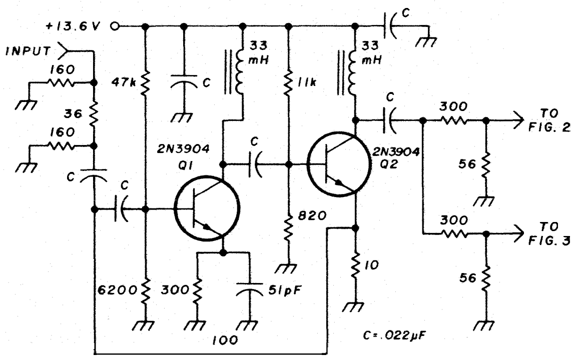

fig. 1. Schematic of the front end showing the input pad. preamplifier, and power divider. The two outputs deliver about 100 mV rms. Output circuit is designed to prevent interaction. The output resistors aren't critical; 330-ohm units could be used.

input to give the same effect; alternatively, you can connect a dc voltmeter along with the scope, although this is not as dramatic.)

Now, the peak-to-trough ratio of the envelope of a 50 per cent amplitude-modulated signal is exactly three; that is, the instantaneous power changes 10 dB over a modulation cycle. Hook up the stuff (using the scope internal sweep) and see what you see. The log of a sine wave isn't that much different from a sine wave, so you'll see a "sort of" sine wave of 400 Hz. Note the peak-to-peak size of the display and the average height.

Now crank in 10-dB attenuation. The shape of the display should be unchanged, because the instantaneous power is still varying 10 dB. The height of the display should have dropped exactly 10 dB; namely, the peak-to-peak size! Run this test over a range of input levels and frequencies.

I find the box works very well from 1-10 MHz and from −10 dBm input to −60 dBm (unmodulated value). At the bottom end of input levels, the display shrinks vertically, indicating loss of linearity, whereas at the highest levels, it first balloons then collapses because of saturation. The bandwidth is well above 400 Hz, because, if it weren't, the displayed waves would "lean" from vertical symmetry. I wish more of my test junk had such a simple "alive-and-well" test.

circuit description

Fig. 1 shows the front end. It's important, of course, that the target present a constant load to the filter. The feedback amplifier is designed for 50-ohm input, and a 6-dB pad makes extra sure.

Levels for operation are arbitrary. I chose −10 dBm as the maximum input because that's about what any old signal generator will produce. The amplifier was chosen so that, at this input level, the two outputs are about 100 mV rms, which is the maximum input level for the following devices.

The 300-ohm resistors keep the two outputs from interacting. All values in all diagrams are the ones I used, mostly because of the Mt. Everest principle (they were there). Obviously, 330 ohms would do as well, and the same goes for many other places where I used values available in 5 per cent tolerance.

Log detector. Fig. 2 shows the logarithmic detector, which uses the agc output from an LM373. The LM373 is connected in the a-m mode precisely as recommended by the manufacturer, with the exception of the 1-μF capacitor at pin 1. Normatty, a much higher value is used to prevent the agc from following audio; we ***want*** the agc to follow audio.

I experimented to find the smallest usable value for this cap; with lower values the agc loop is fast

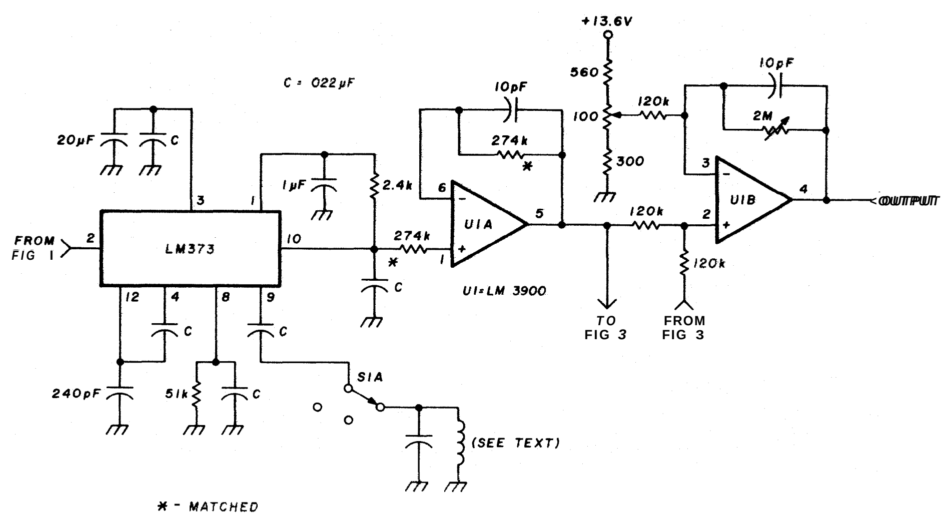

fig. 2. Log detector circuit. The 1-μF capacitor at pin 1 of the LM373 determines respond speed. Useful bandwidth is about 1 kHz.

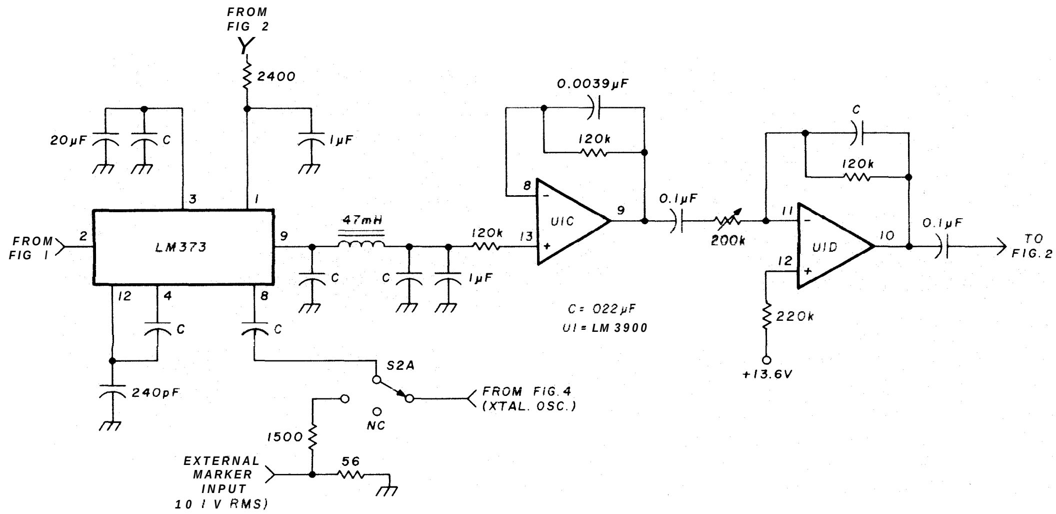

fig. 3. Post-injection marker generator. An LM373 is used in the product-detector mode with BFO input from an external source or crystal oscillator.

enough to lock onto noise and oscillate. This item determines the speed of the whole works. Adding 4 µF in parallel here gives a perceptible effect at 400 Hz, so I expect the useful bandwidth is about 1 kHz.

The tuned circuit at pin 9 could have been eliminated. I set up a bandswitch with circuits at 1.25-, 4.8-, and 10.6 MHz (where I had filters in the works), but I found little difference when I switched to an unconnected position left in reverse. A loss of about 6 dB occurred at the low end, and that's it. Of course, there's a huge difference if I switch to the *wrong* filter.

At about 100 mV rms input, the agc system loses control and the output soars; this is easily spotted and reminds you to cut down the input level.

Opamp U1A is an isolator. I took care to match its two resistors (by using a couple of 1 per cent resistors of the same value) to keep its output very close to the input for reasons discussed later.

Op amp U1B permits adjustment of the output range (my voltmeter has a 3-volt scale, so I selected the range shown on the right of **fig. 5**), and also accepts a "birdie" input from the marker generator. Each op amp has a 10-pF feedback cap to roll off unwanted high frequencies. The units of **figs. 1** and 2 are complete in themselves; the marker (discussed next) is optional.

marker generator

Fig. 3 shows the marker generator. A second LM373 operates in the product-detector mode with BFO input from an onboard crystal oscillator. The agc line of this LM373 is driven by the output of U1A (**fig.** 2) at precisely the level being used on the other LM373. This keeps the marker size independent of where it falls on the filter response curve — just a frill, but the line was available, so why not?

U1C provides a heavy lowpass filter to keep the marker narrow. I selected the 0.0039-µF cap by trial to give a width of about 300 Hz, my personal preference. U1D allows adjustment of the marker height by the 200-kilohm pot. Mine is adjusted for 1 volt p-p, or about 20 dB, which is another item of choice.

I never use markers and filter at the same time; I prefer to precalibrate the horizontal scale, then get the markers out of the way. It used to be that, with a rig like this, one would want several crystal markers available. But nowadays, with counters easy to come by, it seems simpler to connect an external generator and move it around while reading its position from a counter.

I did opt for one internal marker, which is handy for making sure I'm connecting the correct tuned circuit (S1 does both). Switch S2 selects internal, ex-

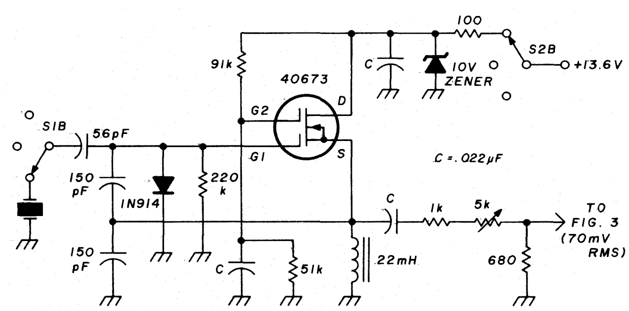

fig. 4. Crystal-oscillator schematic. Circuit is from W2YM. The 1N914 generates "grid-leak" bias.

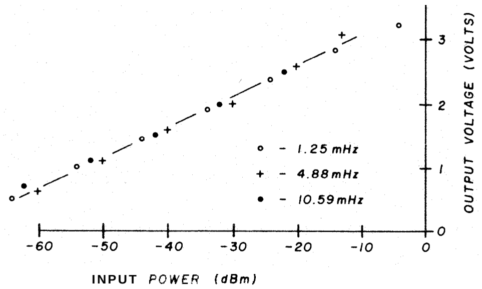

fig. 5. Log detector response showing linearity over its 50-dB range. The absolute level is temperature dependent; relative levels are not (see text).

ternal, or no marker; and, in the internal case, it powers the crystal oscillator, **fig. 4.** This is just the good old W2YM circuit using a diode to generate "grid-leak" bias.

BFO injecton-level is far from critical. The levels shown are what I use, but 10 dB up or 20 dB down from there should be fine. One nice thing about the LM373 is the low level required; this allows a good, solid termination for the external generator without requiring excessive drive. As I mentioned, the marker output goes to U1B for combining with the logarithmic level.

detector response

Fig. 5 shows the linearity of this gadget over its limited range. The absolute level in **fig. 5** is temperature dependent; that is, the curve might go up or down a few tenths of a volt between a cold start and temperature stabilization. The relative levels are not temperature dependent; 10 dB is 0.5 volt hot or cold. All this means is that, if you want to use the target as a "dBm meter," say to measure the output of an oscillator, you must be sure it's warmed up and calibrated. For filter responses you can jump in cold.

concluding remarks

Speaking of secondary uses for the target, note that you don't have to connect a scope to the output. I plugged in a 2000-ohm headset in series with a 0.1-μF blocking cap and used the thing in the external marker mode as a direct-conversion receiver to listen for chirp on the main rig. Worked fine!

I had a small circuit board for the crystal oscillator, having made up a batch of them long ago. I mounted the rest of this thing on a 6 x 6 inch (153 x 153 mm) single-side board, which had room for a second LM3900. Each IC was bypassed at B+ as at pin 3 of the LM373s.

ham radio

log-periodic fixed-wire beams for 75-meter DX

Extensive tests with overseas Amateurs have resulted in an LP antenna with excellent characteristics and performance

This article describes log-periodic antennas made of wire elements, fixed in position, for the 75-meter Amateur band. It includes test results based on contacts between W4AEO in South Carolina and an Amateur in New Zealand, ZLIBKD. During the test period (late 1975 through early 1976), performance of the 75-meter LP, in various configurations, was compared with that of other antennas including dipoles, delta loops, slopers, and verticals.

background

I first became interested in 75-meter DX while talking to Colin, ZL1BKD. Colin had read some of my articles on LP beams[1-5] and asked if I'd tested one on 75 meters. Most of the referenced articles describe the construction of LPs for 10, 15, and 20 meters, giving test results. In one or two of the articles I'd furnished dimensions for a 5-element monoband LP for 75 meters of the log-periodic dipole (LPD) type, but it was never tested. (The dimensions for the 75-meter LPD were merely scaled up from one that worked well on 40 meters.)

As a result of my on-the-air talks with ZL1BKD and an exchange of correspondence, we agreed to conduct a test program involving a 75-meter LP and several popular Amateur antennas.

reference antenna

The antenna used as a reference in the tests was a log periodic consisting of five elements about 18 meters (60 feet) high. This antenna was modified several times during the tests. It was used as an LP Yagi, then as a 5-element Yagi. Test data in the form of operating-log sheets are provided to show on-the-air results **(table 1).**

environmental test conditions

I'm fortunate to have enough space to erect several 75-meter antennas at the same time. Pine trees abound for supports, with heights of up to 21 meters (70 feet). But my location in South Carolina is subject to severe thunderstorms. Lightning took its toll in the summer of 1976: two test antennas were destroyed.

I have also found that vertical antennas don't perform well at this location. I believe this is because of poor ground conductivity in my area,[14] limited clearance between verticals and trees,[15] and an extremely high noise level.

overseas tests

Between July, 1975, and March, 1976, ZLIBKD and I compared over a dozen different antennas with

The 75-meter LP beam antenna described here requires a considerable amount of real estate as well as many high supports. The minimum area required for the 75-meter antenna suggested by W6PYK (3-element LP plus director) is about 0.3 acre (1500 square meters, or 16,131 square feet). This doesn't include space for running supporting lines between antenna elements and trees, which requires another 988 square meters (10,620 square feet). Thus the antenna isn't practical for Amateurs limited to small city lots. **Editor**

By George E. Smith, W4AEO, in collaboration with **Paul A. Scholz, W6PYK**. Mr. Smith's address is 1816 Brevard Place, Camden, South Carolina 29020. Mr. Scholz's address is 12731 Jimeno Avenue, Grenada Hills, California 91344

the 75-meter LP, which was aimed west from my location. Comparision antennas included:

1. Three 75-meter halfwave dipoles at 15, 18, and 23 meters (50, 60, and 75 feet) above ground

2. Three 75-meter delta loops

3. Several 75-meter slopers and phased slopers using various beam configurations

4. Two quarter-wave 75-meter verticals with various numbers of buried radials

5. One half-wavelength 75-meter vertical suspended from a balloon, voltage fed at the bottom with an antenna tuner

6. Two 75-meter half waves in phase (collinear horizontal dipoles) at 23 meters (75 feet) high, oriented broadside to New Zealand

7. One 2-wavelength 75-meter horizontal quad element up 23 meters (75 feet). One lobe was toward New Zealand

8. A two-element, 75-meter bobtail curtain (two phased quarter-wavelength vertical radiators with one-half wavelength spacing). An inverted groundplane was used. Antenna height was about 21 meters (70 feet). The pattern was bidirectional, broadside to New Zealand

9. One 75-meter long-wire antenna (229 meters, or 750 feet long) mounted on tree tops at about 18 meters (60 feet) high. The main lobe was oriented west.

10. A Shakespeare (commercial marine) vertical antenna, center loaded, 7 meters (24 feet) long covering the 4-MHz marine band. The antenna was tuned to 3808 kHz and mounted at 12 meters (40 feet). Four one-quarter-wavelength sloping radials were used.

a note on delta loops

Delta loops for 75 meters were popular during the time of these tests and were used by several 75-meter DXers. Of the three delta loops used in the tests, two were arranged with the horizontal section at the top and with the apex pointed toward ground (delta loops 2 and 3, table **1**). The third delta loop (delta loop 1, table **1**) was in the opposite configuration: apex up and horizontal section about 3 meters (10 feet) above ground.

Two deltas were first fed at bottom or center (horizontal polarization), then changed to corner feed (vertical polarization). The latter configuration was best for the U. S. — VK/ZL path.

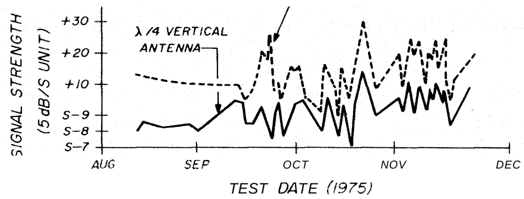

fig. 1. Data showing day-to-day differences between the LP and the quarter-wave vertical used at W4AEO during the test period (August, 1975, through November. 1975.) These data were compiled by W6PYK from reports by ZL1BKD.

Anyone considering a quad or delta loop for 75 or 40 meters is urged to read reference 16, in which the author describes his tests of deltas and quads and shows lobes, radiation angles, and other data. (A reprint of this article appears in reference 17.) Another source appears in the April, 1976, issue of ***ham** radio.*[18]

test results

Overseas tests began in July, 1975. A condensed reproduction of my log (table **1**) shows representative data taken while running tests with ZLIBKD. The vertical, delta loops, and quad were first compared directly with the 5-element LP at about 18 meters (60 feet) above ground, aimed west. Note from table 1 that the quad, erected in October, 1975, and delta loop 3, erected in November, 1975, were mounted over a pond.

The LP, LP-Yagi, and Yagi under test were the only true unidirectional antennas used during the test period. Also note that the LP had been modified several times — it was used as an LP-Yagi for a time, then later modified as a 5-element Yagi.

test note

My first 75-meter LP was completed on August 1, 1975. I made contact with New Zealand stations the following morning. Reports indicated that the LP was at least 10 dB better than the dipole or quarter-wave vertical antenna.

Colin, ZLIBKD, later added an external dB meter to his receiver, which gave more accurate readings. This method was used during the tests between August 21, 1975, and March, 1976, when the tests were completed. Contacts were made several times per week during this period.

Tests were run on 75 meters near 3808 kHz for U.S. and ZL stations, with 350 watts PEP. (VK stations operate split frequency and are received between 3690-3700 kHz.)

During some days there was little if any propagation, so no tests were run because of low signal strength. Little fading was noticed during the test periods. If there was any, it was quite slow in contrast to that on the higher-frequency bands. Signal buildup occurred just before sunrise, usually 5-10 dB. Then a gradual signal-strength decay occurred for about 30 minutes to an hour until the DX signals faded into the noise. We repeated the antenna tests at different times during the 1000-1200 GMT opening.

The unidirectional beams. From reports furnished by ZL1BKD during the 75-meter tests **(table 1)**, the LP and LP-Yagi beams showed a 10-15 dB increase over the other antennas tested. This doesn't mean that the 75-meter beams had a 10-15-dB gain. Theoretically, a truncated LP of the type tested, using only three to five elements and a boom length of only 0.35 wavelength, would probably have no more gain than 5 dB over a dipole at the same height. Increasing elements to 9 or 10, and increasing boom length to about 1.3 wavelength, would probably result in a gain of about 10 dB. But the boom length would be about 103 meters (337 feet), which is impractical for most Amateurs.

I believe that the reported differences, 10-15 dB in favor of the unidirectional LP beams, were caused by the inefficiency of the other antennas tested — possibly by power wasted in lobes in undesired directions.

Comparative reports on reception at W4AEO were about the same. However, a high noise level plus heavy interference at times made direct comparison difficult on some days. (World noise charts show that ZLs and VKs generally experience much less noise than I do in my area.)

During high noise conditions I used several Beverage receiving antennas, which helped to improve reception. I am now erecting several Beverage antennas for 160, 75, and 40 meters. They are of the two-wire type with direction-reversing capability.

More on the delta loops. Table 1 shows that the quarter-wave vertical and delta loop 1 (apex up; horizontal portion near ground) were used for most test comparisons during which data were taken. Delta loop 2 (horizontal section up; apex toward ground) was tested only a few times, as it made a poorer showing than delta loop 1. Delta loop 2 was supported at about 23 meters (75 feet) over a pond. I used a 183-meter (600-foot) length of RG-8/U cable to feed this antenna. Here are some additional interesting observations.

table 1. Condensed log of W4AEO showing representative data on antenna tests with ZL1BKD during the latter half of 1975 and early 1976. Delta loop 1: apex up; delta loop 2: apex down; delta loop 3: apex down. Readings taken by ZL1BKD.

date (1975)	time (GMT)	remarks	LP or LP Yagi	quarter-wavelength vertical	delta loop 1	delta loop 2(1)	two-wavelength quad(2)	delta loop(3)
July		First contacts with ZL1BKD using half-wavelength dipole at 15 meters (50 feet) high						
2 Aug		ZL1BKD reports LP ≈ 10 dB better than dipole or vertical						
2 Aug later		ZL1BKD and ZL2BT report LP now ≈ 15 dB better than dipole or vertical						
21 Aug*		dipole S8	+10-15 dB	S8	S9-S9+10 dB			
25 Aug			+10-15dB	S8	S9-S9+10 dB			
6 Sept			+12 dB	S8	S9+5 dB			
1 Oct		poor conditions	S7 (ref)	−12 dB	−8 dB	−10 dB		
2 Oct			S9 (ref)	− 4dB	−2dB	−10 dB		
8 Oct	1030		+18 dB		+4 dB		+ 6dB	
8 Oct	1200		+26 dB	− 8dB	−6dB		−10 dB	
11 Oct	1115	all sigs. down 10 dB today	5dB	− 7dB	−7dB		− 8dB	
9 Nov		reworked LP	30 dB	15dB	18dB		12 dB	
19 Nov	1038		24 dB	8dB	10 dB		4 dB	21 dB
31 Dec	1115		26dB	21 dB			6-8 dB	
(1976)								
6 Jan	1125		6 dB	0	0 (ref)			4 dB
24 Jan	2120		30 dB	S9				20 dB

Notes: (1) Erected 24 Sept., 1975
(2) Erected 8 Oct., 1975]
(3) Erected 29 Nov.. 1975] These antennas were erected over a pond 1/10 mile from shack

*ZL1BKD improved method of taking readings, which was used for remainder of tests.

1. Delta loop 3 (horizontal part up; apex down) was erected over a swimming pool. The transmission-line length to this antenna was 76 meters (250 feet) of RG-8/U cable.

2. Delta loops 1 and 2 were tried using both end-feed and center-feed for vertical and horizontal polarization respectively. The latter was best for short distances; the former was best for DX. Delta loop 3 was used with vertical polarization only. (The ZL reports shown for the delta loops were for vertical polarization.)

3. Each delta loop had the same length of wire, which was cut by formula to resonate at 3800 kHz. However, I noted that, with vertical polarization, the antenna resonant frequency decreased to 3700 kHz. The following table shows SWR readings taken for the delta loops when fed for vertical polarization:

	standing wave ratio		
f(MHz)	DL1	DL2	DL3
3.5	2.8	2.2	2.7
3.6	2.4	2.8	3.0
3.7	2.0	2.3	2.9
3.8	1.2	2.0	2.0
3.9	1.05	1.07	1.05
4.0	1.8	1.1	1.5

notes on the vertical antenna

The quarter-wave vertical used in most of the tests was a 19-meter (61.5-foot) length of wire suspended by a nylon line between two high pine trees. The vertical was fed from a 61-meter (200-foot) length of RG-8/U coax cable buried in the ground. Ten radials were used originally; the number was later increased to thirty.

Another vertical antenna, consisting of a half-wavelength wire suspended from a balloon, was tested several times for comparison with the quarter-wavelength vertical. Little improvement was noted.

the beam antennas

The original 75-m LP beam first erected for the tests was a 4-element LP with one parasitic director in front. This antenna was later modified to an LP-Yagi using one parasitic director, three driven (LP) elements, and a parasitic reflector. Little difference was noted between these two configurations. Later the LP-Yagi was converted to a Yagi using only one driven element. (This beam was soon destroyed by lightning.)

The last beam used during the test was the Yagi. When comparing it with the quarter-wave vertical, the difference between the Yagi and the vertical was less than that in the previous reports covering the LP

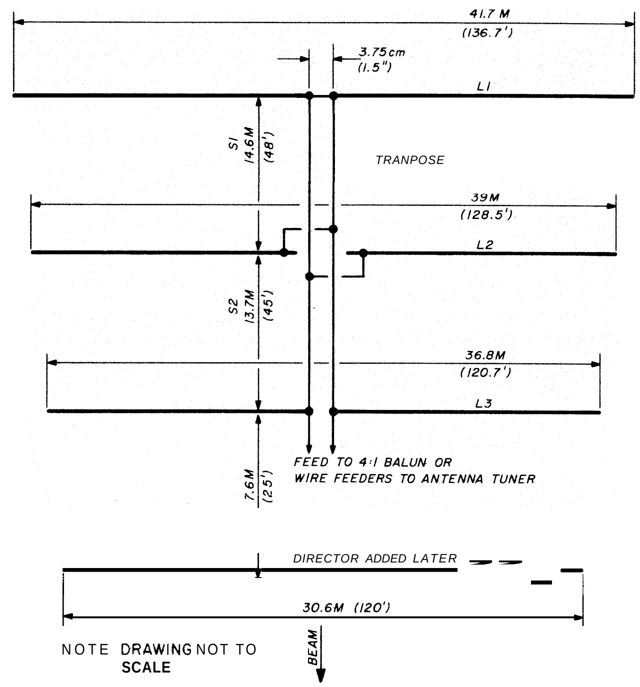

fig. 2. The 75 meter LP design suggested by W6PYK, which is for 3808 kHz. Taper factor, τ, = 0 94; spacing factor, σ, = 0 175.

or LP-Yagi. It's possible that the Yagi gain could have been improved by carefully adjusting the element lengths and spacing, since a Yagi is critical of adjustment. As the Yagi was destroyed by lightning, tests were not completed.

analysis of test results

The beam antennas gave surprisingly consistent day-to-day reports. Average reports were 05, S9+ 10 dB average, with low readings about S9. At times readings peaked to S9+25 dB. These were about the same reports given on the same day to other Eastern U.S. stations running the legal power-input limit, but using only an inverted V or dipole antenna. I therefore feel that the beams did a fair job (especially at a height of 18 meters, or 60 feet). The average report on the beams was about 10 dB better than most of the conventional antennas tested at the same time at my location. I used a coaxial switch for antenna selection, so the readings taken during the tests were made within a second.

Table 1 shows differences among the same three antennas on different days. I believe that this is probably because of the differences in the vertical radiation patterns. **Fig. 1** was compiled by W6PYK from the data taken by ZL1BKD to illustrate the day-to-day difference of the LP antenna and the quarter-wave vertical. The data were taken between August and mid November, 1975. Note that the LP, the LP-Yagi,

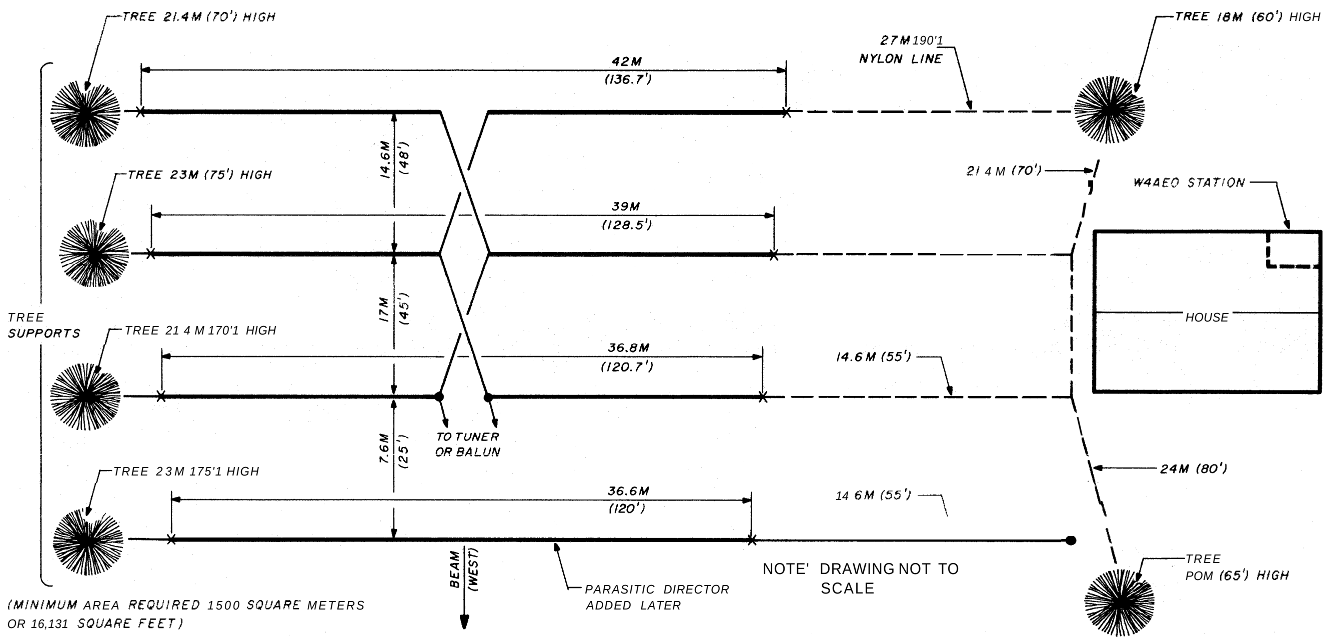

fig. 3. Plan view of the 3-element LP plus director for 3808 kHz (design suggested by W6PYK).

and the Yagi were the only *true unidirectional* beams tested.

Observations:

1. The delta loops (fed at corner's vertically polarized) have bidirectional lobes [plus the smaller high-angle center lobe (90 degrees), straight up]. Thus more than 50 per cent of the radiation is lost in unused lobe(s). Ref. 16, Fig. 10; Ref. 17, Fig. 7; or Ref. 18, Fig. 3.

2. The two-wavelength horizontal quad has four equal-spaced lobes, with only one pointing southwest.

3. The quarter- and half-wavelength verticals are omnidirectional and have maximum radiation at low angles. This may be a doubtful advantage if the antenna is located in a high-absorption environment. Interference from manmade noise was very evident.

a 3-element fixed wire beam LP for 75 meters

The 75-meter Yagi under test was destroyed by lightning in June of 1976. I wanted to replace it with another LP giving more gain, if possible. So Paul, W6PYK, suggested a 3-element, wide-spaced, truncated LP designed with a higher taper factor, τ, and spacing factor, a. Using only three elements would limit the boom length so that it would fit into the available space previously used for the 5-element antenna.

W6PYK had observed a number of the ZL1BKD/W4AEO tests. As he is also interested in antenna design, we became acquainted on 75 meters. We also had a great deal of correspondence, comparing notes. W6PYK made a number of excellent suggestions during this period.

Paul describe; his QTH as a typical California residential lot. He does not have space for a 75-meter beam, but was using a unique 40175 roof-mounted vertical of his own design (which is quite effective considering its size — see *ham radio,* September, 1979, page 44) for working the ZLs and VKs. I agreed to construct and test some of his LP designs. He has written a number of papers on antennas, his latest being "L/P Antenna Design," which appeared in *ham radio,* December, 1979, page 34.

Paul furnished complete dimensions for the 3-element LP: element lengths and spacing **(fig. 2)**

table 2. Standing-wave ratio as a function of frequency for the beam antennas used in the test.

	standing-wave ratio			
frequency (MHz)	**3-element LP (τ 0.94, σ 0.175)***	**3-element LP plus director***	**5-element LP**	**5-element Yagi†**
3.5	1.35	1.35	1.1	2.9
3.6	1.3	1.2	1.25	2.4
3.7	1.2	1.1	1.1	2.0
3.8	1.1	1.1	1.07	1.4
3.9	1.1	1.05	1.15	1.2
4.0	1.5	1.6	1.15	2.0

"The 3-element LP and LP plus director were the last LPs tested. Design suggested by W6PYK using τ = 0.94 and σ = 0.175. This beam was tested after the tests covered by Table 1, which used the previous 4- and 5-element LPs. The improved 3-element LP was the one Bob Tanner, ZL2BT, reported the best tested here to date.

†Note that the Yagi is much sharper than the LPs

designed to $\tau = 0.94$ and a = 0.175. This design gives an overall array length (boom length) of 28.3 meters (93.0 feet). Paul advised that this configuration should provide good gain for the space available.

Note that the element lengths are slightly longer than those given by the formulas. Paul suggested this to allow for ground effect, since the height of the beam above ground would be limited to about 18.3 meters (60 feet), or less than a quarter wavelength. This was evidently correct from the SWR data (**table 2**), which was taken after the beam was completed. A plan view of the beam, **(fig. 3)**, shows method of support — several trees.

During the tests on this new 75-meter beam, the only other antenna available at the time for comparison, was a dipole sloper. The other 75-meter antennas outlined above had been dismantled to make room for several beams needed for 20-meters. Bob Tanner, ZL2BT, reported on this new LP, the best tested to date.

acknowledgment

I especially thank W6PYK for his suggestions on LPs, Beverages, and other antennas. A number have already been tested and several more are still to be tried. I also thank Colin, ZL1BKD, for his many hours of test reports.

references

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15. Edmund A. Laport, *Radio Antenna Engineering,* McGraw-Hill, New York, 1952. Fig. 3.13.
16. L. V. Mayhead, G3AZC, "Loop Aerial Close to Ground," *Radio Communications,* May, 1974, (British ham magazine).
17. Bill Orr, W6SAI, "Antennas: Quads and Delta Loops," *CQ,* August, 1975, page 36.
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ham radio

More **Details?** CHECK — OFF **Page 94**

more conversions of surplus cavity bandpass filters

Follow-up to a previous article in *ham radio* — adapting surplus filters for 220 and 440 MHz

In an earlier issue of *ham radio*,[1] I described the procedure for converting surplus cavity bandpass filters for operation in the 2-meter band. These filters are dual-resonant cavities, gold plated for high conductivity, and are available at low cost in the surplus market."

The previous article thoroughly covered the theory and application of these cavity bandpass filters; therefore, I suggest that you refer to that issue. I'm sure that copies are available.

*Fair Radio Sales Co., Inc., P.O. Box 1105, Lima, Ohio45802

For convenience, the filters used in the 417/GRC receiver and the frequency ranges covered are:

filter	frequency (MHz)	filter	frequency (MHz)
F-2381U	50.0-58.5	F-196/U	184-205
F-239/U	58.5-67.0	F-197/U	205-226
F-2401U	67.0-76.0	F-1991U	224-254
F-241/U	75.0-84.0	F200/U	254-284
F-242/U	84.0-92.5	F201/U	284-314
F-192/U	100.0-121.0	F202/U	314-344
F-193/U	121.O-142.0	F203/U	344-374
F-194/U	142.0-163.0	F204/U	374-404
F-195/U	163.0-184.0	F-236/U	550-600

For conversion to the 220-225 MHz band, I selected the F-195/U and the F-196/U filter assemblies; both are lower in frequency than the desired band. Likewise, I selected a lower-frequency filter for the 420-450 MHz band: the F-202/U. The advantage of selecting the lower-frequency filters for conversion was discussed in reference 1.

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conversion procedure for 220 MHz

The F-195/U and F-196/U filters can be converted to the 220-MHz band by making the following modifications: remove the rear hex nuts from the dual-cavity assembly and carefully slide out the stationary portion of the cavity and the cylindrical housing.

The stationary portion of the cavity consists of the fixed center conductor, the fixed capacitance cup, and the pickup loops. The fixed capacitance cup provides fixed capacitance with respect to the cavity housing and is also the stator of the variable capacitor provided by the movable plunger capacitor cup (fig. 1).

To increase frequency the size of the fixed capacitance cup must be reduced by trimming as indicated by the dotted line in fig. 1, thus reducing the fixed capacitance to the cylindrical cavity wall. To increase frequency to the 200-230 MHz range, cut off no more than 7 mm (9132 inch) on the F-195/U and no more than 5 mm (3/16 inch) on the F-196/U. If a slightly higher range is desired, for example 210-240 MHz, file off an additional amount until the desired range is obtained. A hacksaw or any convenient method can be used. Be sure to remove all burrs for a smooth finish.

Modify one section of the cavity assembly at a time and check with a grid-dip meter at its terminal. About a 12.5-mm (½-inch) loop at the terminal, coupled to the grid-dip meter, should provide a sharp dip at resonance. When the desired frequency is obtained, duplicate the other half of the assembly.

The dual-cavity assemblies can be separated electrically into two individual-bandpass, single-section cavities by removing the connecting jumper and providing two additional terminals using a small BNC type connector.

If two individual series-resonant "suck-out" cavity traps are desired, simply remove the jumper and its pickup loops. You'll then have two single-terminal traps.

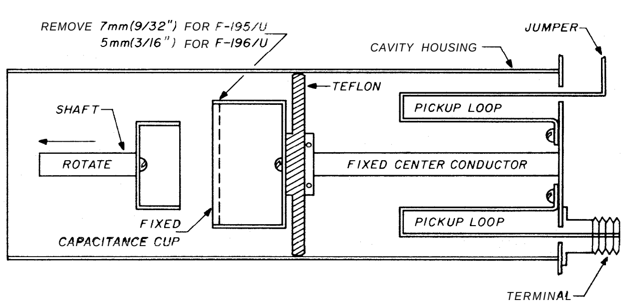

fig. 1. Modifications to the F-195/U and F-196/U filters for 220-225 MHz. The fixed capacitance cup must be trimmed as shown to increase frequency.

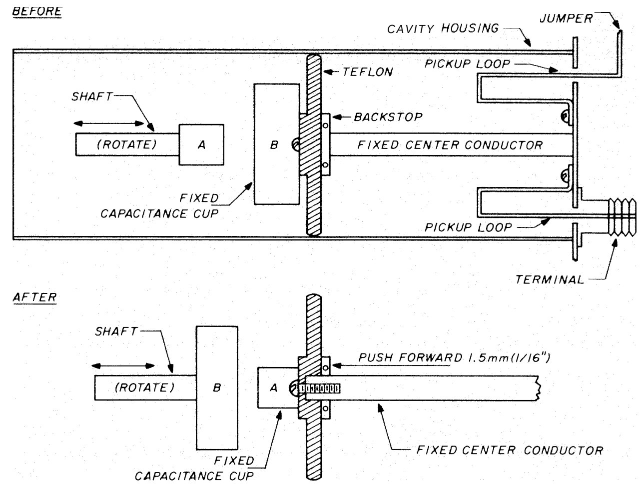

fig. 2. Modifications to the F-202/U filter for 420-450 MHz. Reverse A and B and push forward fixed capacitance cup A as shown.

conversion procedure for 440 MHz

The F-202/U filter assembly can be converted to the 440-MHz band in the following manner. Slide out the stationary portion as previously indicated and refer to fig. 2 "before" and "after." To increase frequency to the 400-530 MHz range, the fixed-capacitor size must be reduced and the movable section size increased by reversing the **A** and B capacitance cups as shown. Note that the small cup, **A,** is pushed forward 1.5 mm (1/16 inch) by loosening the allen set screw in the backstop and pushing it forward.

The frequency range can now be checked by any convenient method such as a transmitter or receiver or a signal generator. You can also use your 2-meter transceiver in the low-power position by feeding the output into either terminal of the dual cavity through a germanium diode to obtain a good third harmonic. You can then read the output as you tune through resonance with a dc microammeter in series with a similar diode at the other terminal of the cavity assembly, Instructions for obtaining two individual bandpass or band reject (suck-out) cavities are the same as previously described.

All the cavities listed for the 417/GRC receiver probably can be converted for use on other frequencies as required. The methods described in this and the preceding article[1] serve as a guide; all that's required is a little patience and some simple test equipment.

reference

1. William Tucker, W4FXE, "How to Modify Surplus Cavity Filters for Operation on 144 MHz," *ham radio,* February, 1980, page 42.

ham radio

LED tuning indicator for RTTY

A novel circuit that takes the guesswork out of tuning RTTY stations

This project was started after I built a modified ST-4 radioteletype (RTTY) demodulator to interface between my EICO 753 transceiver and my model 28 page printer. Since RTTY operation permits only a very small tuning error, and my EICO often drifts freely, my oscilloscope was in constant use to monitor the cross-loop pattern. This pattern shows any drift immediately before the printer starts to garble and therefore was essential for keeping the receiver on frequency.

tuning indicator

It wasn't long before I decided that running my scope just for tuning was a waste of power. It uses 285 watts. Also, the scope's fan is almost as noisy as the teletypewriter. What was needed was something small with low power consumption to provide the tuning indication. This article describes the circuit I used to replace the scope for observing the cross-loop pattern. The display uses two bar graphs, each having seventeen LEDs. The right graph (see photos) shows, bottom to top, the segment **A** to B for mark. The left graph shows, from bottom to top, the segment **A** to **C** for space. See fig. **1.**

As shown in fig. **1,** each loop crosses each segment once. The points where the segments are crossed is directly related to how well the station is tuned in. This circuit samples these points and displays their relative positions on the bar graphs. This circuit also requires that the oscilloscope display of the demodulator output be loops – not straight lines.

configuration

The circuit is connected as shown in fig. 2. The inputs are connected to the same points in the demodulator where an oscilloscope would be connected. U1D, an LM3900N, a unity-gain inverter. (One amplifier in one of the LM3900N quad amps is not used.) Q4 reduces the +12 volts from the demodulator to 6.1 volts to supply the circuit. This voltage is a compromise between LED intensity, system performance, and a desire to reduce power consumption. In the setup shown (fig. **2**), two identical circuits are used: one for the mark input and the other for the space input.

sample-and-hold circuit

Fig. 3 consists of input buffer amplifier U1A, a zero-crossing detector, U1B and Q1, and a sample-and-hold circuit, Q2, Q3, and U1C. When the input signal goes positive through the zero-voltage point, a low-going pulse at Q1 collector turns on Q2 for 5 to 20 microseconds. The voltage on Q2 emitter, which comes from the second sample-and-hold circuit, is stored in C4. Q3 acts as a buffer between C4 and U1C.

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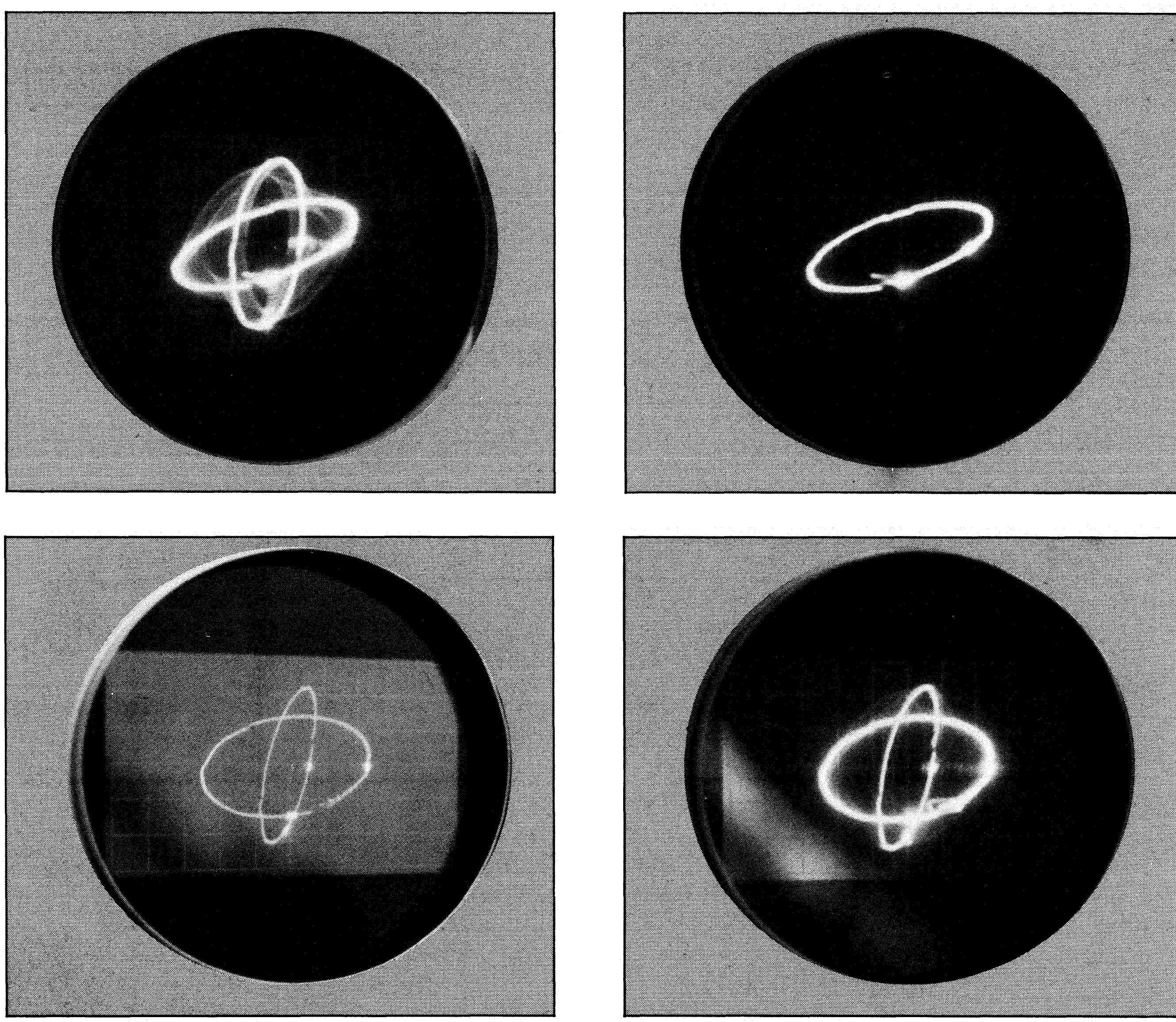


Top left*: *A properly tuned-in mark and space signal with the accompanying bar graph indications. ***Top right:*** A properly tuned-in mark signal with the accompanying bar graph indications. ***Bottom left:*** The four intensified points on the oscilloscope display indicate where sampling is being accomplished. Both sample pulses were coupled to the Z axis on the oscilloscope to help clarify the sampling points in these photos. ***Bottom right:*** A slightly mistuned signal and resulting bar-graph display.

table 1. Parts list for the tuning indicator.

quantity		description	quantity		description
2	C1	0.02-μF, 100V disc	2	R5	180k
4	C2,C3	0.01-μF, 25V disc	2	R6	2.2M
2	C4	0.002-μF, 1000V disc	2	R7	1M
1	C5	50-μF, 25V electrolytic	4	R8,R16	25k pot
1	C6	0.1-μF, 25V disc	2	R9	330k
2	CR1	silicon diode	2	R10	l k
34	CR2-CR18	LED	2	R11	2.2k
1	CR19	6.8V zener, 1W	2	R12	10k pot
2	Q1	2N2222 or equivalent	1	R20	10k
2	Q2	general-purpose PNP	64	R21-R35,	
2	Q3	MPF102, RS276-2035 or equivalent		R37-R53	470k
1	Q4	MPS U01 or equivalent	4	R36,R54	4.7k
2	R1	500k pot	2	U1A-U1D	LM3900N quad op amp
8	R2,R13,R15		32	U3-U18	LM741CN op amp
	R17,R18	100k	1		integrated circuit perfboard 114 x 152 mm
4	R3,R14	560k			(4½" × 6")
3	R4,R19	1.8M			RS276-1394 or equivalent

resistors are ¼ watt

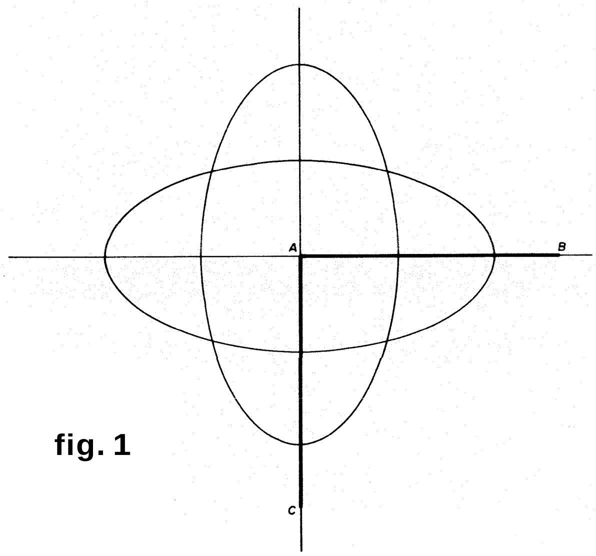


fig. 1

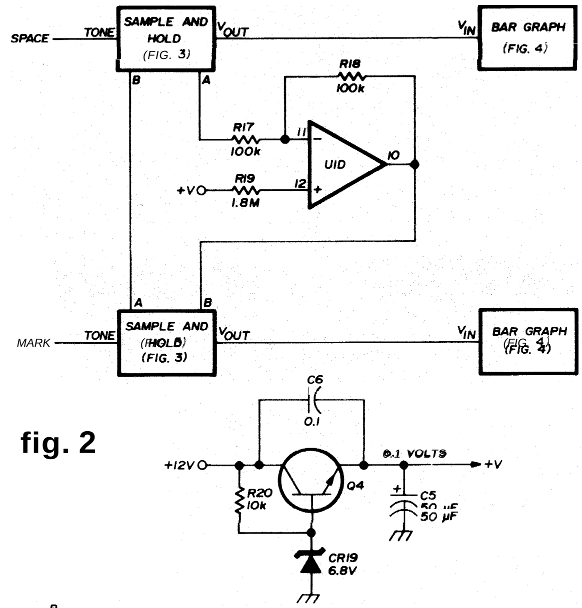


fig. 2

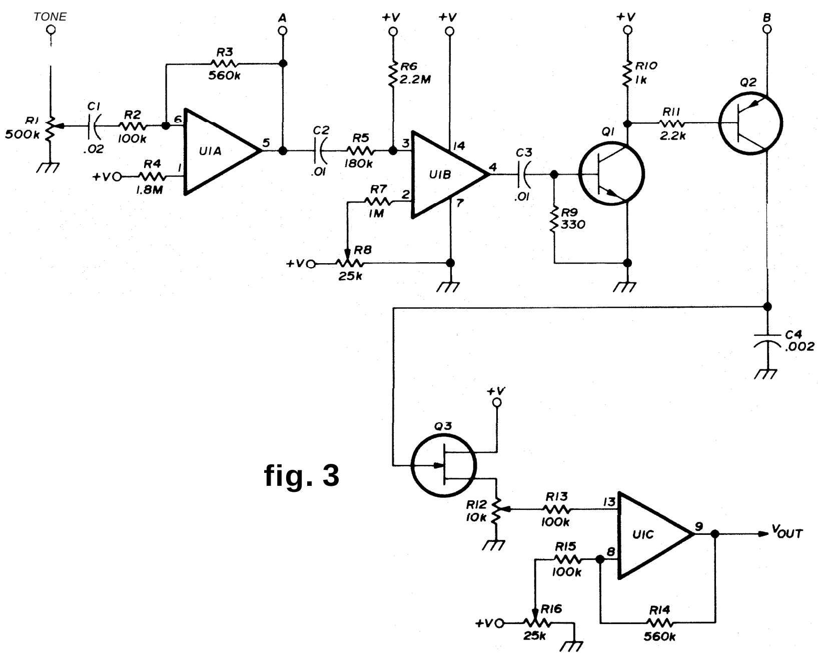


fig. 3

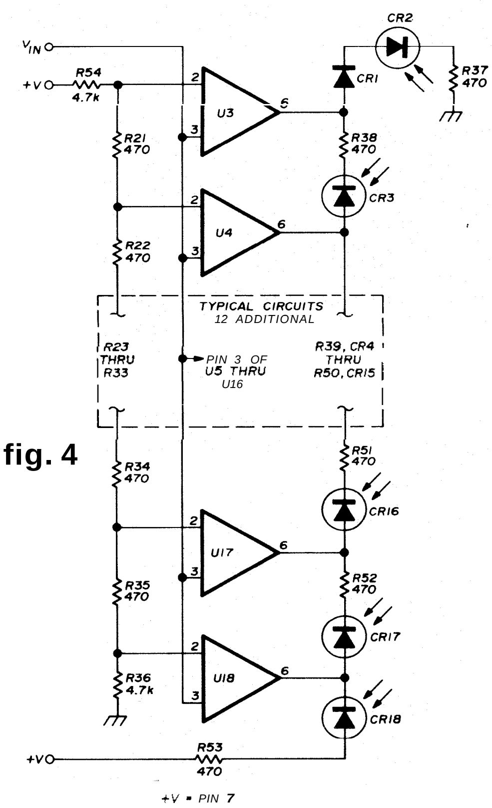


fig. 4

fig. 1. Cross-loop pattern.

fig. 2. Configuration of the complete tuning unit.

fig. 3. Sample-and-hold circuit.

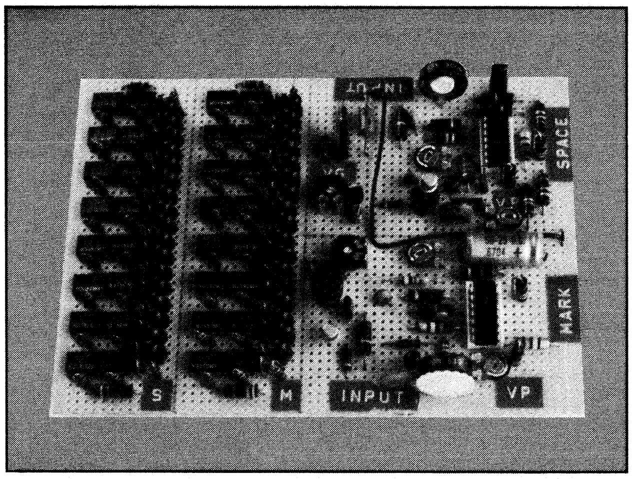
fig. 4. Bar-graph circuit.

bar-graph circuit

Fig. 4 consists of a group of LM741CN comparators that light the appropriate LED, depending on the value of the input voltage at V_{in}. Each 741 has a different reference voltage on pin 2, obtained from voltage divider R21-R36. The sampled voltage is on pin 3. The output of pin 6 of any 741 will depend on the relationship of the reference voltage to the input voltage. When two adjacent 741s have a different output, the LED between them will light. In this way only one LED will be illuminated for any level of input voltage.

construction

A parts list for the tuning indicator is shown in **table 1.** Neither circuit layout nor components were critical in the unit I built. Point-to-point wiring was used throughout. Half-watt resistors may be substi-



The completed project.

tuted if space permits. The power supply voltage can be changed to adjust for LED brightness. Using smaller resistors in series with the LEDs will make possible a lower supply voltage. On my unit, at 5 volts, the upper and lower LEDs could not be lit by adjusting R16 and R12. Using different values of R54 and R36 or adjusting U1C gain will compensate for insufficient or excessive voltage swing from U1C. Using only thirteen or fewer LEDs in each bar graph will still provide a good display if cost is a factor.

adjustments

An oscilloscope is needed in adjusting and trouble-shooting the circuit. Adjust R1 so that U1A puts out a maximum voltage swing with little distortion. Adjust R8 for a symmetrical square-wave output of U1B at maximum input signal. A compromise adjustment may be needed to get a square wave at low input-signal conditions. If Q1 doesn't go low enough, increase either C3 or R9 until Q1 produces a low-going pulse of 5 to 20 microseconds. R12 is the display amplitude control, while R16 is the display position control. Note that R12 also affects the display position.

The displayed pattern on the LED is of sufficient quality to tune in an RTTY station by observing the LEDs alone. Another possible use for the circuit is to interface the LEDs with a computer and have the computer tune the receiver.

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ham radio

improvements to the simplified capacitance meter

Repackaging and improved circuitry make a cap meter published previously even better

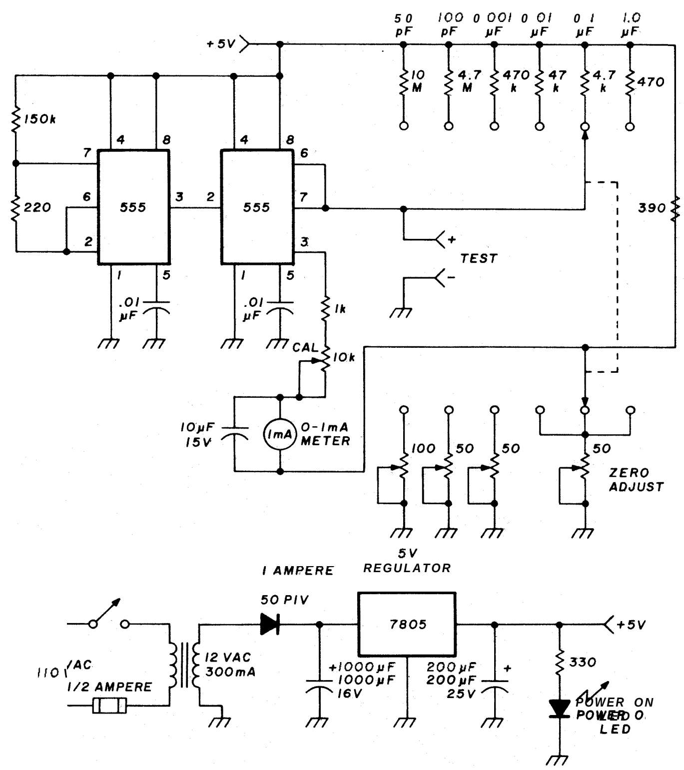

fig. 1. Capacitance meter schematic. All components mount on a single PC board except for the power supply and test terminals.

I've always been suspicious of capacitance meters. They were difficult to read accurately, had extraneous readings, and were nearly useless below 100 pF. But the meter described by WA5SNZ[1] is different. After building the meter, I found that it was easy to read and it measured to 0.5 pF. All those unmarked caps, variable caps, and even my homemade bypass caps could be measured easily. It was fun to put a variable cap on the meter, swing the cap from end-to-end, and watch the meter read from minimum to maximum value. WA5SNZ's cap meter is as easy to use as an ohmmeter.

After discovering what a remarkable meter it was, I designed a packaging scheme so that the meter could be built as a club project. The packaging worked out so well I decided to pass it on to those interested in building this valuable piece of test equipment.

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construction

The cap meter schematic is shown in fig. **1.** All components mount on a single circuit board (**fig. 2**) except for the power supply and test terminals. The cap meter was built into a box measuring 134 x 76 x **149 mm (5-1/4** x 3 x **5-7/8** inches), available from Radio Shack (catalog no. **270-253**).

Wiring the switch on **my** prototype unit presented the biggest problem. There were just too many parts directly to the circuit board so that the unit can be checked out before it's mounted in the box. A bench power supply is used.

The front panel (fig. **3**) was drilled and labeled. The power supply was then built into the box. An LED indicates when power is on.

The meter was removed after tests and mounted onto the front panel. The circuit board was then placed over the meter. The switch was extended through

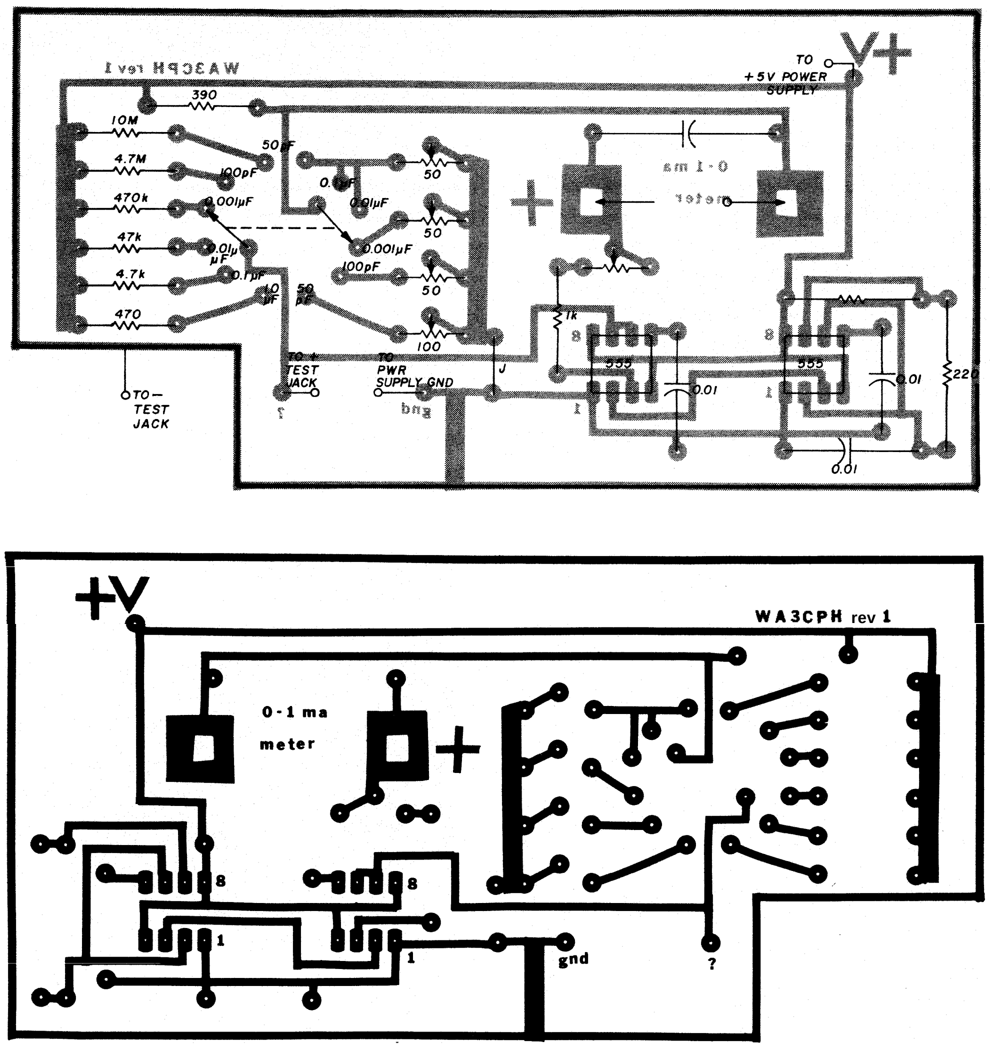


fig. 2. Component layout for the cap meter. above. Foil side of PC board, below.

hanging from it. So I mounted the switch directly to the circuit board with small lengths of bare wire. I attached the wires to the switch first, fed the wires through the holes in the board, and then soldered the switch to. the boards. Be very careful to orient the switch correctly, as it's easy to get it out of place by one position. The switch is mounted on the copper side of the board opposite the components. Holes were provided in the board for mounting the meter the front panel and bolted on (photo). A small length of bare wire was soldered to the + unknown terminal then soldered to the board hole marked (**?**). The ground-side test terminal was connected to both the circuit board and box to prevent ground loops.

modifications to the original unit

The switch I used had six positions (it's available

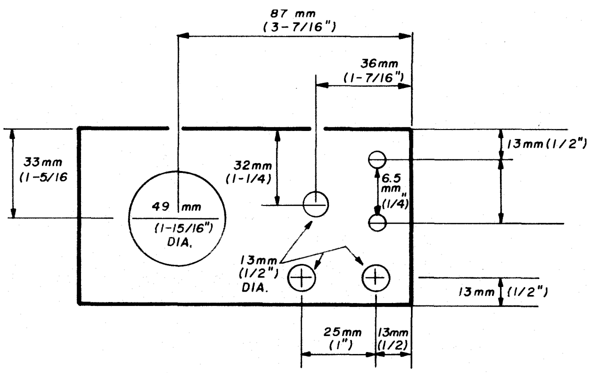


fig. 3. Layout of the cap meter front panel.

from Radio Shack). Rather than leave one switch position blank, I added a 0-50 pF range. On this range, meter readings are divided by two, which is useful when testing small vhf and uhf components.

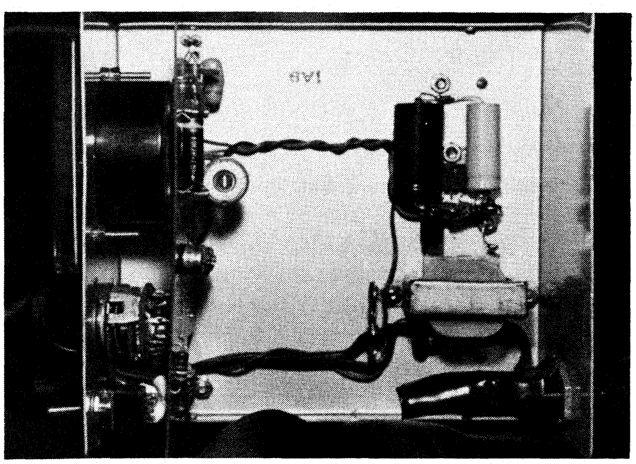
The 500-ohm, zero-adjust pots on the original unit were difficult to set, so they were changed to 50-ohm pots, which made adjustments easier. I used a 100-ohm pot for the 0-50 pF range. I used a 5-volt regulated power supply (**fig. 1**) rather than the original supply, which was 6 volts. No effect was noted on meter operation.

calibration

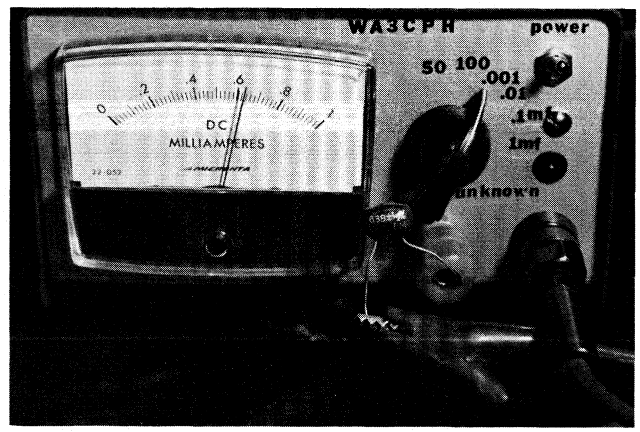
Calibration is the same as for the original unit. Set each range to zero with its respective pot. Then place a 0.001-μF, 5 per cent or better capacitor across the unknown terminals. Set the calibration pot so that the meter reads full scale on the 0.001-μF range. Calibration is then complete.

operation

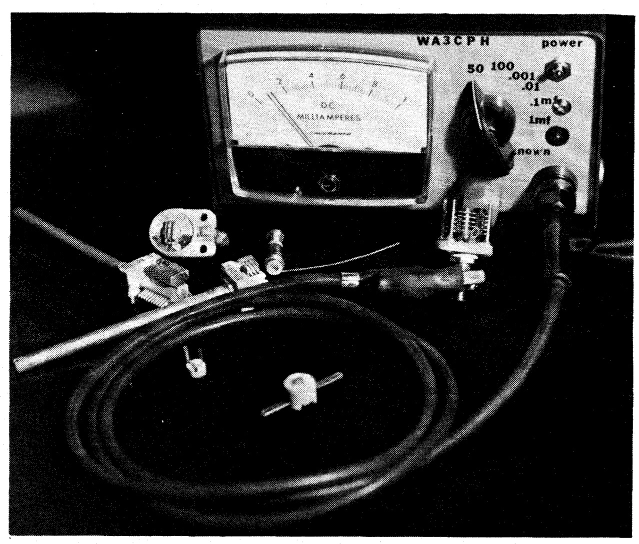
The easiest method of operation is to place a test lead in the ground side test terminal and clip it to the unknown capacitor. Then touch the other side of the capacitor to the plus test terminal (photo). In this way, transient capacitances are minimized and readings are most accurate. Always start on a high range and work down until the best reading is obtained. With variable caps, clip the ground lead onto the cap shaft and touch the other side to the plus test terminal. Vary the cap from minimum to maximum while reading the meter.



Underside of chassis showing component layout in the modified version of the simplified cap meter.



Cap meter with an unknown fixed cap connected to test terminals.



A variable capacitor connected to the cap meter.

reference

1. Courtney Hall, WA5SNZ, "Simplified Capacitance Meter," ham *radio,* November, 1978, page 78.

ham radio

auto-product detection of double-sideband

Novel system for DSB detection which automatically generates the correct reinserted-carrier frequency

The detection of carrierless signals requires very careful control of the transmitter and receiver oscillator frequencies. Mobile communication equipment has the added problems associated with mechanical stability, widely varying supply voltage, and even Doppler shift, which is more troublesome at higher frequencies. Automatic means of correcting to the reinserted carrier frequency in voice single-sideband suppressed carrier systems have met with limited success.[1] In terms of power, a suppressed-carrier double sideband (DSB) transmission offers a highly efficient means of communication, and with DSB you know that the carrier should be halfway between the two sidebands.[2]

double-sideband carrierless detection

The carrier frequency for a pair of sidebands is one half their sum; to derive the carrier, it is necessary to select its twice-frequency component from the second-order products obtained from a nonlinear amplifier (detector). The relationship of the second-order products to an amplitude-modulated signal with carrier is shown in the appendix.

The block diagram, fig. 1, shows the essentials of a system for generating the reinserted carrier to a double-balanced modulator. The carrierless DSB input signal is centered around the typical receiver i-f frequency, 455 kHz, and is applied to two groups of circuits: a nonlinear amplifier or detector, and a balanced modulator or product detector. The nonlinear amplifier is followed by a filter to select the desired sideband-sum component. The waveforms of signals at these points are shown in fig. 2.

The input signal at point **A** is shown at the top of

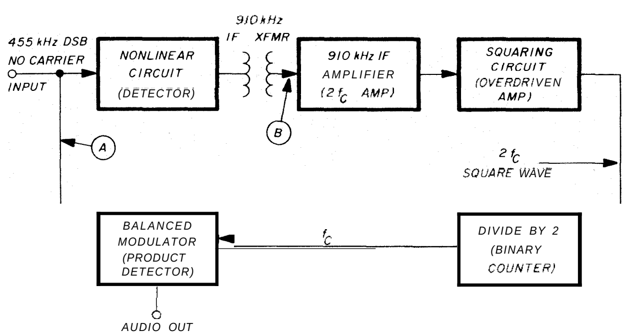
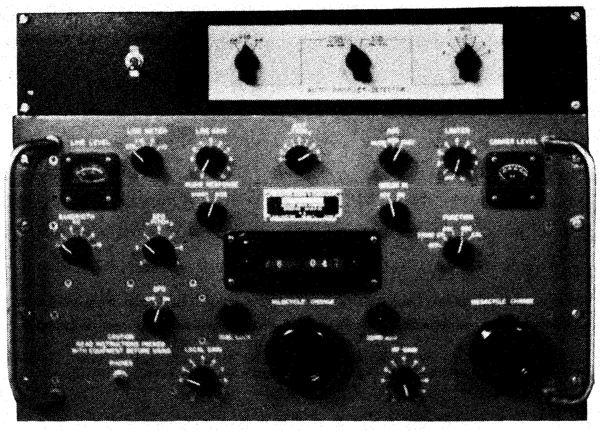
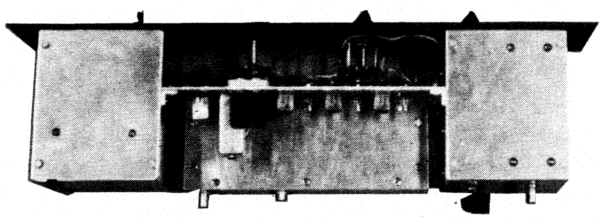
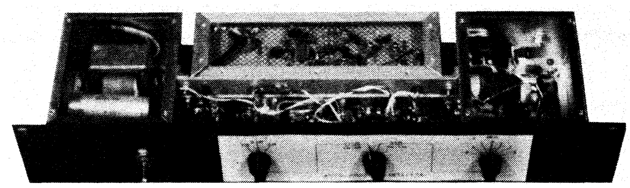


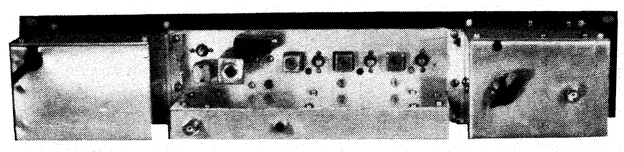
fig. 1. Block diagram of the double-sideband no-carrier demodulation system or auto product detector. Waveforms at points A and B are shown in fig. 2; a schematic diagram of the system is presented in fig. 3.

By H. F. Priebe, Jr., K4UD, 5040 Wickford Way, Dunwoody, Georgia 30338









Left: Top, the auto product detector built by K4UD and used with a Collins R-390A receiver. Directly beneath is a bottom view of the auto product detector showing the twice-carrier frequency amplifier circuit. Above: Top view of the auto product detector with shield covers removed. The power supply is to the left, and balanced modulator with squaring circuit and binary counter are in the center. The BFO for SSB reception is at the right. Directly beneath is rear view of the unit. The twice-carrier circuits are located in the center along the bottom; the knob on the left is a BFO calibration adjustment.

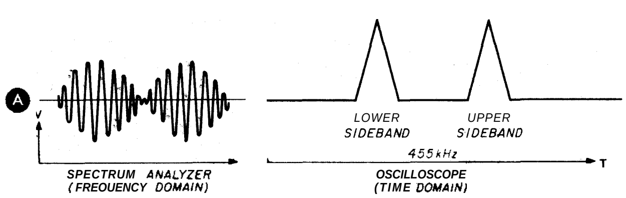
fig. 2 (the time-domain display is the familiar two-tone signal used for testing SSB transmitters). The output of the nonlinear amplifier following the tuned circuit is shown at the bottom. This signal is a 100 per cent modulated amplitude-modulated 910 kHz carrier with sidebands at twice the input sideband frequencies. The 910 kHz i-f amplifier and crystal filter supply additional selectivity to reduce the amplitude of the sidebands and thus reduce the resultant equivalent modulation percentage. The remaining sidebands or modulation is reduced by the squaring circuit in **fig. 1** which provides an output square wave at twice the carrier frequency. This $2f_c$ signal is divided by two in the binary counter stage and supplied as the demodulating carrier to the balanced modulator product detector.

Detection of signals with carrier is similar because the level of the demodulating carrier has several times the amplitude of the original. Reversing the phase of the 910-kHz signal is equivalent to shifting the phase of the carrier by 90 degrees; therefore, a simple phase-reversing switch permits copying amplitude or phase-modulated signals.

circuit diagram

The schematic of the carrierless double-sideband detector is shown in **fig.** 3. The nonlinear input amplifier is based on a dual-gate MOSFET with a double-tuned circuit load. A three-stage tuned amplifier with a crystal filter provides selectivity to reduce the amplitude of the second harmonics of the input sidebands; this ensures that an amplitude squaring circuit will extract the sideband sum component. The amount of selectivity required is related to the squaring circuit sensitivity and the lowest frequency sideband separation.

For voice signals with a low frequency component of a few hundred hertz the 910 kHz i-f is approximately 1.5 kHz wide; this permits a tolerance to signal instability of approximately 1000 Hz.



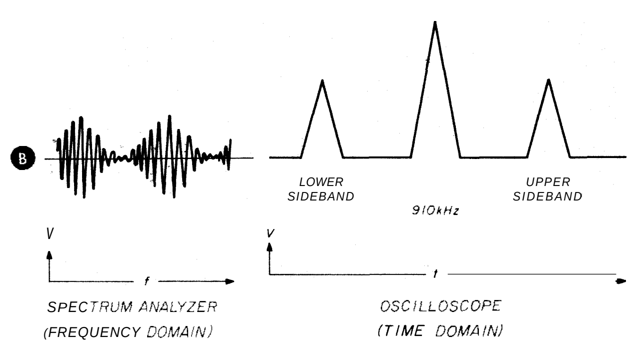


fig. 2. Auto product detector waveforms as displayed in the time domain (oscilloscope) and frequency domain (spectrum analyzer). Signal at point A (fig. 1) is similar to a two-tone SSB test signal; signal at point B has 910 kHz carrier.

The balanced modulator (product detector) uses a MC1496L IC; the type 5596A is similar but the pin-out is different. A diode and transformer balanced modulator could also be used, of course. All three were tested in the circuit and performed about the same.

The **BFO** operates at 910 kHz and uses the familiar Colpitts circuit; its output is divided by two as was done with the derived carrier for **DSB** detection

conclusion

The double-sideband no-carrier detector provides auto product detection of double-sideband signals with or without a carrier. There is no reinserted carrier oscillator — the reconstructed carrier is derived automatically from the two sidebands of the incoming signal. This makes receiver tuning of no-carrier **DSB** signals as easy and simple as tuning in a-m signals.

The detector is compatible with a-m and provides exalted-carrier a-m detection with none of the disadvantages of carrier oscillator control. The detection of phase-modulated signals is accomplished with a switch position that alters the phase of reconstructed carrier.

A **BFO** is included so that **SSB** signals can also be detected. Thus, the described circuit can demodulate all the popular modulation modes as well.

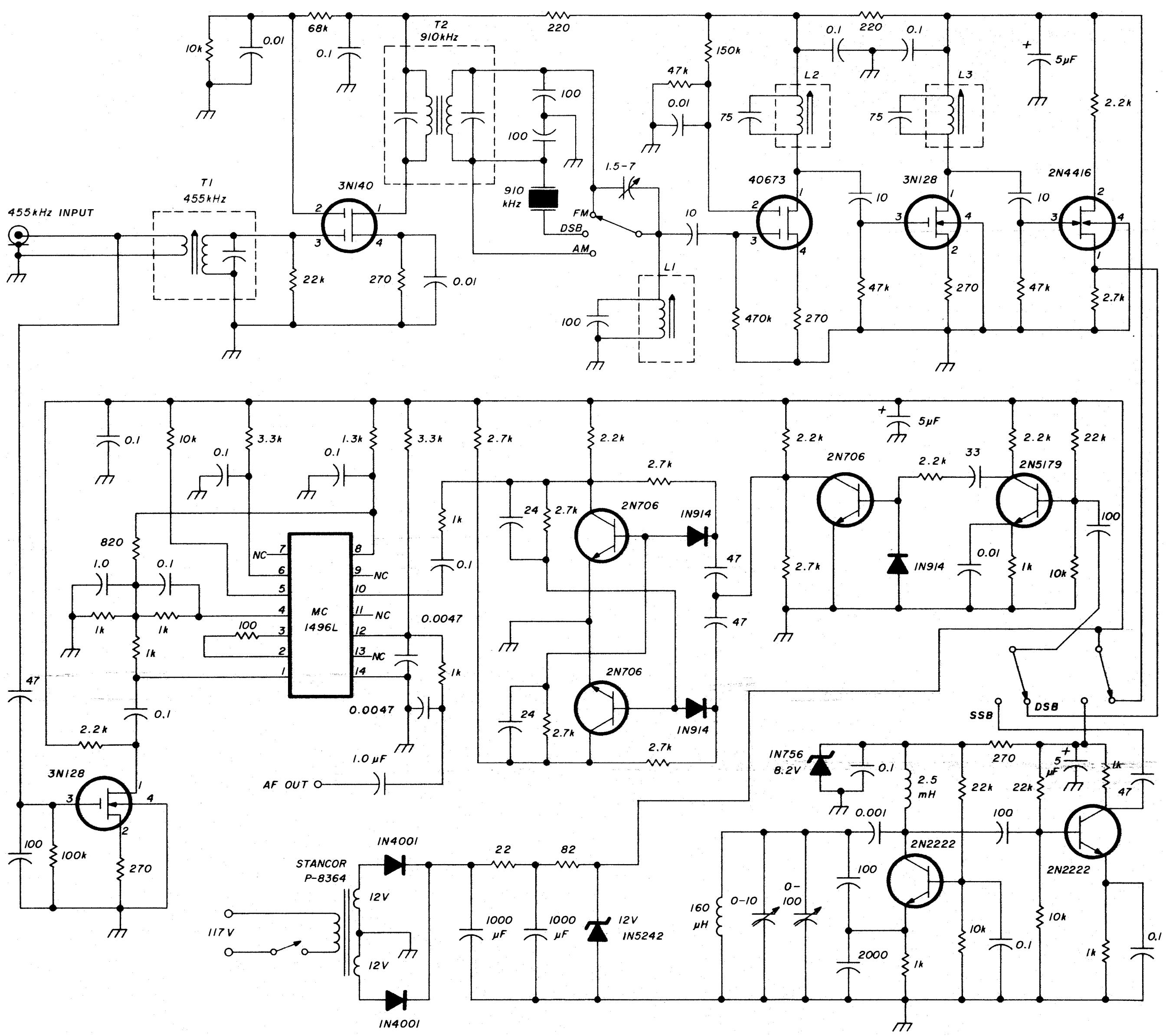


fig. 3. Schematic diagram of the auto product detector for DSB and a-m; an optional BFO is provided for the detection of SSB signals. Inductors L1, L2, and L3 are 290-650 pH (J. W. Miller 9057). Transformer T1 is a 455-kHz i-f transformer; T2 is a tube-type 455-kHz i-f transformer with turns removed.

references

1. *O.G. Villard,* Jr., *"Sideband-Operated Automatic Frequency Control for Reception of Suppressed-Carrier SSB Voice Signals," IEEE Transactions on Communications Technology, October, 1971.*
2. *W.K. Squires and* E. *Bedrosian, "The Computation of Single-Sideband Peak Power," Proceedings of the IRE, January. 1960.*

appendix

The output of the nonlinear amplifier stage is analyzed from the curvature of the transfer characteristic as expressed in terms of a power series.' Such a curvature is characteristic of rectifiers and detectors. When the input e_g is made up of two sine waves E_a and E_b:

$$e_g = E_a \sin \omega_a t + E_b \sin \omega_b t \tag{1}$$

The second-order components are:

$$e_g^2 = \frac{E_a^2 + E_b^2}{2} - \frac{E_a^2}{2} \cos 2\omega_a t$$

$$- \frac{E_b^2}{2} \cos 2\omega_b t$$

$$+ E_a E_b \cos (\omega_a + \omega_b)t$$

$$+ E_a E_b \cos (\omega_a - \omega_b)t$$

With a filter circuit tuned to twice the i-f, the dc term $\frac{E_a^2 + E_b^2}{2}$ is of no concern to this mode of operation, nor is the difference term, $E_a E_b \cos (\omega_a - \omega_b)t$. The important components are then:

$$e_g^2 \cong E_a E_b \cos (\omega_a + \omega_b)t$$

$$- \frac{E_b^2}{2} \cos 2\omega_b t$$

let $\omega_a + \omega_b = 2\omega_c, \quad \omega_a = \omega_c + p, \quad \omega_b = \omega_c - p, \quad E_a = E_b = E$

then $e_g^2 = E^2 \cos (2\omega_c)t$

$$- \frac{E^2}{2} \cos 2 (\omega_c + p)t$$

$$- \frac{E^2}{2} \cos 2 (\omega_c - p)t$$

And, for an amplitude modulated wave: †

$y = A, \cos \omega_o t$ — carrier

$+ \frac{a_m}{2} \cos (\omega_o + p_m)t$ — upper sideband

$+ \frac{a_m}{2} \cos (\omega_o - p_m)t$ — lower sideband

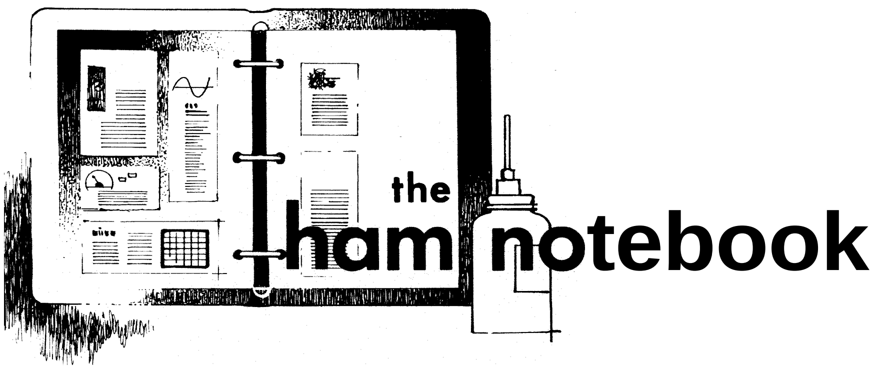
where p_m = the angular frequency of the modulating signal
$a,$ = degree of modulation
$a,$ = $A,$ for 100 per cent modulation

Therefore, the output of the two-times i-f filter is a 100 per cent modulated a-m signal with a modulating frequency of twice the original.

*F.E. Terman, *Electronic and Radio Englneenng,*McGraw-Hill, New York, 1955, page 332.

†*Reference Data for Radio Engineers,* 3rd edition, Federal Telephone and Radio Corporation, New York, 1953. page276

ham radio



improving the Drake R-4C product detector

The single-ended diode product detector used by Drake is typical of the type designed into many present-day receivers. Its simplicity and small number of parts make it a good performer in the Drake R-4C receiver.

However, the 1N270 diode used in this circuit creates large harmonic currents because of its nonlinear nature. This harmonic energy is generated by the BFO and appears as a constant hissing sound in the audio output. It's not noticeable on fairly strong signals but can become annoying if you're listening to a weak signal.

Some time ago I replaced the PN diodes in another receiver with hot-carrier diodes and noted an improvement in performance. Hot-carrier diodes differ from the usual PN diodes in that they switch very fast and don't suffer from the charge storage effect of the junction diode, which creates the high order of harmonics appearing in the audio output.

I replaced the 1N270 diodes in my R-4C with Hewlett-Packard HP5082/2800 hot-carrier diodes **(fig. 1)**. The results were quite pleasing. Although the hiss was not completely eliminated, it was significantly reduced. The audio output level also increased.

A word of caution in replacing the 1N270 diodes: They're difficult to remove from the PC board because they are on the bottom of the board, which is mounted in a vertical position with other parts around.it. A little extra care and a small pencil-type soldering iron should do it. Before replacing, note the polarity of the removed diodes.

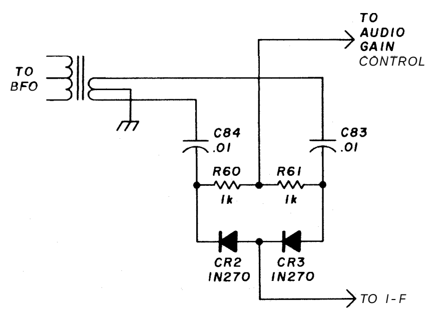


fig. 1. Diodes CR2, CR3 in the Drake R-4 receiver were replaced with H-P 580212800 hot-carrier diodes to reduce product-detector noise.

The HP5082/2800 diodes are very small and have a glass body. They crack easily if the leads are pulled too thightly through the holes in the PC board.

If the HP5082/2800 diodes are difficult to obtain, a suitable replacement is the Sylvania ECG519.

Bernard White, W3CVS

sealing coaxial connectors

Unfortunately, few Amateurs take the necessary precautions to safeguard coaxial cables from water contamination. This is especially true where connections are made to antennas. Often the braid and center conductors are simply fanned out, and no sealant is applied to impede water entry.

Dipole installations are often the worst. The old trick of looping the coax over the center insulator and taping it to provide a strain relief is a good one. However, unless the cable end is carefully sealed before connection to the dipole, water will enter the line, be drawn uphill around the loop by capillary action, and eventually contaminate the entire length of line.

Damage to the line by contamination is permanent. Even when the cable is dried out, the internal corrosion will impede effective shielding, change the characteristic impedance, and increase power losses. Because the braid can act like a metal sponge, putting a liberal application of electrical tape at the line end is about as effective as taping a rope but leaving the end exposed: Moisture will still enter the open end, and capillary action will do the rest.

Frequently, Amateurs attempt to seal line ends with silicone rubber sealants. Two problems exist here. The first is that nearly no adhesion exists between the vinyl or PVC jacket and the silicone rubber. The second problem is that, during curing, the silicone rubber compound releases highly corrosive acid vapors, which can devastate the conductive surfaces of connectors, leading to other problems. The following suggested methods are applicable to the majority of installations and can save you both coax cable and rf.

Before matching coaxial connectors for exterior use, apply a liberal amount of silicone grease, such as Dow Corning DC-4 or General Cement Z-5, to the interior of the connectors to reduce moisture contamination. Any silicone grease which oozes out of the connectors will prevent adhesion of tape or other sealant and should be washed away with a solvent. Aerosol spray solvents that leave no lubricating residue are appropriate for this purpose.

The female SO-239 connector is not waterproof. Care should be taken to seal the rear of the connector. Quick-setting epoxy provides a watertight seal.

Where the cable end attaches directly to an antenna, an effective means of sealing the end is to use epoxy. As shown in **fig. 1,** a plastic pipe cap or plastic chair-leg tip can be used as a form to hold the epoxy as it cures. It can then be left in place. The braid should be expanded so that it's loose enough for the epoxy to flow around all the conductors. After

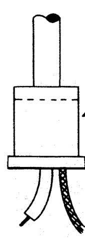


fig. 1. Recommended method for sealing a coaxial cable that connects directly to an antenna. The plastic form holds the epoxy until it cures.

pouring in the epoxy, work the coax around in the pipe cap to promote complete saturation of the braid by the epoxy.

Where a coaxial connector, such as the PL-259, is used to terminate the cable, it's common to attempt to seal it with plastic electrical tape. Depending on the method used, this can be quite effective. However, as often as not, in several days the tape will pull around on the connector as it adjusts its tension. Often this leaves the connector partially exposed to water, a fact not discovered until later when the cable is contaminated with water and some malfunction necessitates antenna service.

The problem can be easily avoided. Behind the connector, wrap enough plastic tape to build up the diameter of the cable gradually to the point where the connector is attached **(fig.** 2). Then, an overwrap of two or three layers of tape covering the entire area can be applied easily, with no wrinkles. This overwrap should be applied tightly, but without stretching the tape, especially during the last several turns. This prevents tension from tearing loose the tape end. Where the connection is to be suspended vertically, the last layer of overwrap

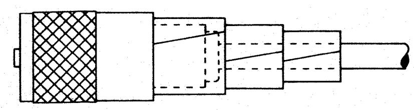


fig. 2. Sealing coax connectors using PVC tape. Proper tape application avoids water contamination (see text).

should be started from the bottom. in this way water running over the surface will run over the tape laps, just as rain runs over shingles on a roof.

Robert Wheaton, W5XW

modifications for the K4JIU frequency counter

A number of people who have built the counter I described in the February, 1978, issue of *ham radio* have experienced problems on the 50- and 500-MHz ranges. The problem goes something like this: On the 5-MHz range the counter will usually count to around 7 MHz. Intersil guarantees only 5 MHz for the ICM7208, but typical performance is much better. However, on the 50-MHz and 500-MHz ranges, the counter is found to perform only to around 28 MHz and 280 MHz respectively.

The problem is not in the prescalers! Most of the counters I've seen and have had the opportunity to test don't have this problem; after some discussion with the Intersil applications and engineering people, there's a simple solution that I've tried, and it works well.

The problem is that the 74196 prescaler puts out a signal having a 20 per cent duty cycle. When inverted by half of the 75452 this becomes an 80 per cent duty cycle. The ICM7208 counter chip, unlike TTL devices, performs best when presented with a 50 per cent duty cycle.

Now for the solution. The Q_C output of the 74196 runs at the same frequency as the Q_D output but has a duty cycle much closer to 50 per cent. Therefore, the circuit trace going to pin 12 of the 74196 should be disconnected from pin 12 and reconnected to pin 2. This requires that the trace to the D data input be switched to the C data input; i.e., disconnect the trace from pin 11 and run it to pin 3. Concerning the last change, rather than disconnecting pin 11, simply jumper it to pin 3. The explanation for this is simple and obvious to anyone familiar with the operation of the 74196.

Much to my embarrassment I also learned that, of all the counters I've tested, the 50-MHz front end sensitivity is about 50 mV or less, flat to 65 MHz. Apparently my own unit has some marginal transistors.

I wish to thank the many hams who have written or called to tell me of their successes. I greatly appreciate the feedback, pro or con. The change described above has been incorporated into the artwork, and the next batch of boards, which I hope to have by the end of May, will have this change.

John H. Bordelon, K4JIU

Collins 32S PA disable jacks

The rear chassis lip of Collins 32S transmitters provides two phono jacks labeled P.A. DISABLE. These jacks disable the final amplifier screen supply and have +275 volts available on the inner conductors during transmit. To avoid equipment damage or operator injury by an incorrectly inserted cable, I inserted phono plugs filled with silicone sealant into each phono jack. Plastic-sleeve-covered plugs, such as Radio Shack 274-451, might be used instead.

With the jacks so covered, no accidental contact will occur. This idea may also be used for all other unused phono jacks for similar protection.

Paul Pagel, N1FB

NEW products

microminiature encoder

Communications Specialists introduces the SS-32 Microminiature Tone Encoder, which produces either sub-audible or burst-tone frequencies.

This encoder measures 0.9 by 1.3 by 0.40 inches and adapts to all mobile units and most portables. It operates on any dc voltage from 6 to 30 volts and may be ordered in either the audible or sub-audible configuration.

The SS-32 is completely field programmable using a dip switch to produce any one of the thirty-two standard EIA sub-audible frequencies or any one of thirty-two audible frequencies which include touch-tones, burst tones, and test tones such as 600, 1000, 1500, 2175, and 2805 Hz. No counter or other test equipment is required to set frequencies.

The output is a low-impedance, low-distortion adjustable sine wave, 5 volts peak-to-peak. In the sub-audible version, the frequency accuracy is ± 0.1 Hz maximum from $-40°$ C to $+85°$ C and the accuracy of the audible tone output is ± 1 Hz.

A remote-mounted rotary switch may be purchased to allow selection of any of the tones within either group. Reverse polarity protection is built-in and all connections to the board are made with color-coded wires supplied with each unit.

A full one-year warranty is provided for factory repair. Price of the SS-32 is $29.95, wired, tested, and with complete instructions.

For more information, write Communications Specialists, 426 West Taft Avenue, Orange, California 92667.

model 299 talking counter

Ten-Tec's Model 299 Talking Counter is a self-contained frequency counter, speech synthesizer, and audio amplifier/speaker system which enhances operating convenience and pleasure for the blind ham operating on the high-frequency bands. It can be used with any high-frequency transceiver, analog or digital, or with any vhf transceiver with an appropriate prescaler. Also, it can be used with any signal generator below 22 MHz as a test instrument. When used with Ten-Tec transceivers employing 9 MHz i-f, special built-in presets allow proper readout of the operating frequency, even though the counter is reading VFO output.

Some operating features are:

1) Synthesized speech readout of any rf voltage applied to the input between 1 MHz and 22 MHz. This includes the 10-meter band on Ten-Tec transceivers since the VFO operates below 22 MHz.

2) Choice of MHz and kHz format, or only kHz portion for a quick-repeat cycle.

3) Choice of one-time or repeat cycling.

4) Counts to four places after decimal (100 Hz). When used with analog transceivers, Model 299 ***increases*** readout accuracy.

5) Self-contained audio amplifier and speaker. No need to tap into transceiver audio system.

6) Only connection required to the transceiver is for the VFO output signal.

7) Runs on 12 Vdc.

Model 299 Talking Counter user price is $290.00. For more information, write Ten-Tec, Inc., Sevierville, Tennessee 37862.

Plexiglas cabinets

Debco Electronics introduces a line of Plexiglas cabinets ideal for LED digital devices. All units feature a clear-red chassis which serves as a filter lens to improve readability. Two sizes are available: Cab-I, measuring 3 × 6% x 5% inches, and Cab-II measuring 2% by 5 by 4 inches.

Both types have a sloped front and friction feet, and are available with black, white, or clear covers. Cabinets are available factory direct. Cab-I costs $9.95, Cab-II $8.95.

For more information. contact Debco Electronics, P.O. Box 9169, Cincinnati, Ohio 45209.

AEA MorseMatic keyer

A computerized electronic keyer is now available that combines virtually all the features of all the other keyers in the marketplace, at a price that is affordable for any true CW enthusiast.

The AEA MorseMatic uses two custom, state-of-the-art micro-computer chips to perform functions that were previously only a CW operator's fantasy.

The MorseMatic can be tailored to the user's needs. Features considered to be great by some users (such as dot and dash memory) are disliked by others. For the first time, the MorseMatic makes a keyer available that will appeal to all users because it can be tailored exactly to each operator's desires with a sixteen-button keypad.

For serious contest enthusiasts, the MorseMatic offers the most flexible automatic serial-number generator on the market.

For serious vhf DXers, the MorseMatic offers the exclusive automatic-beacon mode for precise moonbounce, scatter, or tropospheric DX scheduling. To use the beacon mode, instruct the MorseMatic how long to transmit any selected message and how long to pause before the message is automatically transmitted again. The computers will automatically set the message code speed to fit the desired transmit window. The beacon mode can also be used for contest operating and for vhf beacon transmissions.

The MorseMatic keyer is the first to offer "soft-partitioning" of the memory, unlike the "hard partitioning" in all other keyers. Soft-partitioning means no wasted memory space. All of the memory car! be allotted to one message location, or it can be divided up into as many as ten locations. The memory can be loaded in automatic mode for perfect message formatting, or it can be loaded in the real-time mode for individualizing a message. Memory can also be loaded in automatic-keyer mode (any dot and dash ratio) or in semi-auto (bug) mode. Any message can be played back with any selected dot and dash ratio. Hence, the user can send a sloppily loaded bug-mode message back with perfect 3 to 1 dash to dot ratio. Conversely, a perfectly loaded 3 to 1 dash to dot ratio message can be replayed later with as much as an 8 to 1 dash to dot ratio (sounding like a bug).

Automatic transmit-tune mode. The MorseMatic can be used to key the transmitter for tuning purposes. The operator need only hit any keypad button or the key paddle to defeat the tune mode.

Editing a memory loading mistake is a snap with the MorseMatic. If you are near the end of loading a message into memory and a mistake is made, it only takes seconds to erase the mistake and then continue with an error-free message.

All this, plus the world's best Morse trainer, is included in the basic price of the MorseMatic. It is the only trainer that will automatically increase the speed of the practice characters so that your brain is "fooled" into thinking it is still copying the starting speed. No more need to keep buying practice tapes as you start memorizing old ones, or as you progress in speed. The MorseMatic will take you from 2 to 99 WPM. MorseMatic and Soft-Partitioning are trademarks of AEA.

Introductory 'Amateur net price is $199.95. Write Advanced Electronics Applications, P.O. Box 2160, Lynnwood, Washington 98036.

More **Details?** CHECK – OFF Page **94**

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contents

a second look

by Jim Fisk

The wideiy escaiating precious metais market and Amateur Radio. What you might ask, ***does*** one have to do with the other? As a starter, consider the fact that the basic construction of practically every component in that new transceiver you're thinking about buying uses silver, gold, or palladium. Those inexpensive and innocent looking ceramic bypass capacitors that are used by the hundreds, for example, use thin silver layers deposited on ceramic substrates. Most transistors, diodes, and integrated circuits use gold contact wires and many are built within a tiny gold frame; and palladium is often used in precious monolithic resistors. When you add the precious metals in these common components to the more obvious ones like silver-mica capacitors and silver-plated switch contacts, tank circuits, and variable capacitors, it is suddenly apparent that the transceiver on your operating desk is a source of hidden wealth. More important, it is indicative of the great increases in the cost of Amateur Radio equipment you can expect in the not-too-distant future.

As recently as last year, the cost of precious metals used in the manufacturing processes of electronic components was relatively minor, and the manufacturers simply factored that cost into the selling price of the part. The commodities market was fairly stable, so the manufacturers absorbed any minor fluctuations in material costs. With the recent volatility of the precious metals market, however, the manufacturers are no longer able to absorb the huge cost burden and are beginning to pass it along to their customers in the form of a surcharge. At the present time a 10 to 15 per cent surcharge is not uncommon for many components; it is even higher on some high-grade parts that depend heavily on the use of gold.

And while the soaring costs of gold and silver have been capturing the headlines, costs of other commodities which are important – often vital – to electronics are also going out of sight. Consider for a moment that penny in your pocket; the cost of the copper has now reached the point where the Lincoln penny's monetary value is essentially the same as its copper value. When you translate that into the huge amount of copper used by industry in the manufacture of printed-circuit boards, hookup wire, coaxial cable, and a hundred other electronic products, you are struck with the enormity of the situation – and the great impact it will eventually have on the costs of all electronics equipment.

The costs of equipment will also be greatly affected in the future by the OPEC oil cartel because of the great quantities of petroleum-based materials used in electronics: epoxy-fiberglass circuit boards, thermoplastic insulation, polythylene coaxial cable – the list goes on and on. If you have watched the price of coaxial cable for the past few months, you've probably noticed that the prices quoted in the magazine advertisements seem to be higher in each new issue of the magazine; ditto for rotator cable and hookup wire. Just as one example, when I bought a few feet of Teflon-insulated RG-141A/U coaxial cable (silver-plated conductors) for a W1JR Broadband balun back in the summer of 1978, I paid a bit less than a dollar a foot – that same coax is now about $2.50 per foot in small quantities and the supplier refuses to guarantee the price for more than 30 days! Price increases for RG-8/U type coax have been somewhat less startling so far, but if I were planning a major new antenna installation this year, I think I would order the necessary coax before the soaring price of raw copper has a chance to filter down to the consumer level. Indeed, if you're thinking about buying any new Amateur Radio equipment, this would be a good time to make your final decision; the longer you wait, the more it is likely to cost.

Jim Fisk, W1HR
editor-in-chief

comments

FCC actions

Dear HR:

I agree strongly with your editorial in December *ham radio!* It seems that since a segment of the Amateur world behaves like CBers with regard to self-discipline, the lay public and the FCC tar us all with the same brush, and have in mind the slow attrition of all the non-voice, non-rag-chewing privileges. The part that really concerns me is the loss of our ability to be on the cutting edge of radio technology. The RTTY restrictions and the CW requirement matter are really shameful.

Your editorial has moved me to try to compose a literate letter of objection to send to my governmental representatives on this matter . . . what else should one do?

J.L. Ragle, W1ZI
Amherst, Massachusetts

Dear HR:

Congratulations on your December editorial; that says it all. I recommended to the Dayton Amateur Radio Association members at the last meeting that it was *must* reading if we are to understand the shakey position of Amateur Radio in the hands of the present FCC crowd.

I'm afraid that our success in Geneva is going to develop complacency in the ham ranks at a time when Amateur Radio, as we know it, is really threatened. I hesitate to think what might have happened if the FCC could have slipped through the no code deal. I also feel that if we accept this action by the FCC without challenging it, we will be in for more unhappy surprises in the future.

I sent copies of your editorial to my Congressmen asking that Congress take a look a what is happening in the FCC. It is good to see that there are others as irked about this sellout as I am. We may not get anywhere but they aren't going to turn me into a Citizens Bander without a fight.

Also sent a letter to the ARRL asking that they use the "freedom of information act" to get to the bottom of this. What we really need is a couple of aggressive young lawyers and let them dig. If the kind of information could be developed that I think is there somewhere, Congress or the Chairman would have to do some house cleaning.

Robert R. McKay, N8ADA
Editor, RF Carrier
Dayton Amateur
Radio Association

speed of light

Dear HR:

Harold Tolles, W7ITB, wrote a very interesting article on the speed of light which appeared in the January issue of *ham radio.* May I be another pair of eyes viewing the subject from a different point of view?

In 1675 the Danish astronomer Roemer determined the speed of light at 186,000 miles per second. Considering the crude equipment, it was remarkable that he came so close. Later this was translated into 300,000 kilometers per second. In appreciation of the longhand computations required, this was close enough.

However, in 1926, Michelson was able to refine this speed to 29979 ± 4 kilometers per second. Per Tolles' article, ITT determined in 1970 that the speed was 299,793 kilometers per second.

Joe Reisert, W1JAA, stated in the July, 1976, issue of *ham radio* that the latest revision was determined by several authorities that light travels at 299792.456 kilometers per second. I would be interested in the method used to arrive a? this datum. Was it averaged from several readings or were weighted averages used?

John Kraus, W8JK, wrote *The Big Ear,* in which he reaches out 12 billion light years into space. According to the spectrum red shift, there matter is traveling at 6/10 the speed of light. It is estimated at 16 to 20 billion light years away, objects are traveling at the speed of light. May I pose a question? If the center of the universe is at that distance, then we must be traveling through space at the speed of light, in which case the tip of my little finger would weigh several thousand million tons! It doesn't. Why not?

Another factor in astronomy, the Hertzsprung-Russel diagram, shows the main-sequence of the stars. From these data, it is possible for some smaller stars to be much older than the entire universe, according to the big bang theory. How come?

To answer these two questions, can it be that light slows down after traveling 10 billion light years? This decrement of speed might be the result of light traveling the great distance or of the intrinsic micro-nano-watt of the power left in the light beam. In other words: The speed of light is a variable constant!

This super accuracy is very fine, but practical radio communication and antenna design dictate approximate speed of light (and radio waves) to be roughly 299793 kilometers per second, or 186283 miles per second, or 11803 inches per megahertz for a full wavelength dimension.

Keith Rhodes, WB2AOT
Syracuse, New York

40-meter transceiver

for low-power operation

Design and construction details for a QRP CW transceiver operating in the 40-meter band

Have you seen the many articles that have been published on building simple receivers? Or how about the many QRP transmitter articles? How about a deluxe QRP transceiver that has a superior receiver and a healthy l-watt transmitter, both of which are VFO controlled? Read on.

The excitement of operating QRP accounts for the recent number of articles on the subject, but I feel we've not seen a good transceiver that allows portable, mobile, or fixed operation. So I came up with the rig presented here. I feel sure you'll be amazed by its performance.

The project had some problems getting off the ground. Every direct-conversion (DC) receiver I tried ended up in the scrap bin because it hummed, had tunable hum, overloaded easily, or had microphonics and/or all the above. Then I discovered the design presented here, which has none of these problems.

receiver

This direct-conversion receiver features:

1. Wide dynamic range (resistance to overload)
2. Excellent a-m signal rejection
3. No hum, tunable hum, or microphonics
4. VFO that operates at one-half the desired received frequency

The receiver (**see fig. 1**) has a grounded gate fet rf amplifier, 0101, which is used to bring the typical 40-meter-band noise floor (on a quiet day) above the receiver internal noise floor. This action provides approximately I.&microvolt input level for a 10-dB (S+N)/N. The rf stage is electrostatically and electromagnetically isolated from the VFO by T101 and T102 to prevent rf/VFO interaction. T102 provides rf VFO signals to a pair of detectors, each of which a complete detector: CR101, CR102, and CR103, CR104, which operate differentially.

Detector characteristics. The operation of this detector is the single most important feature of the entire transceiver." From previous descriptions of this detector, I have developed the detector in this rig, which solves the problem of VFO and rf intermodulation and provides translation voltage gain.

The intrinsic characteristics of this detector provide the features stated earlier, because the VFO operates at ***one-half*** the received frequency; therefore, a-m DSB signals contain a modulation envelope that cancels in each diode pair.

'This detector, which is sometimes referred to as a harmonic detector, has been described in an unbalanced configuration by others.

By John L. Keith, WB5DJE, 1633 Dell Oak Drive, **Garland, Texas 75040**

This operation is better understood if you consider that the diodes act as rf switches. When the VFO signal approaches its peak amplitude, positive or negative, it turns on a diode. Therefore, one diode in each pair turns on at every peak of the VFO signal. So to obtain an audio beat note, the incoming rf frequency must be twice that of the VFO. The mathematical expression that represents this detector takes on the form of a cubical parabola, which also verifies its inability to detect a-m signals (for which a square-law function is used).

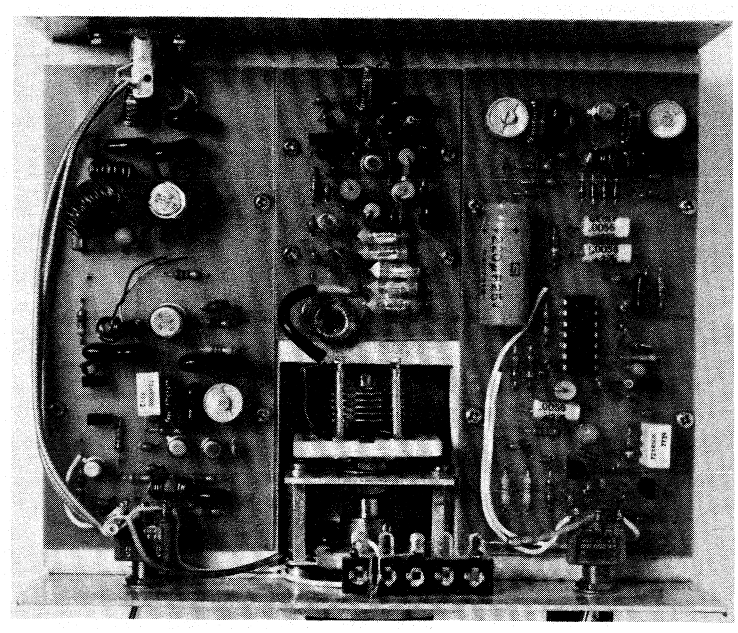
a-rn signal rejection. So why such a big deal about a-m signal rejection? No one can hear your signal under an a-m foreign broadcast station, right? You mighi be surprised — but the big deal is that this feature 1) eliminates tunable hum, 2) reduces static level, and 3) improves microphonic rejection. These parameters benefit from the a-m rejection because all have DSB a-m components that normally go *right through* a product detector. Out-of-band signal rejection is improved because intermodulation is very low in the rf amplifier, VFO, and detector output circuit.

Detector output is dc coupled (since no dc component exists at the detector output) to a differential audio amplifier, U101A, which provides a single-ended output and 46 dB gain. U101B and U101C are 800-Hz filters with a bandwidth of 200 Hz and a gain of 30 dB. The *Q* of these filters is selected to prevent ringing. U101D provides the last 35 dB of receiver gain, picks up the sidetone when transmitting, and drives the headphones.

VFO

The VFO provides an output between 3500 kHz-3590 kHz to transmitter and receiver. On receive the VFO frequency is used directly but on transmit it is doubled. Also in receive the VFO frequency is affset so that a station that returns your call will be shifted in frequency approximately 800 Hz, set by C209, so that it will fall in the audio filter passband.

For a change of pace try operating QRP in the 40-meter Amateur band. What is QRP? It's an operating mode that uses the minimum amount of radio-frequency power that will sustain communications. Some QRP stations use less than 1 watt of input power, sometimes even less than 500 milliwatts. Others use up to, say, 5 or 10 watts. The idea in the QRP world is to see how far you can conduct reliable radio communications with the least amount of radiated power. This article provides construction information on a QRP transceiver that will do a good job on 40 meters with only 1 watt output. The receiver is a notch above most circuits using the direct-conversion process. QRP operation on the Amateur bands today is a real challenge, especially on 40 meters. The circuit and information by WB5DJE will get you started. Good luck and have infinite patience. Editor.

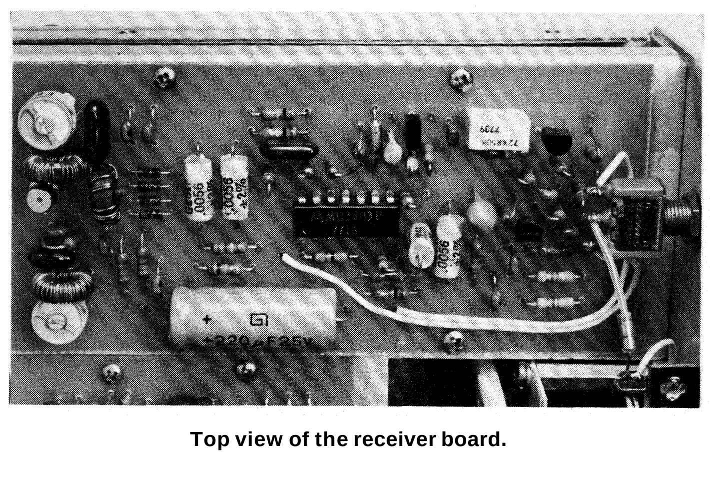


Inside top view of the transceiver, showing three PC boards. The miniature terminal strip secures the dial LED.

Description. The VFO is a Seiler type using a 2N4416 fet followed by an fet buffer and output amplifier. U201 in the VFO is a 5-volt, three-terminal regulator biased to provide +7 Vdc, set by R210. This regulator provides excellent voltage stability far superior to that provided by a zener. The VFO frequency holds within 10 Hz for input supply voltage variations between 15 and 9 Vdc.

The oscillator is very stable, although not temperature compensated, for changes in loading and mechanical vibration. The tank circuit is made of an SF material powdered-iron toroid and polypropylene capacitors. Dipped silver micas will work here but have poorer warmup drift because of the very small rf heating in their dielectrics.

I had an interesting experience with this VFO in the design stages. I started with a Colpitts oscillator,



Top view of the receiver board.

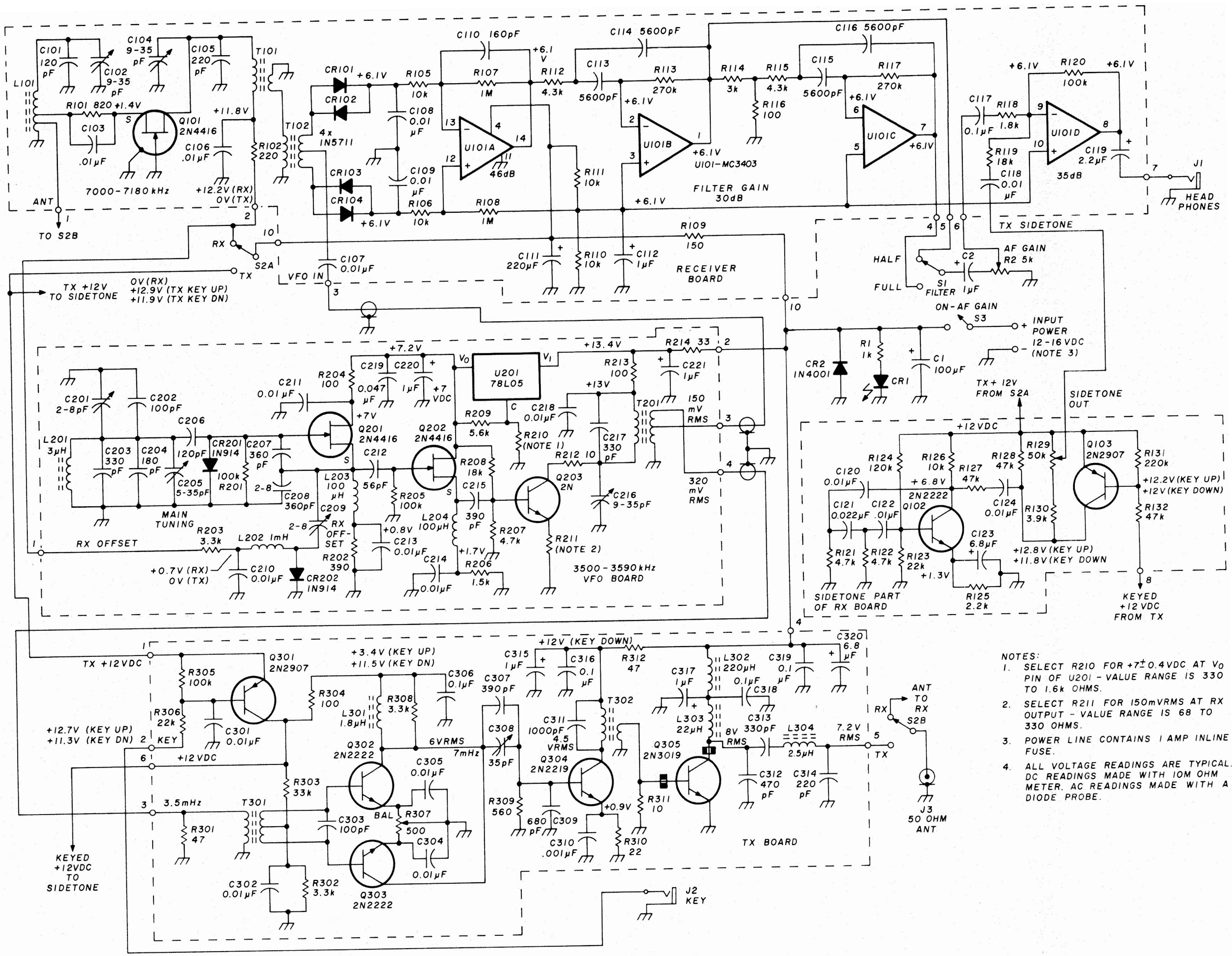
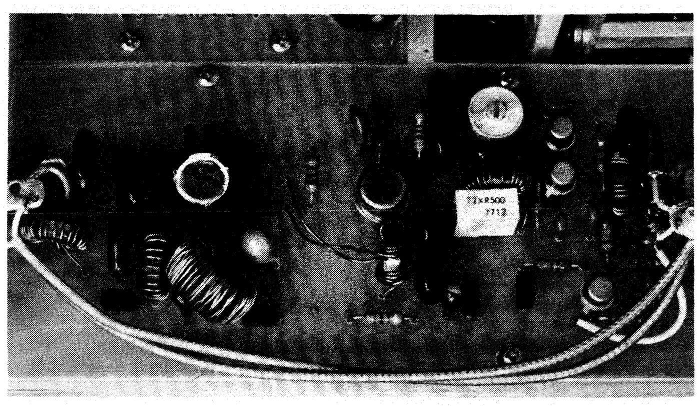


fig. 1. Schematic of the QRP transceiver. The receiver has some novel features not found in most direct-conversion circuits. The VFO is a Seiler type that provides an o put between 3500-3590 kHz. On receive the VFO frequency is used directly; on transmit it's doubled. The transmitter provides 1 watt output and uses transistor keying. As added operating aid, a sidetone circuit is also included.



Close-up view of the transmitter board.

which seemed to work very weii. However, when I keyed the transmitter it shifted about 650 Hz. After working on shielding and buffering, and still stuck with a 300-Hz shift, I threw it out and designed the Seiler, which, without any shielding, shifts only about 25 Hz when the transmitter is keyed and sounds as if it is crystal controlled.

Tank inductor L201 is mounted with a coating of polystyrene Q-dope to secure the turns and the inductor to the board. This prevents VFO frequency shift when the rig is vibrated.

table **1.** Coil and transformer winding data for the **low-**power 40-meter **CW** transceiver.

L101	3.5 *pH* (33 turns no. 28 10.3 rnml on T37-6 toroid core; tapped at 1 turn for antenna; tapped at 3 turns for Q101 source)
L201	3 *pH* (30 turns no. 28 10.3 rnml on T37-6 toroid core). Adjust number of turns and spacing to set center frequency and tuning range; coat with *Q-dope* when final adjustments are completed.
L301	1.8 *pH* (23 turns no. 26 10.4 rnml on T37-6 toroid core)
L302	220 *pH* *(64* turns no. 30 10.25 mm] on FT50-1 ferrite toroid)
L303	22 *pH* (20 turns no. 24 10.5 rnml on FT37-1 ferrite toroid)
L304	2.5 *pH* (27 turns no. 26 10.4 rnml on T37-6 toroid core)
T101	2.2 *pH* primary (25 turns no. 28 10.3 rnml on T37-6 toroid core; secondary is 6 turns no. 26 10.4 mm])
T102	1.3 *pH* *(5* trifilar wound turns no. 26 10.4 rnml on FT37-1 ferrite toroid)
T201	6.4 *pH* primary *(48* turns no. 28 10.3 rnml on T37-6 toroid core; secondary is 6 turns no. 26 10.4 mml, tapped at 3 turns for the receiver)
T301	1.3 *pH* (5 trifilar wound turns no. 26 10.4 rnml FT37-1 ferrite toroid)
T302	0.53 *pH* primary (13 turns no. 26 10.4 rnml on T37-6 toroid core; secondary is 4 turns no. 24 10.5 mm])

Notes T37-6 powdered iron toroid core (SF material) has 318 inch (10 mm) outside diameter, is rated at 30 µH per 100 turns.

FT50-1 ferrite toroid (Q1 material) has ½ inch (25 mm) outside diameter, is rated at 510 pH per 100 turns.

FT37-1 ferrite toroid (Q1 material) has 3/8 inch (10 mm) outside diameter, is rated at 425 µH per 100 turns.

Dial calibration. Before the coil is coated, the VFO tuning range should be set to calibrate the dial, with C201 set at mid position, by adjusting the number of L201 turns and spacing. Once this is set, calibration can be made by adjusting C201, for which a hole is provided in the bottom cover.

I selected a tuning range of 7000-7180 kHz (3500 kHz-3590 kHz VFO frequency) so that my dial would provide 10 kHz per 10 degrees of rotation. This isn't quite right, because the change is not perfectly linear; but the dial can be laid out with a protractor without having to mark it on the tuning capacitor, and the error will be only a few kHz. Just make sure it's correct at 7025 kHz above, which isn't that critical.

You can calculate or measure the actual change if you want a more accurate calibration across the entire band. Also you can increase the range if you desire. A handy equation for this is:

$$L = \frac{1 - \frac{\omega_o}{\omega_1}}{(\omega_o)^2 (\Delta C)} \qquad \textbf{(1)}$$

where L = inductance (henries)

$\omega_o = 2\pi f_o$, f_o being the lowest frequency of interest (Hz)

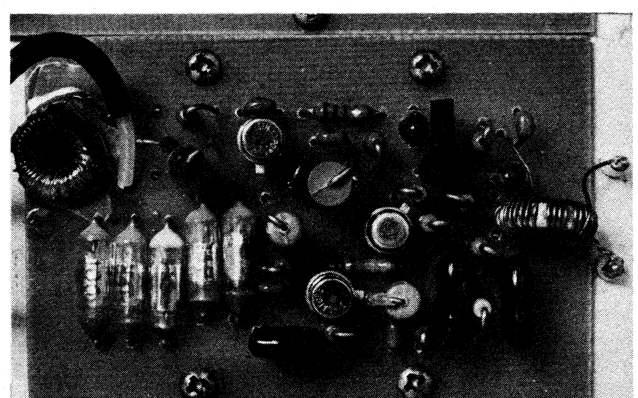
$\omega_1 = 2\pi f_1$, f_1 being the highest frequency of interest (Hz)

AC = change in capacitance available (farads)

Once the value of L is found, the total capacitance required for resonance is:

$$C_o = \frac{\frac{I}{(\omega_o)^2}}{L} \quad and\ C1 = C_o - \Delta C \qquad \textbf{(2)}$$

where CI is the amount of fixed capacitance required and includes the oscillator capacitive loading.



The **VFO** board. Note coax connection to variable capacitor.

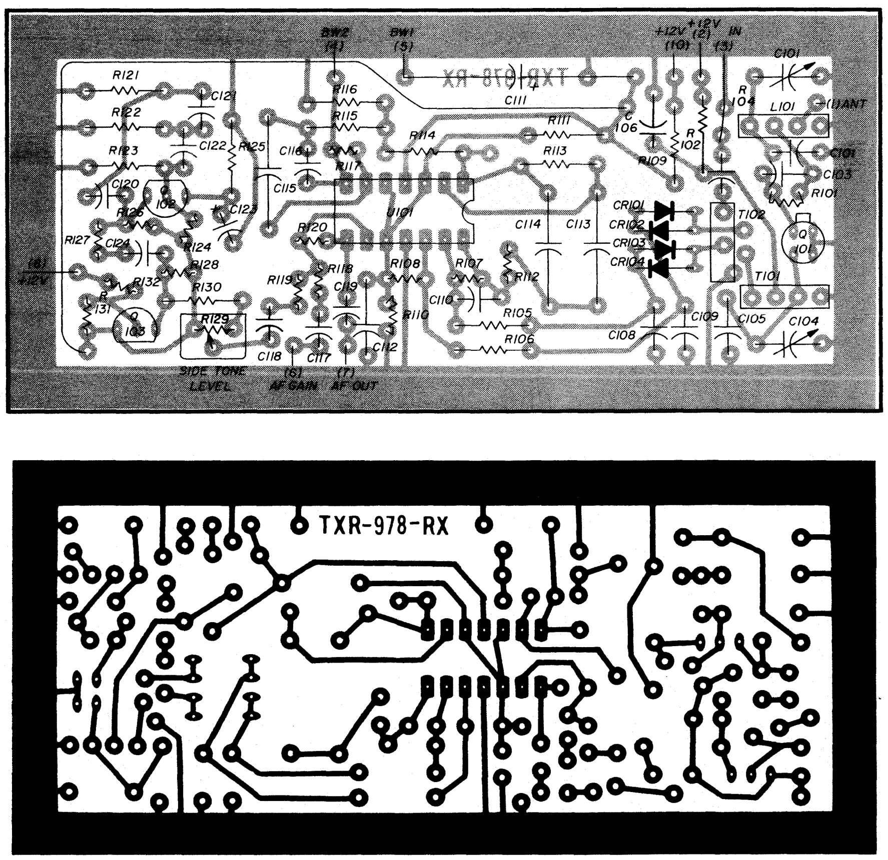


fig. 2. ***Above,*** **receiver-board parts layout.** ***Below,*** **receiver board, foil side.**

Output amplifier 0203 is operated class A into T201, which provides VFO energy to receiver and transmitter continuously. T201 is tuned with C216 to peak the VFO output in the center of the band. It operates with a *Q* that prevents excessive level variation across the band. The VFO output level should be 150 mV rms at the receive output. At this level the detector is optimized for Schottky diodes (for which germanium could be used), and the transmitter doubler is designed to operate at the level provided by the second output of 320 mV rms. (These readings were made with a diode probe.) R211 in Q301 emitter is selected to provide these levels. I prefer this method of level adjustment over dividing the VFO externally because it provides a lower noise floor.

transmitter

The transmitter is straightforward using transistor keying, a frequency doubler, a driver, and a power amplifier. The keying is accomplished by Q301, which turns on when the key is closed. However, Q301 has only +12 Vdc available when in the transmit mode to prevent keying the transmitter without an antenna. Also note that Q301 keys the +12 Vdc to only the frequency doubler stage, because the driver and PA operate class C and don't require keying of the dc supply.

Q302 and Q303 are connected as a push-push doubler with 180-degree base feed accomplished by trifilar-wound T301. R307 in the emitter circuits allows the 3500 kHz fundamental to be balanced out, so that the output waveform contains very little fundamental component. The doubler output is capacitively tapped down to provide the base drive for the driver Q304.

C308 tunes the doubler output and should be peaked in the center of the band. The doubler adjustments should be made carefully to ensure it's stable and operating properly. Do not peak R307 and C308 for maximum amplitude alone, but adjust them for a stable 7090-kHz output that contains a minimum amount of 3545-kHz energy.

Driver Q304 operates class *C* with some self bias, which should not be changed. The bias selected pro-

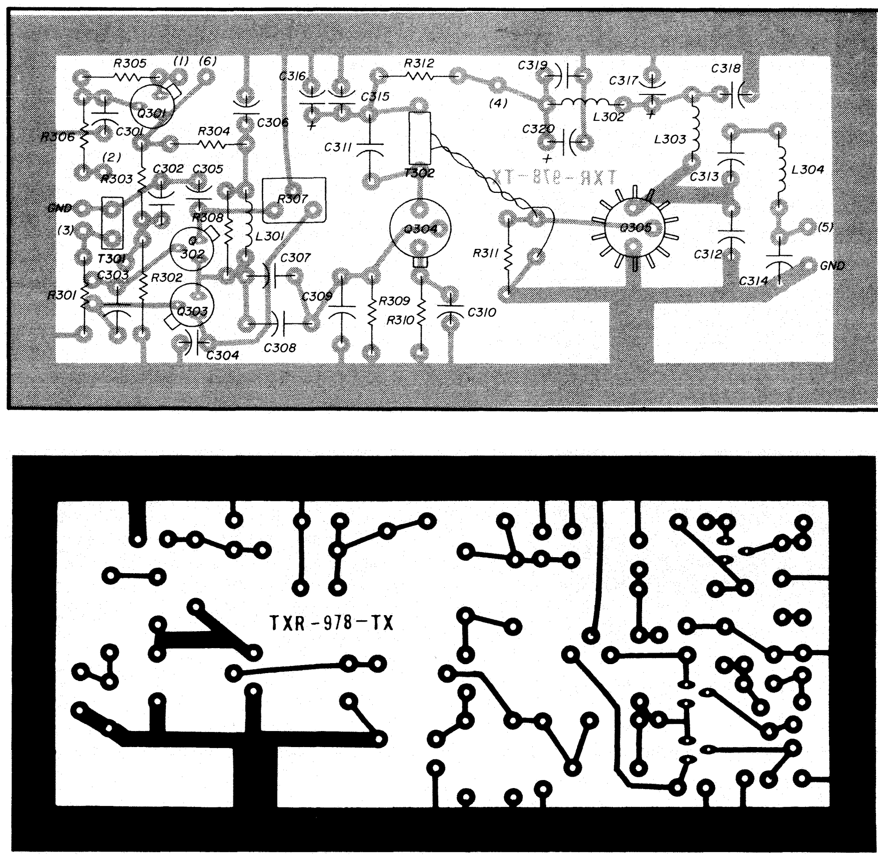


fig. 3. ***Above,*** **transmitter-board parts layout.** ***Below,*** **transmitter board, foil side.**

vides a clean optimum output for the drive level in use. T302 is tuned by C311 and provides a low-impedance drive to power amplifier 0305.

The power amplifier is a 2N3019, for which other 5-watt, 1 ampere transistors could be used. However, if the 2N3019 is not used, select one with an f_T of about 100 MHz. If the f_T is much higher, such as in a 2N3866, the possibilities of VHF components on the output are very great (TVI).

Some component-value selection can be made in the output-matching circuit if desired. The values shown were slightly changed from the calculated values by using a spectrum analyzer to optimize the 7000-kHz-to-harmonic-energy content. The values shown provided 1 watt at 7000 kHz with the second harmonic down 40 dB, the third down 50 dB, and other harmonics down 60 dB or better. The VHF harmonics were down better than 90 dB.

When in the transmit mode, power is also applied to the sidetone, which is keyed by 0301. R129 provides an independent adjustment level for the sidetone regardless of the af gain setting.

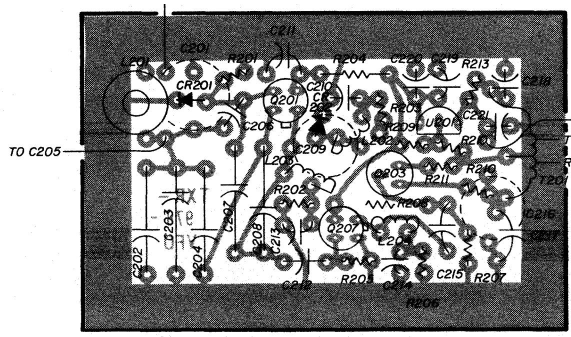
construction

The QRP transceiver uses PC-board construction.* Parts layouts for the receiver, transmitter, and VFO boards are shown in **figs. 2** through **4** respectively. Coil data are given in **table 1.**

General notes. I have built an enclosure for my rig so that I could have the size I wanted and accessibility to both top and bottom of the circuit boards. Other types of enclosures can be used without any problems since the board interconnects are not very critical. I do suggest that 50-ohm cable be used to connect the VFO to the receiver and transmitter as well as to connect the antenna circuitry. The audio circuitry is low impedance and shouldn't require shielded cable.

As you can see from the pictures, I did not shield the VFO. I found that on 40 meters it was not necessary. However, on higher frequency bands it would be a good idea.

*Many components for this transceiver are available from Radiokit, Box 429, Hollis, New Hampshire 03049.



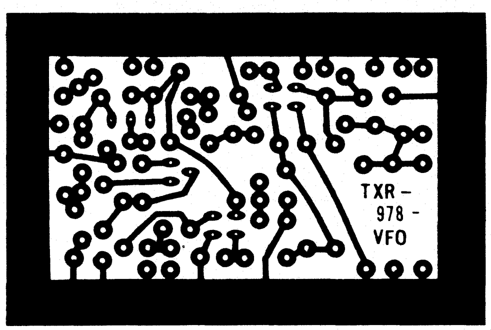


fig. 4. Above, VFO-board parts layout. Below, VFO foil side.

Making the VFO dial. I made the dial by cutting a piece of Plexiglas into a perfect 2-inch (50-mm) circle by pushing a file against the rotating disk.

I then cut a thin sheet of mylar onto which I lettered the markings with dry transfers. I then sprayed an adhesive onto the front of the Mylar and attached it to the *back* of the Plexiglas. This prevents fingernails from damaging the dial markings. To make the pointer, I cut a slit with a saw blade into a thin aluminum plate, which fits behind the dial, and I back-lighted the slit with a green LED. The dial mounts with two screws onto the Jackson Drive (a ball bearing 6:1 reduction drive).

Note that the transceiver is made up of three circuit boards. This allows you to choose the type of packaging that suits your needs — or your can build just the receiver or transmitter if you wish.

I think you will find this project to be well worth while if you have the QRP bug or think you might get it. I've worked coast-to-coast with this rig on 40 meters with very good signal reports.

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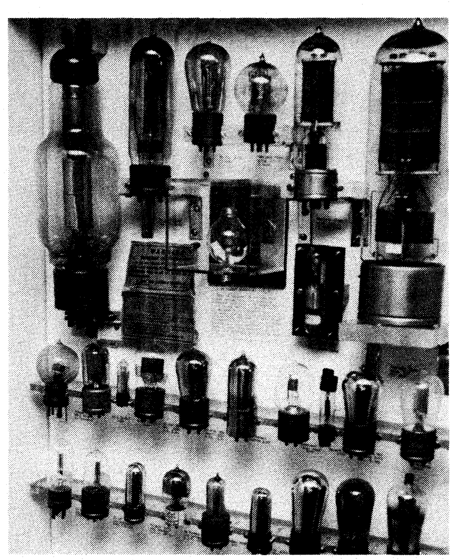
ham radio

Antique Radio Collector

Jim Fisk's editorial in the October ***ham radio*** seemed to be written for me, since I am in the position he described. Having been active in collecting for quite a few years, I have a rather representative collection of antiques, and I am still collecting interesting things. In recent years, I have reduced my collection to a manageable quantity, but the question remains: What will I do with it, when my own time comes?

Although neither of my sons is a ham, one is a physicist with a scientific company in San Diego, and heavily into electronics. We have often discussed the best way to handle my collection. Since he travels extensively — visiting universities in Japan, the British Isles, Europe and Scandinavia — he has rather intimate contact with the academic community. It is his opinion that giving one's collection to a university or college would offer little more probability of its being retained intact than one of several other possibilities. The staffs and administrations of universities change over time and the newer staffs may not value or protect collections as well as those to whom the collection was given.



A portion of W9LC's tube collection.

This phenomenon is true even in commercial establishments. A number of years ago, when I was working for WGN, an erratic personnel manager decided he needed more room, and without consulting the management, junked almost the entire file of 16-inch records of past programs. This was similar to those instances of junking valuable antique equipment, mentioned in the editorial.

My son has promised that my collection will be preserved, although I am not sure how this can be guaranteed. On the bottoms of the more desirable items I have attached notices, stating that these should be retained by the family.

If you learn what can best be done when the collector's final QSO has ended, I shall be most glad to hear.

Paul C. Crum, W9LC
Chicago, Illinois

recent developments in circuits and techniques for high-frequency communications receivers

The recent increase in high-frequency communication traffic and the present height in the sunspot cycle has further crowded the high-frequency spectrum. Because of the vulnerability of communication satellites to jamming and attack by missiles, use of the short-wave communications bands is expected to increase. Therefore, new receivers must have substantially improved large signal handling capability and better frequency resolution. The digital circuitry now in use has made it impossible to implement a number of mechanical solutions, such as tracking filters. A number of new approaches to improve and simplify shortwave receiver design will be presented in this article.

The work described here resulted from a research project and study for RCA Astro Division and a project now underway for the Naval Research Laboratory in Washington. In both cases, new ways had to be developed to increase the performance of a communications receiver; the following areas are of great importance:

- Good input selectivity
- Ultralinear amplifiers
- High level mixers
- Low distortion Thompson VHF crystal filters
- Choice of AGC
- Linear detectors
- New low noise synthesizers
- Microprocessor support

input selectivity

Because of electronic switching requirements, all modern receivers are double-conversion systems with a first intermediate frequency between 40 and 100 MHz. The second i-f is kept as low as possible; frequencies from 10.7 MHz down to 30 kHz are used. As a rule of thumb, the first i-f should be above 60 MHz and slightly more than twice the highest reception frequency. The first i-f filters should consist of a crystal filter with good selectivity and low insertion loss; 72.03 MHz for a second i-f of 30 kHz or 72.455 MHz for a second i-f of 455 MHz are recommended.

Modern mechanical filters are available with shape factors equally as good as crystal filters commonly used at about 10.7 MHz with superior group delay and pulse response. Siemens (West Germany) manufactures 30-kHz mechanical filters to suit all practical purposes; Collins and AEG Telefunken make 200-kHz mechanical filters with similar performance which are slightly less expensive.

The selection of 72...MHz first i-f also permits the use of a second local oscillator at 72 MHz, and modern technology permits the design of low noise, low aging 72-MHz crystal oscillators. **Fig. 1** is a block diagram of such a modern concept where, instead of the usual 5-MHz input for the sythesizer, an internal 72-MHz crystal in a proportional temperature controlled oven is responsible for the stability and can be phase locked against an external 1 MHz standard which has lower sideband noise requirements. Traditionally, 5 MHz crystals have been used because they combine low aging and low noise.[1,2]

In some cases where high-frequency receivers are

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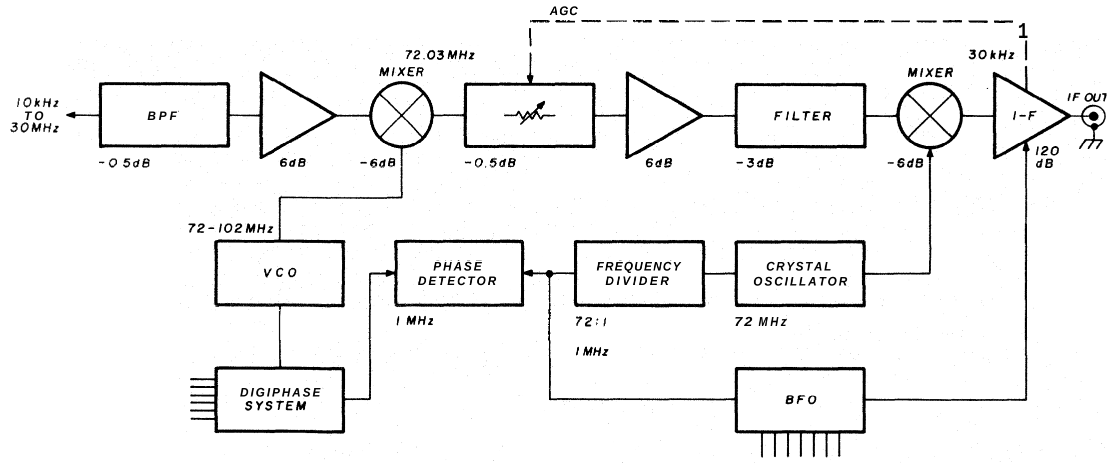

fig. 1. Block diagram of a modern high-frequency communications receiver showing the use of a 72-MHz first i-f and oven stabilized 72-MHz crystal oscillator as the reference oscillator for the phase-locked loop.

used in the vicinity of transmitters, an electronically tuned tracking input filter is required; these preselectors cannot be built with tuning diodes. **Fig. 2** shows such a preselector which uses PIN diodes as switching elements for the capacitors and inductors, and is controlled by a "look-up" table under microprocessor control. This type of input filter has been successfully used for shipboard applications where several 1 kW transmitters were present. **Fig.** 3 shows a newly developed input stage for a similar application in the vhf band; **fig. 4** is a graph of the selectivity curve. It is apparent that this type of input filter exhibits the best image suppression and local-oscillator suppression of its kind with very few components. The mathematics of this filter are presented in reference 3; it has been used in a number of production receivers.

ultralinear amplifier

Since some receiving systems require a noise figure of less than 10 dB, even on the high-frequencies, and some requirements call for very low oscillator radiation through the antenna, it is unavoidable to use an antenna preamplifier with less than 10 dB gain. These amplifiers must be designed to combine very low distortion and low noise figure. A method called "noiseless feedback" has been developed in both Germany[4] and the United States.[5] While conventional feedback techniques use resistive elements like an unbypassed emitter resistor for current feedback and a resistor from input to output for voltage feedback, these methods introduce additional noise. The noiseless feedback technique permits independent choice of input and output impedance and power gain while maintaining the transistor's inherent noise figure. The circuit as shown in **fig. *5*** should be built in a push-pull configuration to further reduce the second-order intermodulation distortion products. In selected cases, it has been possible to achieve an intercept point of 80 dBm with a noise figure of about 3 dB.

high level mixers

Even without the use of a preamplifier, the mixer has always been the weakest link in the chain because the mixing action requires a prescribed nonlinearity and the third-order term will always be apparent. A number of efforts were made to build an

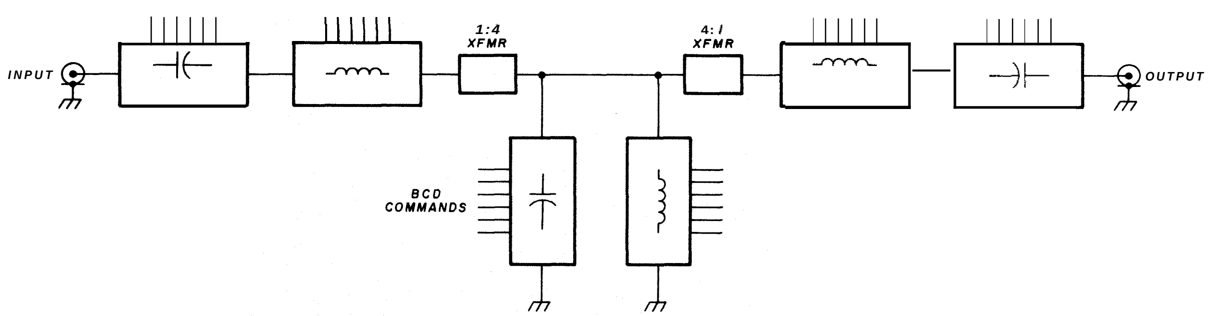

fig. 2. Preselector is microprocessor controlled, using PIN diodes to switch inductors and capacitors in the filter.

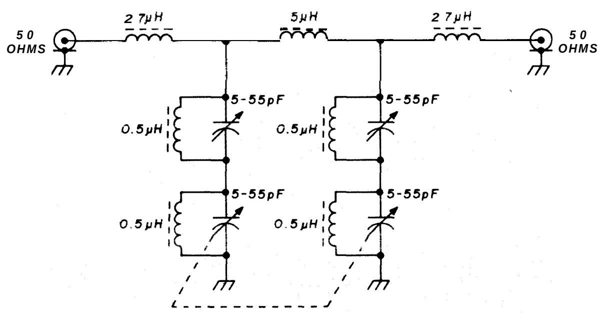

fig. 3. Vhf input filter covers the range from 118 to 136 MHz. Tuning capacitors are 5-55 pF trimmers. Selectivity curve is shown below in fig. 4.

active high dynamic mixer like the one shown in fig. **6**.[6,7] These mixers are extremely sensitive to load changes and cannot be easily built totally symmetrically at the higher frequencies. The fet version of these mixers recommended by Siliconix requires fairly high local-oscillator drive. Since the input of these fets is purely capacitive, the cable requires a termination into 50 ohms and, ultimately, the same drive power as a passive mixer.

A novel high-level mixer circuit developed for the Rohde & Schwarz HF1030 shortwave receiver is shown in fig. **7.** Here a push-pull version of two double balanced mixers is being used; since these mixers require perfect termination, an fet cascode arrangement with a 50 ohm termination is used. Previously, fets in grounded-gate circuits had been recommended, but this has two drawbacks: the drive impedance of about 50 ohms results in a worsening of the noise figure because the fet wants to see a higher drive impedance, and it was next to impossible to apply feedback to such a circuit (the circuit would also oscillate above 1000 MHz). The circuit in **fig. 7** has the advantages that the increased drive impedance of 200 ohms provides a much better noise figure; and the use of feedback more than compensates the differences between the grounded-gate and grounded-source configuration. While more gain can be achieved in the cascode arrangement, the tendency to uhf oscillation has not been observed.

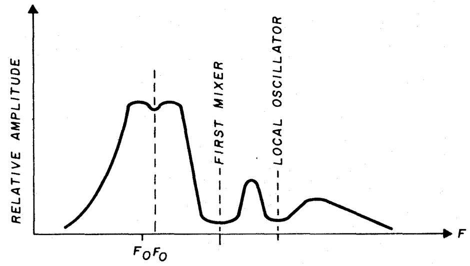

fig. 4. Selectivity curve of the vhf input filter of fig. 3.

A typical noise figure of 2 dB, intercept point of +35 dBm, and gain of 15 dB is possible with this stage in the frequency range from 40 to 120 MHz. The combination of this stage with the mixer maintains its intercept point, reduces the gain to about 10 dB, and reduces noise figure to about 8 dB. A PIN diode attenuator is incorporated in the circuit and will maintain its input and output impedance very closely and provide an agc range of 45 dB with insertion loss of less than 1 dB. It would have been possible to apply agc to the cascode, but measurements indicate that the PIN diode attenuator gives better dynamic range.

Thomson vhf crystal filters

Immediately following the mixer and its amplifier, it's necessary to provide as much selectivity as possible. The type of filter and its bandwidth depends on the receiver's application. In some cases, such as fast data transmission, it is desirable to use Thomson-type8 crystal filters with bandwidths from ±3 kHz to ±8 kHz. The basic disadvantage of a Thomson filter is skirt selectivity, and it may not be possible to obtain 80 dB skirt selectivity 60 kHz away from the center frequency. In those cases where perfect pulse response and group delay performance are required, a higher second i-f between 200 kHz and 500 kHz must be chosen.

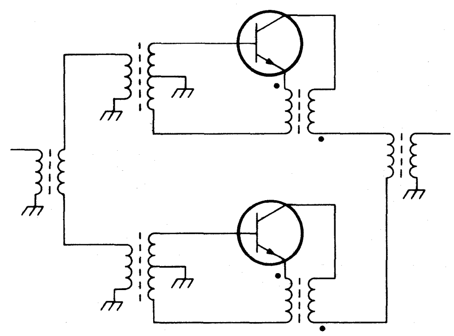

fig. 5. Basic circuit for an rf amplifier with feedback which enhances strong-signal handling ability without increasing noise figure. In selected cases, it has been possible to achieve a +70 dBm intercept point with 3 dB noise figure.

Thomson filters have been recently developed and are available commercially." Such filters exhibit less than 3 dB attenuation even with 10 crystal resonators and have an intercept point in the vicinity of 35 dBm. Applications that require a higher intercept point should be based on helical resonators. These reso-

"Communications Consulting Corporation, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458.

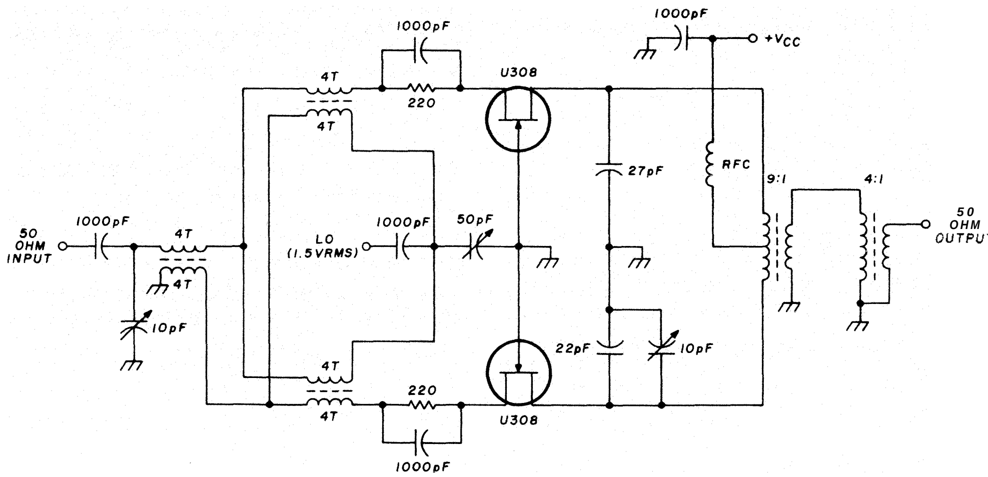

fig. 6. Circuit for a high-level balanced fet mixer. Local oscillator requirement is approximately 1.5 volts rms.

nators can have Q of 1000 or more and will not introduce any noticeable distortion.

One of the most critical characteristics of a receiver is its response to signals of varying strengths – it is not sufficient to optimize the filter for pulse response – the agc system must be present: fast agc is desirable for a-m reception while fast attack and slow decay time constants are required for CW and SSB. Fm requires somewhat intermediate time constants as the limiter is supposed to cancel on a-m components, but it is always desirable to have an S-meter to provide information on the input signal.

agc choice

It appears that most currently manufactured high-frequency receivers suffer from good agc; in most cases the agc is too slow and the attack time produces unpleasant overshoot in the audio circuits. The reason for choosing the insufficient attack time is the requirement of avoiding a peak detector for short pulses which would "hang up" the receiver during the decay time. A better way of handling this problem is through the use of a symmetrical audio limiter (clipper) which permits the audio to rise by about 10 dB over the audio under agc control. This prevents the unpleasant audio bursts and accepts 30-40 ms attack as a perfect choice. A number of agc circuits have been published in the past and described in references 9 and 10.

linear detectors

A-m detectors are frequently required to have very low distortion and because of this, the i-f level must be kept at a very high level. Because of gain distribution, this is not necessarily very desirable, and a feedback a-m detector as shown in **fig. 8** is ideal. The output distortion is substantially less than one per cent up to very high modulation percentages; it

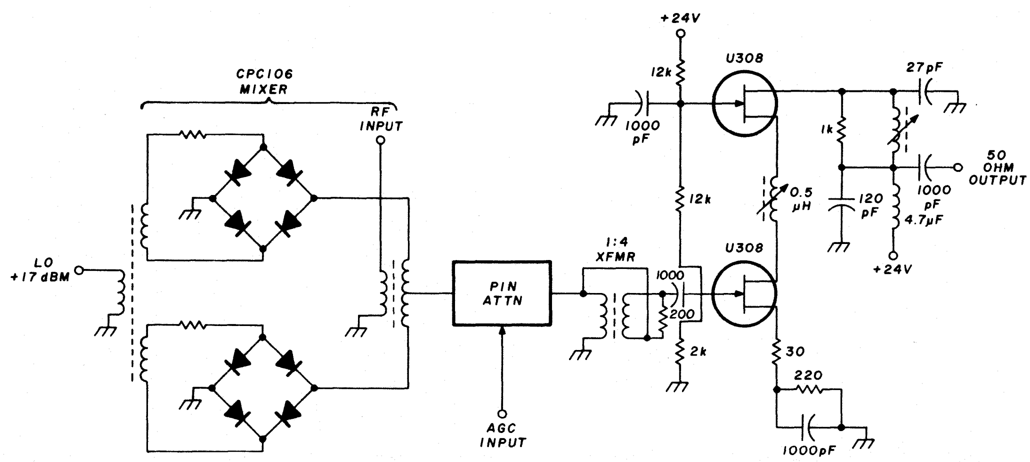

fig. 7. High-level mixer stage used in the Rohde & Schwarz HF1030 communications receiver offers noise figure of 2 dB with intercept point at +35 dBm.

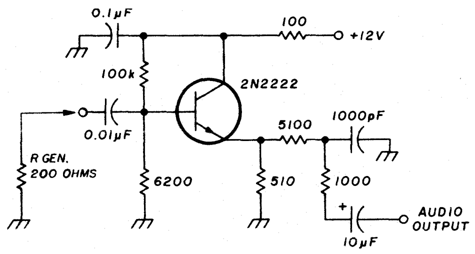

fig. 8. Linear detector is highly recommended for a-m.

should be driven from an impedance of less than 200 ohms which can easily be provided by using an emitter follower.

low noise synthesizers

I previously discussed the impact of nonlinear effects like intermodulation distortion of second, third, and higher order, and will now consider reciprocal mixing or blocking. The so-called blocking effect (which is commonly confused with receiver desensitization) is a result of poor sideband noise performance of the local oscillator.[11,12] Synthesizers are traditionally built with a compromise of three parameters in mind: noise sideband performance; spurious response; and settling time. Some military applications require a switching time of less than 10 microseconds, but practical applications can live with about 1 millisecond switching time. Most current synthesizers are multiloop synthesizers using the techniques described in references 13, 14, and 15. The recently developed *digiphase* system[16] or fractional division N system offers new capabilities and

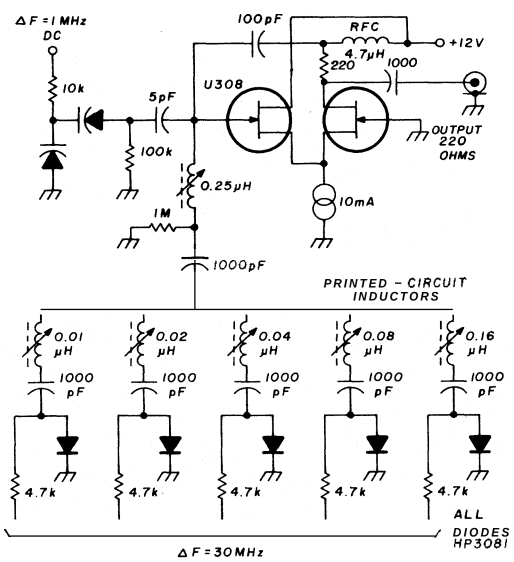

fig. 9. Ultra low-noise oscillator. Tuning range is held to about 1 MHz; coarse steering is accomplished with a *lookup table* under microprocessor control.

substantially higher resolution. The digiphase system is used by Hewlett-Packard[17] and Racal[18] and offers a number of advantages. While the presently published digiphase systems suffer from the noise sideband limitations of this technique, an improved version of this would use the phase locked loop not only as a control mechanism for almost infinite resolutions but also for suppressing these sidebands. If it were possible to build a VCO and use it with a very narrow loop bandwidth of less than 500 Hz, the unwanted sidebands of the digiphase system could be kept under control; if the tuning range of each oscillator could be kept extremely small, the added noise from

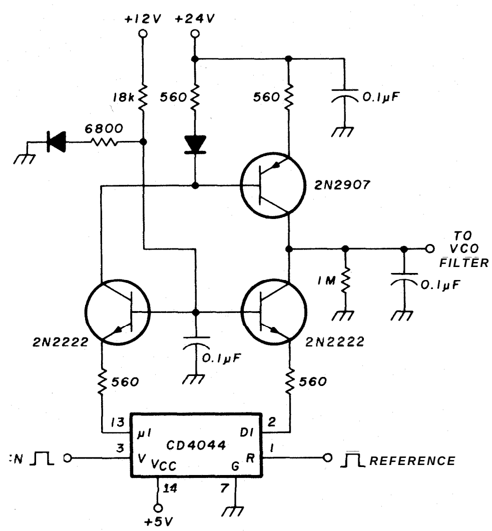

fig. 10. Low noise phase detector for use up to 1 MHz.

the tuning diode would not have to be taken into consideration. **Fig. 9** shows such an oscillator where the tuning range is held to about 1 MHz while the coarse steering is done by a "look-up table" under microprocessor control. Such an oscillator exhibits a noise figure of about 90-100 dB/Hz 1 kHz from the carrier and 140 dB 20 kHz from the carrier.

In the past, a number of discrete oscillators were used for this purpose, but since it may require several hundred milliseconds for the oscillator to settle before it is under control of the phase lock loop, inductor switching is fast enough that it does not degrade the switching performance. Such low noise systems also require very good phase frequency detectors and dc amplifiers; **fig. 10** shows such a circuit that is recommended for use up to a 1 MHz reference frequency.

microprocessor support

Microprocessors can help in a number of ways to improve the performance of an "imperfect" receiver. As mentioned earlier, a microprocessor may be used to control the antenna preselector, pre-steer the sythesizer oscillators, or programmed for sweeping and frequency hopping. A large number of channels can be stored and selected – the microprocessor quickly switches the synthesizer. Thus new applications for automated systems become feasible.

Such a system has been built in a joint development with RCA Astro Division for a satellite sounder system in which a transmitter generates bursts and the receiver, acting as a phase-coherent radar detector, analyzes the signal and provides information about the various layers of the ionosphere.

conclusion

A number of circuit developments have been shown which will substantially improve high-frequency receiver design. Some of these circuits have already been incorporated in equipment. In some cases only the highlights have been shown, and it is apparent that further improvements are possible. Because of this, I am interested in the exchange of ideas and would appreciate your comments and recommendations.

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ham radio

More Details? CHECK—OFF Page 94

log-periodic fixed-wire beams for 40 meters

Extensive tests with overseas Amateurs have resulted in an LP antenna with excellent characteristics and performance

During the testing of the various 75-meter antennas described in an earlier article,' QRN hadn't been too bad arid many overseas contacts with New Zealand resulted in much useful data on antenna performance. During June, 1978, however, propagation conditions deteriorated and QRN on 75 meters became so bad that many of the morning DXers (usually on 3808 kHz) moved to 40 meters.

Because the QRN on 40 meters was generally less than that on 75 meters during the DX window (1000-12000UTC), I decided to remove the three-element 75-meter beam[1] and replace it with a similar beam for 40 meters, also using an optimum LP design.

40-meter LP design

Using W6PYK's LP design data,[2] I built a 40-meter LP using taper factor $\tau_{,} = 0.95$ and spacing factor $\sigma_{,} = 0.18$. This beam would be about 18 meters (60 feet) above ground, or almost one-half wavelength on 40 meters, so it should give about the maximum gain possible. The design was a four-element truncated log periodic with a boom length of 22 meters (71.8 feet). Complete dimensions and VSWR response are shown in **fig. 1.**

I'd hoped to make this 40-meter LP a five-element array, but a tree near the center of the forward-element end of the antenna prevented extending the feeder to include a fifth element. However, two trees

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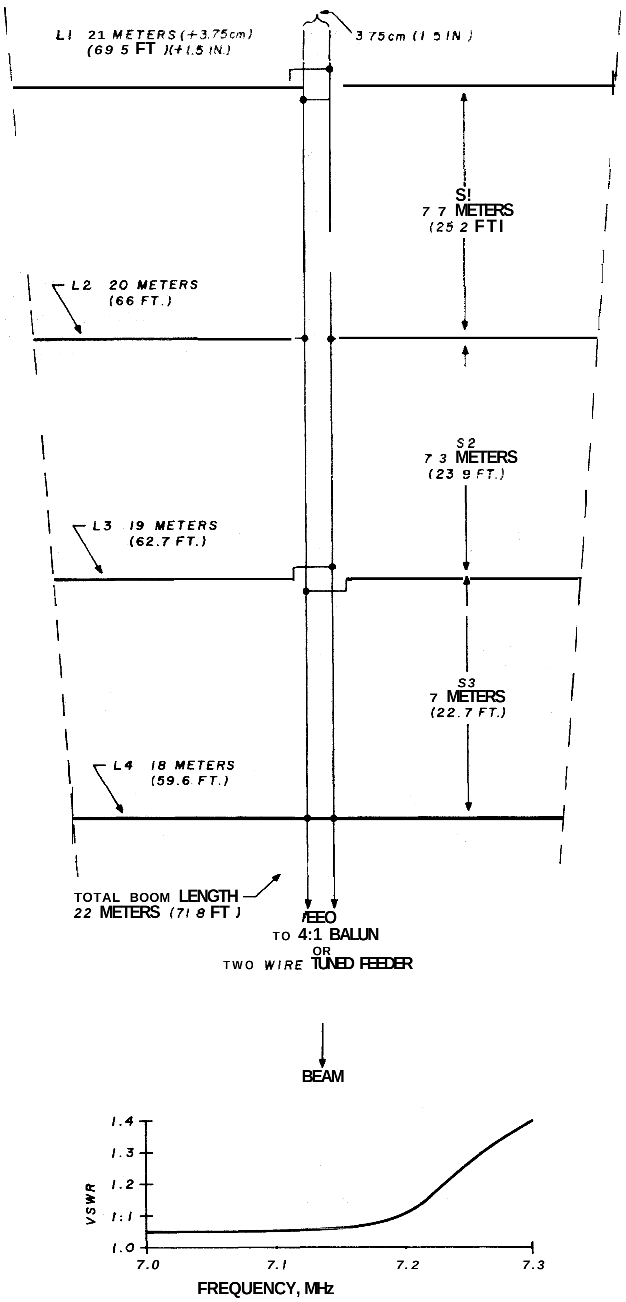

fig. 1. Four-element truncated LP antenna for 40 meters designed from data provided by W6PYK (reference 2). Curve showing VSWR is also provided.

to the sides allowed the addition of a parasitic director as shown in **fig. 2.** Paul, W6PYK, estimated that this director should add about 1 dB additional gain."

Fig. 2 also shows the method of suspending the 40-meter LP. (Also added later was a 40-meter dipole to the side and in line with the director.) This dipole was used as a standard for comparison with the beam. Both antennas were about the same height above ground, exactly parallel, and oriented broadside to the west.

"Before considering the addition of this director element in your design, note that 1 dB represents a power gain of 1.26. Note also that, to double the power, a gain of 3 dB is required. Considering the fact that a 1-dB increase in signal strength is virtually imperceptible at the receiving end of a radio circuit, adding the director hardly seems worthwhile. Editor.

test antennas

For the 40-meter tests I used four antennas for direct comparison with the 40-meter west beam illustrated in **figs. 1** and **2.** The comparison antennas were:

1) a 40/75 meter trap dipole sloper suspended over a pond;

2) a four-band Hustler trap vertical mounted on the roof of my house at about 9 meters (35 feet) above ground. Radial elements were used in the ground system;

3) the dipole shown in **fig.** 2;

4) a 40-meter LP-Yagi consisting of seven elements directed north.

The north-oriented LP-Yagi deserves special mention. I've had many requests for information about its design and performance.

40-meter north beam

This antenna had four driven elements and three parasitic directors. Boom length was 29 meters (93.8 feet). Height above ground was only 12 meters (40 feet). If the height above ground could have been increased to one-half wavelength, and if the antenna could have been beamed to the northeast, it probably would have been a good DX antenna for Europe. However, trees on my property aren't properly spaced for that direction.

The 40-meter north beam was constructed in an inverted-V configuration, with the center supported by a nylon line between two trees. Element ends were supported by two side catenary lines attached to trees at either end. An inverted-V configuration is shown in reference 3. This design is used by a number of commercial LP-antenna manufacturers.

The north beam was aimed about 90 degrees north of the west beam. It was interesting to switch from west to north when monitoring ZLs and VKs on 40 meters. At times, these stations were almost nil on the north beam but were received strongly on the west beam, which demonstrated the side attenuation of the north beam.

construction notes

The construction information presented in part 1 and its references apply to the 40-meter designs shown here.

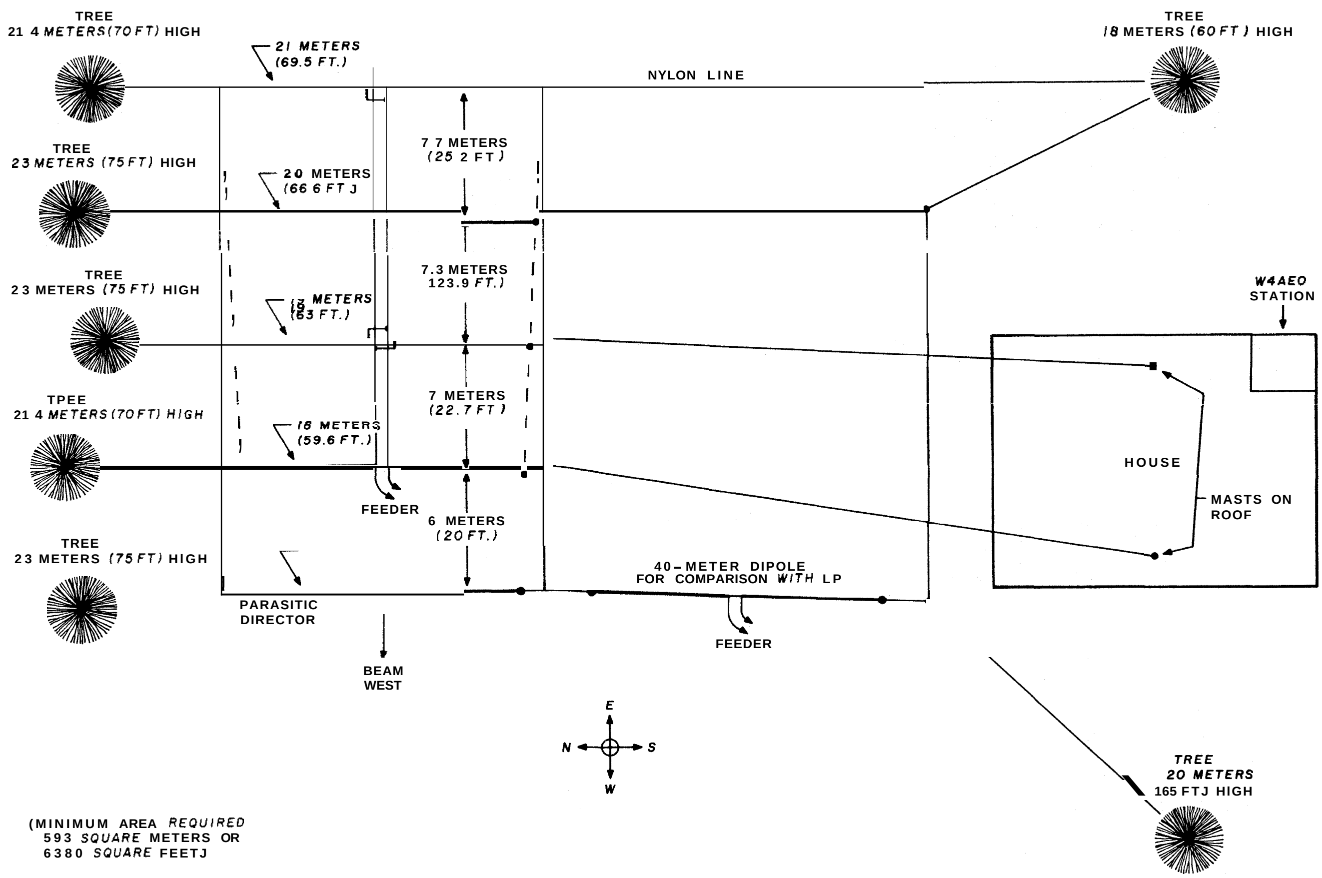

fig. 2. Author's 40-meter LP west beam using four driven elements and a director. A dipole was added for comparison in the 40-meter tests described in the text.

My 75- and 40-meter LPs required no side catenaries because enough trees and other supports were available on either side of the antenna elements for halyards, which made construction and suspension simple and easy — each antenna element could be adjusted separately for proper tension and alignment.

The two lines shown on either side of the 40-meter LP (fig. 2) aren't support catenaries; they're merely spacing lines of nylon to keep the ends of the elements parallel. This gives the same spacing as determined by the center feeder.

If you don't have trees available for supports, you'll need four masts or towers and you'll need two side catenaries to support the one- and two-element sections between the rear (no. 1) element and the forward element.

Feedline considerations. Little information has been published on the spacing of the two-wire intra-array feedline for high-frequency LP wire beams, so I tried various spacing distances from 3.75 to 15 cm (1½ to 6 inches). All LPs constructed here since 1975, including the 75- and 40-meter arrays described, have used a feedline spacing of 3.75 cm (1½ inches) and no. 16 (1.3 mm) insulated wire.

With this wire size and spacing, the feedline impedance appears to be near 450 ohms, which is about right for use with a 4:1 balun feeding 52-ohm coax. The driving-point impedance appears to be approximately one-half that of the feedline characteristic impedance for LP arrays.

Flexible stranded feeder wire is desirable so that the two wires will be straight and parallel with little fore and aft tension required on the center feeder. Insulated wire is used should the two wires touch in a high wind. I use trees for supports, so it's important to reduce element and center-feeder weight to a minimum. Reference 4 gives suggested feed methods for the high-frequency LP beams tested here and may be of interest if one of these LPs is to be assembled.

If you don't want to make the two-wire feeder you might consider the 450-ohm, low-loss, open-wire TV line offered by Saxton Products.* This line has been used for years by Amateurs for feedlines and stubs. It uses no. 18 AWG (1-mm) wire spaced at 25.5 mm (1

*Saxton Products, Inc., 215 North Route 303, Congers, New York 10920 (Catalog no. 2500 or C-4-500-6).

inch) and has molded insulators spaced every 153 mm (6 inches). It's available in standard lengths of 30.5, 76, and 153 meters (100, 250, and 500 feet).

I've received inquiries as to whether standard 300-ohm TV line (ribbon) can be used for the intra-array center feedline. The answer is absolutely no. This is because the 0.82 velocity factor of TV line would not be compatible with the required element spacing, as given by the LP formulas. To confirm this I removed the two-wire center feeder from one of the LPs here and replaced these sections with a good grade of 300-ohm TV feeder. The LP immediately showed a loss in gain, both on transmission and reception. Therefore, some types of two-wire open feeder (air dielectric) ***must*** be ***used.*** A velocity factor of at least 95 per cent or better is recommended, which rules out any solid-dielectric 300-ohm feeder, including the tubular-shaped 300-ohm uhf "low loss" TV line (velocity factor of 0.82).

W6PYK mentions the requirement of air dielectric "to be used to prevent excessive phase shift within the array feed. Any other dielectric has the effect of increasing the spacing factor, *a,* in a complex manner."[2]

Now, the above remarks don't rule out the use of 300-ohm solid-dielectric line between the 4:1 balun and the feed point of the intra-array center feeder (LP feed point at the short element, or front of the array). I've used this method of feeding many of my LPs; some use 31-61 meters (100-200 feet) of good grade 300-ohm TV line.

Most of my LPs are supported by trees, so weight must be kept to a minimum. The weight of a 4:1 balun plus the weight of RG-8/U, or even RG-58/U, coax would cause the front end to sag, resulting in a height loss of the forward end (above ground).

Even the best, or rather highest gain, LP used here to date and described in reference 6 was fed by about 76 meters (250 feet) of 300-ohm TV line between this 17-element LP feed point and the 4:1 balun, which was located at about 3 meters (10 feet) above ground to the rear of the LP. From the balun I used RG-8/U coax, buried to the station. The 300-ohm feeder was suspended from the forward element and draped under the full length of the 17-element array.

Insulators. I've been unable to locate four-hole "off-the-shelf" insulators suitable for the two-wire center feeder-spacer insulators, so homemade Lucite insulators[5] are used. Reference 5 also shows the best method for securing these insulators to the open-wire feedline as well as an assembly sketch of a seven-element LP showing the transposition method of feed to alternate elements.

If the open-wire TV line described above is used, the homemade Lucite center insulators can be replaced by standard 64-mm (2.5-inch) ceramic or porcelain ribbed insulators. These insulators are available from dealers selling antennas for shortwave listeners.

The two outside ribs of these SWL-antenna insulators are spaced at about 25.5 mm (1 inch), and the 450-ohm TV line can be secured to, and suspended below, these insulators. The two insulator holes secure the element centers. Connect short jumper wires between element-center ends and the feedline.

As I mentioned in previous articles, small strain insulators (Johnny Ball) are suggested for the center and end insulators used on the long, rear element (S1) and the short forward element.

a higher-gain LP

Should you have the available space and necessary supports and want a 40-meter wire beam having a gain of 10 dB over a dipole, you can build a monoband LP giving this gain. Referring to W6PYK's article2 **table 1** ($B = 1$), and using $\tau = 0.972 - 0.978$ and $a = 0.180 - 0.181$, will give about maximum gain for an LP.

Referring to the four-element 40-meter LP described above, **fig. 1,** for which I used ($\tau = 0.95$ and $a = 0.18$); this beam can be extended to seven elements and will require a length of only 40.7 meters (133.4 feet). This will, of course, increase gain and bandwidth over the four-element model tested here, which was only 22 meters (71.8 feet) long. Thus by using an open space about 30 meters (100 feet) wide by 46 meters (150 feet) long in the desired beam direction, an excellent 40-meter wire LP beam can be erected. **Fig.** 3 illustrates this LP with dimensions for element lengths and spacing.

If the length of the open space can be extended to about 69 meters (225 feet), a 40-meter LP beam having 10.6-dBd gain can be erected, as shown by W6PYK's article. This requires ten elements. Parameters are: $\tau = 0.978$ and $a = 0.181$, resulting in an array length of ($/\lambda = 1.48$), or overall LP length of $\lambda O = 984/7\ (MHz) = 140.6\ feet \times 1.48 = 208\ feet$ *(63.39 meters)* boom. It's assumed that this array would be at least 18 meters (60 feet) above ground to provide maximum possible gain.

summary of 75- and 40-meter LP antenna tests

These tests were made to determine if there is any type antenna or beam best suited for long-haul, multi-hop DX on 75 or 40 meters.

At my location, the last 75-meter LP, designed for 3808 kHz with $\tau = 0.94$ and $a = 0.175$ (reference 1), appeared to be the best of the various beams tested. Second best were the first 75-meter LPs and the

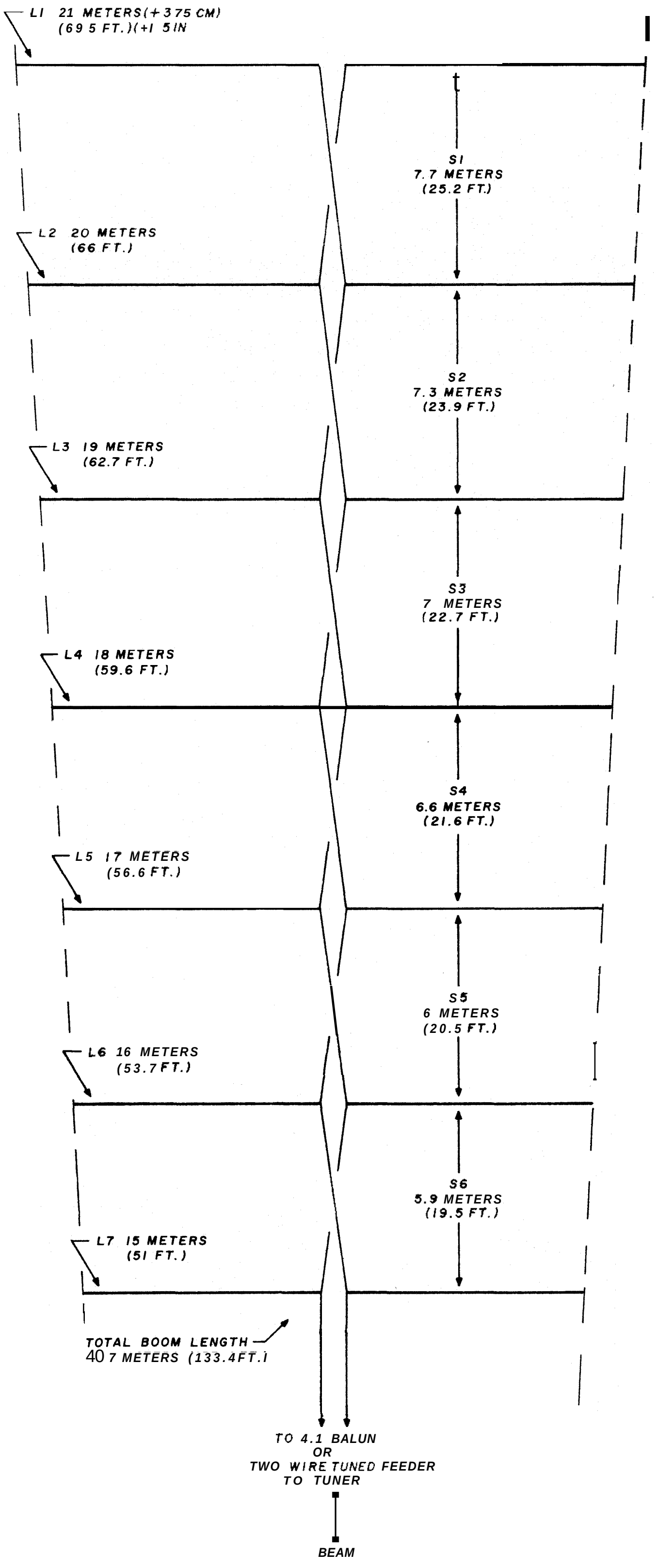

fig. 3. Seven-element LP for 40 meters with improved performance. You'll need an open space about 30 meters (100 feet) wide by about 46 meters (150 feet) long in the desired beam direction.

Yagi. These were compared with the more common antennas.

The LPs and the Yagi were the only unidirectional beams tested. There was little difference between the LPs (prior to the last) and the Yagi; however, the Yagi had the disadvantage of smaller bandwidth and it didn't cover the entire 75-80 meter band, which the LPs did. The Yagi was used only for a short time before being destroyed by lightning.

The other 75-meter antennas tested were nongain. However, some were bidirectional, as mentioned in part 1, and were thus no more than 50 per cent effective because of half the power, or radiation loss, in the undesired direction.

One of the three delta loops performed fairly well and was used throughout the test. One of the verticals was also used during most of the test; however, both of these antennas were generally about 10 dB below the LPs.

The 75-meter beams were all less than one-quarter wavelength above ground, so their radiation angle was probably far from optimum for the DX path. However, multi-element end-fire arrays should tend to lower the takeoff (and arrival) angle, compared with a dipole at the same height. The latter, of course, has most of its radiation straight up when only one-quarter wavelength above ground.

It appears that, for my location, a radiation angle of about 35 degrees for 75 meters and 25-30 degrees for 40 meters is about optimum for the early morning (local time) DX path. At another location a lower angle could possibly be more effective.

No single-type antenna is best suited for ***all*** locations. An antenna that may perform well at one location may give poor DX performance at another. Anyone desiring a good antenna for a long-haul DX circuit on 40 or 80 meters should first try, at least, two ***entirely different*** types of antenna; possibly a quarter or half-wavelength vertical, with ***at least*** 50 radials to start and a good dipole at least 22 meters (72 feet) above ground. Then compare these antennas directly for a few days, preferably with the same DX station. Then repeat the test several times during the DX opening for that day.

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5. George E. Smith, W4AEO, "Log-Periodic Beam for 15 and 20 Meters," ham radio, May, 1974, page 6.
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ham radio

homebrew hardline-to-uhf coaxial cable connectors

Easy method for making your own coaxial connectors for CATV cable

While the CATV industry has made **hardline** cable more readily available, it has not been as generous with the required cable terminations; the Amateur without CATV friends has little choice but to make his own. The inexpensive method described here requires no machine work or CATV hardware and will accommodate any cable with a 12.7-mm (0.500-inch) OD shield and a center conductor of 2.92 mm (0.115 inch) OD or less. It consists of a standard PL-259 (Amphenol 83-ISP) type uhf plug mated to a common brass plumbing fitting known as a "1/2 inch OD x 3/8 inch OD compression union," typically $1.50 at plumbing supply houses.

assembly

Open the hole in the smaller union hex nut with a 12.7-mm (112-inch) taper reamer until it's a tight press fit over the cable end of the PL-259. Screw the nut onto the union firmly by *hand,* omitting the small compression ring. Temporarily fit a spare SO-239 chassis connector to the PL-259 to sink heat away from the pin insulation and to keep the threaded coupling barrel out of the way when soldering. Press the PL-259 through the hex nut hole until its end abuts squarely with the body of the union. Apply a small amount of water-soluble acid flux (oleic acid, available at most stained-glass hobby shops) to the joint and sweat solder, using moderate heat from a pencil-flame torch played onto the nut. After it cools, wash the connector thoroughly and dry it.

installation

Prepare the end of the hardline as in **fig. 1.** A tubing cutter is generally used to cut the solid shield; unfortunately this method often results in metal being swaged into the foam core, severely reducing the ID of the aluminum shield at this point. The use of a new, sharp cutter may minimize the problem. An alternative is to lightly score the shield with the cutter, then chase the mark with a rat-tail file or fine-tooth hobby saw until the metal just parts. Admittedly, the latter is a chore.

Clean off any coatings found on the inner and outer conductors with lacquer thinner and lightly polish both with fine steel wool. With the large hex nut and compression ring slipped on, insert the cable fully into the connector. This will require firm yet gentle pressure, as the core OD and union ID are pretty much the same. Tighten the hex nut with the appropriate wrenches. Soldering the center conductor to the pin completes the installation.

By Bud Weisberg, K2YOF, 62 Harriet Avenue, Bergenfield, New Jersey 07621

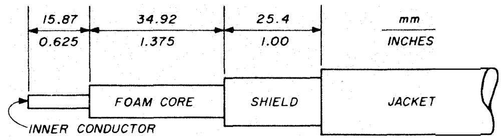
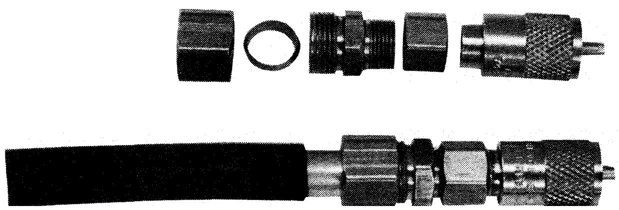
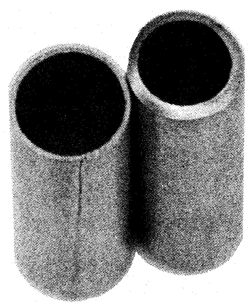


fig. 1. Stripping dimensions for hardline. Inner conductor may he left longer and trimmed after installing connector.

The final connection should be waterproofed with heat-shrink tubing or Mylar tape wrap. Although this is unconfirmed, I expect that the compression union will add little or no impedance disturbance to the line, because the union bore conveniently approximates the shield ID for its entire length. Compression unions are available in straight and reducing configurations, from 114 inch OD to 518 inch OD in 118-inch steps, suggesting that similar connectors could be made for other hardline sizes.





Above, component and assembled view of hardline connector. Smaller compression ring, not shown, is not used. At right, result of trimming shield with tubing cutter. Original ID of 11.43 mm (0.450 inch) was reduced to 8.39 mm (0.330 inch) at right.

A slightly divergent yet related final comment: should you be tempted to regard hardline as "state of the art" coax, consider for a moment a 78-ohm cable with a 2.1-mm (no. 12) tinned solid copper center conductor and a solid drawn-copper shield, both separated by a continuous string of polystyrene beads. The round nose of each bead fitted the concave base of the adjacent one, allowing a bending radius of 101.6 mm (4 inches). Rated at 700 watts rms to 100 MHz, it cost 50C per foot. That was Amphenol 72-12C, marketed to hams in the mid 1930s! And they offered the connectors, too.

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ham radio

high-frequency diversity receiver from the 1930s

A report
on the development
of the Hallicrafters DD-I,
the first dual-diversity
receiving system
for Amateur use

It was a monster – but a very friendly monster. It weighed 102 kg (225 pounds), measured 112 cm wide x **48** cm deep x 30 cm high **(44** x 19 x 12 inches) and occupied a giant parking place on my operating table. It had twenty-five vacuum tubes, four meters, a seven-gang variable tuning capacitor, no transistors, and it received the same signal twice! What was it? A Hallicrafters Dual Diversity receiver, model DD-1, which Hallicrafters presented to the world in June of 1938 with a two-page spread in QST. It took two pages, too, to do justice to the receiver.

the monster

To the best of my knowledge, the DD-1 was the only commercially available diversity receiver at that time; it was designed for Amateur as well as commercial use. **Fig. 1** shows the tuner portion only of the receiver without the external power supply and audio amplifier chassis. Inside and bottom views are shown in **figs. 2** and **3** respectively. All three photographs add up to an impressive piece of equipment.

The DD-1 failed to become popular for several reasons, the most important probably being its cost. The receiver cost $422.00 complete at a time when the top-of-the-line HRO was $179.00, and a man who made $20 a week was considered moderately successful. Another reason may have been its many advanced engineering features: like those of the Airflow Chrysler, they weren't appreciated at that time, and some of them are not available even today.

diversity reception

To fully appreciate the DD-1, you must first understand diversity reception. A more detailed description is given in a previous article,' but briefly, diversity reception is a technique for reducing the adverse

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effects of multipath fading by receiving the same signal on two or more diverse, or different, antenna-receiver combinations with a means of choosing the combination with the strongest signal.

Under some ionospheric conditions, signals in the shortwave bands may travel between the transmitter and receiver over more than one path. If the lengths of two of these paths differ by an odd multiple of half wavelengths (only 10 meters [30 feet] for the 20 meter band), the two signals will arrive out of phase at the receiving antenna. The signal will appear to be in a fade even though either of the two signals, if received separately, would be strong. The short, sharp fades that characterize high-frequency propagation are caused by this effect.

While there are several different forms of diversity reception, what is known as polarization diversity reception appears to be the most practical for Amateur use because it doesn't require additional real estate. With polarization diversity two antennas are used: one vertical, the other horizontal. Each antenna is connected to its own receiver. Polarization

fig. 1. The tuner section only of a Hallicrafters DD-1 receiver. There was also a power supply chassis and an audio amplifier chassis. Dual S-meters sat on top of the receiver, all of which added impressiveness to the system.

diversity is effective because, in general, the horizontal and vertical components of a signal will not travel over the same paths and therefore will not fade at the same time. By choosing the antenna-receiver combination with the stronger signal, communications can be carried on when signals on a single receiver would be in a fade.

Although two separate receivers can be and have been used,[2,3] operating convenience and performance improvements can be had by designing a receiver specifically for diversity operation.

brief history of diversity reception

The development of the DD-1 is closely tied to the development of diversity reception, which in turn is tied to the exploitation of the shortwave bands by

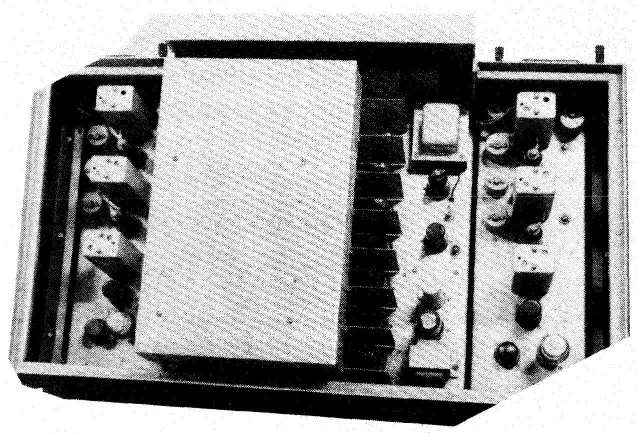

fig. 2. Top inside of the DD-1. The pushbutton bandswitch is under the cover.

commercial operators more than 50 years ago.

Following World War I, the commercial radio interests forced Amateurs out of what are now called low- and medium-frequency bands and into the shortwave or high-frequency bands. After Amateurs demonstrated that the shortwave bands could be used for communications over long distances much more economically than could the longwave bands, the commercials began to investigate shortwave propagation phenomena.

The fading problem. Particular attention was paid to fading, since fading considerably lowered the reliability of the circuit and could not be tolerated in a commercial system. Around 1925-26, Drs. Beverage and Peterson of RCA Communications, Inc., began experimenting with high-frequency communications and had installed a shortwave receiver at the RCA receiving station at Riverhead, Long Island, in New York. They were using a Beverage longwave antenna system that was over 14 km (9 miles) long!

Fading was present. Asecond receiver was installed

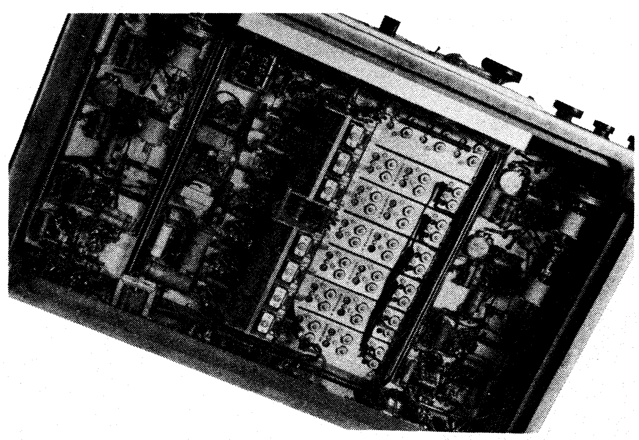

fig. 3. Underside of the receiver. Notice the seven-gang variable capacitor: they don't make them like that anymore.

in Dr. Peterson's home about 1.6 km (1 mile) away; fading, of course, was present there, too. Correlating the fading by telephone, the two experimenters observed that there was no relationship in the fading between the two locations. Hence space diversity was born. Beverage and Peterson jointly hold the basic patent in this art.

Multipath. By the late 1920s the causes of short-term fading were reasonably well understood; the villain was multipath propagation. It was also determined that fading did not occur uniformly over the earth's surface or uniformly with polarity; that is, a signal received on a horizontal antenna might be in a fade when a signal received on a vertical antenna at the same location could be loud and clear — or vice versa. Rice in a 1927 QST article4 describes this phenomenon.

Early commercial system. Before 1930 a triple

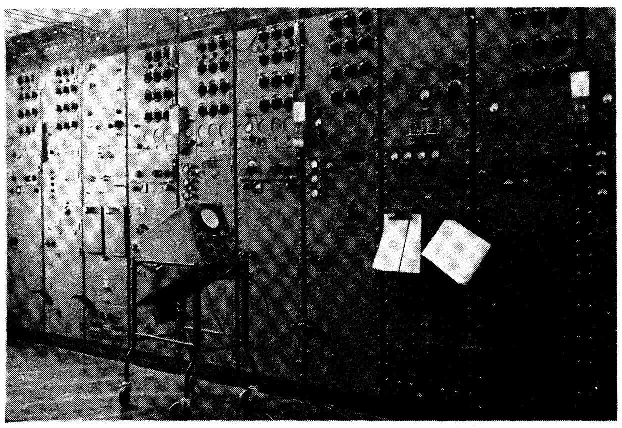

fig. 4. The diversity receivers used by RCA Communications, Inc., Riverhead, Long Island, New York. Notice the clipboards in the center; these held the receiver log. The log clipped at the top, in the usual way, indicates that receiver is in use. The log held by the corner indicates that that receiver is on standby. The index file on each receiver gives dial settings for commonly used frequencies. (Photo courtesy Robert McGraw, W2LYH, and Harlod Moore.)

diversity receiving system (three antennas and receivers) was in operation at RCA Communications, Inc., Riverhead, Long Island. Two papers by Beverage et al[5] describe this installation. The receivers used at Riverhead were truly mammoth, as shown in the photograph of fig. 4. An installation of this type was suitable for commercials because they operate on one frequency for many hours continuously; but it wa uslу not ti ı mateur use I trying to tune across 40 meters with this equipment!

Diversity reception for Amateur use. About this time, James L.A. McLaughlin began experimenting

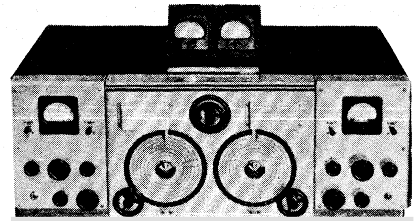

fig. 5. The prototype of the DD-1 as described by McLaughlin and Miles (from reference 8).

with diversity reception using two similar receivers. He was iater joined by James Lamb, then technical editor of QST. Their early experiments were successful in proving the advantages of diversity reception for Amateur use, but were not successful in developing a practical receiver for Amateurs, mostly because of the lack of suitable components, especially vacuum tubes. By the middle 1930s, when improved tubeswere available, McLaughlin and Lamb designed and constructed a single-tuning control, dual-diversity receiver for Dr. James Hard, XE1G, an American who operated a pharmaceutical factory in Mexico 1936.[6] A photograph of Dr. Hard's very elaborate station including this receiver is shown in *Radio*.[7]

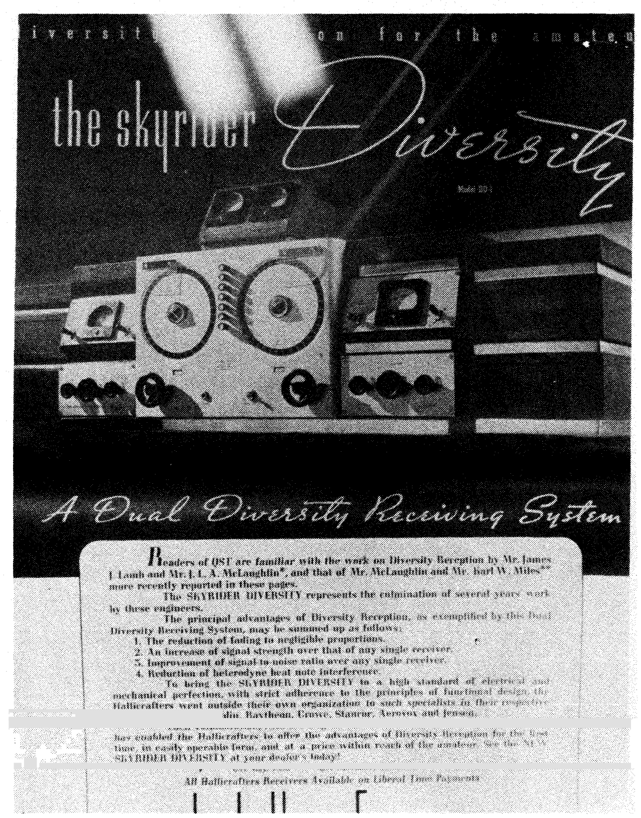

fig. 6. Advertisement announcing the DD-1 in the June, 1938, issue of *QST*. The complete receiver was pictured.

fig. 7. An early Hallicrafters advertisement for the DD-1 tuner as well as several accessories. (Photo courtesy Steve Sullivan, N4GZ.)

The results obtained by Dr. Hard under actual operating conditions were very encouraging proof of the practicality of diversity reception for Amateur use.

Hallicrafters design. McLaughlin later joined Hallicrafters as a consultant and collaborated with Karl Miles, then Chief Engineer of Hallicrafters, in redesigning the earlier receiver. Some electrical improvements were made, but the major changes appear to be mechanical; that is, repackaging. This receiver is described in ***QST,*** December, 1937.8 It was also made under the sponsorship of Dr. Hard and is what I call the prototype model of the DD-1. A photograph taken from the McLaughlin-Miles article is shown in **fig. 5.**

Although early Hallicrafters advertisements show this receiver, I believe Dr. Hard's model was the only one made of this exact design. Later advertisements feature the receiver shown in **fig. 6**, which I call the production version. **Fig. 6** is a reproduction of the two-page spread in *QST* announcing the receiver. A better picture of this version is given in **fig. 1.** The differences between this and the prototype appear to be cosmetic. I'll briefly mention these for the benefit of collectors and historians.

Mechanical developments. On the prototype, the pointer on the tuning dials moves up and down as the band switch is rotated, much like the pointer on the Model SX-16, -17, and -18 receivers. The production version, shown in **fig. 4,** shows a simple piece of plastic as the tuning dial markers.

A second difference is in the band-changing mechanism. The prototype used a rotary switch, while **fig. 1** shows a row of pushbuttons. The change in the band switch may explain the change in pointers because, without a rotary bandswitching mechanism, there's nothing the pointer cord can wind up on to move the pointer.

The third difference is in the cabinet. The center portion of the left- and right-hand front panels has a slight slope, while these panels are vertical in the prototype. The cabinet used for the production receiver appears to be the same as that used for the Hallicrafters HT-1 Amateur transmitter, the HT-3 marine transmitter, and the upper half of the famous HT-4, all of which came out at about the same time. The

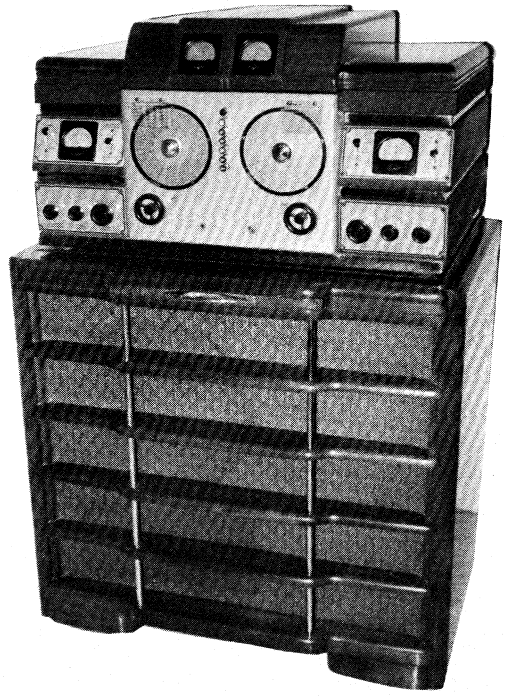

fig. 8. The late production or "jazzed-up" version of the DD-1 showing the special hi-fi loud speaker made by Jensen especially for this receiver. Note that the bandspread dial is blank. (Photo courtesy Charles Dachis, WD4EOG.)

change was probably made to achieve manufacturing economy as well as to jazz up the appearance for commercial sales.

It should also be pointed out that fig. **1** shows only the tuner. In addition to the tuner, there were two other chassis, a high-fidelity audio amplifier using a pair of 2A3s and a power-supply chassis. These went on the right and left sides of the tuner respectively. Diversity S-meters, which measured the signal strength in each receiver, were also available and mounted in a separate housing, which sat on top of the receivers. These accessories all increased the size and impressiveness of the receiver. A copy of a Hallicrafters advertisement showing the various options and prices is shown in fig. **7.** This advertisement was obtained through the kindness of Steve Sullivan, N4GZ.

A later change in cabinet design made the receiver even more stylish. The top center of the cabinet was raised and rounded off, while the diversity S-meters were enclosed in the cabinet instead of sitting on top. This version is shown in fig. **8;** I call this version a late production model. This particular receiver belongs to Charles Dachis, WD5EOG, of Austin, Texas. The photograph also shows the special speaker cabinet that was made by Jensen for the DD-1. The cabinet has compartments in the back that hold the audio amplifier and power-supply chassis. WD5EOG has these items, but they are not visible in the photo. The two receivers shown in figs. **1** and **8** are the only two copies of this receiver that are still in existence that I'm aware of. I'd be happy to hear from any reader who knows of others.

Technical description. A simplified schematic of the DD-1 is shown in fig. **9.** It consists of two identical superheterodyne receivers with a common first local oscillator and a common BFO. Provision is made at the output of each second detector for switching to either receiver separately or combining the two outputs in a summing network for diversity reception. It's not too apparent, but the agc buses (avc in those days) of each receiver are also tied together. The receiver with the stronger signal and higher agc voltage will thereby tend to mute the "down" receiver, so that it doesn't put out a lot of noise.

Since many of the electrical features are conveniently identified by the front-panel controls, I show three detailed photographs of the left, center, and right panels in figs. **10, 11,** and **12** respectively.

Left-hand panel. The upper half of the left-front panel (fig. **10)** shows the send-receive switch. The

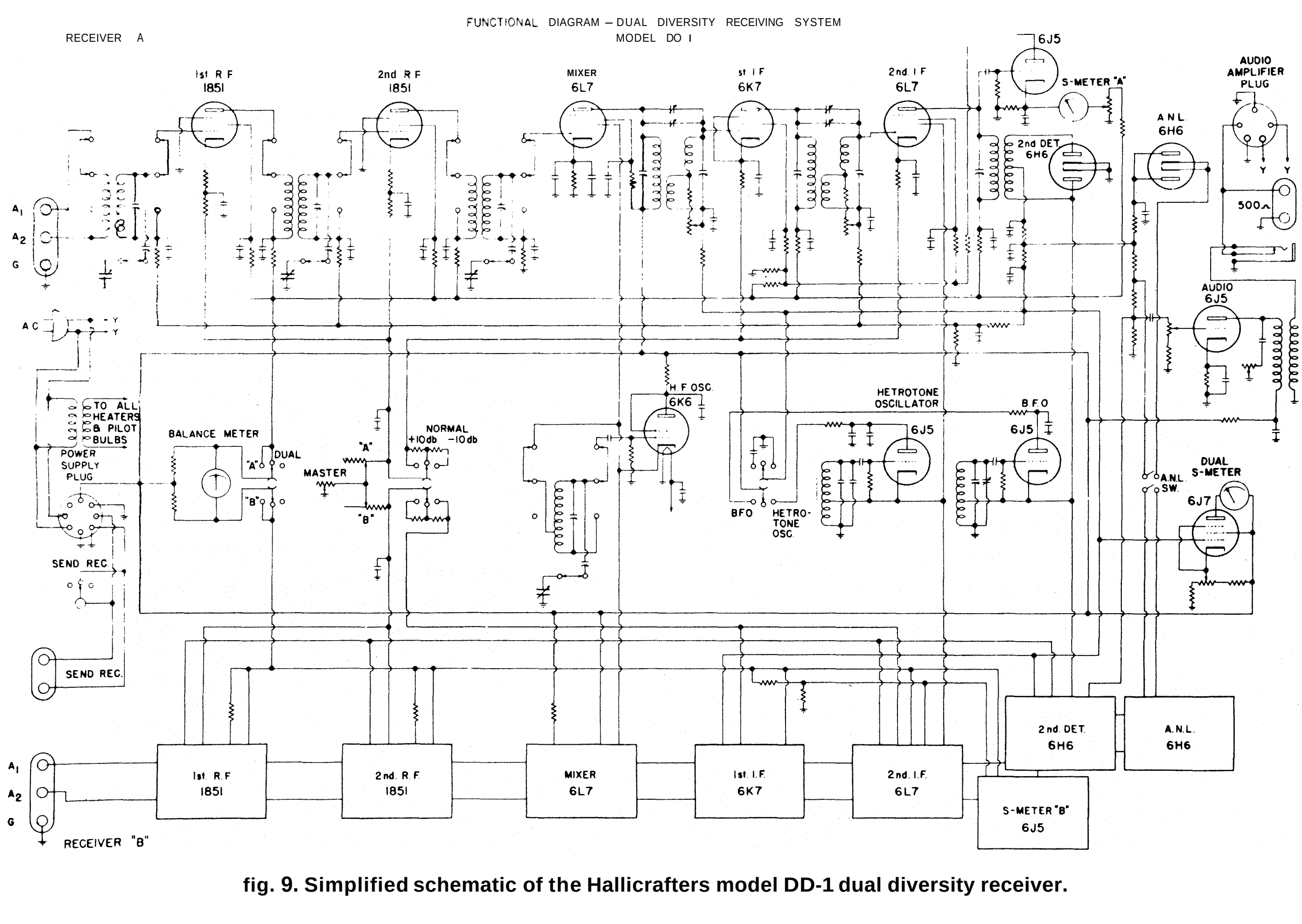

fig. 9. Simplified schematic of the Hallicrafters model DD-1 dual diversity receiver.

fig. 10. The left-hand control panel of the DD-1.

upper send position of the lever-type switch has a spring return for a quick transmit; the lower send position has a holding position. Separate contacts on the switch are wired to terminals on the rear apron, which may be used to turn on the transmitter. These are conveniences not found on many receivers today.

The S-meter on the left-hand is the combined S-meter; that is, it shows the sum of the two received signals, and as such will usually read high. To the right of this meter is another lever switch marked "Heterotone Oscillator" (up) and "Heterodyne Oscillator" (down). The heterodyne oscillator is the conventional *BFO* with a pitch control knob in the lower left-hand corner of the panel. The heterotone oscillator is a unique and interesting circuit I'll describe in detail later. For the present, I'll only say that it serves the same purpose as the BFO except that the pitch of a CW signal remains constant as the receiver is tuned through the signal instead of decreasing to zero and increasing, as in a conventional *BFO.*

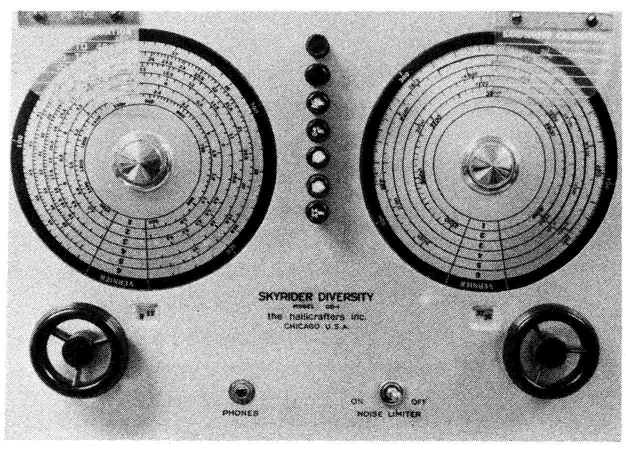

fig. 11. The main tuning dial of the monster. Note the "writ by hand" calibration on the bandspread dial.

The tone control in the center of the lower left-hand panel is conventional. The knob on the lower right controls the i-f bandwidth through a very interesting system developed earlier by the designers of this receiver. I describe this, too, in greater detail later.

Center panel. The center panel, **fig. 11,** has the main tuning dial and control on the left and the band-spread dial and control on the right. The controls tune both receivers simultaneously. The small windows just above each control are a vernier scale that is used in conjunction with the outer scale on the tuning dial for logging. The bandspread dial was originally supplied blank, to be calibrated by the owner according to his desires. The scale on my set was calibrated by the previous owner for the Amateur bands. The phone jack and noise switch, again, are conventional.

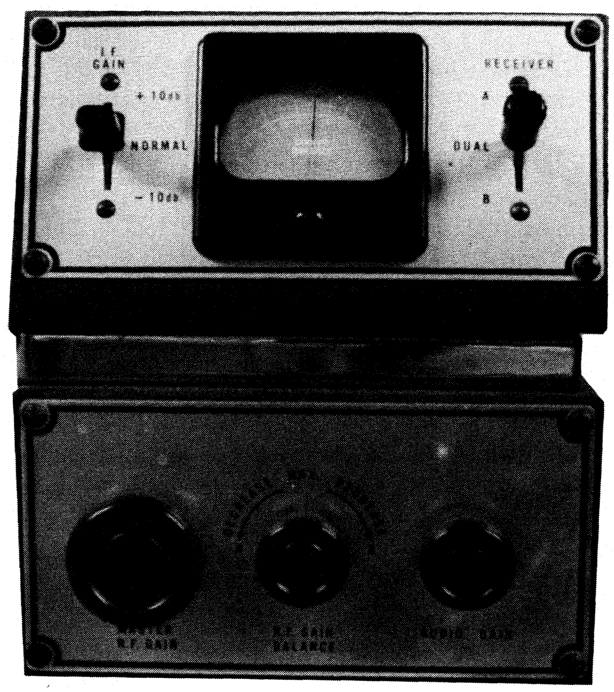

fig. 12. The right-hand control panel. The S-meter is a differential type.

The column of pushbuttons is a combined band-

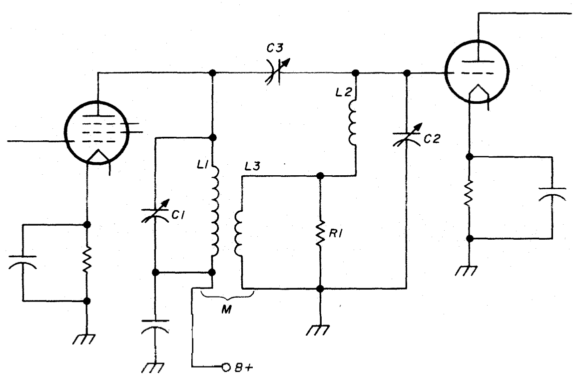

fig. **13.** Basic schematic of the infinite off-frequency rejection coupling system used on the **DD-1**.

switch and primary power switch. The top button turns the receiver off; any of the remaining six buttons simultaneously select the desired band and apply ac power.

Right-hand panel. The right-hand control panel, **fig. 12,** has a zero-center meter used as a differential S-meter; it reads the difference in signal strength between the two receivers. Under poor propagation conditions it's fascinating to watch this meter swing back and forth as first the vertical then the horizontal signal predominates, while the combined S-meter remains relatively constant. At times like this the advantages of diversity reception become most apparent.

The selector switch to the right of the differential S-meter selects the audio output of either receiver separately or the combined output for diversity reception. An i-f gain control switch to the left of the meter increases or decreases the gain of both i-f amplifiers by 10 dB. I'm not certain what the purpose of this control is but have found it convenient when signals were either very strong or very weak.

The lower left-hand knob on this panel is the master rf gain control; it controls the rf gain of both receivers. The center knob is the rf gain balance control. It is used to roughly balance the signal levels in the two receivers. With this control in the center, the rf gain of the two receivers is approximately equal. Adjusting the knob off-center increases the gain in one channel while decreasing it in the other. The control is adjusted until the differential S-meter reads zero on the average.

The audio gain control is common to both channels and needs no comment.

The DD-1 featured several engineering advances, two of which may be of interest to receiver designers today. The first was a continuously adjustable bandwidth control that did not use a crystal filter; the second was the heterotone oscillator.

The off-frequency rejection system. This circuit was described by McLaughlin and Miles[9] a month before the DD-1 itself. The basic schematic is shown in **fig.** 13. To quote the author:

> It will be observed that coupling is provided by the mutual inductance, M, between L_1 and L_3 and the capacitive coupling, C_3 . . . Mutual inductance, M, and the capacity coupling, C_3, are so chosen that at some determined frequency off resonance the voltage induced through M is opposite in sign to the voltage induced in C_3 and will therefore cancel it out. In other words, no coupling exists at this particular frequency. In order to achieve infinite rejection at this undesired frequency, correction for power factor in the circuit must be made. Resistor R_1 in the diagram is the power factor corrector. The rejector control, C_3, can be made variable and tuned over a fairly wide frequency range of rejection without noticeable interlocking effect on the i-f. For proper operation, resistor R_1 should be variable; but once the infinite rejection point has been found, it need not be touched again . . .
>
> [Fig. 14 shows the response of] a single rejector circuit set to reject a frequency 4.5 kHz off resonance. The rejector slot, as is apparent, goes to infinity at this frequency and its action is very similar to the rejection in a crystal filter. . .

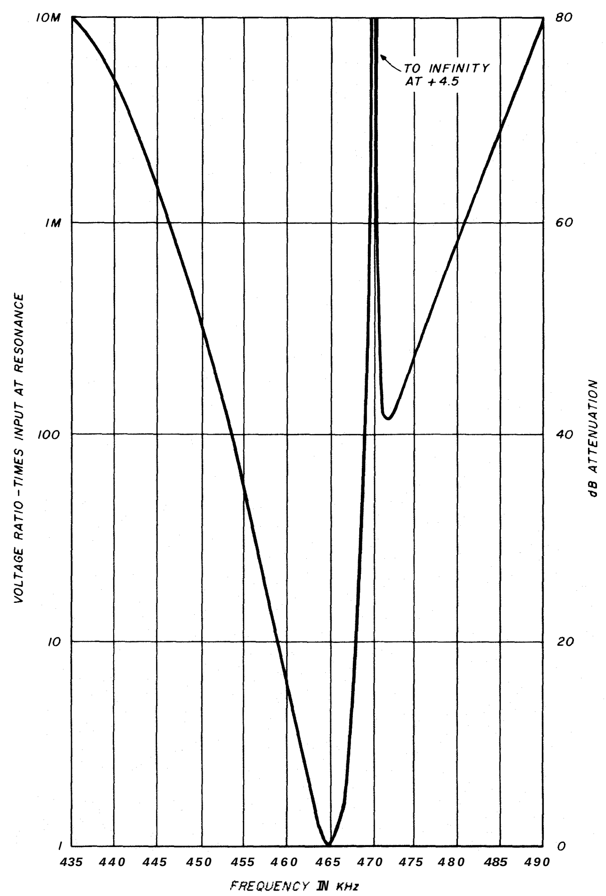

fig. **14.** Frequency response of a single rejector circuit set for infinite attenuation at **4.5 kHz** above the resonant frequency (from reference **9**).

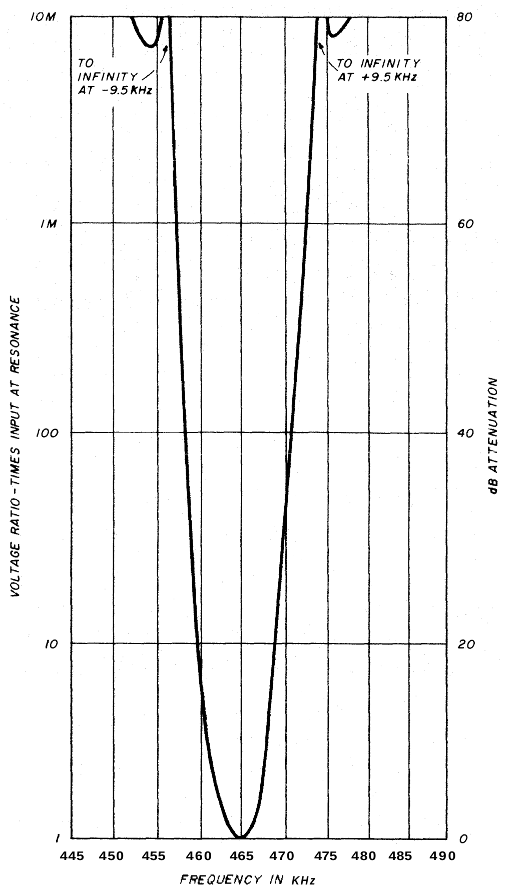

fig. 15. Bandpass of two rejector circuits set at 9.5 kHz above and below the resonant frequency (from reference 9).

To obtain a symmetrical response, two such circuits replaced the i-f transformers for two consecutive stages. The coupling capacitors, C_3, were chosen so that one infinite rejection slot was above the desired center frequency and the other below it. These variable capacitors were also ganged so that one increased the slot frequency while the other decreased it. Thus the bandwidth could be made continuously adjustable by means of tuning the two capacitors.

Figs. **15** and **16** show the resulting bandpass with the rejector control set at $\pm$**9.5** kHz and $\pm$5 kHz above and below resonance.

The schematic of fig. **17** can be used to obtain a continuously adjustable bandwidth for a single receiver. In the DD-1, it's necessary to gang two rejector pairs, one for each receiver. Although this system is relatively simple and straightforward, I'm not aware of its use on present receivers. One variation might be to use a varicap for remote bandwidth control. The circuit seems to be a forerunner of the so-called single-sideband mixer used in the microwave region today.

Heterotone feature. Another interesting feature of the DD-1 is the heterotone oscillator used to receive diversity CW signals. Receiving diversity CW presents a problem not found in diversity phone reception; that is, in phone reception the combining of the two signals occurs after all rf phase information is lost. With CW reception, using a conventional heterodyne BFO system, the phase information is still present in the audio output, since the audio signal is a CW signal itself.

When we discussed short-term fading, we said it was caused by two signals from the same transmitter traveling over two different paths varying in distance by one-half wavelength and arriving at the receiving antenna out of phase. Assuming polarization diversity, it's entirely probable at times that the signals arriving at the horizontal and vertical antennas would travel over slightly different paths and arrive at their respective antennas out of phase. The audio output of each channel would also be out of phase, as the

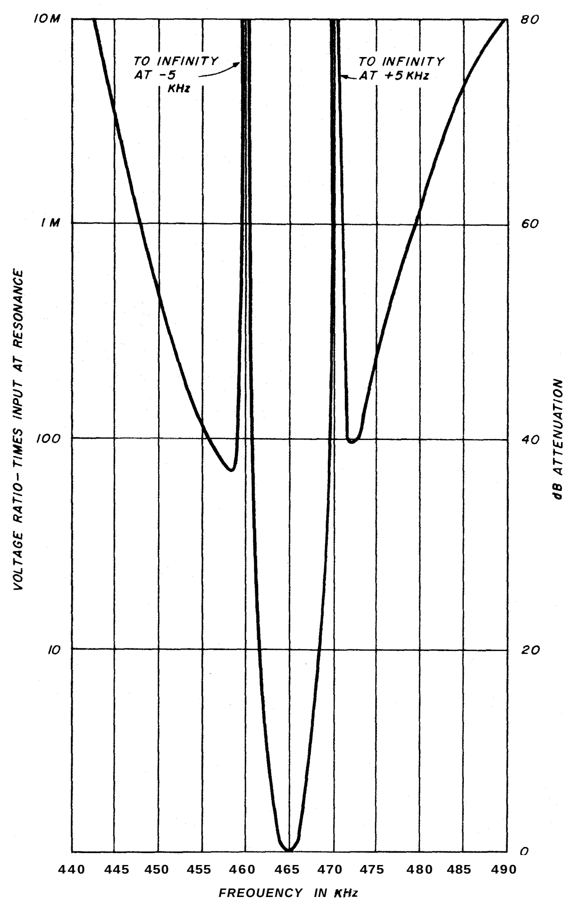

fig. 16. Same as fig. 15 except that rejector circuits are 5 kHz above and below resonance (from reference 9).

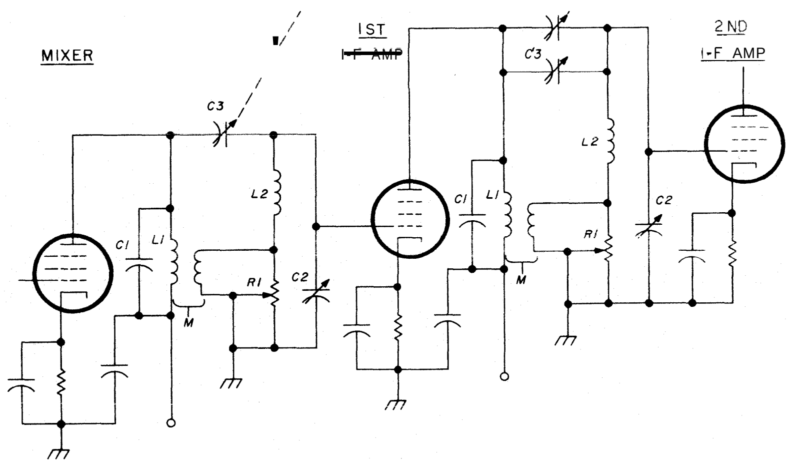

fig. 17. Two infinite rejection circuits ganged to give continuously adjustable bandwidth. In the DD-1, the i-f amplifier of each receiver has its own infinite rejection pair, and all four circuits are ganged together.

same heterodyne oscillators are used for both channels, so that the combined audio signals would null. We have then, in effect, generated a fade inside the receiver, when both signals are strong at their antenna terminals. But this is just what we're trying to avoid! The answer is, of course, to remove all phase information before combining the signals. Two ways of doing this are generally accepted.

The method used by the commercials is to feed the output of the i-f amplifier for each channel into its own diode detector without using a BFO. Each detector output becomes a series of dc pulses, in Morse code, from which all phase information has been removed. The detector outputs are then summed in a conventional summing network; as this sum is the total of all channels, it is relatively fade-free. The summed output is used to key a local audio oscillator, which the operator hears in his headset.

Another method is the heterotone circuit as used in the DD-1. With this circuit, no BFO is used. Instead, an audio-frequency oscillator is used to amplitude-modulate the i-f signal in each receiver. A conventional diode (envelope) detector is used in each channel. The detector output is therefore an audio tone when a carrier is present and no tone when one is not present. All rf phase information is lost, so the two signals can be combined without phase effects.

It's interesting that the heterotone type of CW reception was originally developed by James Lamb[10] and has some advantages over heterodyne detection, even in single-receiver applications. This is especially valuable for operators who spend long periods of time at the key, such as in contests or moving traffic.

Let's go back some forty-odd years and read what James Lamb had to say on the subject:

> In heterodyne reception of pure dc telegraph signals there is a monotony, an exasperating tiresomeness about that piercing beat-note that makes old-time operators wish for the good old days and makes those who haven't had modulated mcw or icw experience wish they could do something besides change the beat note to just another single tone that drills a hole into the hearing system. This fatigue and monotony from listening to a pure dc beat-note isn't all imagination, either. It's quite real and demonstrable by authentic scientific proof . . .

Although I don't do a lot of operating, my own limited experience with heterotone reception has been very favorable. This appears to be another engineering feature not available that perhaps ought to be.

closing comments

As a writer interested in the history of diversity reception, particularly for Amateur applications, I've been frustrated in researching this article. There is so much I have not been able to find out!

For instance, I'd like to know more about Dr. Hard, XE1G, and how he became interested enough in diversity reception to finance the development of two early diversity receivers. Where are his receivers now? What became of Dr. Hard's two very elaborate stations? What became of Dr. Hard? I assume he's long since become a silent key. The Mexican Embassy in Washington, D.C., has no record of a Hard chemical works today.

How many DD-1s were actually manufactured? In the August, 1976, issue of QST I ran an "I would like to get in touch with" notice for present or former owners of the DD-1. I received three responses. One said he understood "about a dozen" sets were manufactured. Bill Halligan tells me he thinks it was more like 200 sets. Who bought them? How were they used? Where are they now? I only know of two, my own and Chuck Dachis's.

In addition to being the first and only diversity receiver available commercially, the DD-1 featured many engineering advances as well. i beiieve it deserves more prominence in the annals of Amateur Radio than it's been given to date. If it's going to assume its rightful place, questions like these should be answered. I'd be pleased to correspond with anyone who knows these answers.

I've haard the DD-1 referred to as the Edsel of the radio industry; I'd rather think of it as the Airflow Chrysler of the industry, since it was so far ahead of its time.

With the development of solid-state components and PC board techniques, a diversity receiver can now be built that's considerably smaller than the "monster"; smaller, in fact, than a modern transceiver. Diversity reception has proved itself in over fifty years of commercial service. Isn't it time Amateurs tried it?

acknowledgments

Thanks to Harold Moore and Robert McGraw, W2LYH, both former employees of RCA Communications, Inc., for the photograph of the RCA diversity receivers. They also provided much additional information which I'd have included in my other diversity article,' except that that article was too far down the editorial path to change. My thanks, too, to Steve Sullivan, N4ZG, for his Hallicrafters advertisements, and to Charles Dachis, WD5EOG, for his much-needed encouragement and photograph of his own "monster."

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9. Karl W. Miles and J.L.A. McLaughlin, "A New I.F. Amplifier with Infinite Off-Frequency Rejection," *QST,* November, 1937, pages 19-23.

10. James J. Lamb, "Heterotone C.W. Telegraph Reception," *QST,* November, 1936, pages 16-18.

ham radio

More Details! CHECK—OFF Page 94

capacitance-measurement accessory for your frequency counter

Still another use for the NE-555 timer IC — this time in a unit of test equipment for measuring capacitance

Here is a very useful accessory for your counter that will allow you to measure capacitance from about 5 pF to 50 μF in two overlapping ranges. Resolution is 1 pF or 100 pF, provided the counter can display 1 Hz. Accuracy depends chiefly on standards used for calibration or comparison with a digital capacitance meter of known accuracy. Using all new parts, it costs about $18.00.

This is a modification of the W6ALF[1] circuit using the 555 timer IC in the one-shot mode to produce an output pulse of length proportional to the capacitance used in the RC timing network. The pulse is used to gate a transistor switch, which passes 1-MHz pulses obtained from the counter time base. The counter displays the number of 1-MHz pulses passed through during the timing period.

The basic circuit has the disadvantage that the 555 timer will produce an output pulse with no external capacitor; thus the counter will display anywhere from about 9 to over 30 depending on layout, strays, etc. The improved circuit uses another 555 to output a pulse that can be adjusted to the same length, which delays output of the 1-MHz pulses so that the count can be zeroed when no capacitor is connected to the test terminals.

This unit needs no external power. It will work with TTL- or CMOS-level signals, requiring only a 1-second gating pulse and 1-MHz clock pulses from the counter. Current drain from the internal 9-volt alkaline battery is 16 mA maximum. I tend to be forgetful, so I used a momentary push-to-test switch.

circuit operation

Refer to **fig. 1.** A positive-going gating pulse from the counter at El is differentiated by R1C1, inverted by Q1, and the required short negative-going trigger pulse is applied to U1 pin 2, the main timer, and also to U2 pin 2, the delay timer. Pin 3 of each 555 very rapidly goes high. U1 pin 3 back biases CR1 for the duration of U1 output pulse. When S1A is in the low-range position, U2 pin 3 switches Q2 emitter high, which cuts off Q2 and back biases CR2. When both CR1 and CR2 cathodes are high, Q3 turns on and its collector goes low to the output at E3.

When U2 times out, its pin 3 goes low, which enables Q2 to switch 1-MHz pulses coming in from the counter time base at E2. CR2 is then switched on and off, which turns Q3 off and on at a 1-MHz rate. When U1 times out, CR1 cathode goes low, disabling Q3 so the output at E3 ceases. The counter accumulates the 1-MHz pulses and displays the number.

The delay timer, therefore, inhibits the count for a

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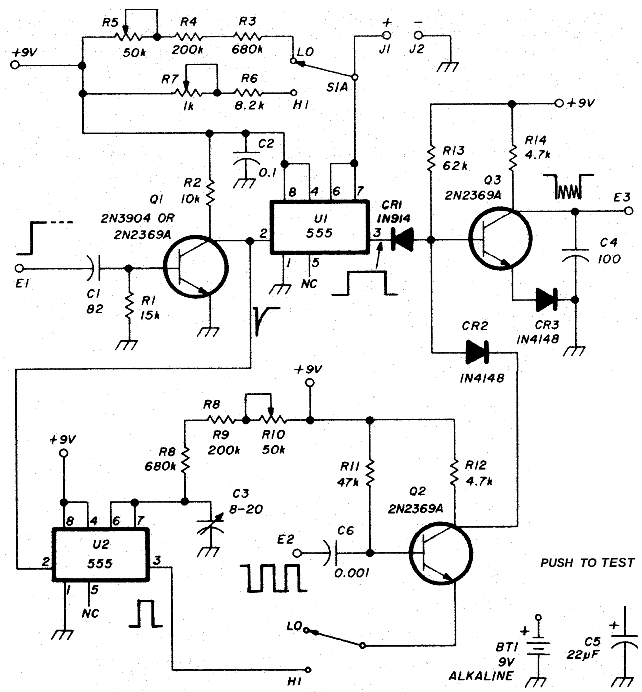


C1,C4,C6	mica or ceramic
C2	disc ceramic
C3	miniature ceramic trimmer
C5	tantalum
E1	gating pulse in
E2	1-MHz pulses in
E3	gated 1-MHz pulses to counter
S1	range switch, dpdt toggle
All resistors	¼-watt, 5 per cent carbon composition; pots 10 mm (0.4nch) square PC-board mount

fig. 1. Schematic of the capacitance-measurement accessory for use with a frequency counter. Circuit is a modification of that in reference 1. It uses a second NE-555 timer IC, which allows the count to be zeroed when no capacitor is connected to the test terminal. Circuit works with TTL- or CMOS-level signals, requiring only a 1-second gating pulse and 1-MHz clock pulses from the counter.

time interval that is adjustable by R10 and C3. That interval is made equal to the open-circuit time of U1 — the pulse length that occurs when no external capacitor is connected to the test terminals. The delay circuit is not needed on the high range. S1B grounds Q2 emitter on that range.

The diode in Q3 emitter circuit ensures that the transistor is turned off when either CR1 or CR2 cathode is low. Pin 3 of the 555 goes to less than 100 millivolts when low; but when this voltage is added to the drop across either CR1 or CR2, Q3 base may be sufficiently positive to cause some conduction in Q3.

Bypass capacitors C2 and C5 were included for good design practice. No jitter on the timer output pulse was seen using a 15-MHz scope. Capacitor C4 suppresses a vhf oscillation in the counter input circuit when the tester is connected but not in use.

Battery voltage is not at all critical. The unit works over a range of 5-15 volts. Stability was slightly better at 9 volts than at 5 volts, but no noticeable improvement occurred at 13.8 volts. I used the 9-volt battery for convenience. It should last a long time in this service. It can be checked easily by loading with a 470-ohm resistor. If its voltage drops below 8 volts after a few seconds, replace the battery.

component selection

The total resistance needed for the timer circuit can be calculated from $t = 1.1RC$. When $t = 1$ *second* and $C = 1\ pF$, $R = 909$ *kilohms.* The series string, R3, R4 and R5, allows adjustment from about 880 to 930 kilohms. For the high range, which is one hundred times the low range, R *calculated* $= 9.09$ *kilohms.* R6 and R7 provide adjustment from 8.2 kilohms to 9.2 kilohms.

The low range will allow measurement from typically 2 or 3 pF to nearly 1 µF on a six-digit counter, or up to 0.1 µF on a five-digit counter, provided the counter will resolve 1 Hz. The high range is useful from about 0.01 µF (count of 100) to over 50 µF.

The resistor string for the delay timer uses the same values as the main timer. The two should track reasonably well with changes in temperature and humidity.

There's no point in using precision resistors in this circuit except for improved long-term stability that might be obtained. Likewise, multiturn pots aren't really needed.

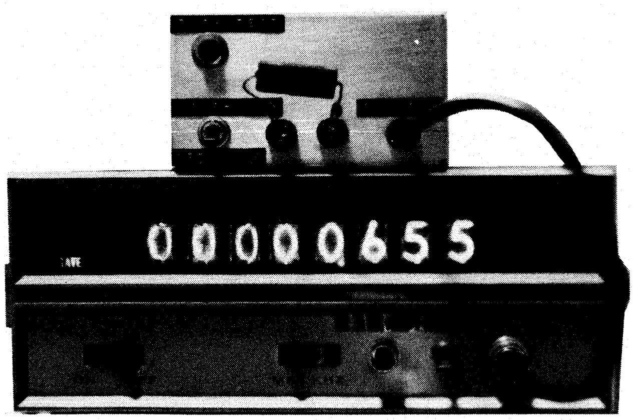
Battery voltage change from 8.0-9.5 volts made less than 0.1 per cent change in counter reading.

Transistor Q1 can be any medium-speed switch, but Q2 and Q3 should be fast-switching. The 2N2369A works well at low collector current. Some 2N3904s tried as Q2 and Q3 were found not to switch reliably.

Capacitor C1 was the smallest value that allowed reliable triggering of the 555s. Pin 2 of the 555 should be driven close to ground; but if it's held low too long, the timer will be fired again when the output pulse is short.

counter interface

This accessory unit was built to be used with a Heathkit IB-1102 counter. A pin jack and a BNC connector were mounted on the counter back panel, from which unshielded leads were connected to J4-1 and J4-3 respectively on the timebase-board socket, which are the gate (I-second or I-millisecond) and 1-



Accessory unit in use with the Heath IB 1102 counter. The extra BNC socket to the right of the counter *MHz-kHz* switch is for input to a divide-by-10 scaler.

MHz outputs. Connections to these points didn't disturb the normal operation of the counter in any way.

Other counters may be used. The gate-pulse line is loaded very lightly. The 1-MHz pulses should be taken from a buffer rather than directly from the timebase oscillator. In the IB-1102, a 7473 divide-by-four flipflop supplies the 1-MHz signal to the string of decade dividers, which provides excellent isolation.

It should be possible to use the 3.579545-MHz clock in counters using a TV color-oscillator crystal time base, although I have not tried it. The resistor string should then total 254 kilohms for the low-C range, for 1 pF and 100 pF resolution respectively.

If your counter has a 0.1-second gate time – and resolves to 10 Hz – you can increase the resistor string to ten times values given and still obtain desired resolution. You may enounter some jitter and drift, however, because of varying leakage associated with resistance values that are as great as 9 megohms.

Actually, the gate-time length in the counter isn't important provided it's long enough to allow full count of the 1-MHz pulses. For instance, the IB-1102 has selectable I-second and I-millisecond gate times. The switch is labeled MHz and kHz. Used as a frequency counter, the gate time must be 1 second to count and display to the closest 1 Hz; and for 1 second of counting time it will, of course, display 1,000.000 kHz (1,000,000 Hz) for an input signal of 1 MHz. For I-millisecond counting time, the display is 1.000 MHz.

As a pulse counter or totalizer, the displayed digits are the same regardless of the MHz-kHz switch position if the number of pulses is less than allowed by the gate time. For example, a 910-pF 5 per cent mica cap measured 877 pF on the low-C range using a 1-second gate (1-MHz position) and exactly the same using a 1-millisecond gate (I-kHz position). This is an interesting observation, because it shows that the pulse length is essentially independent of the duty cycle of the 555 timer. But a 1500-pF cap, which measured 1506 pF using I-second gate, caused a display of 970 using the I-millisecond gate. The reason for 970 instead of 1000 is that the delay circuit suppressed the first 30 pulses, and the shorter gate time eliminated counting the last 506 pulses.

construction

The entire circuit was put on a 50- by 75-mm (2- by 3-inch) single-side board with room to spare. Lines and pads were made with a high-speed hand grinder using dental burrs – used burrs are free from your friendly dentist (thanks to Murray, WB2DXD). It could just as well have been done on perf board or etched if you wish.

A single 16-pin socket accommodates both 555s. The board fits inside the 57 x 57 by 100 mm (2% by 2% by 4 inch) aluminum box. The dpdt HI-LO range switch, momentary NO pushbutton switch, and the banana jacks are on the face of the box. A length of RG-58/U comes through a grommet, and a UG-88/U BNC connector on its end goes to the counter input jack. A single angle bracket holds the board in position. The 9-volt battery fits at the end of the board; it's held in place by a bit of polystyrene foam.

Another length of RG-58/U with BNC plug, and a single unshielded wire with a phone tip plug, pass through a grommet in the back of the box. These go respectively to the 1-MHz output and to a pin jack connected to the counter gate pulse line.

calibration and use

An acceptable calibration can be made with an assortment of 5 per cent silver mica and 5 per cent polyester or polystyrene caps. If you have access to a good bridge or digital capacitance meter, or some known 1 per cent caps, so much the better.

First, remove the delay 555 from the socket, and put a wire jumper from socket pin 3 to ground. This grounds Q2 emitter and disables the delay circuit. Do not ground pin 3 directly when the 555 is in the circuit, because this will short-circuit the 555 output and exceed its current rating.

checks for low-range operation

Set the range switch at LO. Set main timer pot R5 at about 50 per cent resistance. Do not connect anything to the test jacks. Connect unit to counter and push the test switch. The counter should display some number — approximately 30. If it doesn't, check the board for cold solder joints and solder bridges. Use a scope to check at U1 pin 3 for an output pulse. Unless you have a very fast scope, you can't see the trigger pulse at pin 2; but if there's no pulse at pin 3, the 555 may be bad or simply not triggered. Also check at Q1 base for a 1-second gate and at Q2 base for a 1-MHz pulse.

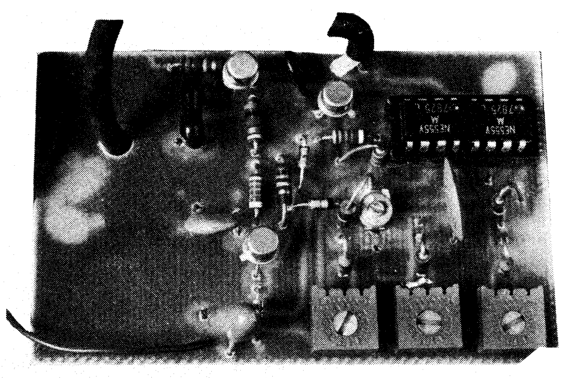
When all is well, connect a known value of capacitor to the test jacks — something in, say the 50-200 pF range. Adjust the 50-kilohms pot, R5, to make the display read a known value of test capacitor plus the open-circuit reading. Remove the test cap to see whether the "open" reading changed. If it did change, repeat the above procedure until the open reading does not change.

Remove the jumper in the delay 555 socket, replace the chip, and replace the test capacitor. Now adjust the delay timer pot, R10, and C3 trimmer until the display is correct. Check with the smallest value cap you have for which capacitance is accurately known.

high-range checks

Switch to HI range. Insert a known value capacitor of about 0.5 μF. Adjust the 1-kilohm pot, R7, for correct reading. There is no delay adjustment on this range.

Check readings on various capacitors (observe polarity). Electrolytics generally show increasing readings. This can be caused by gradual reformation of the dielectric, especially if the cap is old. Note that capacitance values obtained by measurement with this or any other low-voltage tester may be quite different from the effective values if working voltage is much higher. Also, measurements on very high K ceramic caps may be erratic. That's not the fault of the tester — it simply indicates that the retrace



Circuit board and parts. The three square pots are at lower right; the two 555s are in one sixteen-pin socket. Q1 is at top left; Q2 at top right, which is nearest the 555s. Q3 is at bottom left. (Some resistors are standing on end.)

(charge and discharge) times for such dielectrics aren't constant, so it's a useful check on whether an unknown cap is stable enough for some intended use.

final checks

Check linearity by measuring two caps separately and in parallel. There should be good agreement. I measured two 650-pF, 0.5 per cent micas as 647 pF and 652 pF and together 1300 pF. Two 0.1-μF polystyrene caps were 0.1001 μF and 0.1026 μF; together they were 0.2025 μF.

Small-value caps should be checked by insertion directly into the test jacks with no extra leads. I made some small spring clips soldered to bananna plugs to allow good contact. Test leads can be used for connection to larger value caps with negligible error — or the actual capacitance across the leads can be noted and readings corrected.

Another useful check can be made on a capacitor of about 0.01-1.0 μF. Measure it on both ranges. Very close agreement should occur if the calibration is accurate. But note that some metallized and ordinary paper caps may have enough leakage so that the readings on the low range are higher than on the high range. Only the best quality polyester or other high-grade caps will consistently give equivalent readings on both ranges.

On the high range, the counter display with test terminals open is 10. According to a note in the National Semiconductor data for the 555, this occurs because comparator storage time limits pulse width to 10 microseconds minimum when pin 2 is driven fully to ground for triggering. I made a series of tests using caps of 100 pF-1500 pF in approximately 100-pF steps. From 100-700 pF, the counter displayed an increasing count, as expected, up to 17 for a 700-pF cap. But with an 800-pF cap the count dropped to 10, and at 1500 pF the display was 16; at 10,000 pF the count was 100. Evidently, with a large enough capacitor in the test position, the minimum 10-microsecond pulse is swallowed. This particular bit of serendipity was accepted gladly, for it made delay compensation unnecessary on the high range.

To me, the most useful feature of this accessory is that I can match caps closely, using the full resolution capability of the counter, which can't be done on the usual four-digit capacitance meter, and the cost is a lot less. Besides, it was fun to design and build.

reference

1. Ray Krarner, "Using a Frequency Counter as a Capacitance Meter," *QST*, August, 1977, Page 19.

ham radio

600-MHz prescaler for use with electronic counters

A simple prescaler for LSI chips that covers frequencies to 600 MHz

The production of LSI integrated circuits has allowed complete frequency counters on a single chip. The disadvantage of these devices is that the maximum input frequency is limited, usually to about 6 MHz. However, the frequency range can be extended by using a prescaler. This article presents a frequency prescaler usable to 600 MHz.

To achieve maximum input frequency, emitter-coupled logic (ECL) is used because of its high speed. (MECL is Motorola's trade name for ECL.) The prescaler has three separate amplifiers, one for direct counting and the other two for inputs to be prescaled. Exclusive ORs automatically switch the outputs of each amplifier, which eliminates the need for front-panel rf switching. None of the amplifier inputs were diode protected. Diodes add extra input capacitance, which degrades frequency response.

design

The sensitivity of the uhf range is shown in **fig. 1,** input impedance in table **1.** As shown in **fig. 2,** the uhf front end is an Amperex ATF-417 broadband amplifier. The gain of this device is approximately 25 dB, with a bandwidth of 960 MHz (40 MHz - 1 GHz).

The Fairchild 11C90 divides the output of the ATF-417 by ten. The Motorola MC10107 exclusive OR gates this signal to the next divide-by-ten, which an MC10138. The signal level from the MC10138 is ECL and the output (74S86) is TTL, so the two levels must be interfaced. The interface circuit is a Signetics SD211 DMOS fet, which shifts the ECL level to TTL through the 74S86 exclusive-OR to the output buffer, which completes the dividing chain.

The vhf front-end amplifier is a MECL triple-line receiver. Using the MC10116 as an amplifier produces a gain of approximately 20 dB. The vhf-input sensitivity is shown in **fig. 1** and input impedance in table **2.** Since this signal must be divided by ten once, the MC10107 exclusive-OR gates the output of this amplifier directly to the MC10138. The signal is divided by ten and gated to the output buffer, which completes the vhf dividing chain.

The schematic of the prescaler is shown in **fig.** 3. To achieve a higher input impedance in the hf front end, the amplifier has a fet input. The RCA 3028 differential amplifier is used as an amplifier and limiter with a gain of approximately 30 dB. The sensitivity of the hf input is shown in **fig. 4** and input impedance in table 3. The output of the amplifier is a limited sine-wave but not directly compatible with TTL. The fet at the amplifier output converts the signal to a TTL level. This signal isn't divided, so the 74S86 gates the signal to the output buffer, completing the hf section.

Power requirements for the prescaler are 15 Vdc, 35 mA and 5 Vdc, 325 mA. A Motorola MWA120 broadband amplifier can be used in place of the ATF-417. The MWA120 has a gain of 14 dB to 600 MHz; the cost is much lower than that of the ATF-417.

table 1. Uhf input impedance.

frequency (MHz)	magnitude (ohms)	phase (degrees)
60	88.6	−16.3
80	85.4	−20.5
100	80.7	−21.8
200	58.7	−20.4
300	41.1	−13.3
400	41.2	+14.0
500	53.8	+33.2
600	91.7	+29.3

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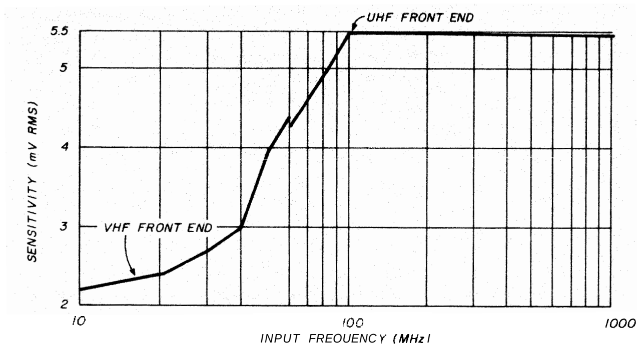


fig. 1

fig. 1. (top). Sensitivity of the vhf and uhf ranges.
fig. **2.** (center). Functional block diagram of the 600-MHz prescaler. Three separate amplifiers are used, one for direct counting and two for inputs to be prescaled.
fig. 3. (bottom). Prescaler schematic: circuit was constructed on a PC board. Power requirements are 15 Vdc at 35 **mA** and **5** Vdc at **325 mA**. **L1** is six turns no. 30 **(0.6-mm)** on **3B** ferrite bead.

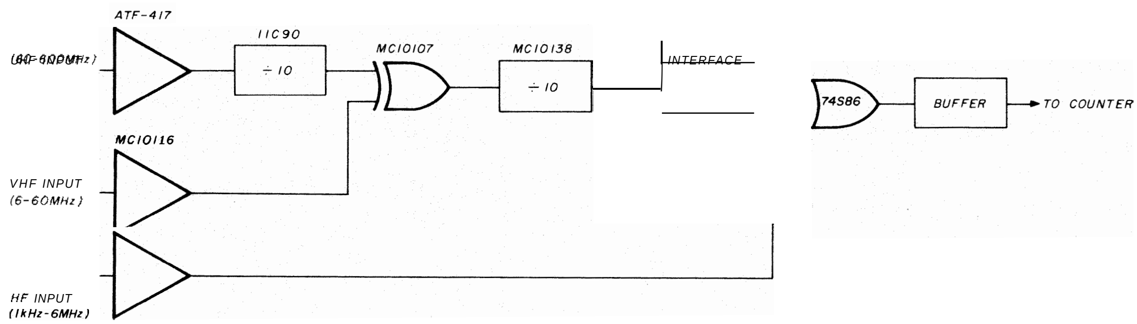


fig. 2

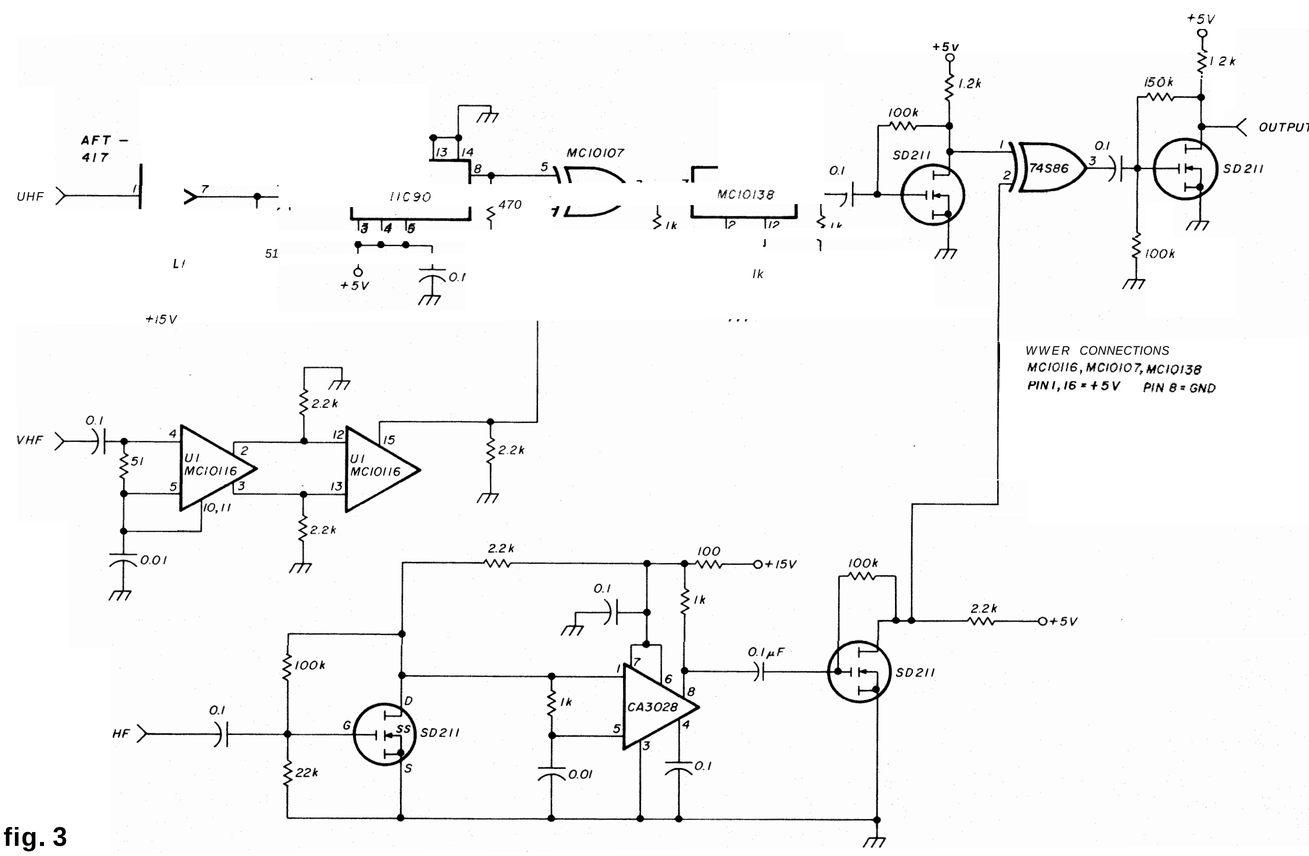


fig. 3

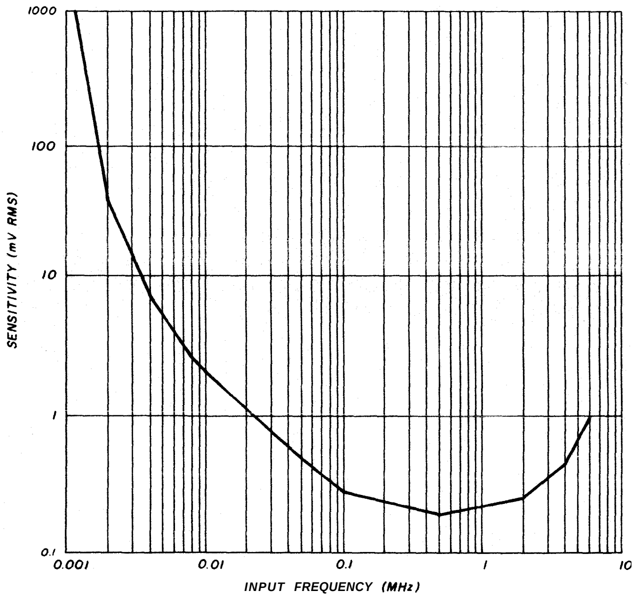


fig. 4. Measured sensitivity of the high-frequency range.

The advantage of high sensitivity is that a small loop soldered to a piece of coax can be used as a probe. This method of taking measurements won't load down a circuit. This prescaler is presently being used with the Intersil ICM 7028 counter chip.

table 2. Vhf input impedance.

frequency (MHz)	magnitude (ohms)	phase (degrees)
6	54	−2
10	54	0
20	52	0
30	51	+3
40	51	+7
50	52	+8
60	54	+10

table 3. Hf input impedance.

frequency (kHz)	magnitude (ohms)	phase (degrees)
500	12,500	−30
600	11,000	−34
700	11,000	−38
800	10,000	−42
900	9,800	−46
1000	9,000	−50
2000	5,300	−62
3000	3,800	−69
4000	2,900	−74
5000	2,300	−76
6000	2,000	−78

acknowledgment

I would like to thank Eric Blomberg, N1BF, for his time and advice.

ham radio

FCC study guide

Study Guide for Amateur Radio License Examinations

We're very happy to present, on the following pages, the complete text of the FCC Study Guide for all classes of Amateur License. This is FCC Bulletin 1035, dated January, 1980. You should use this material to find areas of study in each subject listed under the class of license you are trying for.

Note that the FCC lists two publications available from the Government Printing Office. Previously issued license-study manuals will still be helpful, but you'll have to do considerable interpreting to be sure that the subjects mentioned in this syllabus are thoroughly covered in those books.

As this goes to press, we've just learned of an electronics textbook that has been specifically revised to include study material listed in this FCC bulletin. It is *Electronic Communication,* by Robert *L.* Shrader, published by McGraw-Hill Book Company. This book is one of the best all-around electronic texts we've seen, and the inclusion of new material for the Amateur licenses can only make it more useful. This new fourth edition should be available soon after you read this, so watch for advertisements or write to Ham Radio's Bookstore for availability and price.

This Bulletin contains syllabi for the FCC Amateur radio examinations.

Why Are Amateur Radio Operator Examinations Required?

The examinations determine if you are qualified for the privileges conveyed by an Amateur radio license. Those privileges are many and diverse. As an Amateur radio operator, you will be allowed to build, repair, and modify your radio transmitters. You will be responsible for the technical quality of your station's transmissions. You will be allowed to communicate with Amateur radio operators in other countries around the world and, in some cases, send messages for friends. As you upgrade to the higher operator license classes, you will be allowed to communicate using not only telegraphy and voice, but also teleprinting, facsimile, and several forms of television. For such a flexible radio service to be practical, you and every other Amateur radio operator must thoroughly understand your responsibilities and develop the skills needed to operate your Amateur radio station properly.

What Subjects Do The Amateur Radio Examinations Cover?

The examinations cover the rules, practices, procedures, and technical material that you will need to know in order to operate your Amateur radio station properly. Each examination element is composed of questions which will determine whether you have an adequate understanding of the topics listed in the corresponding syllabus. For example, all Element 3 examination questions are derived from the Element 3 syllabus, which appears later in this Bulletin. To properly prepare for an examination, you should become knowledgeable about all of the topics in the syllabus for the element you will be taking. Every examination covers nine general subjects:

- Rules and Regulations
- Electrical Principles
- Signals and Emissions
- Circuit Components
- Practical Circuits
- Operating Procedures
- Antennas and Feedlines
- Radio Wave Propagation
- Amateur Radio Practice

Periodically, the syllabi are updated to reflect changing technology and Amateur radio practices. Comments on the study guide contents are welcome. Mail them to:

Personal Radio Branch
Federal Communications Commission
Washington, D.C. 20554

Where Can Study Manuals Be Obtained?

A study manual can be helpful in preparing for an examination. Several publishers offer manuals or courses based upon the material in this Bulletin. These may be found in many public libraries and radio stores. The FCC does not offer such manuals, nor recommend any specific publisher. However, you will find two FCC publications, *Part* 97 – *Rules and Regulations for the Amateur Radio Service* and *How to Identify and Resolve Radio-TV Interference Problems,* useful when preparing for the Amateur radio examinations. Copies are sold by the Superintendent' of Documents, U.S. Government Printing office, Washington, D.C. 20402. Specify stock number 004-000-00357-8 for Part 97 and stock number 004-000-00345-4 for the Radio-TV interference booklet.

STUDY TOPICS FOR THE NOVICE CLASS AMATEUR RADIO OPERATOR LICENSE EXAMINATION

(Element 2 Syllabus)

A. RULES AND REGULATIONS

Define:

(1) Amateur radio service 97.3(a)
(2) Amateur radio operator 97.3(c)
(3) Amateur radio station 97.3(e)
(4) Amateur radio communications 97.3(b)
(5) Operator license 97.3(d)
(6) Station license 97.3(d)
(7) Control operator 97.3(o)
(8) Third party traffic 97.3(v)

Novice Class Operator Privileges:

(9) Authorized frequency bands 97.7(e)
(10) Authorized emission (A1) 97.7(e)

Prohibited Practices:

(11) Unidentified communications 97.123
(12) Intentional interference 97.125
(13) False signals 97.121
(14) Communication for hire 97.112(a)

Basis and Purpose of the Amateur Radio Service Rules and Regulations:

(15) To recognize and enhance the value of the Amateur radio service to the public as a voluntary, non-commercial communication service, particularly with respect to providing emergency communications. 97.1(a)
(16) To continue and extend the Amateur radio operators' proven ability to contribute to the advancement of the radio art. 97.1(b)
(17) To encourage and improve the Amateur radio service by providing for advancing skills in both the communication and technical phases. 97.1(c)
(18) To expand the existing reservoir within the Amateur radio service of trained operators, technicians, and electronics experts. 97.1(d)
(19) To continue and extend the radio Amateurs' unique ability to enhance international good will. 97.1(e)

Operating Rules:

(20) U.S. Amateur radio station call signs 2.302 and FCC Public Notice
(21) Permissible points of communications 97.89(a)(1)
(22) Station logbook, logging requirements 97.103(a), (b); 97.105
(23) Station identification 97.84(a)
(24) Novice band transmitter power limitation 97.67(b), (d)
(25) Necessary procedure in response to an official notice of violation 97.137
(26) Control operator requirements 97.79(a), (b)

B. OPERATING PROCEDURES

(1) R-S-T signal reporting system
(2) Choice of telegraphy speed
(3) Zero-beating received signal
(4) Transmitter tune-up procedure
(5) Use of common and internationally recognized telegraphy abbreviations, including: CQ, DE, K, SK, R, AR, 73, QRS, QRZ, QTH, QSL, QRM, QRN

C. RADIO WAVE PROPAGATION

(1) Sky wave; "skip"
(2) Ground wave

D. AMATEUR RADIO PRACTICE

(1) Measures to prevent use of Amateur radio station equipment by unauthorized persons

Safety Precautions:

(2) Lightning protection for antenna system
(3) Ground system
(4) Antenna installation safety procedures

Electromagnetic compatability — identify and suggest cure:

(5) Overload of consumer electronic products by strong radio frequency fields
(6) Interference to consumer electronic products caused by radiated harmonics

Interpretation of S.W.R. readings as related to faults in antenna system:

(7) Acceptable readings
(8) Possible causes of unacceptable readings

E. ELECTRICAL PRINCIPLES

Concepts:

(1) Voltage
(2) Alternating current, direct current
(3) Conductor, insulator
(4) Open circuit, short circuit
(5) Energy, power
(6) Frequency, wavelength
(7) Radio frequency
(8) Audio frequency

Electrical Units:

(9) Volt
(10) Ampere
(11) Watt
(12) Hertz
(13) Metric prefixes, mega, kilo, centi, milli, micro, pico

F. CIRCUIT COMPONENTS

Physical appearance, applications, and schematic symbols of:

(1) Quartz crystals
(2) Meters (D'Arsonval movement)
(3) Vacuum tubes
(4) Fuses

G. PRACTICAL CIRCUITS

Block Diagrams:

(1) The stages in a simple telegraphy (A1) transmitter
(2) The stages in a simple receiver capable of telegraphy (A1) reception
(3) The functional layout of novice station equipment, including transmitter, receiver, antenna switching, antenna feedline, antenna, and telegraph key

H. SIGNALS AND EMISSIONS

(1) Emission type A1

Cause and cure:

(2) Backwave
(3i Key clicks
(4) Chirp
(5) Superimposed hum
(6) Undesirable harmonic emissions
(7) Spurious emissions

I. ANTENNAS AND FEEDLINES

Necessary physical dimensions of these popular high frequency antennas for resonance on amateur radio frequencies:

(1) A half-wave dipole
(2) A quarter-wave vertical

Common types of feedlines used at Amateur radio stations

(3) Coaxial cable
(4) Parallel conductor line

STUDY TOPICS FOR THE TECHNICIAN/GENERAL CLASS AMATEUR RADIO OPERATOR LICENSE EXAMINATION

(Element 3 Syllabus)

A. RULES AND REGULATIONS

(1) Control point 97.3(p)
(2) Emergency communications 97.3(w); 97.107
(3) Amateur radio transmitter power limitations 97.67
(4) Station identification requirements 97.84(b), (f), (g); 97.79(c)
(5) Third party participation in Amateur radio communications 97.79(d)
(6) Domestic and international third party traffic 97.114; Appendix 2, Art. 41, Sec. 2
(7) Permissible one-way transmissions 97.91
(8) Frequency bands available to the technician class 97.7(d)
(9) Frequency bands available to the general class 97.7(b)
(10) Limitations on use of Amateur radio frequencies 97.61
(11) Selection and use of frequencies 97.63
(12) Radio controlled model crafts and vehicles 97.65(a); 97.99
(13) Radioteleprinter emissions 97.69

Prohibited practices:

(14) Broadcasting 97.113
(15) Music 97.115
(16) Codes and ciphers 97.117
(171 Obscenity, indecency, profanity 97.119

B. OPERATING PROCEDURES

(1) Radiotelephony
(2) Radio teleprinting
(3) Use of repeaters
(4) Vox transmitter control
(5) Full break-in telegraphy
(6) Operating courtesy
(7) Antenna orientation
(8) International communication
(9) Emergency preparedness drills

C. RADIO WAVE PROPAGATION

(1) Ionospheric layers; D, E, F1, F2
(2) Absorption
(3) Maximum usable frequency
(4) Regular daily variations
(5) Sudden ionospheric disturbance
(6) Scatter
(7) Sunspot cycle
(8) Line-of-sight
(9) Ducting, tropospheric bending

D. AMATEUR RADIO PRACTICE

Safety precautions:

(1) Household ac supply and electrical wiring safety
(2) Dangerous voltages in equipment made inaccessible to accidental contact

Transmitter performance:

(3) Two tone test
(4) Neutralizing final amplifier
(5) Power measurement

Use of test equipment:

(6) Oscilloscope
(7) Multimeter
(8) Signal generators
(9) Signal tracer

Electromagnetic compatibility; identify and suggest cure:

(10) Disturbance in consumer electronic products caused by audio rectification

Proper use of the following station components and accessories:

(11) Reflectometer (VSWR meter)
(12) Speech processor — RF and AF
(13) Electronic T-R switch
(14) Antenna tuning unit; matching network

(15) Monitoring oscilloscope
(16) Non-radiating load; "dummy antenna"
(17) Field strength meter; S-meter
(18) Wattmeter

E. ELECTRICAL PRINCIPLES

Concepts:

(1) Impedance	(4) Inductance
(2) Resistance	(5) Capacitance
(3) Reactance	(6) Impedance matching

Electrical units:

(7) Ohm
(8) Microfarad, picofarad
(9) Henry, millihenry, microhenry
(10) Decibel

Mathematical relationships:

(11) Ohm's law
(12) Current and voltage dividers
(13) Electrical power calculations
(14) Series and parallel combinations; of resistors, of capacitors, of inductors
(15) Turns ratio; voltage, current, and impedance transformation
(16) Root mean square value of a sine wave alternating current

F. CIRCUIT COMPONENTS

Physical appearance, types, characteristics, applications, and schematic symbols for:

(1) Resistors
(2) Capacitors
(3) Inductors
(4) Transformers
(5) Power supply type diode rectifiers

G. PRACTICAL CIRCUITS

(1) Power supplies
(2) High-pass, low-pass, and band-pass filters
(3) Block diagrams showing the stages in complete am, ssb, and fm transmitters and receivers

H. SIGNALS AND EMISSIONS

(1) Emission types AØ, A3, F1, F2, F3
(2) Signal; information
(3) Amplitude modulation
(4) Double sideband
(5) Single sideband
(6) Frequency modulation
(7) Phase modulation
(8) Carrier
(9) Sidebands
(10) Bandwidth
(11) Envelope
(12) Deviation
(13) Overmodulation
(14) Splatter
(15) Frequency translation; mixing, multiplication
(16) Radioteleprinting; audio frequency shift keying, mark, space, shift

I. ANTENNAS AND FEEDLINES

Popular Amateur radio antennas and their characteristics:

(1) Yagi antenna
(2) Quad antenna
(31 Physical dimensions
(4) Vertical and horizontal polarization
(5) Feedpoint impedance of half-wave dipole, quarter wave vertical
(6) Radiation patterns; directivity, major lobes

Characteristics of popular Amateur radio antenna feedlines; related concepts:

(7) Characteristic impedance
(8) Standing waves
(9) Standing wave ratio; significance of
(10) Balanced, unbalanced
(11) Attenuation
(12) Antenna-feedline mismatch

STUDY TOPICS FOR THE ADVANCED CLASS AMATEUR RADIO OPERATOR LICENSE EXAMINATION

(Element 4A Syllabus)

A. RULES AND REGULATIONS

(1) Frequency bands available to the advanced class Amateur radio operator and limitations on use 97.7(a); 97.61
(2) Automatic retransmission of Amateur radio signals and signals from other radio services 97.3(x); 97.113; 97.126
(3) Amateur radio stations in repeater operation 97.3(1); 97.85; 97.61(c)
(4) Amateur radio stations in auxiliary operation 97.311); 97.86; 97.61(d)
(5) Remote control of Amateur radio stations 97.3(m)(2); 97.88
(6) Automatic control of Amateur radio stations 97.3(m)(3)
(7) Control link 97.3(n)
(8) System network diagram 97.3(u)
(9) Station identification 97.84(c), (d), (e)
(10) Station log requirements 97.103(c), (d), (e), (f), (g)
(11) Height limitations for Amateur radio station antenna structures, including FAA notification criteria, and calculation of height above average terrain 97.45; 97.67(c); Appendix 5

B. OPERATING PROCEDURES

(1) Facsimile transmission
(2) Slow-scan television transmission

C. RADIO WAVE PROPAGATION

(1) Sporadic-E
(2) Selective fading
(3) Auroral propagation
(4) Radio-path horizon

D. AMATEUR RADIO PRACTICE

Use of test equipment:

(1) Frequency measurement devices
(2) Grid-dip meter; solid state dip meter
(3) Performance limitations of oscilloscopes, meters, frequency counters; accuracy, frequency response, stability

Electromagnetic compatibility:

(4) Intermodulation interference
(5) Receiver desensitizing
(6) Cross modulation interference
(7) Capture effect

E. ELECTRICAL PRINCIPLES

Concepts:

(1) Reactive power
(2) Series and parallel resonance
(3) Skin effect
(4) Fields, energy storage, electrostatic, electromagnetic

Mathematical relationships:

(5) Resonant frequency, bandwidth, and ''Q'' of R-L-C circuits, given component values
(6) Phase angle between voltage and current, given resistance and reactance
(7) Power factor, given phase angle
(8) Effective radiated power, given system gains and losses
(9) Replacement of voltage source and resistive voltage divider with equivalent circuit consisting of a voltage source and one resistor (an application of Thevenin's theorem, used to predict the current supplied by a voltage divider to a known load)

F. CIRCUIT COMPONENTS

Physical appearance, types, characteristics, applications, and schematic symbols for the following:

(1) Diodes; zener, tunnel, varactor, hot-carrier, junction, point contact, pin
(2) Transistors; NPN, PNP, junction, unijunction, power, germanium, silicon
(3) Silicon controlled rectifier, triac
(4) Light emitting diode, neon lamp
(5) Crystal lattice ssb filters

G. PRACTICAL CIRCUITS

(1) Voltage regulator circuits; discrete and integrated
(2) Amplifiers; Class A, AB, B, C; characteristics of each type
(3) Impedance matching networks; PI, L, PI-L
(4) Filters; constant K, M-derived, band-stop, notch, modern-network-theory, pi-section, T-section, L-section (not necessary to memorize design equations; know general description, characteristics, responses, and applications of these filters)
(5) Oscillators; various types and their applications; stability

Transmitter and receiver circuits – know purpose of each, and how, basically, each functions:

(6) Modulators; am, fm, balanced
(7) Transmitter final amplifiers
(8) Detectors, mixer stages
(9) RF and IF amplifier stages

Calculation of voltages, currents, and power in common Amateur radio oriented circuits:

(10) Common emitter class A transistor amplifier; bias network, signal gain, input and output impedances
(11) Common collector class A transistor amplifier; bias network, signal gain, input and output impedances

Circuit design; selection of circuit component values:

(12) Voltage regulator with pass transistor and zener diode to produce given output voltage
(13) Select coil and capacitor to resonate at given frequency

H. SIGNALS AND EMISSIONS

(1) Emission types A4, A5, F4, F5
(2) Modulation methods
(3) Deviation ratio
(4) Modulation index
(5) Electromagnetic radiation
(6) Wave polarization
(7) Sine, square, sawtooth waveforms
(8) Root mean square value
(9) Peak envelope power relative to average
(10) Signal to noise ratio

I. ANTENNAS AND FEEDLINES

(1) Antenna gain: beamwidth
(2) Trap antennas
(3) Parasitic elements
(4) Radiation resistance
(5) Driven elements
(6) Efficiency of antenna
(7) Folded, multiple wire dipoles
(8) Velocity factor
(9) Electrical length of a feedline
(10) Voltage and current nodes
(11) Mobile antennas
(12) Loading coil; base, center, top

STUDY TOPICS FOR THE AMATEUR EXTRA CLASS AMATEUR RADIO OPERATOR LICENSE EXAMINATION

(Element 4B Syllabus)

A. RULES AND REGULATIONS

(1) Frequency bands available to the U.S. Amateur radio operator and limitations on their use including variations for regions 1 and 3 97.61; 97.95
(2) Space Amateur radio stations 97.3(i)
(3) Purity of emissions 97.73
(4) Mobile operation aboard ships or aircraft 97.101
(5) Races operation Part 97, Subpart F
(6) Points of communications 97.89

B. OPERATING PROCEDURES

ii) Use of Amateur radio satellite
(2) Amateur fast-scan television

C. RADIO WAVE PROPAGATION

(1) EME; "moonbounce"
(2) Meteor burst
(3) Trans-equatorial

D. AMATEUR RADIO PRACTICE

Use of test equipment:

(1) Spectrum analyzer; interpret display; display of transmitter output spectrum, such as commonly found in new product review articles in Amateur radio magazines
(2) Logic probe; indication of high or low state, pulsing state

Electromagnetic compatibility:

(3) Vehicle noise suppression; ignition noise, alternator whine, static
(4) Direction finding techniques; methods for location of source of radio signals

E. ELECTRICAL PRINCIPLES

Concepts:

(1) Photoconductive effect
(2) Exponential charge/discharge

Mathematical relationships; calculations:

(3) Time constant for R-C and R-L circuits (including circuits with more than one resistor, capacitor or inductor)
(4) Impedance diagrams; basic principles of Smith chart
(5) Impedance of R-L-C networks at a specified frequency
(6) Algebraic operations using complex numbers; real, imaginary, magnitude, angle

F. CIRCUIT COMPONENTS

Physical appearance, types, characteristics, applications, and schematic symbols for:

(1) Field effect transistors; enhancement, depletion, MOS, CMOS, N-channel, P-channel
(2) Operational amplifier and phase-locked loop integrated circuits
(3) 7400 series TTL digital integrated circuits
(4) 4000 series CMOS digital integrated circuits
(5) Vidicon; cathode ray tube

G. PRACTICAL CIRCUITS

(1) Digital logic circuits; flip-flop, multivibrator, and/or/nand/nor/gates
(2) Digital frequency divider circuits; crystal marker, counters
(3) Active audio filters using integrated operational amplifiers

High performance receiver characteristics

(4) Noise figure, sensitivity
(5) Selectivity
(6) Dynamic range

Calculation of voltages, currents, and power in common amateur radio oriented circuits:

(7) Integrated operational amplifier; voltage gain, frequency response
(8) F.E.T. common source amplifier; input impedance

Circuit design; selection of circuit component values:

(9) L-C preselector with fixed and variable capacitors to tune a given frequency range
(10) Single stage amplifier to have desired frequency response by proper selection of bypass and coupling capacitors

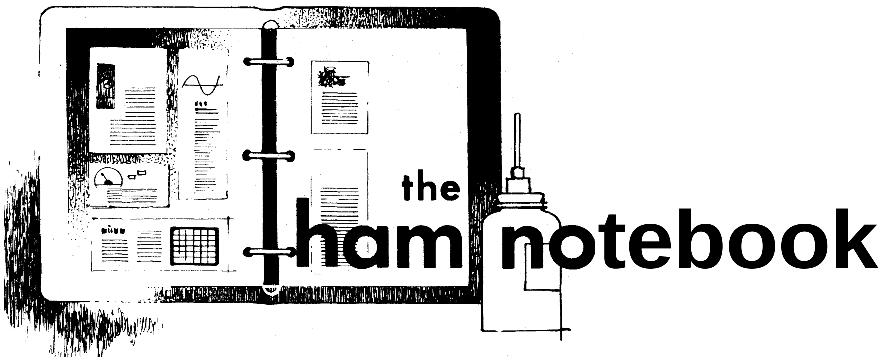
H. SIGNALS AND EMISSIONS

(1) Pulse modulation; position, width
(2) Digital signals
(3) Narrow band voice modulation
(4) Information rate vs. bandwidth
(5) Peak amplitude of a signal
(6) Peak-to-peak values of a signal

I. ANTENNAS AND FEEDLINES

(1) Antennas for space radio communications; gain, beamwidth, tracking
(2) Isotropic radiator; use as a standard of comparison
(3) Phased vertical antennas; resultant patterns, spacing in wavelengths
(4) Rhombic antennas; advantages, disadvantages
(5) Matching antenna to feedline; delta, gamma, stub
(6) Properties of 118, 114, 318, and 112 wavelength sections of feedlines; shorted, open

ham radio



counter control pulses

In a counter, the necessary control pulses can be generated from the crystal-controlled clock train without much additional circuitry. The most difficult problem is to find a suitable time into which the strobe and reset pulses can be inserted after the count enable pulse.

By inverting the 2-Hz output from the divide-by-five section, the divide-by-two section of the 7490 is triggered 0.1 second earlier, thus creating a 0.1 second blanking pulse by the difference in time between the negative and positive edges of the 2-Hz pulse. The 7421 quad-AND gate, shown in fig. **1,** is wired to combine the appropriate pulses for the 50 ms strobe and reset pulses, after the 1.0 second count enable pulse and in the alternate second of the counting sequence. The frequencies shown are for a one-second count, one-second strobe and reset, but the principle can be used for most counting sequences.

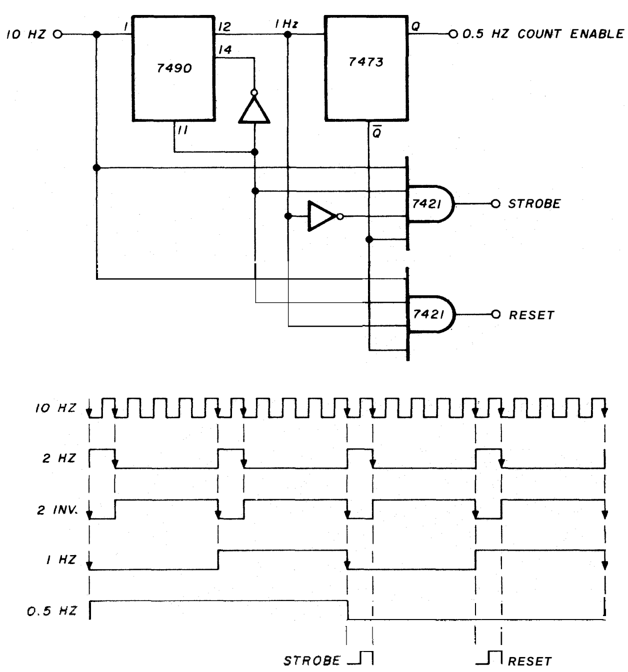


fig. 1. Schematic and timing diagram that shows the generation of the strobe and reset pulses.

R. S. Naslund, W9LL

dc-dc converter increases Gunnplexer frequency swing

Microwave Associates' Gunnplexers are easily tuned if a varying dc voltage of 5-20 volts is applied to the varactor tuning diode. Field operation from a 12-volt battery limits tuning range somewhat. The circuit in **fig.** 2 allows maximum use of the tuning varactor.

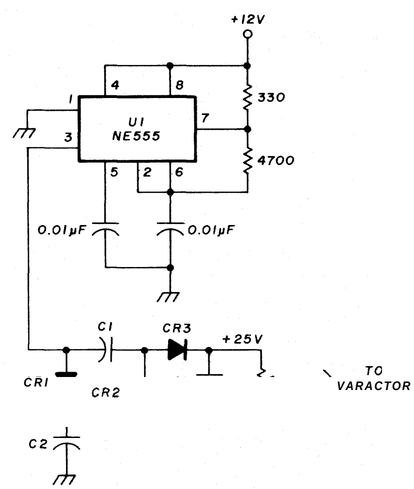


fig. 2. The dc-dc converter. C1-C3 aren't critical: values from 0.47 to 1μF were successfully tested. CR1-CR3 are 1N4148 diodes or equivalent.

U1 generates a high-frequency ac voltage, which is rectified by a voltage-tripling circuit composed of C1-C3 and CR1-CR3. Output voltage is approximately 25 volts.

Jim **Kearman, W1XZ**

noise figure relationships

In my radio class the question of noise figure and its relationship to noise temperature and sensitivity comes up time after time. Confronting the Amateur are terms such as:

1. Noise figure of a receiver in dB.

2. Receiver sensitivity in microvolts to produce a given signal-plus-to-noise ratio (usually 10 dB).

3. The dBm at different bandwidths ($B\ kHz$).

4. The equivalent noise temperature (T_e in degrees Kelvin).

Without going into their theory or derivation, the following formulas provide a convenient way to determine the relationship between these terms.

1. The input noise power at a standard temperature of $T_0 = 290$ K and a bandwidth of 1 kHz is 4×10^{-18} watt.

2. $NF = 10 \log \frac{e^2 \times 10^6}{4RB}$ **(1)**

where $e = \mu V$ to produce the desired

$$\frac{S\&N}{N} \ ratio$$

R = input resistance
B = bandwidth in kHz

example: $e = 0.3\ \mu V$
$R = 50\ ohms$
$B = 2.7\ kHz$

$$NF = 10 \log \frac{(0.3)^2 \times 10^6}{4 \times 50 \times 2.7} = 22\ dB \quad \textbf{(2)}$$

3. Given the NF in dB, B in kHz and the input resistance, R, the μV sensitivity can be determined by

$$e = \sqrt{\frac{10^{\frac{NF}{10}} \times 4RB}{10^6}}\ \mu V \quad \textbf{(3)}$$

example: $NF = 21$ dB
$B = 2\ MHz$
$R = 50\ ohms$

$$e = \sqrt{\frac{10^{2.1} \times 4 \times 50 \times 2000}{10^6}} = 7\ \mu V$$

4. The equivalent noise temperature, T_e, is more convenient to use with receivers having a very low NF:

$$T_e = (F - 1)\ 290\ K$$

$F = 10^{\frac{NF}{10}}$, where NF is in dB **(4)**

example:
$NF = 1.95\ dB$ or
$F = 10^{0.195} = 1.567$

$$T_e = (10^{0.195} - 1)\ 290 = 164.4K$$

5. Given the T_e of a receiver, NF may be determined as follows:

$$NF = 10 \log \frac{T_e + T_0}{T_0} \quad \textbf{(5)}$$

example: $T_e = 190\ K$
$T_0 = 290\ K$ standard

$$NF = 10 \log \frac{190 + 290}{290} = 2.19\ dB$$

6. At $T_0 = 290\ K$, the noise power in dBm (dB below 1 milliwatt) is

$$dBm = 10 \log \frac{1 \times 10^{-3}}{4 \times 10^{-21}} = 174\ dBm \quad \textbf{(6)}$$

at a B of 1 Hz

To determine dBm at the desired bandwidth, B, in kHz

$$dBm = 144 - 10 \log B \quad \textbf{(7)}$$

example: $B = 3\ kHz$

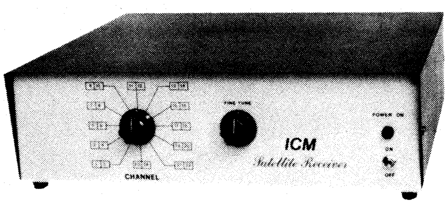
$$dBm = 144 - 10 \log 3$$
$$= 144 - 4.77$$
$$= 139.2\ dBm$$

I.L. **McNally, K6WX**

More Details? CHECK—OFF Page 94

NEW products

ground station satellite receiver



A new satellite receiver, covering 3.7-4 GHz, is available from International Crystal Mfg. Co., Inc.

The TV 4200 receiver is fully tunable and provides standard dual audio outputs of 6.2 and 6.8 MHz, with other outputs available. The receiver has a built-in LNA power supply and output levels are compatible with video monitor or VTR input. It's priced at $1,995.

For more information, write International Crystal Mfg. Co., Inc., 10 North Lee, Oklahoma City, Oklahoma 73102.

Bird rf power analyzer

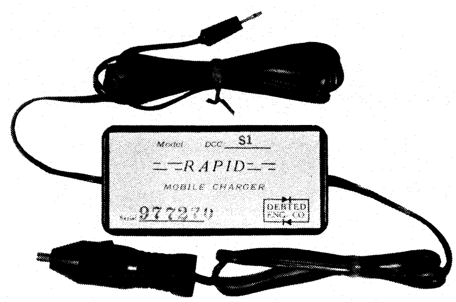
A new era in rf power measurement was announced by THRULINE® Wattmeter designer, Bird Electronic Corp., with the introduction of the new series 4380 RF Power Analyst™. First of the series, the portable model 4381 is a multi-purpose, digital, directional rf wattmeter for power levels from 1/10 watt to 10,000 watts, and from ½ to 2300 MHz. CW or fm power in both forward or reflected directions is displayed in watts or dBm at the push of a button. VSWR is calculated continuously, and indicated through a fifth button, as is dB return loss. Button seven and eight are for peak envelope power (as in SSB transmissions) in watts, and the ninth button calls up per cent modulation. The final set of three buttons make tuning a transmitter, matching an antenna or tweaking rf components a fast and simple task. A deita (Δ) function identifies either rise or fall in displayed values, while a minimum or maximum memory recalls optimum conditions during adjustments. Other models in the 4380 series measure to 250 kW, or are panel mounted.

This new generation of rf wattmeters with nine-mode system versatility was designed around existing Bird Plug-in Elements, which determine full-scale power and frequency range. Once a set of two elements is chosen (for incident and reflected power), the large LED display correctly places the decimal point, making mental notes of multipliers unnecessary. Overranging of up to 120 per cent in watts, and 400 per cent in dBm, often obviates changing to a higher-power element, and retains "up-scale" accuracy.

The RF Power Analyst™ is the first uniquely different directional rf wattmeter system for gauging and analyzing rf power since the Bird THRU-LINE@model 43 was designed 25 years ago. It calculates parameter products that formerly required consulting a graph or chart, reveals whether an undesirable hum is present and – if so – how much, and permits minimum/maximum power searches even with closed eyes. Accuracy of model 4381 is ± 5 per cent of nominal full scale and VSWR is a low 1.05 max to 1 GHz in 50-ohm systems.

Price of Model 4381 RF Power Analyst is $590. Delivery is 90 days after receipt of order, from Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139.

mobile rapid charger



DebTed Engineering introduces a line of 12-volt operated rapid chargers for Amateur and commercial use, available exclusively through Debco Electronics. The rapid chargers come with a cigarette lighter plug on the input side and the appropriate charging plug on the output side. Models are currently available for the Tempo S1, Wilson Mark II, and Wilson Mark IV with direct plug-in capabilities. Units are also available for other transceivers.

A fully discharged battery can be recharged in 4-6 hours and the unit can be used during transmit, receive, and off periods. It will not damage batteries if left connected for prolonged periods of time, due to automatic shut-off circuitry. Cord lengths will allow convenient use of radio while charging. Further applications include rapid charging from 12-volt power supplies in motor homes and during emergencies.

Price of the rapid charger is $29.95. For more information, write Debco Electronics, P.O. Box 9169, Cincinnati, Ohio 45209.

Hamtronics 1980 catalog

Hamtronics, Inc., has announced a new 1980 catalog, which is yours for the asking. The 24-page catalog features many types of kits for the Radio Amateur or two-way shop. Exciting new products in the catalog in-

ham radio magazine

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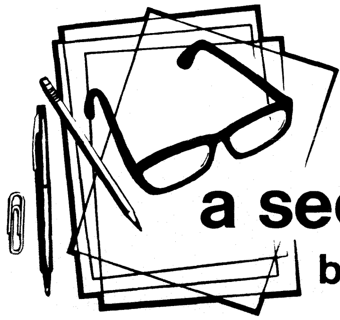


a second look

by Jim Fisk

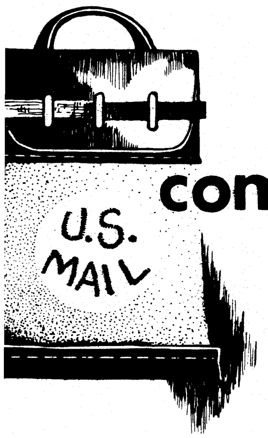
From time to time I am taken to task for editing a monthly Amateur Radio magazine that is "much too technical" and is written for radio engineers, not hams. I do not disagree with that premise in principle, only in degree – I feel strongly that most of our feature articles can be understood and applied by Amateurs who are interested in the technical aspects of radio: Amateurs who are still interested in the subtle details of designing their own circuits and building some of their own station accessories. Even with the great influx over the past couple of years of new Radio Amateurs who are primarily communicators with little technical knowledge, I believe the majority of licensed Amateurs still spend a large portion of their hobby time in their home workshops.

Subscribers who have been regular *ham radio* readers since I put together the first issue more than twelve years ago know that, from the beginning, *ham radio* has always placed the emphasis on radio theory and technique; operating news and views were left to others. Over the years we have tried to stay abreast of the state of the art, although at times this has been nearly impossible because of the great and rapid advances in developing technology. And, in general, when viewed in terms of the technology of the day, a better technical background is required now than it was ten years ago to understand a comparable level of circuit theory. The corollary, of course, is that what we consider to be unduly complex and difficult to understand in 1980 will likely seem relatively simple in 1990. This presents an interesting problem because as we continue to publish up-to-date radio circuits and projects, we run the danger of leaving a few dedicated readers behind. That is not our intent, and we shall make every effort to continue to appeal to as wide an audience as possible.

The microprocessor revolution has also greatly affected the technical content of *ham radio,* although not in the way you would expect; when most of the magazines jumped on the computer bandwagon, we continued to stress analog circuitry and presented only those computer topics that were closely related to radio communications. This has been highly popular, but many radio engineers and technicians who previously depended upon the industrial magazines for up-to-date design information became regular readers of *ham radio;* these are the same people who design the frequency-synthesized solid-state hf transceivers that are currently popular. Thus a dilemma: to publish up-to-date design articles that will ultimately improve the type of communications equipment we all have available, or continue to present decade-old technology to a 1980 world? Our answer has been a carefully chosen mix of established older techniques and complex new ones; when we stray too far one way or the other, we hear about it!

The continuing series on Yagi antenna design by W2PV which returns this month after a two-month hiatus is a good example of the type of article some readers feel is out of place; it explores antenna design and performance at a level normally reserved for engineering journals. If you've seen any engineering magazines lately, though, most of their articles are centered around microprocessor circuits; antenna articles are few and far between unless they're for satellite earth terminals. Those editors are appealing to a completely different market, so W2PV's Yagi series would be out of place – more important, it would not have been read by those people who are designing and building new high-gain antennas for the Amateur market. I feel we made the right decision when we accepted it for publication – it's likely to have more impact on high performance high-frequency antenna design than anything published in the last 20 years.

Jim Fisk, W1HR
editor-in-chief



comments

Hallicrafters story

Dear HR:

"The Hallicrafters Story" did not end with World War 2! Several years later, in Korea, the BC-610 transmitters — tired old rigs that they were — still carried the brunt of our radio communications during the early days of the Korean conflict. We of the Military Advisory Group in Korea inherited several BC-610s and a few SCR-399 rigs when the American Army of occupation left in 1949. When the invasion of south Korea started on June 25, 1950, we fired-off a SC-399 and established an emergency communications link with Tokyo.

When the Hahn river bridge was blown sky high two days later, one of our SCR-399 trucks was jammed in heavy traffic — on the north ramp of the bridge. On the south side of the river, I was operating another SCR-399 in contact with Tokyo. I heard the operator on the north side of the river and he was in contact with Tokyo, too (all on CW) but for some reason he couldn't hear my signals! He was telling Tokyo that the bridge had blown up in his face, the traffic jam was so bad that he might have to destroy the radio, and there were enemy tanks wandering around the city of Seoul! Then he went off the air; I figured we could scratch one radio, but Sgt. Francisco was not giving up so easy!

A few hours later he was back on the air — reporting that he had gotten the truck and trailer of his SCR-399 across the river on a barge! Now he was out of gasoline and the enemy was lobbing mortars at him! We rushed a jeep up the river with a couple of jerry-cans of gas, and a bit later the truck I never expected to see again pulled into our camp!

A few nights later we made a crash retreat south; there was a report that enemy tanks had gotten behind us and cut the roads, so we drove down trails so narrow that we scraped brush on both sides of the road. I operated my BC-610 mobile in motion all that night, and a radioman in Yokohama kept contact with me straight through!

During the battle for Taejon the choke in the high-voltage supply of my BC-610 went down to ground. We replaced it with a resistor and went right back on the air — with a real pretty note and a few more watts of power! (A Korean mechanic rewound the choke for us and it was still working two years later.)

The roads were so rough in Korea that wheel-bolts would sometimes crystalize and break off the trucks. Once a modulation transformer broke loose and wiped the whole audio deck clean — right down to reducing everything on the deck to dust! So what! We were operating all CW anyway, so we didn't miss the audio deck in our BC-610.

One night there was some kind of blackout, and the whole high-frequency band was just a hiss — not a signal to be heard; of course, that was the night the big brass had an urgent message that just had to get through! We fired off both SCR-399 rigs, one with a horizontal antenna, the other vertical, and we adjusted the frequency of one about 800 hertz off the frequency of the other, then keyed them both together; what a God-awful signal that made, but we got the message through. The only message that got through from all Korea that night!

Our SCR-399s served all through the Pusan perimeter days, went north when the breakout came, and made the long retreat back down the peninsula when the Chinese clobbered us. By the time the cease-fire came, we had added another ten-thousand hours to those old war-weary rigs left over from World War 2!

James Houldsworth, W1TVN
Pittsfield, Massachusetts

Dear HR:

I feel compelled to let you and Bill Orr know how much I enjoyed the "Hallicrafters Story" in the November *ham* ***radio.*** To most of us hams over the age of 50, the mere mention of the HT-4 brings back fond recollections. Some of us also spent time around the BC-610 during and after WWII, and during the golden years of military surplus that followed.

It was the kind of article that I could not lay down until I finished it, and I have read it a second time. Bill Orr should be commended for yet another very fine article.

Thanks again and keep up the good work.

Marshall B. Turner, KØADM
Parkville, Missouri

talking digital readout

Dear HR:

I want to congratulate you on the fine article, "Talking Digital Readout for Amateur Transceivers," in the June, 1979, issue. Pete Tanner, N5EJ, built one for Jerry Thomas, KA5GBP, who is blind, and it works great with his TS 120S. You are to be commended for providing this assistance to handicapped Amateurs.

Sammy Neal, N5AF
Cleveland, Texas

three-element quad

for 15-20 meters which uses circular elements

Development of a circular-element quad beam from conception to final result — the *Dream Beam*

This project started in the winter of 1977-78 when I became active after having been off the air for over 40 years. I wanted a good antenna for 15 and 20 meters, but my location precluded a big beam. This article describes a quad antenna using circular elements rather than the usual square element configuration. Advantages of using circular rather than square elements are described together with construction details for building your own antenna. Development of the idea is discussed, beginning with a single-element circular antenna for 15 meters. The final version, a three-element circular quad, has given a good account of itself.

The idea was inspired by a bicycle wheel. Structural rigidity for the circular quad elements is provided by "spokes" radiating from the element hubs to the elements. Another bonus: the circular loop has a 0.9 dB gain over a square or diamond.'

May I now present the ***Dream Beam,*** its early development, model tests, construction, and performance.

early antennas

The first attempt was a very small model. The "bicycle" rim or tire (conductor) was made by springing a length of small-diameter stiff plastic tubing into a circle about two feet (0.6 meter) in diameter. The wheel hub was a 6-inch (153-mm) length of %-inch (6.5-mm) wood dowel with small plywood flanges on each end. Holes were drilled in the flanges all around, spaced at 45-degree intervals. Eight pairs of "spokes" were made from kite string connecting the "tire" in pairs to the holes in the hub flanges. Sure enough there was the wheel! It proved to be what I hoped for – very lightweight, surprisingly strong, resilient, simple, and a near perfect circle.

One-element circular quad loop for 15 meters. This work led to an attempt at a single-element, full-size antenna for 15 meters. The conductor for this antenna was 318-inch (9.5-mm) aluminum tubing lengths spliced together to a total length of about 46 feet (14 meters). This assembly was easily sprung into a circle, and the ends were attached to a feed-point insulator that included an SO-239 coax connector. The hub was a 30-inch (765 mm) length of l-inch (26-mm) PVC plastic pipe to which some end flanges had been fitted.

Eight pairs of spokes were used, which were made from 40-pound (18-kg) test monofilament nylon fish line. It did indeed look like a big bicycle wheel! It was about 15 feet (5 meters) in diameter, and I wondered if I could ever get it up into the vertical position. I gingerly picked it up by the hub and, to my pleasant surprise, found it quite stable and easy to handle. The whole element weighed only about 2½ pounds (1 kg) – so light that I could carry it up a ladder to the roof alone using one hand for myself and one for

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the antenna. The 15-meter antenna performed beautifully, and I used it for several months. It proved to my satisfaction that the construction principle was sound and would work.

One-element circular quad loop for 15-20 meters. A single-element 15 and 20 meter antenna was the next step. It was similar to the 15 meter version but quite a bit larger. About 70 feet (22 meters) of %-inch (12.5-mm) aluminum tubing was sprung into a 22-foot (7-meter) diameter circle and connected with a similar feed point insulator. The hub was 4-foot-&inch (1.4-meter) length of 1½-inch (38-mm) PVC pipe with flanges on each end. The number of spokes was increased to twelve pairs, and these were made from neavier 80 pound i36 kg) test monofiiament nylon. The 15-meter element of no. 16 (1.3 mm) copper antenna wire was attached to the spokes in much the way a spider spins a web: rather than a true circle, it was a regular twelve-sided polygon. This antenna was a success and I worked much DX with it.

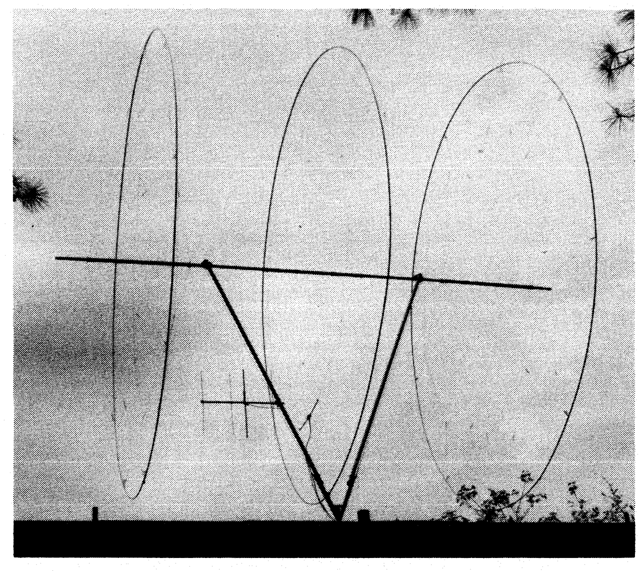
two-element beam for 15-20 meters

The two-element beam was a natural and easy development. Reflector elements were made just like the driven elements, except there were no feed-point insulators, of course. And did it work! I received excellent reports and many compliments on my signal from all over the world. I could usually contact any station I could hear. Europeans, ZLs, and VKs were worked with the greatest of ease and almost at will. It was a quiet receiving antenna and seemed to have excellent directional properties.

three-element beam for 15-20 meters

Then I began to think of installing a director. Would the antenna be further improved with a third element, a director? If so, could I get one up there? Yes, there might be a way, and I dreamed up the basics of the three-element version.

I was about to begin building when nagging doubts began to creep through my mind. The message said, "Look OM, you have a fine antenna now, but you really know very little about it. Do you know what the beamwidth is, what the pattern looks like, and whether you have the optimum reflector length and spacing? If you put up a director, what length will be best? What spacing will be optimum? What will the front-to-back ratio be? You really don't know these things. Enlarging the beam will take a lot of time and much work. You may fall flat on your face!" I had to agree. I decided to postpone the director and



The *Dream Beam* for *15* and 20 meters: like three giant bicycle wheels on one axle. The small 2-meter array (lower left center) gets a free ride.

embarked on a three-month period of model testing to find out where I was, where I wanted to go, and how I was going to get there.

model tests

In my backyard antenna range, a one-watt 2-meter carrier from a dipole illuminated the model antenna under test. Induced currents were observed and recorded and patterns plotted. After literally hundreds upon hundreds of patterns, I felt I had the answers to all the questions – plus much other valuable and interesting data.

One particularly interesting result concerns a very closely spaced two-element (driver and reflector) circular loop beam. If the reflector is made 1.018 times the length of the driven element and spaced at 0.065 wavelength, a very nice beam results. At 146 MHz this spacing is only about 5% inches (140 mm). A beautiful 2-meter mobile beam antenna could be made using an aluminum loop driver. The reflector could be bracketed right to the driver with nonconducting material. You can't add a director to this arrangement; it will not work that way.

The model tests of the two-element configuration showed that I had been lucky. My earlier guesses at reflector length and spacing were reasonably close to optimum.

The bulk of the work was with the three-element configuration. Director and reflector lengths were varied, and the effects of various spacings investigated. The goal was to zero in on the optimum antenna. This was eventually done and the *Dream Beam* was made to the following dimensions:

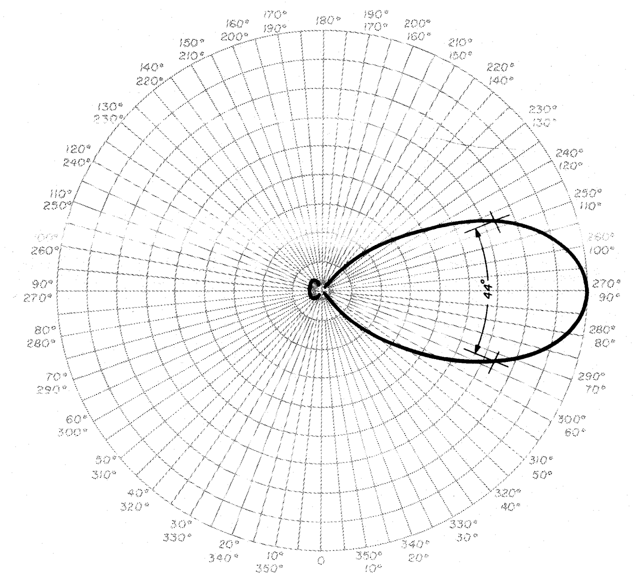


fig. 1. Horizontal radiation pattern of three-element, full-size circular loop beam at 145 MHz.

element lengths

14,275 kHz	wavelength	feet	(meters)
director element	0.977	67.34	(20.5)
director spacing	0.123	8.50	(2.6)
driven element	0.990	68.23	(21.0)
reflector spacing	0.123	8.50	(2.6)
reflector element	1.008	69.46	(21.2)
21,350 kHz			
director element	0.974	44.90	(14.0)
director spacing	0.141	6.50	(2.0)
driven element	0.987	45.50	(14.0)
reflector spacing	0.184	8.50	(2.6)
reflector element	1.005	46.30	(14.0)

You'll notice that the proportions of the 15-meter elements don't agree with those of the 20-meter elements. This comes about by mechanical considerations and the fact that it was necessary to tune the 15-meter element to resonance at 21,350 kHz. While the 15-meter array doesn't quite meet the optimum dimensions, only minimal harm comes from the dimensions used. The turning radius is only 14 feet (4 meters).

Fig. 1 is the horizontal pattern from model experiments. Numbers of vertical patterns were also made, and it always turned out that they were almost identical to the horizontal patterns.

impedance

The feedpoint impedance of a fullwave loop has been estimated to be in the range of 100-130 ohms. For the earlier one- and two-element antennas a very close match to 52-ohm coax was made with the use of quarter-wave matching sections of 75-ohm coax. In the three-element configuration the feed point impedance becomes much reduced. The impedance of the 20-meter driven element turned out to be about 20 ohms; the 15-meter element about 27 ohms. Excellent matches were made between the transmission lines and the antenna elements using matching stubs cut for these impedances.

circle versus rectangle

In general character there is much similarity between the *Dream Beam* and the familiar quad. Construction arid configuration aside, there are some electrical differences. The resonant length of the circular configuration is somewhat less than that of the square or diamond. The resonant length of the circular loop is about *974/f*. The usual formula for the quad is *1005/f*. The parasitic element lengths are nearer to the length of the driven element in the case of circular loops. The director element is 1.3 per cent shorter; the refiector element is 1.7 per cent longer.

construction

To start at the bottom: A bearing and rotator are installed in the attic of my house. The installation is remarkably similar to the one described in detail by WØYBV.[2] He and I were perhaps cutting holes through our roofs at about the same time! The lower part of the Y-base (fig. 2) is a piece of 1 %-inch (38-mm) steel pipe. To the top were welded two U-shaped steel members from a junk pile. They were just the right size and the V-struts were attached to them with U-bolts.

The V-struts are 12-foot (4-meter) lengths of 2-inch (51-mm) aluminum tubing. They are connected to the two 1 %-inch (38-mm) diameter boom halves with U-bolts using tie plates cut from 118-inch (3-mm) aluminum sheet. This assembly must be carefully laid out so everything is in good alignment.

The 20-meter elements are made from lengths of 314-inch (9-mm) type 6061-T6 aluminum tubing. This tubing was flattened to make it something like an elliptical section about 15/16 inch (24 mm) wide and ½ inch (12.5 mm) thick. For the joints between sections, solid aluminum inserts were used. These inserts, 318 x % x 3 inches (9.5 x 12.5 x 77 mm), were fixed and connected by using no. 6 (M3.5) stainless steel machine screws. Spoke attach eyes are 3/32 x 1 inch (25 x 25.5 mm) stainless steel cotter pins in holes in the plane of the loops at proper intervals. Points are bent sharply back around the outside of the elements. The feedpoint insulator for the 20-meter driver is placed at the bottom. The SO-239 fitting mounted underneath was potted in silicone to make it watertight.

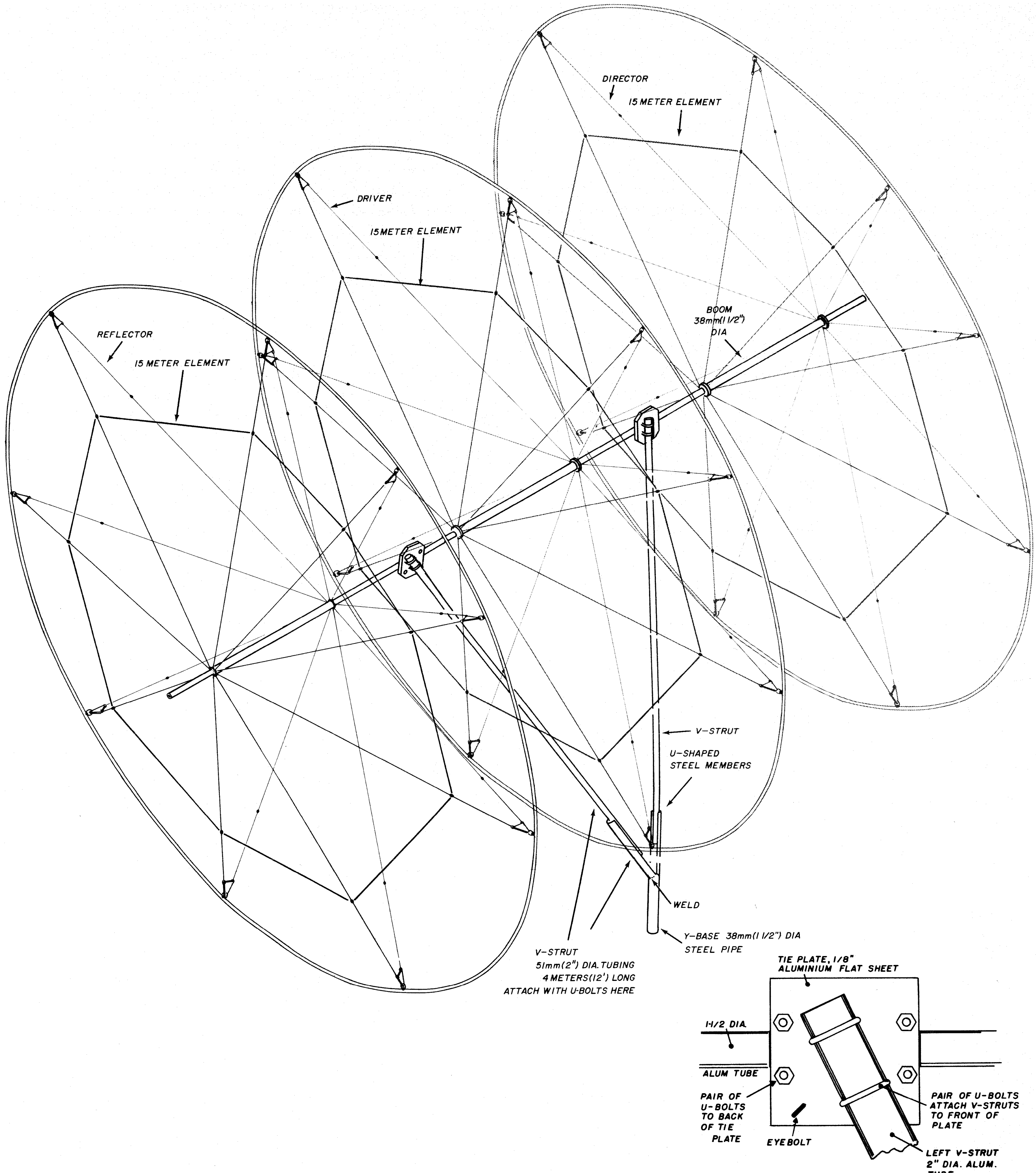


fig. 2. Above, the ***Dream Beam***. Although this antenna is similar to **the** quad, there are some electrical differences because the resonant length of this circular configuration is **somewhat** less than that of a square or diamond: for a circular loop, it's about **974/f, as** opposed to 1005/f for the quad. Below, details of the **boom/V-strut** mounting bracket.

The hubs are 6-foot (1.8-meter) lengths of 1½-inch (38-mm) schedule 40 PVC plastic pipe with spoke-attach flanges cemented on each end. Spoke-attach eyes are 1/8 x 1 inch (3 x 25.5 mm) stainless-steel cotter pins inserted in holes drilled through the flanges. Holes are parallel to the axis of the hub.

Nylon monofilament spokes in earlier antennas were not completely satisfactory. The ***Dream Beam*** elements have spokes of no. 20 (0.8 mm) stainless-steel wire. These are insulated, and nine pairs are spaced around at 40-degree intervals. Their lengths were caicuiated. They were made fairly accurately on a simple jig. This jig was a 12-foot (4-meter) length of 2 x 4 lumber with small finishing nails in it corresponding to the several points on the spokes. The element end of each pair of spokes is fitted with a 4-inch (102-mm) triangular insulator cut from a sheet of ¼-inch (6.5-mm) Lucite. The insulator is attached by a %-inch (19-mm) "key ring" to its spoke-attach eye. ("Key ring" is used for lack of knowing a better term. It consists of almost two turns of stiff, springy stainless wire and may be easily threaded onto and off of the attach eye. I got the key rings at a sailboat supply house, where they're called "cotter rings.")

In toward the hub another small plastic insulator is inserted in each spoke. These are ¼ x ½ x 4 inches (6.5 x 6.5 x 102 mm) long. There is a hole in each end to which the spoke is attached, and a hole in the center. The hole in the center is the eye through which the 15 meter element of no. 16 (1.3 mm) copper antenna wire is threaded. These elements become nine-sided regular polygons. These insulators must be accurately located (a nail in the jig) so that neat polygons of the correct perimeter result. The hub ends of the spokes are simply fixed to their attach eyes on the hubs. The flanges and hubs serve as insulation here.

handling

The elements may be assembled where there's sufficient room handy to the antenna location. Simply put them together. No jig or other special tooling is needed. They end up, completed, lying on the ground.

Handling and transport of the assembled elements at first appeared to present a trickv and complicated challenge. However, it turned out to be ridiculously simple, easy, and safe. A carrier was made like a grossly elongated T. The "up-and-down" portion is a 12-foot (4-meter) length of I %-inch (32-mm) light steel tubing. At the top, a 4-foot (I-meter) piece of tubing is attached across. Two clips were made from a piece of PVC pipe (somewhat larger then the hubs). The clips were cut so they could be snapped on and off the hubs with ease; they were bolted to the top of the T.

Simply clip the carrier onto the hub, pick up the element, and lift it to the vertical position. Superhuman strength and balance are not needed; after all, an element weighs only about 9 pounds (4 kg). I must admit to being a little frightened when I tried it for the first time. The 22-foot (7-meter) "wheel" looks gigantic when towering over your head! The purpose of the carrier, of course, is for transporting the element and slipping it onto the boom. The hub ID is about 3/32 inch (2.5 mm) greater than the 1½-inch (38-mm) boom, so it may be slipped on and off readily. When the hub is on the boom, give the carrier a downward jerk. The clips open and the carrier is separated. Elements went up and on, and off and down, many times during development for changes, adjustments, and pruning. Not the slightest difficulty was ever encountered.

assembly

Assembly of the entire array is really not so complex as it may at first seem. It was quite fun — a great satisfaction to see it up there, in place and "flying." I'm not exactly a spring chicken and don't claim the vision, balance, and agility of 30 or more years ago. The whole antenna was, however, assembled from the parts on the ground to their places in the rooftop array in about four hours. This work was done entirely alone with no assistance whatsoever.

Fig. 2 shows the various parts in relative positions. The key to the assembly of the structure is 20-foot (6-meter) length of 3132-inch (2.5-mm) stainless-steel ***flexible*** cable. This is obtainable at most marine hardware dealers, particularly those handling sailboats and supplies. It is permanently attached to the ***inside*** of the left boom tube at the right end. Steps are as follows:

1. Position the steel Y-frame in tne bearing and rotator.

2. Run a 25-foot (8-meter) messenger of stout cord or fish line through the eyebolt in the left tie plate (used later for the reflector preventer cord). Erect the left V-strut and boom half, securing it to the Y-frame with U-bolts.

3. Run a 25-foot (8-meter) messenger through the hub of the driven element. Bring the driven element into position, ready to slip onto the left boom half. Connect the flexible cable to this messenger and pull it through the hub of the driven element so it comes out on the right. Slip the driven element onto the left boom half.

4. Drill a 318-inch (9.5-mm) hole in the bottom of the right boom half near where it is attached to the V-strut. Run a 25-foot (8-meter) messenger (for the flexible cable) through this hole so that it comes out from this boom half on the left.

5. Run a 25-foot (8-meter) messenger through the eyebolt in the right tie plate (used later for the director preventer cord). Connect the flexibie cabie to its messenger. Pull the flexible cable into the right boom half so that it comes out through the 318-inch (9.5-mm) hole, Slip the right boom half into the driven-element hub and secure this V-strut to the Y-frame with U-bolts.

6. Pull the flexible cable tight and fasten it securely to the Y-frame. The structure is now erected. The hub of the driven element acts as a sleeve connecting the boom halves. The flexible cable holds them tightly together inside the hub.

7. Slip on the reflector element. Connect the preventer cord (118-inch or 3-mm high-grade nylon parachute cord) to its messenger and pull it through the eyebolt in the tie plate, securing it to the Y-frame. This preventer cord simply restrains the element from sliding off the boom.

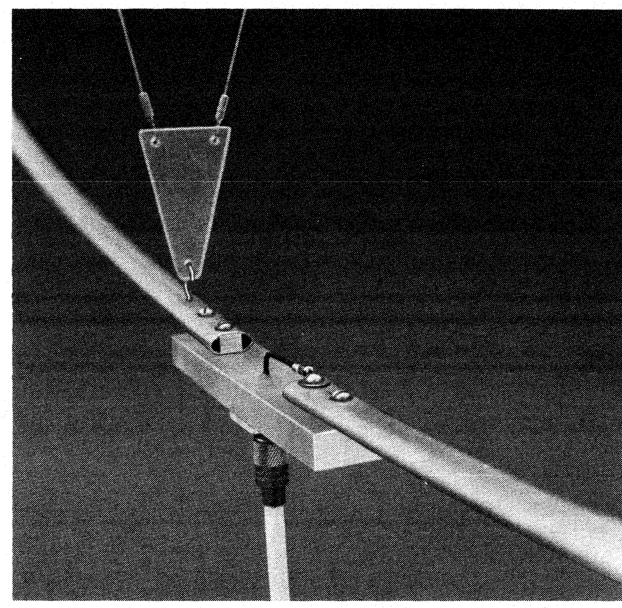
8. Slip the director element on in the same manner as the reflector.

disadvantages

The only disadvantage of this array is in the sheer size of the element assemblies. Once it's put together in your yard or patio, the only possible thing you can do with them is put them up where they belong. They are too large to ship by any means. They won't fit in your garage or basement, so they can't be stored away – unless you happen to have a vacant airplane hangar! They could be disassembled for shipping or storing, but not nearly so readily as with other beams.

serviceability

Only high-grade corrosion resistant materials were used. The **PVC** hubs were painted with polyurethane to avoid deterioration by sunlight. The antenna has satisfactorily survived two rather severe winter icings. Being somewhat resilient, the elements swing and sway in a gusty wind, but the array has withstood quite a number of very high winds in thunderstorms. As the array is mounted just above the rooftop, it's centered only about 30 feet (9 meters) above thk ground. Some protection occurs from numerous tall trees in the area. I can't sav how the arrav would do atop an 80-foot (24-meter) tower. The exposed area of each loop is 3 square feet (0.3 square meters). Further experience may show up weaknesses not anticipated.



How the feedpoint insulator is made and connected to the 20-meter driven element. The triangular-shaped insulator is between the circular conductor and the wire spokes. Portions of the Y-frame are visible, and the lower end of the V-strut can be seen.

performance

Evaluation of antenna performance is both difficult and perilous. The difficulty lies in the large number of uncontrollable variables, which render numerical comparisons very questionable. The peril is in one's ability to enforce strict self-discipline and maintain a truly objective viewpoint. It's easy and tempting to overrate something which is your own baby, your own creation.

In the past year every opportunity for evaluation and comparison has been seized; this process is still going on. Reports and results have been extremely encouraging, and I become more pleased and confident as the hard evidence comes in day by day and week by week. Much of the time I get reports such as: "You are very, very strong;" You have the strongest signal on the band;" and, "Your signal is 15 to 20 dB over S9." Some of the reports have been so good as to be not believable. I can't remember when another station couldn't read me if I could read him. Being picked out the first time in DX pileups has become fairly common. Though many long months in coming, the *Dream Beam* is now a reality.

reference

1. Frank Witt, W1DTV, "Simplified Antenna Gain Calculations," *ham radio*, May, 1978, pages78-85.
2. Charles J. Ellis, WØYBV, "A Novel Way to Mount a Rotary-Beam Antenna," QST, May, 1979, pages32-33.

ham radio

Yagi antenna design:

performance of multi-element simplistic beams

A Yagi antenna can be characterized by one or more driver elements and a number of parasite elements, all supported on a boom.

For each element we must specify X and Y coordinates, a length (LE), a radius (RO), all measured in terms of (central) wavelength, and in the case of each driver, the excitation potential (V) or current (I) and its phase referred to some time standard.?

It is instantly apparent that with all of these variables an exhaustive investigation into all possible configurations is impractical! Instead I shall begin with an initial consideration of "simple" or simplistic Yagi antennas and will subsequently discuss a variety of departures from this simplistic design. I will define this (simple) class of Yagi antennas as those involving a single driven element with one or more parasitic elements. No more than one "reflector" will exist, and all "directors," if any, will be uniform, i.e., they will have identical lengths and diameters. Moreover, all elements will be uniformly spaced along the boom. Additionally, the antenna will be in ***free space;*** we shall initially investigate only free-space performance properties.

These restrictions may seem at first sight to be quite severe, but I hasten to remark that free-space performance will relate to actual performance over ground or earth (to be discussed in a later article) and the simplistic Yagi antenna, as defined above, can, in many instances, provide performance levels fully as good as those from more sophisticated designs. Furthermore, we can learn a great deal about Yagi antenna performance from studying these simple designs and, as we shall see, will develop useful conceptual ideas about Yagi behavior and ideas for "best" design.

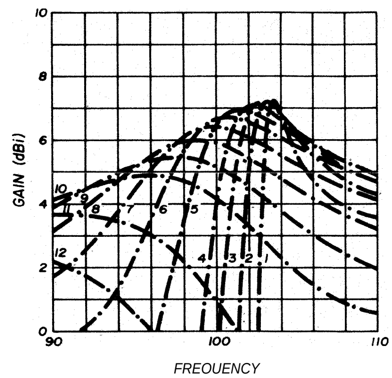
Throughout this investigation of simplistic antennas I will choose element dimensions (radii) characteristic of "normal" 14 MHz construction ($RO = 0.000526\ \lambda$). The results can be translated to any other element dimension by proper scaling calculations; scaling rules will be given later.

two-element beams

I shall begin with a 2-element Yagi beam involving one parasite which can act either as a "reflector" (for frequencies above its resonance) or as a "director." For such a beam there are only two fundamental variables: The physical separation of the two elements along the boom and the physical length of the parasite! The exact length of the driven element is of little consequence as far as gain and pattern are concerned; it does, however, affect driving-point impedance (especially reactance) which is considered later. Since we shall be interested first in a frequency swept plot of the gain and F/B properties, the physical length of the parasite can be fixed; as we increase the frequency from well below to well above the parasite free-space resonant frequency we can observe the properties of the beam first where the parasite behaves as a director and secondly as a parasitic reflector.

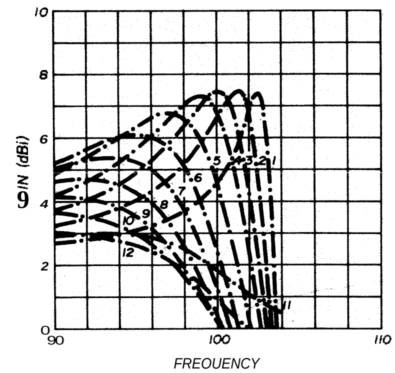
The computation methodology I shall use is that given explicitly in an earlier article.[1] In all cases of the 2-element beam I have used a cylindrical length, $LE = 0.48167\ \lambda$ and a radius, $RO = 0.0005260\ \lambda$, for both parasite and driver; this makes each element's isolated free-space resonant frequency, FR, equal to unity on the normalized frequency, F, scale, where $F = f/f_o = 1$. The frequency itself is varied in steps

By James L. Lawson, W2PV, 2532 Troy Road, Schnectady, New York 12309



Curve	S	Curve	S	Curve	S
1	0.025	5	0.150	9	0.350
2	0.050	6	0.200	10	0.400
3	0.075	7	0.250	11	0.500
4	0.100	8	0.300	12	0.600

fig. 1. Frequency swept gain plots of 2-element Yagi antennas with different element spacings, S, where the parasitic element acts as a *reflector.*



Curve	S	Curve	S	Curve	S
1	0.025	5	0.150	9	0.350
2	0.050	6	0.200	10	0.400
3	0.075	7	0.250	11	0.500
4	0.100	8	0.300	12	0.600

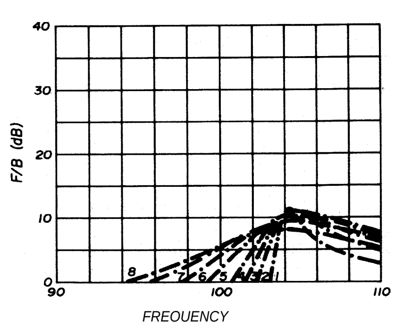
fig. 3. Frequency swept gain plots of 2-element Yagi antennas with different element spacings, S, where the parasitic element acts as a *director.*

from about 90 per cent ($F = 0.9$) to 110 per cent ($F = 1.10$) of the central frequency ($F = 1.0$); these steps were made sufficiently small to fully show the behavior of the beam. Element separation, S, measured in wavelengths at the central frequency, f_o, is varied from 0.025 to 0.5, again in steps sufficiently small to bring out essential behavior.

Fig. 1 shows the frequency swept gain of the 2-element Yagi antenna for several element spacings, *S,* where the gain is positive in the direction of parasite towards driver, i.e., where the parasite acts like a reflector! Each curve represents a particular spacing, S, appropriately keyed in the legend. **Fig.** 2 is a similar frequency swept plot of the free space Front-to-Back ratio, F/B, for the same series of element separations, *S.* **Figs.** 3 and **4,** in the same format, show the gain and F/B where the parasite acts like a director, i.e., where the gain is positive in the direction of driver towards parasite!

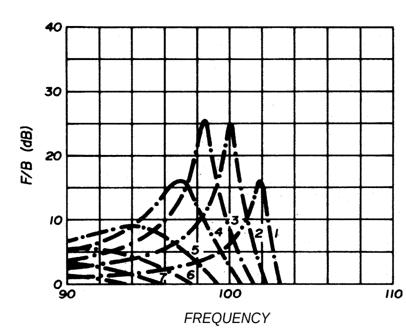
Examination of these performance plots, together with additional information on computed driver input impedance, reveals a number of interesting facets of the behavior of 2-element Yagi beams. First I show in **fig. 5** a plot of the gain at central frequency ($F = 1$) only as a function of element spacing, *S.* This is similar but not identical with a plot shown on page 147 of the *ARRL Antenna Book*;[2] it is possible that the differences, particularly at small spacings, are due to greater precision in the new calculations. In any case, I believe the implications of this plot may be somewhat misleading!

You can easily see that the *maximum* gain(s) obtainable at the "best" frequency(s) in **figs. 1** to **4** look somewhat different! These are shown in **figs. 6**



Curve	S	Curve	S
1	0.025	4	0.100
2	0.050	5	0.150
3	0.075	6	0.200

fig. 2. Free-space front-to-back ratio for the 2-element Yagis vs different element spacings, S (parasitic element acting as a *reflector*).



Curve	S	Curve	S
1	0.025	4	0.100
2	0.050	5	0.150
3	0.075	6	0.200

fig. 4. Free-space front-to-back ratio for 2-element Yagis vs different element spacings, S, (parasitic element acting as a *director*).

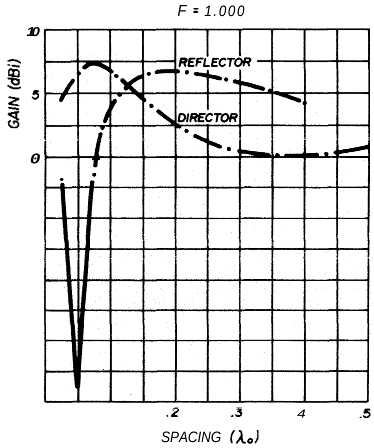


fig. 5. Plot of gain of a 2-element Yagi at central frequency ($F = 1$) as a function of element spacing S.

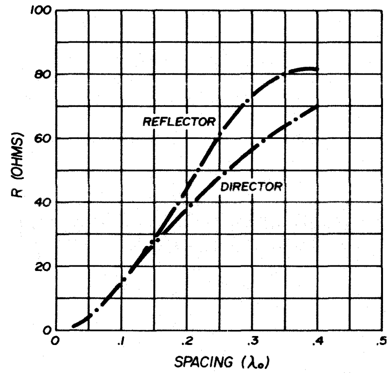


fig. 8. Driving point resistance of a 2-element Yagi vs element spacing.

and 7, where for reference the curve for $F = 1.0$ is also shown. You can see that the obtainable gain does *not* depend greatly on whether the parasite is a reflector or a director as implied in fig. **5;** moreover, the largest gain is obtained at very *small* spacings! This is a result which is not intuitive. Figs. ***6*** and 7 can be compared with the early analysis by Brown,[3] and there appears to be good agreement. However, the result shown on page 146 of the *ARRL Antenna Book* which cites Brown's analysis is somewhat different. The fall off in maximum gain for low spacings shown by the ARRL reference does not appear to agree with Brown nor does it substantiate the calculations I have made (figs. ***6*** and 7).

If you examine the maximum gain(s) shown in figs. **1** and 3 for best frequency and corresponding (driver) driving-point impedance, you obtain the values shown in **table 1.** A plot of R is shown in fig. **8** which can be compared with a similar diagram shown on page 147 of the *ARRL Antenna Book;*[2] except for low values of ***S,*** agreement is fairly satisfactory.

You can see from **table 1** and fig. **8** that element spacing affects (driver) driving-point resistance, and, therefore, circuit loaded Q (Q_L) over a *very* large range! This factor, as well as the gain curves shown in figs. **1** and 3 set a practical limit to the achievable gain over a desirable bandwidth, e.g., perhaps 4 per cent in F. Moreover, the higher values of (radiation) loaded Q will, in practice, cause circuit resistive losses to be large and therefore the antenna efficiency to be low! Thus, in practice, really short booms are *not* very desirable; one must choose between efficiency and bandwidth on the one hand, and gain and F/B ratio on the other!

Long booms, however, also appear undesirable because gain really falls off (primarily due to reduced excitation of the parasite). Furthermore, for booms longer than 0.3λ a new phenomenon can be seen from a detailed computational analysis (not shown here). The front lobe of radiation begins to "dimple" in the forward direction, resulting in a pattern where the gain maximum occurs at an elevation angle other than zero with respect to the boom direction. (The gain shown in figs. **1** to **4,** however, is just the energy flux in the direction of the boom referenced to

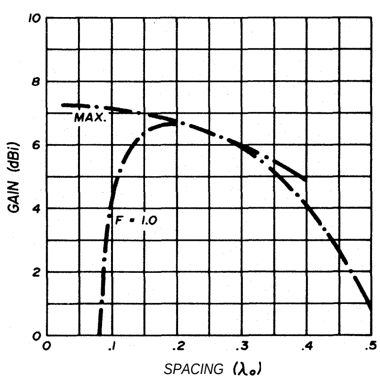


fig. 6. Gain of a 2-element Yagi vs element spacing; parasite a reflector.

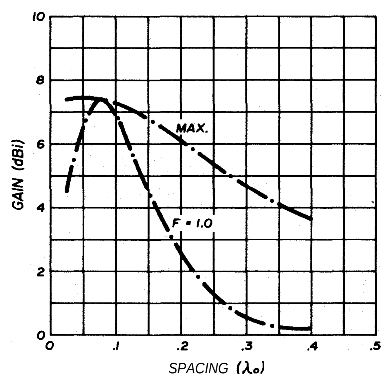


fig. 7. Gain of a 2-element Yagi vs element spacing; parasite a director.

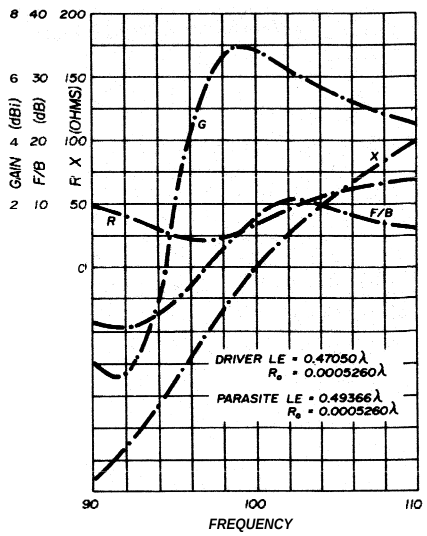


fig. 9. Gain, front-to-back ratio (F/B), and feedpoint impedance (R and X) as a function of frequency for a 2-element Yagi (parasite as a reflector). S = 0.15λ at central frequency.

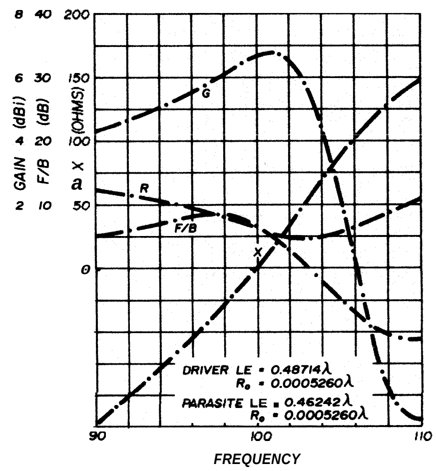


fig. 10. Gain, front-to-back ratio (F/B), and feedpoint impedance (R and X) as a function of frequency for a 2-element Yagi (parasite as a director). S = 0.15λ at central frequency.

an isotropic radiator.) This pattern effect was predicted by Brown3 and shown in Kraus, page 294.4

Note that the 2-element Yagi gives respectable performance in gain for a wide range of element separations! However, the F/B figures are not especially impressive; moreover best F/B does not occur at the same frequency as best gain! Thus, in designing a 2-element Yagi beam a practical compromise is necessary. If you wish to obtain good gain with at least a fair F/B ratio over a bandwidth of say **4** per cent, you can determine by inspection of figs. **1** through **4** and table **1** that a 2-element beam should have a boom length of perhaps 0.15 wavelength. For such a boom the gain is essentially independent of whether the parasite is a reflector or a director; moreover, the F/B is about equivalent for either situation.

To move the peak of the gain curve(s) in figs. **1** to **4** to center frequency, the parasite length is adjusted commensurately; to reduce central frequency (driver) reactance, the driver length is adjusted.

These characteristics of 2-element beams are shown in figs. **9** and **10** with a frequency-swept plot of each design. Note that each of these figures show gain, F/B, R and X of the driver. They illustrate the kind of design compromises which must be made. They also show the frequency-swept behavior of the main performance parameters.

The "best" central frequency is a matter of choice and is a compromise between gain and F/B ratio; it is adjusted by the length of the parasite. The (driver) driving point resistance and reactance vary significantly with frequency! Note that you cannot generally specify "a" resistance except at a single frequency such as the central design frequency; also note that

table 1. Maximum gain and feed-point impedance of a 2-element Yagi at various element spacings.

	reflector					director				
S	dBi max. gain	at freq.	R ohms	X ohms	Q_L	dBi max. gain	at freq.	R ohms	X ohms	Q_L
0.025	7.244	1.036	1.115	11.163	924.90	7.410	1.026	1.091	– 10.029	1003.00
0.050	7.218	1.032	4.216	20.371	237.30	7.461	1.014	4.108	– 17.556	265.50
0.075	7.158	1.030	9.679	31.834	99.13	7.417	1.000	9.801	– 29.917	104.74
0.100	7.122	1.025	15.551	35.778	59.66	7.300	0.990	15.253	– 34.122	63.03
0.150	6.964	1.015	29.577	38.574	28.90	6.800	0.970	27.452	–44.157	30.56
0.200	6.722	1.005	44.625	34.700	17.38	6.115	0.955	37.979	– 49.407	20.64
0.250	6.406	1.000	61.400	31.612	11.01	5.383	0.940	48.183	– 59.362	15.28
0.300	5.999	0.990	72.656	16.936	8.37	4.722	0.920	57.242	– 80.457	12.27
0.350	5.491	0.980	80.527	– 0.378	6.89	4.159	0.900	64.194	– 104.165	10.62
0.400	4.913	0.960	81.957	– 29.152	6.86	3.688	0.900	70.739	– 103.310	9.24

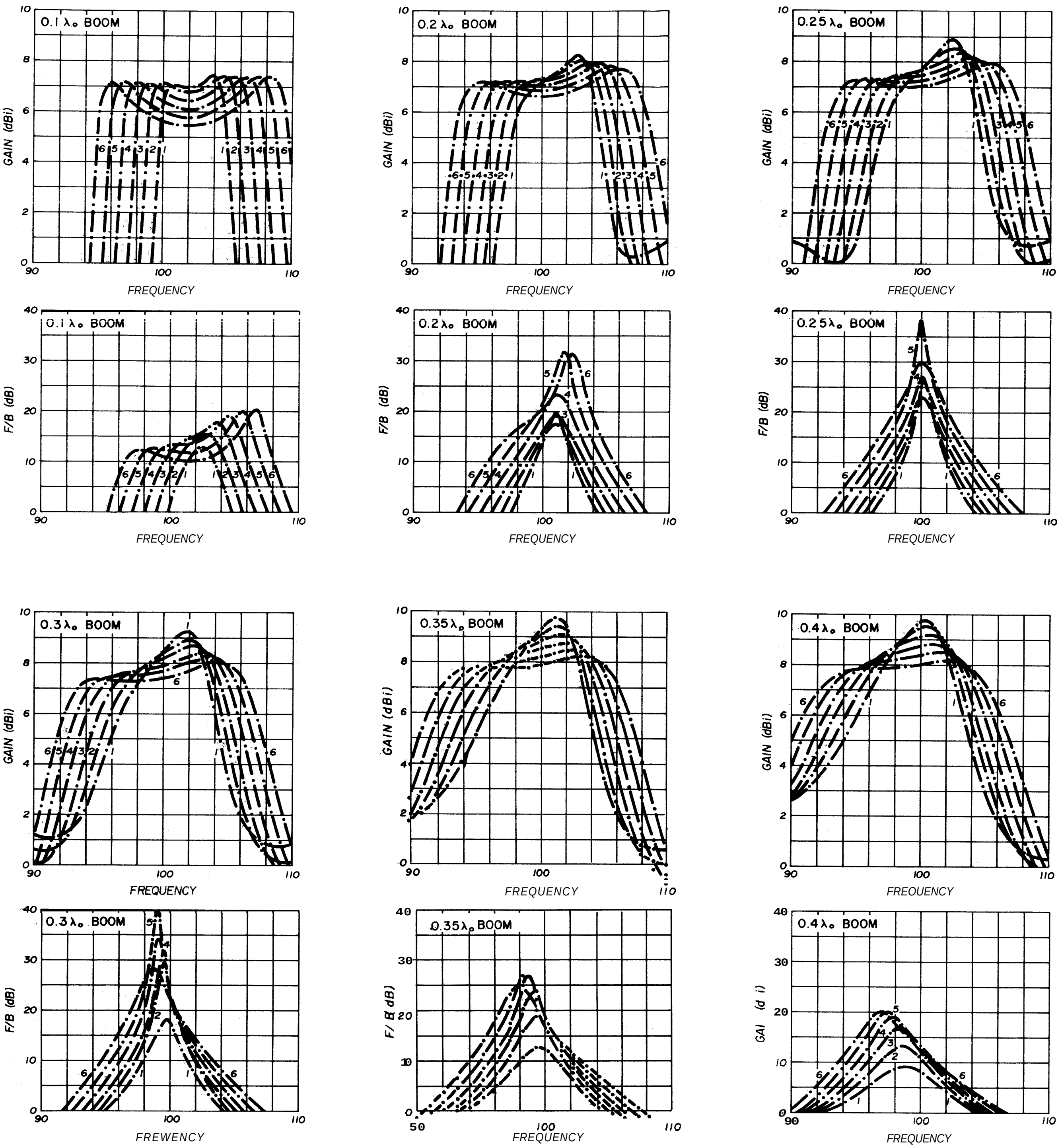
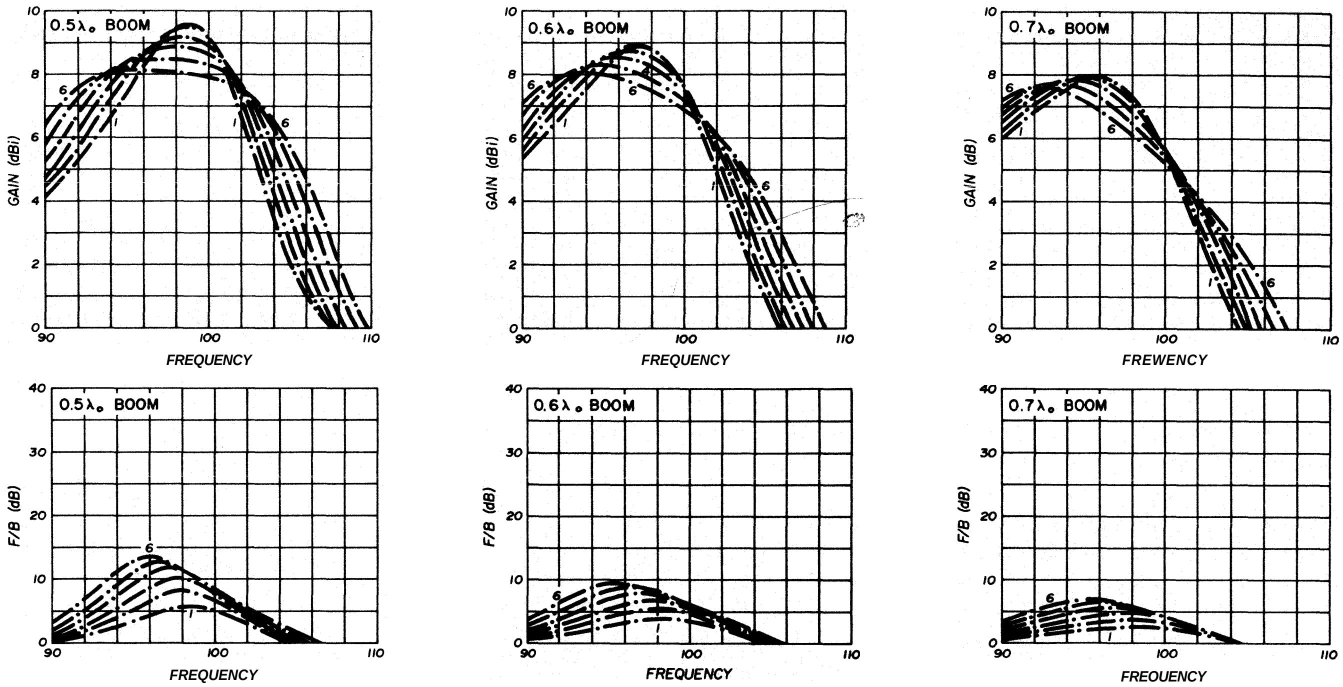


fig. 11. Gain and front-to-back (*F/B*) ratio for three-element Yagi beams with varying boom lengths, and changing reflector and director lengths (see table 3 for complete data). Boom lengths from $0.1\lambda_o$ to $0.4\lambda_o$. See next page for boom lengths from $0.5\lambda_o$ to $0.7\lambda_o$.

feedpoint reactance is not a linear function of frequency. This is caused by the combined effect of self and mutual impedance of the elements.

You can adjust the frequency of the zero reactance point by the exact length of the driver; this has been done only approximately in **figs. 9** and **10.** However, note that the adjusted driver lengths are quite *different* for the two cases, and each driver length is also different than the length for a single isolated resonant dipole in free space! These differences are again caused by mutual reactance coupled into the driver by the parasite.



Gain and front-to-back (F/B) ratio for three-element Yagi beams; boom lengths from $0.5\lambda_o$ to $0.7\lambda_o$.

table 2. Yagi antenna parameters and range over which one has been varied for this study.

parameter	range of exploration/display
Number of elements, N	N = 3, 4, 5, 6, 7
Boom length, ℓ_B	Generally up to 1.5λ
Reflector	f_R = 0.98t00.93
Director	f_R = 1.02 to 1.07

more than two elements

I will now turn to an analysis of simplistic Yagi antennas having more than one parasite. In all cases there will be only one reflector and all directors will have identical lengths. All elements are uniformly spaced along the boom. I shall display for clarity only the essential frequency-swept gain and frequency-swept F/B behaviors; the driving point impedances, all of which were computed, are of secondary interest at this point. For these Yagi antennas there are a number of parameters which should be systematically explored. **Table 2** shows these parameters and the range over which each has been varied. To display results in a consistent way I have chosen for each frequency-swept plot a fixed number of elements and a fixed overall boom length, ℓ_B, measured in wavelengths at the central frequency. On each plot there are six numbered curves; each number designates a particular parasite "tuning" combination; these combinations are shown in **table 3**. The lengths and free space resonant frequencies of parasites are shown for each numbered combination.

The curves of **fig. 11** show the results for 3-element beams as boom lengths, ℓ_B, are varied from 0.100 to 0.700 wavelengths. It is apparent from an inspection of these plots that the performance is superior to that of the 2-element beams; this is especially true in the F/B ratio. As you increase boom length the maximum gain increases (unlike that for 2-element beams); the F/B increases spectacularly, then decreases again. For this class of Yagi antennas there seems to be a ***best*** boom length; we shall see this kind of result for all of the simplistic Yagis and the physicai explanation will soon be apparent!

Note that the chief parameter controlling the bandwidth over which gain remains high is the (resonant)

table 3. List of parasitic lengths and resonance in terms of the center frequency for the six numbered curves on each of the graphs of antenna gain and F/B ratio (figs. 11 and 12).

curve	reflector length	reflector resonance	director(s) length	director(s) resonance
1	0.49150λ	0.98	0.47223λ	1.02
2	0.49657λ	0.97	0.46764λ	1.03
3	0.50174λ	0.96	0.46314λ	1.04
4	0.50702λ	0.95	0.45873λ	1.05
5	0.51241λ	0.94	0.45441λ	1.06
6	0.51792λ	0.93	0.45016λ	1.07

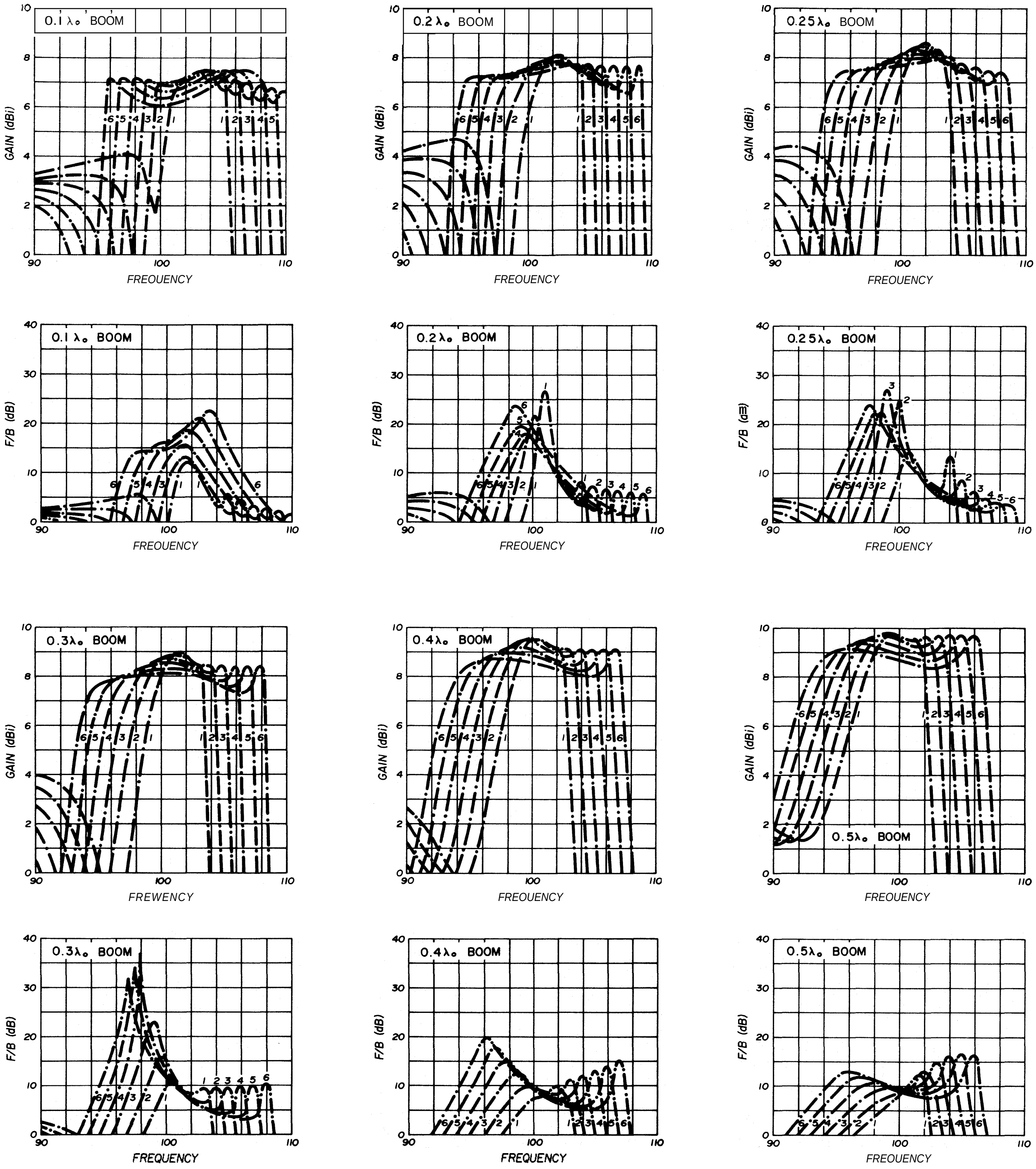
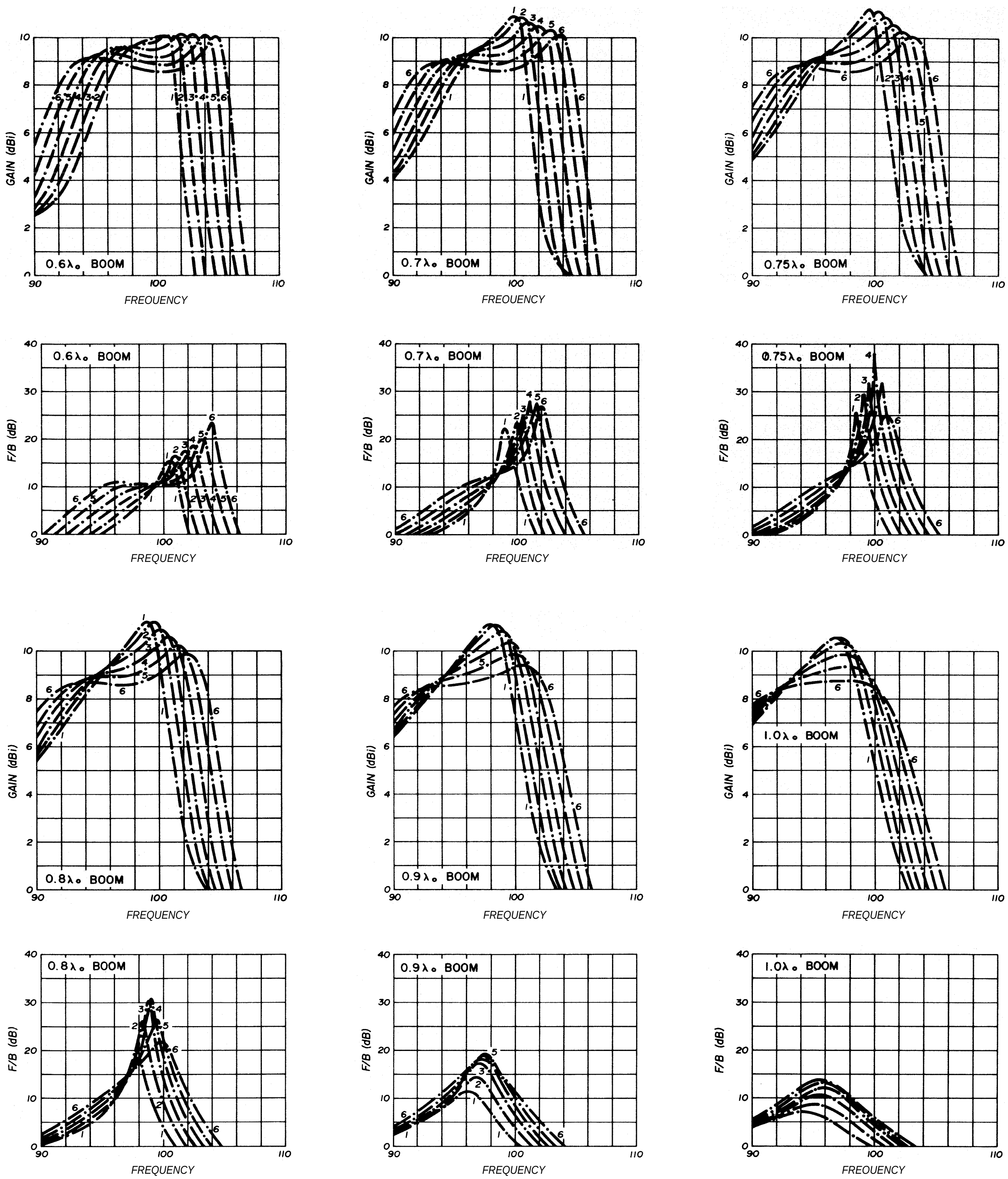


fig. 12. Gain and front-to-back (F/B) ratio for four-element Yagi beams with varying boom lengths, and changing reflector and director lengths (see table 3 for complete data). Boom lengths from $0.1\lambda_o$ to $0.5\lambda_o$. See next page for boom lengths from $0.6\lambda_o$ to $1.0\lambda_o$.

frequency separation of reflector and director; this observation will also prove to be generally true for all simplistic Yagi antennas! The bandwidth of the F/B performance (when the F/B is very high) is small; this is due to the critical nature of low back radiation.

Back radiation is very low only when there is vectorial cancellation of field in the back direction; this comes about only where element complex currents



Gain and front-to-back (F/B) ratio for four-element Yagi beams; boom lengths from $0.6\lambda_o$ to $1.0\lambda_o$.

are accidentally favorable for such cancellation. When this happens very small changes in those currents, e.g., by shifting frequency slightly, will destroy the favorable vectorial cancellation. This general result is inherent in all Yagi antennas; if the F/B is exceptionally high it will be so only over a very narrow frequency band!

Similar results for 4-element simplistic beams are shown in **fig. 12;** results for 5-, 6-, and 7-elements, although not plotted here, show increasing complexity with number of elements of the frequency-swept plots; this is caused primarily by the larger number of resonances in the system. The gain "cutoff" at high frequencies is also increasingly abrupt due no doubt to the large number of directors (which shift over to "reflectors" at high frequencies). F/B curves become very complex. Moreover, really high values of F/B (greater than 30 dB) are quite rare; it is very difficult to find combinations where vectorial cancellation in the rear direction is nearly complete!

summary

Simplistic Yagis with two, three, and four elements with boom lengths to 1.0λ have been systematically explored, and it has been shown that the gain function is generally not flat and the F/B ratio varies greatly from one example to another. Next month I will continue this discussion with a series of graphs which show gain and front-to-back ratio for 6-element Yagis with boom lengths up 1.5λ. I will also compare the performance characteristics of Yagis with up to seven elements and present interesting new data on the subject of front-to-back ratio.

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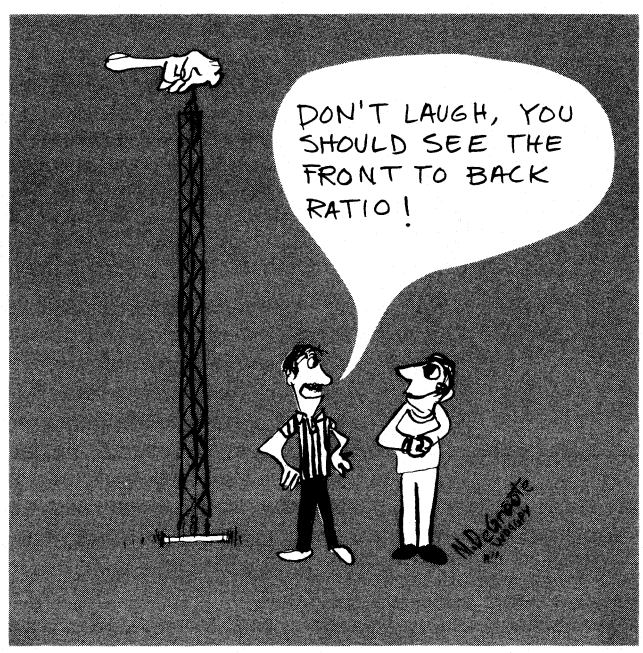
ham radio



notes on ground systems

How to provide an effective ground system for your station and how to measure ground-system resistance

The old timers took their ground systems very seriously. Quoting from ***Practical Wireless Telegraphy*** by Elmer E. Bucher, (1921):

> The earth plate is sometimes very elaborate and may consist of a great number of copper or zinc plates buried in moist ground to a depth of several feet.

Most ham stations don't have a proper ground, or, in many cases, none at all. The ground system must have very ***low*** resistance for good performance of receiver and transmitter and for reducing radio-frequency interference.

A good ground system makes as good a contact as possible with the earth. A large surface area, as well as depth to moist earth, is essential. Several ground rods of large-diameter copper (at least 518 inch or 16 mm) will meet these requirements if driven in 6 to 8 feet (1.8-2.5 meters).

In this article I discuss the delta-Y ground-rod configuration, which is efficient and far superior to a single ground rod, utility ground system, or water pipe. Also discussed is a straightforward method of measuring net resistance of the ground system.

system considerations

If desired, three rods may be used in a delta arrangement without too much increase in the net ground resistance. The delta or Y configuration isn't mandatory but does provide a convenient way of measuring the net ground resistance and thereby a means of determining the quality of the ground system.

An ac voltage is employed to measure the resistance to avoid dc electrochemical effects such as battery action and polarization. If available, an ac megohmmeter could be used to measure the resistance between the respective ground rods.

The ground rods should be copper or copper-clad steel and 6-8 feet (1.8-2.5 meters) long. Ground rods using 318-inch (9.5-mm) copper clad steel are generally available. Five-eighths inch (16-mm) copper tubing would be better but in some soils it may be very

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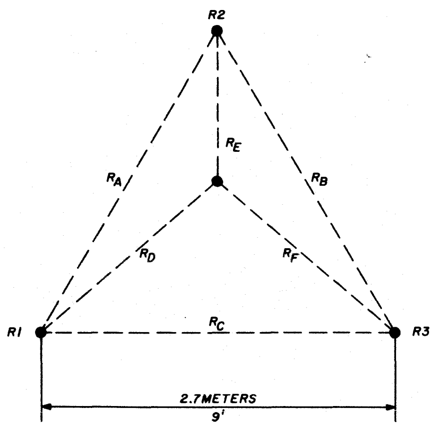

fig. 1. Delta configuration of ground rods for measuring the net resistance of the system to ground. Ground rods are designated R1, R2, and R3; R_A, R_B, R_C, R_D, R_E, and R_F are the ground resistances between the designated rods.

difficult to drive. Aluminum should be avoided, and galvanized steel is not very satisfactory.

Do not use the electric utility ground bus, as the common impedance will introduce the noise and interference on that ground wire. Water pipes are not very satisfactory for a variety of reasons.' Many times, if copper plumbing is used in the house, the copper pipe extends only about 10 feet (3 meters), and from there on plastic pipe is used. With galvanized pipe, the resistance of the coupling joints as well as the surface resistance may be quite high. The fact that the pipe is full of water does not contribute to its effectiveness as a ground.

The antenna tower or mast should not be connected to the station ground because of the hazard of lightning; furthermore, never ground the tower down through the concrete foundation; a lightning strike would probably shatter the concrete. All ground leads from the equipment to the ground bus should be as short as possible. One-half inch (12.5-mm) or wider copper braid, tinned at each end and drilled for strapping to the ground terminal, makes a satisfactory and flexible connection.

The connecting wires should be as large as possible: no. 10 or no. 8 (2.6 or 3.3 mm) copper to keep resistance and reactance as low as possible. Two wires in parallel are equivalent to three sizes larger; hence,

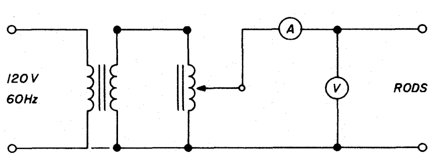

fig. 2. Test setup for measuring ground resistance. When the voltage is adjusted to provide 1 ampere of current, the value of the voltage is equal to the resistance in ohms.

two no. 10 (2.6 mm) wires would be equivalent to a single no. 7 (3.7 mm) wire. The dc resistance of no. 7 (3.7 mm) copper wire is 0.5 ohm per 1000 feet (305 meters), so for connecting lengths less than 10 feet (3 meters) the dc resistance would be about 0.005 ohm. Because of skin effect, the rf resistance will be higher; at 14 MHz it would be about 0.35 ohm.

Avoid connecting lengths that are a half wavelength or multiples thereof at the operating frequency, because they would act like a half-wavelength antenna. Also avoid quarter wavelengths, as they present a very high impedance at the resonant frequency.

Do not let a bare ground wire to the equipment touch any metal, as the intermittent contact will introduce serious noise.

resistance measurements

The following procedure may be used to determine the net resistance of the delta-Y configuration to

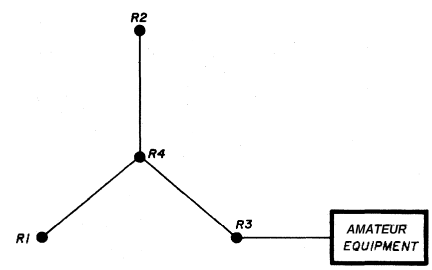

fig. 3. Proper method for connecting the ground rods together for the station ground.

ground (see **fig. 1**). In the following example, the delta sides are 9 feet (2.7 meters).

Using an isolation transformer, Variac, ac voltmeter, and ac ammeter as shown in **fig. 2,** determine the resistances between each pair of ground rods. Use the Variac to adjust the voltage until 1 ampere is obtained. Then the value of the voltage will be equal to the resistance in ohms. Determine R_A, R_B, R_C, R_D, R_E, and R_F in this manner.

Calculate R_1:

$$R_1 = \frac{1}{2}(R_A + R_C - R_B) \qquad \textbf{(1)}$$

which is one half of the sum of the two adjacent legs of the delta minus the opposite leg.

Likewise calculate:

$$R_2 = \frac{1}{2}(R_A + R_B - R_C) \qquad \textbf{(2)}$$

$$R_3 = \frac{1}{2}(R_B + R_C - R_A) \qquad \textbf{(3)}$$

$$R_4 = \frac{1}{2}(R_D + R_E - R_A) \qquad \textbf{(4)}$$

R_1, R_2, R_3, and R_4 should be essentially the same, and if so, the net resistance of the configuration to ground will be

$$R = \frac{R_1}{4} \tag{5}$$

if the values of R_1, R_2, R_3 and R_4 are significantly different, calculate the net resistance of the four in parallel. The value of the individual rod-resistance-to ground is about 10 to 20 ohms and depends on the type of soil, moisture content, depth, and rod size.

$$R_X = \frac{R_1 R_2}{R_1 + R_2} \quad R_Y = \frac{R_3 R_4}{R_3 + R_4} \quad R = \frac{R_X R_Y}{R_X + R_Y} \tag{6}$$

The net resistance of the connected rods should be less than 5 ohms. The rods are connected as shown in **fig.** 3.

calculating wire inductance

The self-inductance of a single wire may be calculated by the following formula (National Bureau of Standards Circular No. 74):

$$L = 0.002\,\ell\,(2.303 \log \frac{4\ell}{d} - 1)\,\mu H \tag{7}$$

Example:

$\ell = 9\ feet\ (274\ cm)$

$d = no.\ 7\ AWG = 0.14\ inch\ (0.36\ cm)^2$

$L = 0002 \times 274\,(2.303 \log \frac{4 \times 274}{0.36} - 1)$

$= 4\ \mu H$

At 4 MHz, $X_L = 2\pi \cdot 4 \cdot 4 = 100\ ohms$

Contrast this with the reactance of a 25-foot (8-meter) length of no. 16 (1.3-mm) ground wire, which would be over 200 ohms. The length and size of the wire are the determining elements of the resulting inductance, so it's readily seen that the length must be as short as possible and the wire size as large as possible.

The procedure used in determining the ground resistance is covered on page 257 of the Government Printing Office publication DCAC 330-175-1 addendum 1.

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ham radio

dual quad array

for two meters

Design and construction of a quad array that challenges a Yagi-Uda with the same number of elements

Several months of work on two meters with quad antennas of various designs and configurations have resulted in an improved design using all-metal construction. The driven elements are of the closed-loop type,[1] employing an improved feed method that provides ease of adjustment and an excellent match to the feed system.

This construction method offers several advantages, both mechanical and electrical, over the usual insulated-spreader type of quad layout. The all-metal structure will withstand severe environmental conditions such as high wind or ice loading. It also presents a grounded system for electrical charges that may be induced by a severe electrical storm.

description

The dual quad array consists of two 4-element quads mounted on a common cross boom spaced approximately 5/8 wavelength apart. The quads are connected **by** 314-wavelength phasing sections of RG-59/U foam-filled coax. {More on this later.)

Basically, a quad antenna is a one-wavelength conductor that may take the form of a square, diamond, or round loop. Regardless of configuration, the quad antenna's electrical characteristics remain essentially the same. In this article I refer to it as a quad loop.

The two halfwave dipoles diagrammed in **fig. 1A** show the formation of a quad loop. They have two low-impedance points, so they may be mounted to a metal support without affecting their electrical characteristics. The voltage curve along a halfwave dipole shows maximum voltage at the ends, with the center at zero potential (low impedance), this being points **A** and B of the dipoles.

If the dipoles are spaced 1/4 wavelength apart and their ends folded over at the 1/8-wavelength points and joined together, a cubical quad is formed **(fig.**

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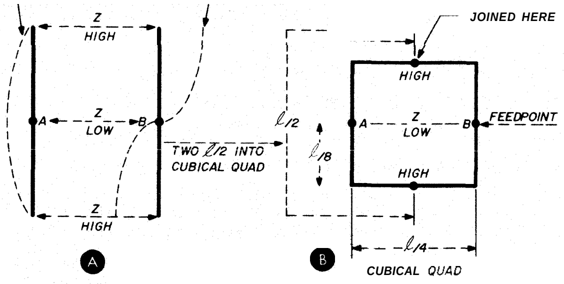

fig. 1. Principles of a quad antenna. Two halfwave dipoles form a quad loop, (A). A cubical quad is formed by folding the ends at the 1/8-*wavelength* points, (B). Mounting points A and B are at zero rf potential. Polarization depends upon feedpoint: vertical when fed from side, horizontal when fed at top or bottom.

1B). Points **A** and B remain at zero rf potential so long as the feedpoint is at either **A** or B.

feedpoint considerations

The feedpoint to a quad loop determines the voltage and current distribution around the loop as well as the polarization of the emittted wave front.

The quad loop looks and performs like two half-wave dipoles connected back-to-back. If the quad loop is fed at either side, maximum current flow occurs in the vertical sides; hence it's vertically polarized. If fed at either the top or bottom, the voltage and current nodes are shifted around the loop by 90 degrees and the quad becomes horizontally polarized.

loop configuration

Circular loops are used in the dual quad array described here. They are easily formed and perform slightly better than the other configurations at the higher frequencies. (High-frequency currents don't like sharp bends or abrupt changes, and the circular configuration offers a more uniform transition from the low-to-high impedance points around the loop.)

The circumference of the driven loops can be determined by a simple equation. Circumference of the loop is:

$$C = \frac{12{,}060}{f\,(MHz)} \qquad (1)$$

where C is the loop circumference in inches. For centimeters, the numerator in **eq.** 1 is replaced by ***30.624.*** The driven loops were cut to 82 inches (208 cm), which is near the center of the 2-meter band. The reflector is 2 inches (51 mm) longer, and the directors progressively 2 inches (51 mm) shorter. Spacing is 16 inches (406 mm) on the reflector and 12 inches (305 mm) on the directors.

matching system

Common practice is to open the loop at the desired feedpoint and attach a 50-ohm feedline. This practice gives an acceptable match, but it has been found that a more desirable method is to leave the loop closed and feed it with a modified gamma match as shown in fig. **2.** This method makes for an easy adjustment of SWR and gives an excellent match to the phasing harness. The gamma rods are 8 inch (203 mm) lengths of 1/4-inch (6.5-mm) copper tubing. The gamma capacitor is a miniature Johnson variable. A cap with 15-20 pF maximum capacitance is sufficient. The capacitor is mounted in a plastic tube and sealed for weather protection. [I used 1 inch (22.5 mm) diameter plastic pill boxes in the original construction.]

construction

Construction of the array is quite simple. Hand tools will be adequate for the job. Most of the materials are off-the-shelf items obtainable in any hardware store.

The main framework (fig. **3**) is constructed of thin-wall 1/2-inch (12.5-mm) electrical conduit. Pipe straps, as used for mounting 1/2-inch (12.5-mm) tubing, are used throughout the assembly for mounting the loops and securing the 112-inch (12.5-mm) conduit to the mounting points.

The loops (fig. **4**) are constructed of no. 8 (3.3-mm) aluminum wire sold as TV ground wire. Two strands of this wire are used in each loop. The strands are tightly twisted together to form a semi-rigid conductor.

1. Clamp the two ends of the wire in a bench vise and chuck the other ends into an electric drill motor. Keep the wires evenly spaced by applying a little pressure.

2. Turn on the drill motor and the wires will be tightly and evenly twisted together.

3. Cut the wires about 6 inches (152-mm) longer

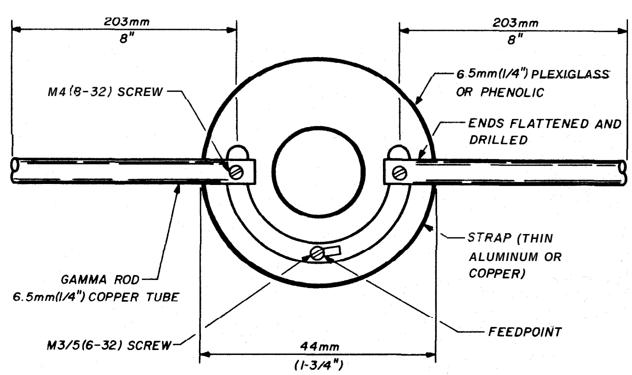

fig. 2. Matching system used in the dual quad antenna. A modified gamma match makes for easy SWR adjustment and gives an excellent match for the phasing harness.

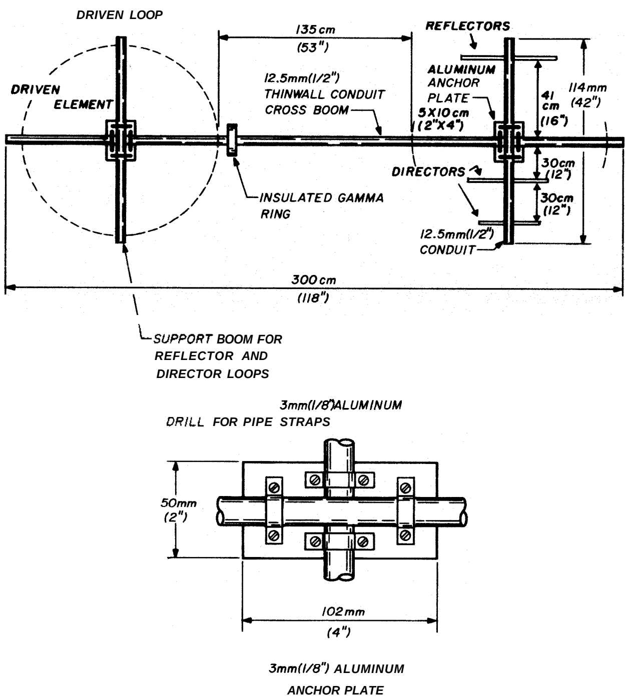

fig. 3. Construction of the main framework of the dual 2-meter quad.

than the desired loop, as they will be shortened by the twisting action.

4. Hand form them into a loop that will be rigid when mounted to a metal cross member.

The insulated mounting rings to which the gamma rods are attached are cut from 114-inch (6.5-mm) plexiglass or phenolic. Use a 1 x 314 inch (25.4 x 19 mm) hole saw.

1. Drill the centers to slip over the 112-inch (12.5-mm) conduit cross member.

2. Flatten the copper tubing gamma rod ends in a bench vise and drill for mounting to the insulating rings. A half-circle strap connects the ends of the gamma rods together; this may be made of copper or thin aluminum.

3. Drill the center of the strap for a 6-32 (M3/5) screw, which secures a solder lug at this point.

This solder lug is the feedpoint and connects to the inner conductor of the phasing harness. The coax braid is being attached directly to the cross boom and secured by another screw and solder lug.

phasing harness

The phasing consists of two equal lengths of RG-59/U coax cut to an electrical 3/4 wavelength and terminated in the center to an SO-239 connector to which the main 50-ohm feedline connects. Because of the coax velocity factor, the actual length will be shortened by this factor. Unfortunately, the propagation velocity will vary between cables of different coax brands.

Foam-filled cable comes out to be near 47 inches (119 cm) for 3/4 wavelength at 2 meters, while the solid type comes out near 36 inches (91 cm).

The recommended method of determining the electrical length is to cut a length of cable slightly longer than the estimated length required. Short one end with a small pickup loop and couple this loop to a grid-dip oscillator. Carefully trim off the free end until you get a dip near the center of the 2-meter band.

tuneup

Each quad section is tuned separately.

1. Connect a 50-ohm feedline from the transmitter to the solder lug feedpoint. (The coax shield is grounded directly to the cross boom.)

2. Set the shorting strap that connects the lower gamma rod to the radiator four inches (102 mm) below the feed point. This dimension may be adjusted if unity SWR is not obtained by adjusting the gamma capacitor.

3. Insert an SWR meter into the line, preferably near the antenna where it may be easily observed.

4. Apply power to antenna at a frequency near the

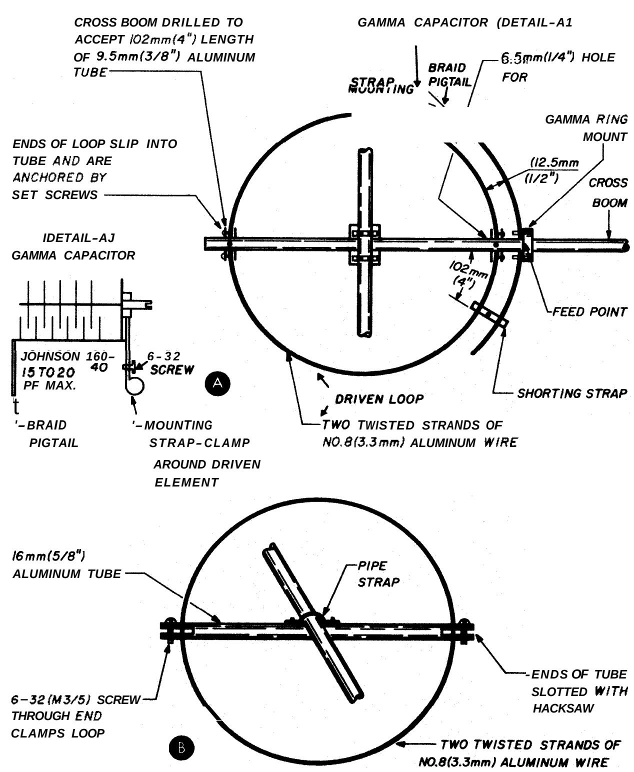

fig. 4. Construction of the driven loop and gamma-match capacitor (A). Sketch (B) shows reflector and director mounting details.

center of the 2-meter band. Use reduced power to prevent interference.

5. Note meter reading. With an insulated tool, adjust the gamma capacitor until a reading of near unity SWR is obtained. Use the same procedure for the other quad.

6. Now connect the phasing harness and main feedline. A slight adjustment of the gamma capacitor will again bring the SWR reading to near unity.

You must, of course, observe the usual precautions of having the antenna in the clear while making these adjustments. Height above ground doesn't affect the quad as much as the Yagi for tuneup adjustments. You are now ready for on-the-air tests with that distant station you've been unable to work.

in retrospect

Those familiar with the Swiss quad have no doubt noted that it's also an all-metal design, with closed loops and a modified gamma match.[2] The Swiss quad has been popular in Europe for several years and is now making an appearance in this country. It's a very good antenna, but I must add that the dual quad array shown here has out-performed it at my location.

Over the years, there's been much controversy concerning the relative merits of the quad versus the Yagi.[3] My only comment is this: Compare this dual quad array with your favorite Yagi, with a comparable number of elements, using on-the-air tests with a distant station. I think you'll find that on-the-air tests don't always agree with gain measurements made on an antenna range under controlled conditions.

This quad array has a low takeoff angle of radiation and will work quite well at low heights above ground. It's very tolerant of variations in dimensions and very easy to match. Directivity is very sharp; a variation of a few degrees can mean the difference between an S1 and an S9 signal. If you wish to use the array for horizontal polarization, rotate the array 90 degrees so that the quads are stacked one above the other.

references

1. *Frank Witt,* W1DTV, *"Simplified Antenna Gain Calculations," ham radio, May, 1978, pages 78-85.*
2. *Jim Fisk,* W1DTY, *"Unusual Cubical-Quad Antennas," ham radio, May, 1970, pages 6-10.*
3. *Wayne Overbeck,* N6NB, *"Quads vs Yagis Revisited," ham radio, May, 1978, pages 12-21. Also Comments, ham radio, October, 1979, page 80.*

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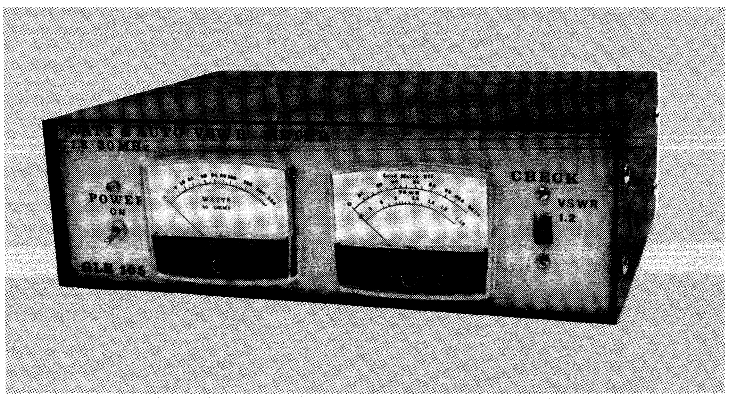
ham radio

More Details? CHECK—OFF **Page 94**



automatic VSWR and power meter

Design and construction of an instrument that indicates rf power and VSWR simultaneously

One of the most common station accessories used by the Amateur Radio operator is the VSWR meter, which determines the voltage standing wave ratio between transmitter and antenna. VSWR is a measure of the quality of the system match, or may be looked at as a measure of system efficiency. Many articles have been written stating that a high VSWR doesn't usually seriously degrade transmission-line performance. Also the articles have correctly stated that reduction in radiation efficiency caused by antenna nonresonance is barely noticeable.

Now that most Amateurs are convinced that their 80-meter dipole will work from 3.5-4 MHz even though the VSWR may be high at either end, they're faced with the VSWR limitation of their transceiver. To solve this problem, a transmatch or antenna tuner is placed between transmission line and transmitter.

In light of the previous comments, it's not as important to use a VSWR meter to monitor the transmission line VSWR as it is to monitor the VSWR between transmitter and antenna tuner.

The antenna tuner introduces more knobs to be adjusted during tune up and can cause some delay in tune up because of interaction between loading adjustments. Also, because of the moderate *Q* involved in the tuner, you can easily encounter VSWRs from infinity to 1:1 while adjusting the tuner. Thus, it's usually desirable to do initial tuning at low power and final tuning at full power. The wide range of power and VSWR encountered during tuner operation makes use of the conventional VSWR meter difficult, because the REF or CAL point changes, which causes inaccurate VSWR readings. This is where an automatic meter can greatly speed up the tuning procedure, because the VSWR-meter readings are accurate and independent of power level down to some minimum power level.

automatic VSWR meter

The automatic VSWR meter described in this article is unique, versatile, and easy to use. The basic circuit design is covered by U.S. Patent **4,110,685.** It's battery operated and has an automatic ON/OFF feature for extended battery life. Two meters that display power and VSWR simultaneously give a greater feel for how well the transmitter is operating on a continuous monitoring basis. Also, the two-meter display of power and VSWR as independent

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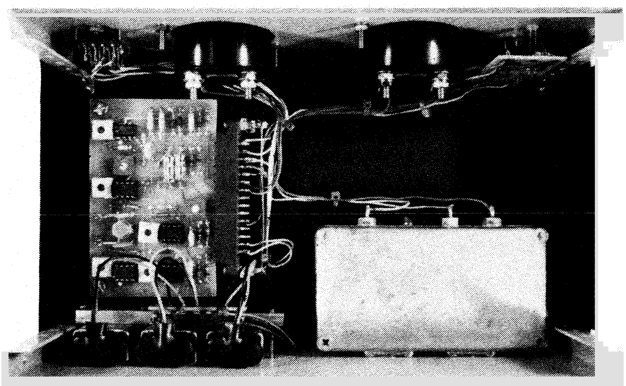


Inside view – complete instrument.

parameters greatly speeds up antenna loading adjustments and gives greater confidence that the transmitter is operating properly without any switching, reference setting, or mental calculations.

how it works

As in all VSWR meters (**fig. 1**), an in-line directional coupler senses and develops a voltage proportional to the forward and reflected voltage on the transmission line. The directional coupler in this instrument also has a diode compensation network to provide additional linear range at low levels. It also senses when rf is present to saturate a transistor, which is used to turn the instrument on.

The signal from the directional coupler that senses rf power is fed to a circuit that uses transistors to turn the supply voltages of +16 and –9 volts to the analog computing circuits ON or OFF.

The directional-coupler dc outputs are connected to fet input buffer amplifiers to provide a high-impedance load to the directional coupler outputs and low impedance outputs for the logarithmic amplifier and wattmeter.

Dc outputs V_F and V_R are linearized to provide outputs that can be used directly. Thus V_F can be used to drive a wattmeter. Power is displayed in watts on a scale constructed by using the formula

$$P = \frac{V_F^2}{R} \qquad (1)$$

Since dc output voltages V_F and V_R are developed by peak-detecting diodes, the wattmeter scale displays two decades of power range; *i.e.*, the 250-watt full-scale meter displays power from 2-250 watts. This can be seen in the photo of the meters. A peak detector whose output is calibrated to read volts rms or power can have errors if the signal has large harmonic content. However, in the rf transmitter applications such as this, the signal usually has a very low harmonic content; thus the peak detector gives accurate results.

The logarithmic amplifier develops a voltage at the output with a relationship proportional to the logarithm of the input voltage. At the log amp outputs are two voltages that are the logarithm of V_F and V_R. An expression called return loss, which is used by the telecommunications and instrument industry, is defined as follows:

$$\begin{aligned} \text{return loss} &= 20 \log 1/\rho \\ &= 20 \log V_F/V_R \end{aligned} \qquad (2)$$

where ρ = reflection coefficient

By mathematical equality, **eq. 2** can also be expressed as

$$\text{return loss} = 20\,(\log V_F - \log V_R) \qquad (3)$$

It can now be seen that if we take the difference of the two voltages, log V_F and log V_R, the result will be a voltage proportional to the return loss in dB:

$$E_{RL} = \log V_F - \log V_R \qquad (4)$$

Return loss is related to VSWR by

$$RL_{dB} = 20 \log (VSWR + 1)/(VSWR - 1) \qquad (5)$$

By scaling construction the meter can be made to display VSWR (see the VSWR-meter photo). Voltage E_{RL} is the difference between voltages log V_F and log V_R, which is always present from the log amps. Thus VSWR is automatically and continuously displayed with no need to set a reference and operate a switch.

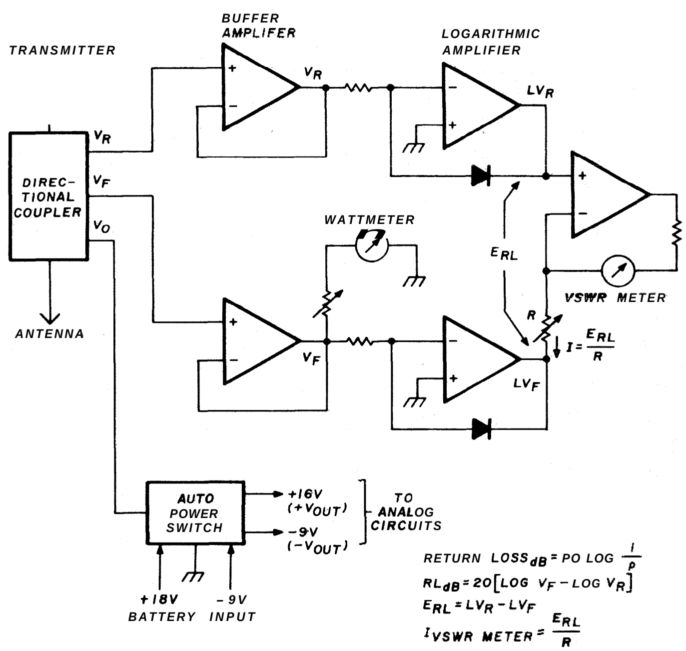


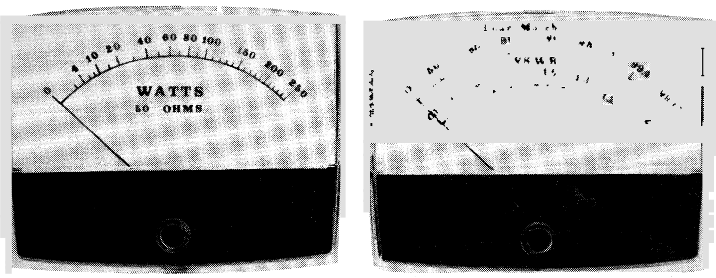
fig. 1. Automatic VSWR/power meter block diagram.

VSWR-meter scale

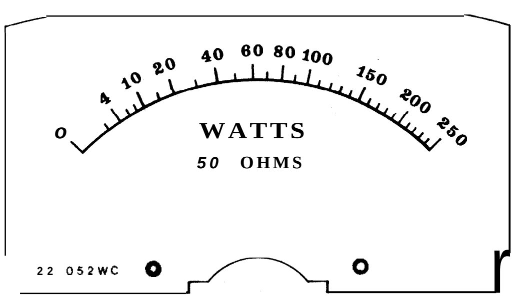
One of the things you'll notice when you look at the VSWR meter scale (photo) is that it's "backwards." With this meter you peak the VSWR meter as well as the wattmeter for best loading.

Having $VSWR = \infty$ at the left-hand end of the meter is explained by the way VSWR is computed. Note table 1, in which the three parameters, reflection coefficient, VSWR, and return loss, are tabulated.

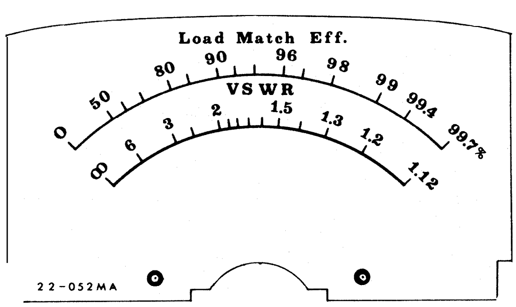
From this table you can see that when $VSWR = \infty$, return loss = 0 dB; and conversely, when $VSWR = 1$, return loss $= \infty$ dB. In this instrument return loss is computed so the display of $VSWR = \infty$ is no problem, because the subtraction of two equal voltages is easily made electronically. However,



Wattmeter and VSWR meter scales.



Wattmeter scale — actual size.



VSWR **scale — actual size.**

table 1. Reflection coefficient, p, VSWR, and return loss.

ρ	VSWR	return loss (dB)
1.000	∞	0
1.891	17.391	1
1.794	8.724	2
1.708	5.848	3
1.501	3.010	6
1.398	2.323	8
1.316	1.925	10
1.200	1.499	14
1.126	1.288	18
1.100	1.222	20
1.063	1.135	24
1.056	1.119	25
1.032	1.065	30
1.010	1.020	40
1.003	1.006	50
1.001	1.002	60
1.000	1.000	∞

there a p[illegible] i[illegible] the display of $VSWR = 1$, because i[illegible] value is equal to infinite return loss. Obviously, a d[illegible] of zero t[illegible] infinite return loss in dB is impossible. A[illegible] a result, a practical limit must be selected, so I've chosen to limit return [illegible] to 25 dB, or a VSWR of 1.12.

This limit was chosen for practicality. A VSWR of 1.12 represents an efficiency of 99.7 per cent of the available power being delivered to the load. Ivory Soap long ago convinced the consumer that 99.44 per cent pure soap was good enough, and I feel that 99.7 per cent of my transmitter power delivered to my antenna is good enough.

With the VSWR scale is a scale called LOAD EFFICIENCY in per cent. This scale greatly enhances your feel for system efficiency. (There's always the question, Just what does VSWR mean and how much should I have?) The scale is also a constant reminder that there's a reasonable limit to how *low* you must keep the VSWR. Another practical reason for the limit in the displayed VSWR is that a directional coupler with isolation good enough to measure low VSWR numbers of high return-loss numbers is difficult to construct.

directional coupler

The directional coupler is of conventional design and works on the principle described by Bruene;[1] however the design of this directional coupler for specific applications wasn't covered. I've developed some simple formulas that can be used in predicting, to a good approximation, the values needed to obtain the desired operation. Reference 1 covers the theory of operation and some considerations for the directional coupler design. I urge you to review this fine article. The basic circuit is shown in **fig.** 2 and is used to develop the formulas.

Bridge balance. The basic principles of the directional coupler is that, at bridge balance, the voltage presented to reflected diode detector CR_R is zero and the voltage presented to forward diode detector CR_F is E_D. The diode-detector voltage output is determined by two voltages proportional to the magnitude and phase of transmission-line voltage E_V and current I_i. The voltage proportional to transmission line current I_i is made to appear in phase with I_i on the reflected diode detector by the center-tapped resistor load for the current-to-voltage transformer.

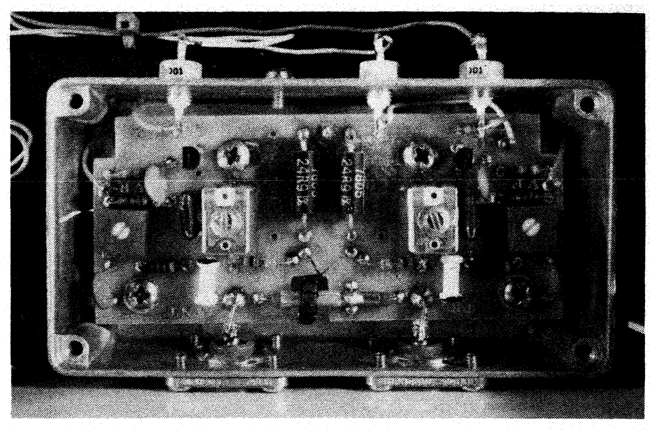
At bridge balance, and with a reference resistive load, $E_O = E_D/2$ is in phase on the reflected side and −180 degrees out of phase on the forward side. With these conditions the forward diode detector voltage is $E_O + E_D/2 = E_D$, and the reflected diode detector voltage is $E_O - E_D/2 = 0$.

When a load other than the resistive reference value is present, the bridge will not be balanced, and the voltages presented to the diode detectors will be determined by both magnitude and phase of the voltages E_O and $E_D/2$. The amount of unbalance is detected in the diode detectors. Their dc outputs result in the outputs, voltage forward V_F and voltage reflected V_R, from which the ratio of these two voltages determines the VSWR.

There's no unique solution that will completely design the coupler. **Table 2** lists the main parameters and guidelines to design this type of directional coupler. From the previous paragraph note that, at bridge balance, $E_O = E_D/2$ and also that the voltage presented to the diode is E_D. Since voltage E_D across

table 2. Design guidelines.

parameter	minimum	maximum
diode voltage	diode threshold voltage	diode breakdown voltage
insertion resistance	none	power in resistor and design maximum
value of $C1$ for loading transmission line	none	$XC1 \geq 10 \cdot Z_O$ at highest frequency of interest
number of turns on toroid core	$X_L \geq 5 \cdot R$ at lowest frequency of interest	length of wire $\leq 1/20\lambda$ at highest frequency of interest
value of $C2$	$C2 \geq 25 \cdot C_D$, C_D = diode capacitance	



Inside view — directional coupler.

the diode is a parameter to keep track of, I've found it convenient to develop my formula using E_D.

Current transformer. Now for the expression to use for the current transformer. Assume the conditions for a valid transformer are met. Then $E_D = E_i \cdot N$, since the primary in this case has only one turn. Also $E_i = R_I I_i$; $I_i = \sqrt{P/R_L}$; and $R = R_I N^2$. Putting all this together, the expression to work with for design is

$$E_D = \frac{R}{N}\sqrt{P/R_L} \tag{6}$$

From this expression we see that, at a given power level, E_D is increased with increasing values of R and decreased with increasing number of turns.

For the above expression to be true, current-transformer action is assumed. To obtain this, the inductive reactance of the transformer secondary must be $\geq 5 \cdot R$ at the lowest frequency of interest.

To compute the inductance of a toroid inductor, the expression $L = N^2\, LT/NT^2$ is used. LT/NT is the A_L or ***inductance index*** of the toroid. If A_L is expressed in inductance per unit turn, the inductance of a toroid is the number of turns squared times A_L or $L = N^2 A_L$. After calculating the toroid inductance, the reactance at the lowest frequency of interest can be determined by

$$X_L = 2\pi f L \tag{7}$$

As for E_D, the value of R and N also determines the directional coupler insertion resistance. Insertion resistance is equal to R divided by N^2; $R_I = R/N^2$. Insertion loss can also be expressed in dB for a 50-ohm system:

$$dB_I = 20 \log \frac{100 + R_I}{100} \tag{8}$$

The current transformer load, R, will have power applied during operation, and the power in R can be calculated by

$$P_R = E_D^2/R \qquad (9)$$

The voltage at the other side of the diode is determined by capacitive divider $C1$ and $C2$. Voltage-divider output E_O is equal to E_v times $C1$ divided by $(C1 + C2)$. If we let $E_v = \sqrt{P \cdot R_L}$ and $E_O = E_D/2 = \frac{R}{2N}\sqrt{P/R_L}$ (condition for bridge balance) and solve for $C2$, we obtain

$$C2 = \left[C1 \quad \frac{2R_LN}{R} - 1\right] \qquad (10)$$

This expression is convenient because it's in terms of the two parameters usually adjusted to obtain the desired performance.

In summary, here are the expressions needed to solve for the values of the components for a directional coupler:

$$N = \frac{R}{E_D}\sqrt{P/R_L} \text{ (turns)}$$

$$L = N^2 \cdot A_L \; (A_L \text{ is inductance per unit turn})$$

$$R_I = R/N^2 \text{ (ohms)}$$

$$P_R = E_D^2/R \text{ (watts)}$$

$$C2 = C1\left[\frac{2R_LN}{R} - 1\right]$$

Design example. To gain a feel for using the formulas, let's work out the values used for the directional coupler in this article. (See the basic circuit, fig. 2). In this case I'm operating the diodes in a peak-detecting mode, so I'll select the diode voltage to be 10 volts rms when the maximum power of 250 watts is applied through the directional coupler. Also I'll select R to be 50 ohms and the frequency range to be 1.8-30 MHz. These initial design parameters will allow the component values to be calculated.

The number of turns, N, for the toroid is

$$N = \frac{R}{E_D}\sqrt{P/R_L}$$

$$= \frac{50 \text{ ohms}}{10 \text{ volts}} \; \frac{\sqrt{250 \text{ watts}}}{50 \text{ ohms}} = 11.18 \text{ turns}$$

Since we can't wind fractional A_L turns, the number of turns are rounded off to 11, which will be used in the following calculation.

Now we check to see if the number of turns is okay for inductance. Calculate $L = N^2A_L$ where $A_L = 2.1\mu H$ per turn for the Ferroxcube toroid core number 266T125 (fig. 3) and we find $L = 11^2 \cdot 2.1 = 254\mu H$. At 1.8 MHz, the inductive reactance is 2872 ohms. The inductive reactance is easily greater than 5 times 50 ohms, which is okay. Empirically, by winding the toroid core, we find the length of wire to be 18 cm. The maximum length of wire is approximately

$$1 = \frac{300}{20f} \text{ in meters or } 1 = \frac{1500}{f} \text{ in centimeters.}$$

In this case, 1 max $= \frac{1500}{30} = 50$ cm, which means my 18-cm length of wire is okay.

The insertion resistance is $R_I = R/N^2 = \frac{50}{112} = 0.4$ ohm, and the power in the resistors at full scale is

$$P_R = E_O2/R = \frac{102}{50} = 2 \text{ watts, or 1 watt per}$$

resistor. The values obtained are acceptable. A 25-ohm, 1-watt resistor is a practical value, and an insertion resistance of 0.4 ohm, which translates to a 0.04-dB loss, is reasonable. If, during this step, the power in the resistor is too high or insertion loss is unacceptable, you'd adjust either R or N or both and try again until an acceptable answer is obtained.

Capacitor values are determined by first selecting the value of $C1$, which should have an impedance of approximately $10 \cdot Z_O$ or 500 ohms at 30 MHz. The maximum value of capacitance is 10 pF. Since there are two $C1$s, each $C1$ has a maximum value of 5 pF. To obtain some margin, I've selected 3.3 pF for $C1$. From this selection of $C1$, $C2$ can be calculated by:

$$C2 = C1\left[\frac{2R_LN}{R} - 1\right]$$

$$= 3.3\,[22 - 1] = 69\,pF$$

The diode capacitance is about 2 pF maximum; therefore $C2$ should be greater than $2 \cdot 25$ or 50 pF. We must adjust the bridge so that $E_O = E_D/2$ for balance, so either $C1$ or $C2$ is made adjustable.

In summary, we find that for a 250-watt, 1.8-30 MHz directional coupler, the following values will work: $R/2 = 25$ ohms; 11 turns on the toroid core; $C1 = 3.3\,pF$, and $C2 = 69\,pF$. I've verified that the preceding values will give the desired performance. The calculated values are ideal, and in practice the results will be slightly different. However, I've found from building a number of different types of directional couplers that the calculated and measured values are always quite close in agreement.

Dc output voltages V_F and V_R are taken from the junction of $C1$ and $C2$ (fig. 2). The resistor value or impedance must be high enough not to load $C2$, which means the load impedance should be approxi-

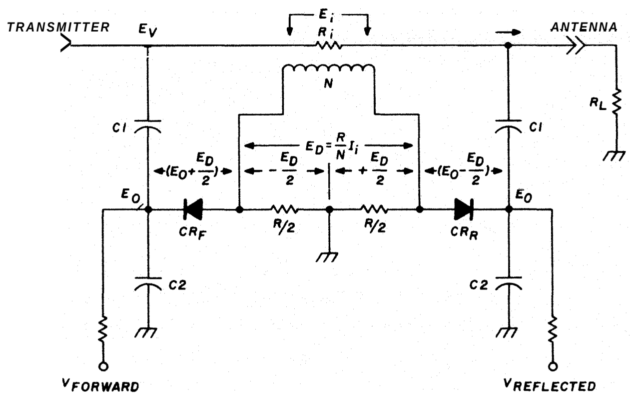


fig. 2. Directional coupler basic circuit.

Construction. The construction can be noted in the photo of the inside of the directional coupler. The mately 20 times or greater than *C2* reactance at the lowest frequency of interest.

In this directional coupler design a diode compensation network uses the base-emitter junction of a transistor to extend the low-end range of the diode detector. Note the circuit in **fig. 3.** The compensation works by the variable attenuator formed by ***R4,*** *R2,* and *Q1.* When the dc voltage from the diode detector is high, the current in R2 is high, thus causing Q1 base-emitter junction to conduct and look like a low-impedance for R2 to ground. Since R2 is grounded, the detector voltage is attentuated by the ratio of R4 to R2. When the dc voltage from the diode detector is low, the current in *R2* is low, causing *Q1* base-emitter junction impedance to increase due to less current in ***R4,*** which, in turn, decreases detector-voltage attenuation to the V_F output. Note also that *Q1,* which compensates for V_F, has its collector brought out so that, when enough signal is present, *Q1* collector saturates and turns on the power to the instrument through the ON/OFF circuit.

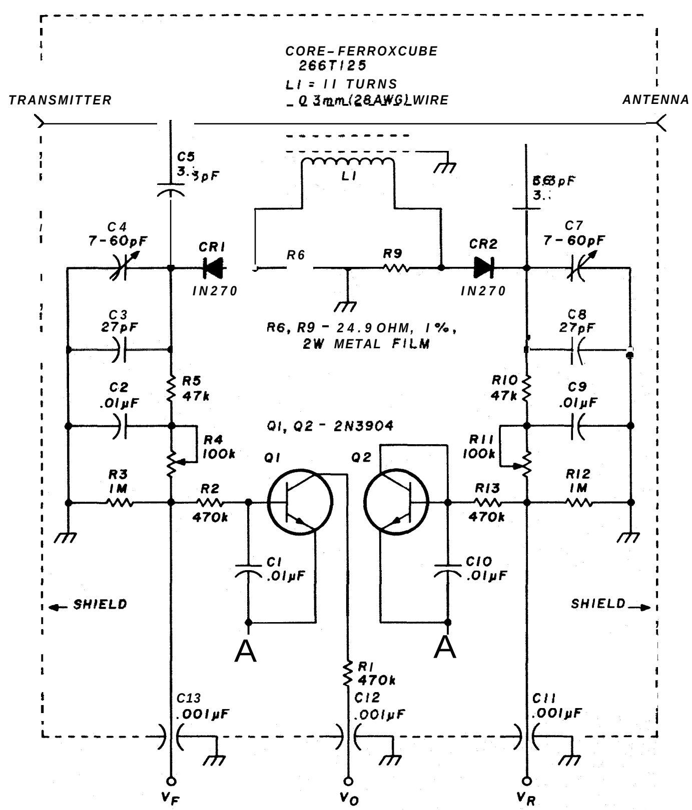
Calibration. The composition circuit is calibrated by applying a known 3 volts rms at some midrange frequency, say 7 MHz, and adjusting ***R4*** (*R11*) for an output of 3 volts dc at V_F (V_R) with a 10-megohm voltmeter. Disconnect the anode of *CR1* (*CR2*) from ***R6*** (*R9*) and ***L1*** for this adjustment. Calibration of the bridge for balance is straightforward. Adjust ***C7*** for minimum voltage at V_R when connected in the normal manner. Reverse transmitter and load for adjusting *C4* for minimum voltage at V_F. The use of a known good 50-ohm load is a must for proper calibration.

Because it's not easy to determine the sense of the toroid output, it may be necessary to reverse connection to the toroid to get voltages V_F and V_R at the proper side. I usually find a small amount of interaction in adjustments, so I repeat the procedure to be sure the directional coupler is calibrated accurately. directional coupler was constructed in a die-cast aluminum box, Bud no. CU-124. Feedthrough capacitors were used to bring out the dc signal voltages. The layout of the PC board isn't especially critical; however, a compact symmetrical pattern usually works best.

power on-off circuit

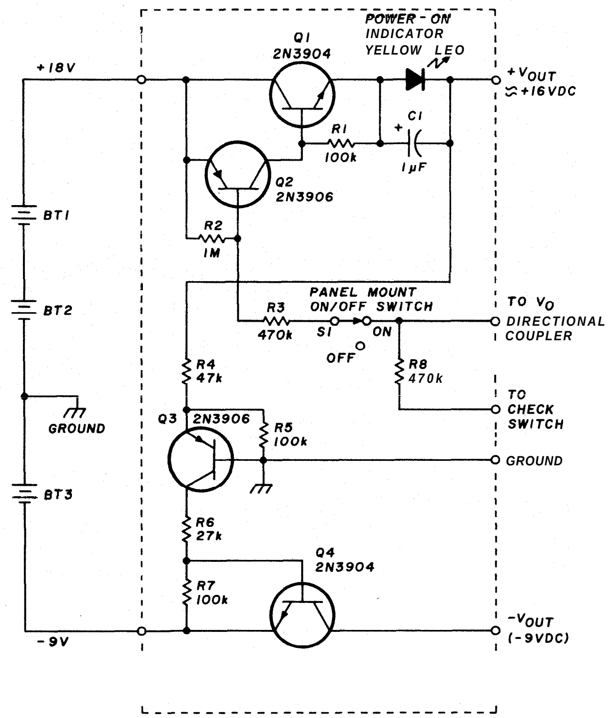
Power for the automatic VSWR meter comes from three standard 9-volt batteries. Two are connected in series to provide +18 volts and the other provides – 9 volts to the analog circuits. The circuit is in **fig. 4.** When ***R3*** is grounded by the saturated collector of *Q1* in the directional coupler, the Darlington transistor pair *Q1, Q2* is turned on so that the +18 volts is supplied to the analog circuits through *CR1,* which is a power-on indicator LED (amber).

When the positive supply goes on, current flows through ***R4,*** which turns on *Q3,* which in turn turns on *Q4.* This connects the – 9 volt battery to the analog circuits. The ON/OFF switch is connected so that when it's in the open or OFF position, the instrument will not come ON even though rf power may be present in the directional coupler. The purpose for this is to allow for longer battery life because it's usually not necessary to monitor power and VSWR



RESISTORS 1/4 WATT 5% CARBON COMPOSITION
C4, C7 VARIABLE MICA NO. 404 (ARCO)

fig. 3. Directional coupler schematic.



BT1.2.3 STANDARD 9V BATTERY NEDA 1604
RESISTORS 1/4 WATT 5% CARBON COMPOSITION

fig. 4. Automatic power switch schematic.

after tune up and may also be distracting to see meters going up and down during normal CW or phone operation.

Expected battery life is about thirty-six hours for continuous operation. However, with the OFF switch and normal transmit-receive duty cycle, the expected life should be at least a year for the average Amateur. Battery operation is very desirable for ease of installation and use of the VSWR meter, especially for tuning mobile antennas.

A connection from the power ON/OFF circuit also goes to the CHECK switch, which turns on the power so that analog circuit calibration can be checked from time to time and also allows a measure of battery condition from the wattmeter. (This will be covered in more detail in the description of the check-switch circuit.)

analog circuits

The analog circuits **(fig. 5)** receive the dc signals from the directional coupler and, when power is applied, process the dc signals to be applied to the wattmeter and VSWR meter. The two dc signals from the directional coupler, V_F and V_R, are sent to the buffer amplifiers, which are connected in a unity-gain configuration. The output of the V_F buffer amplifier *U2* is connected to the input of V_F logarithm amplifier *U5* and also the wattmeter, The output of the V_R buffer amplifier, *U1*, is connected to the input of the V_R logarithmic amplifier, *U3*. Logarithmic amplifiers *U3* and ***U5*** are of a basic configuration using a grounded-base transistor for the feedback diode. Note in the circuit that the feedback diodes, ***U4*** are a matched transistor pair (National Semiconductors LM394BH).

In this application it's very important that the temperature and logging characteristics match, so that the difference in voltage between the two logarithmic amplifiers doesn't vary as a function of temperature and other environmental factors. The absolute voltage of each logarithmic amplifier will change, but that's not an important factor because the meter amplifier, *U6*, is connected to reject the common-mode voltage from the logarithmic amplifiers; thus only their voltage difference is measured. Also note in the diagram that diodes *CR1* and *CR2* are connected to prevent possible reverse voltage on the base-emitter junctions of ***U4*** from becoming too large.

The output voltages of the logarithmic amplifiers are called LV_F and LV_R; their voltage difference results in a new voltage, E_{RL}, which is proportional to return loss in dB. Meter amplifier *U6* is connected as a difference-voltage amplifier. The VSWR meter is connected in series with feedback resistors ***R17*** and ***R18***. In this configuration, the current in the meter is determined by the value of *R15*+***R14*** and E_{RL}. ***R15*** is adjustable, so that the scale factor, or calibration, of the VSWR meter can be set.

Diode *CR3* in the feedback circuit of *U6* limits the forward and reverse current through the meter to protect the meter from damage when VSWRs greater than 1.12 are measured. The reverse-voltage protection is necessary during initial calibration of the unit when a reverse polarity signal can easily be present.

Also note that the inputs to the meter amplifier are not referenced to ground, thus providing common-mode voltage rejection. The operational-amplifier inputs will never be more than plus or minus one diode drop from ground.

The analog board is an analog computer that computes return loss and, by scale construction on the meter, displays VSWR. Because most all operational amplifiers have offset currents and voltages, each operational amplifier has an offset voltage adjustment so that these errors can be compensated for proper operation. A voltage divider network consisting of ***R19***, *R20*, and *R21* provides an accurate voltage ratio for calibration and check.

check-switch function

The check switch serves a very useful function. It

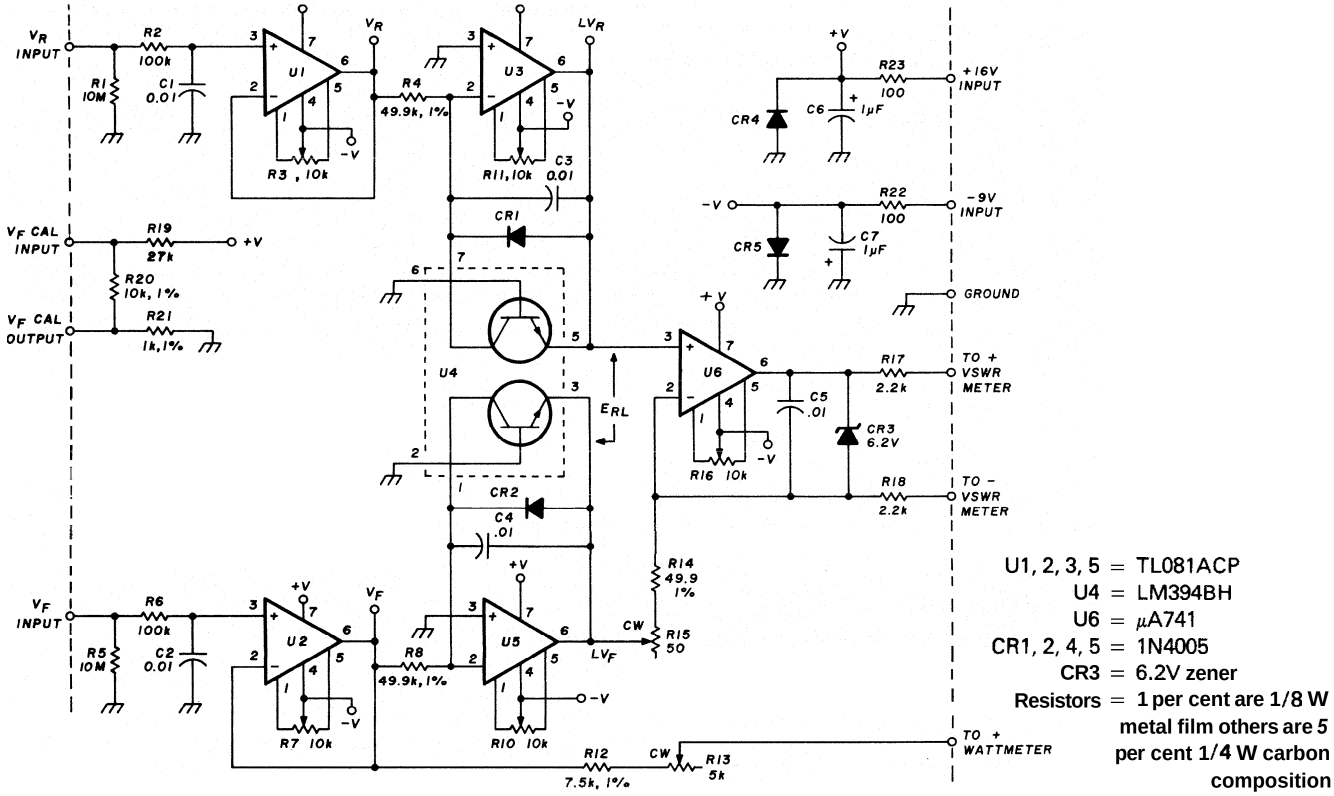


fig. 5. Analog circuit schematic.

allows instrument calibration and operation to be verified at regular intervals. The calibration is verified by the voltage ratio from resistor divider *R20* and *R21*, **fig. 5.** The voltage ratio in this case is 0.091, which is equal to reflection coefficient ρ. A ρ of 0.091 is equal to a VSWR of 1.2:1. During normal operation V_F and V_R inputs come from the directional coupler, but when the check switch is put in the CHECK position, V_F and V_R inputs are connected to the known voltage ratio corresponding to a VSWR of 1.2:1. This known voltage ratio is used to calibrate the instrument.

The V_F *Cal* voltage is a measure of the battery voltage. As such it appears on the wattmeter in watts. Accuracy isn't excellent, because the wattmeter calibration affects the actual *V-to-watts* reading; however, it's an excellent first-order reading of the battery voltage. On this 250-watt scale, a low battery voltage reading is approximately at **40** watts.

construction notes

Construction of this instrument requires no special techniques to achieve success. A wiring diagram appears in **fig. 6.** Construction isn't detailed because anyone experienced in construction should be able to meet the parts and fabrication requirements.

One area of concern is shielding. The analog board and power ON/OFF board have high impedances, so care must be taken to make sure these boards are not exposed to rf. Except for external rf pickup, there are no critical layout areas inside the instrument. The shielded directional coupler is essential so that it may be located inside the instrument (see **fig. 3**).

The meters used are readily available from Radio Shack, part 22-052. I made a three-to-one scale of the meter face, then had it photographically reduced to fit on the scale. The scale photograph was then attached to the Radio Shack meter scale by doubie-sided tape. I removed the original scale from the meter, attached the new scale to it, and then replaced the scale, securing it with the two mounting screws.

Six 1-per cent metal film resistors are called out for the analog board. Availability may be a problem, so here are some possible alternatives with expected performance changes. ***R4*** and ***R8*** must be a matched pair for logarithmic tracking. Thus the actual value is not critical, and they could be carbon film resistors if they are matched to 1 per cent. ***R14*** and *R12* are metal film resistors only to minimize temperature influence and could be carbon film.

R20 and *R21* must be selected to be within 1 per cent of the absolute value, because these two resistors determine the basic accuracy of instrument calibration. I encourage the builder to find 1 per cent

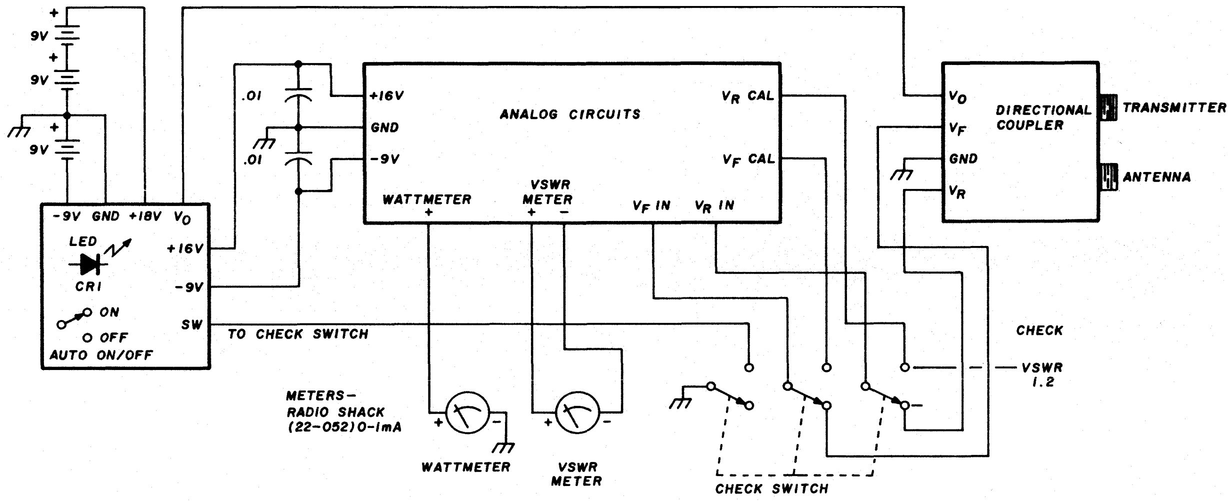


fig. 6. Wiring diagram.

metal film resistors if at all possible. If carbon film resistors are used in any of the l-per cent metal-film resistor slots, some performance degradation will occur as a function of temperature.

Us 3 and 5 must be in sockets to allow for adjusting the offset trim pots for the operational amplifiers. Also test points for V_R, V_F, LV_R, and LV_F will allow for easy testing and calibration of the analog board.

calibration

Calibration of the instrument is in two parts, rf and dc. The directional coupler can be calibrated before installation in the instrument; the procedure was covered in the discussion on the directional coupler. Refer to **fig. 5** for the following procedure:

1. With U3 and 5 removed from their sockets, ground the V_O output of the directional coupler and turn on the instrument. The amber LED pilot light should glow, indicating power is being drawn by the analog board.

2. Connect test points LV_R and LV_F together with a jumper lead.

3. Adjust R16, *U6 offset adjust* for "∞" on the VSWR meter (normal meter zero).

4. Disconnect jumper from test points LV_R and LV_F. Connect voltmeter to V_R and adjust R3, *U1 offset adjust* for zero volts ($< \pm 5$ mV).

5. Connect voltmeter to V_F and adjust R7, *U2 offset adjust,* for zero volts ($< \pm 5$ mV).

6. Turn off instrument and install *U3* and *5.*

7. Connect voltmeter to test point LV_R and ground. Turn instrument on and adjust *R11* for approximately -0.30 volts. Then adjust *R10* for "m" on VSWR meter. There will be some interaction, so repeat adjustments of *R11* and *R10* until voltage test point LV_R is -0.30 volts and VSWR meter is "∞".

8. Set check switch to 1.20 VSWR, check position, and adjust *R16* for a VSWR reading of 1.20.

9. Return check switch to NORMAL or OFF position and note that voltage at test point LV_R is still about -0.30 volt. With no inputs to the analog board from the directional coupler, the voltage at test points LV_R and LV_F will drift around due to the extreme low currents into the logarithmic amplifiers. This is normal. When the dc signals from the directional coupler are present the logarithmic amplifier outputs will be stable.

10. Turn instrument off and disconnect voltmeter and jumper wire that grounded V_O from directional coupler.

11. Connect input side (TRANSMITTER) of directional coupler to transmitter and output (ANTENNA) side of directional coupler to a good wattmeter and dummy load. With an applied rf power of between 100-250 watts, adjust *R13* so that the instrument wattmeter reads the same as the external wattmeter. Calibration frequency isn't critical, but I usually use the 40-meter band as a midrange frequency.

performance

Wattmeter accuracy is primarily determined by the design and construction of the directional coupler. In this instrument, the wattmeter accuracy is ± 5 per cent reading $+1$ per cent full scale between 1.8 and 30 MHz. The VSWR measurement accuracy is difficult to specify, because several factors affect the measurement. For VSWR greater than 2, the directional coupler open-short ratio dominates the VSWR error.

This directional coupler measures an open-short ratio of 0.6 dB, which translates to an error of 7.5 per cent at a VSWR of 6. A complete discussion of the open-short ratio can be found in an instruction book for the ANZAC Model RB-3 standing-wave-ratio bridge, dated March, 1966.[2] For VSWR less than 2, directional-coupler isolation is a dominant factor with the detector diode operating-voltage level.

The isolation of this directional coupler is about 30 dB. The analog circuit dynamic range exceeds that of the detector diodes and isn't a factor in VSWR accuracy. As with any diode detector, a minimum level of ac signai is required before any dc current wiii fiow. Also since the diodes are in the peak detecting mode, there's a minimum level of ac signal where the diode dc current will be accurate with respect to the ac signal peak value. Therefore, a minimum amount of signal is necessary to obtain accurate readings.

In this instrument, the minimum forward power required to obtain an accurate ±10 per cent VSWR at various VSWR numbers is:

1. 2.5 watts for VSWR = 2.0.
2. 5 watts for VSWR = 1.5.
3. 25 watts for VSWR = 1.2.
4. 35 watts for VSWR = 1.12.

You may ask what it means if the VSWR meter reads more than 1.2 with 25 watts input. It means that the VSWR is better than 1.2 but the actual value isn't known. To put it another way, at any power level a minimum VSWR number is accurate, and any lower number is optimistic. This minimum power for an accurate VSWR number is characteristic of all directional couplers using diode detectors; however, it's usually not specified nor mentioned as a limitation in VSWR measurement.

Performance is more than adequate for quick and accurate measurements of a typical transmitter power output and how well it is matched to the load.

VSWR measurement

Numerous articles correctly state that VSWR in a transmission line is not an important parameter to keep low in value, and that only upper-limit numbers are of concern based on transmission line loss and other factors. So why is there still concern about VSWR?

One aspect of the measurement sometimes overlooked is that a ***reference impedance*** is involved. The VSWR measurement is only valid for the reference impedance for which it has been designed, in this case 50 ohms. Common practice is to define system performance with reference to a system impedance. Also common practice for Amateurs is to use a 50-ohm dummy load for tune up. In doing so, the 50-ohm reference is established, and if all other measurements are relative to this 50-ohm reference, the results will be valid. If VSWR measurements are made in which the reference impedance is other than that of the instrument, the results will not be valid, and a correction must be made.

I feel that until a readily available means of measuring power transfer independent of impedance is available, the VSWR meter will continue to be a valid means of determining system performance. Plate-current meters on most transceivers aiiow a first-order means for measuring power transfer. Careful use of these meters can provide successful results. But I've never been quite as satisfied with this method as with the use of the external power and VSWR meter, using a transmatch to couple the antenna and a 50-ohm dummy load for reference.

conclusion

This article has several objectives. One is to give complete details on how my automatic VSWR meter design works so that others can understand its operation and use. Another is to provide design equations and show their use in the design of a commonly used directional coupler. Finally, it was desired to provide construction details for duplication of the instrument.

This automatic VSWR/power meter is a project I've been working on for more than two years. I'd like to thank fellow Amateurs who've encouraged me and given valuable advice. With their help I've developed what I believe is a very useful station accessory.

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ham radio

experimental high-gain phased array

Some interesting test results using phased colinear elements based on the "ZL Special"

Since the construction of the **ZL beam**[2], I've been intrigued by the possibility of producing another noticeable improvement in antenna performance. After reviewing the literature, several options seemed available. First, one could duplicate the antenna and stack it over the present assembly. Secondly, one could stack the antenna in a side-by-side arrangement. Both arrangements should produce a 3-dB change in antenna performance. However, the cost of construction would be enormous, as would the cost of strengthening the support structure.

These options were discarded for economic reasons, and I continued to search the literature. After some thought, the following question came to mind: Why not extend the lengths of the ZL elements into full-size, half-wave elements and feed it as if it were a ZL **Special?**[1,2] The following report gives the results of the experiment.

description

Fig. 1 shows the basic construction. Nine half-wave elements were constructed consisting of four driven half-wavelength elements E_B, E_A, collinear reflector R, and collinear director D_1. Yagi director D_2, 1/2 wavelength long and spaced 1/4 wavelength from D_1, was also constructed. Two phasing stubs were connected to points A, B. This basic configuration was alternately fed at these points for the tests **(figs. 2** and **3**). The parasitic elements were tuned and relative field-strength patterns were obtained as shown.

Reversing the A-B feed changed the action of

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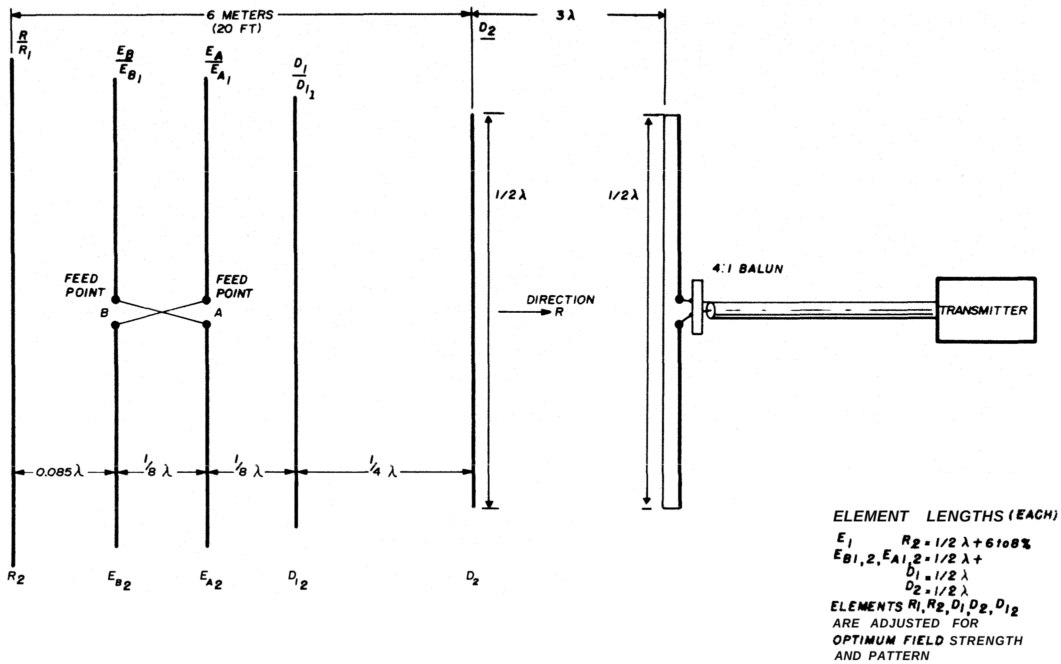


fig. 1. The basic construction used in the experiments with the phased array.

phased elements to such an extent that the action of the parasitic reflector element was overwhelmed, which in **fig.** 3 had been tuned for maximum radiation in direction R.

tests

The test using feed point A **(fig. 2)** produced the best results. A 15-meter model was made, and performance was good. However, it didn't "feel right." Subsequently, a model was made for 10 meters and operated by a friend, who reported results not much better than those obtained from a regular Yagi. Again, the literature was searched and I discovered that element E_A must have *opposite* polarity with respect to element E_B.

Once more the literature was reviewed, and a new question arose: Is there a way to connect the elements so that when a wave is emitted from E_B, it will coincide simultaneously with a wave emitted from E_A when E_A and E_B are 1/8 wavelength apart? The literature suggested that a 518-wavelength piece of coax, fed at the 1/8-wavelength point, would pro-

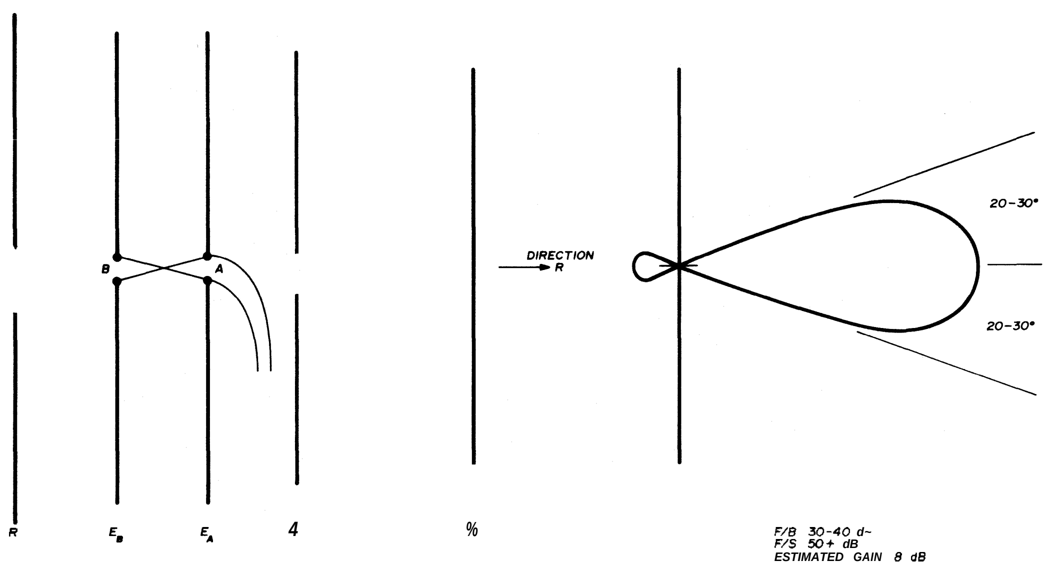


fig. 2. Responsewhen feed point A was used, which produced the best results.

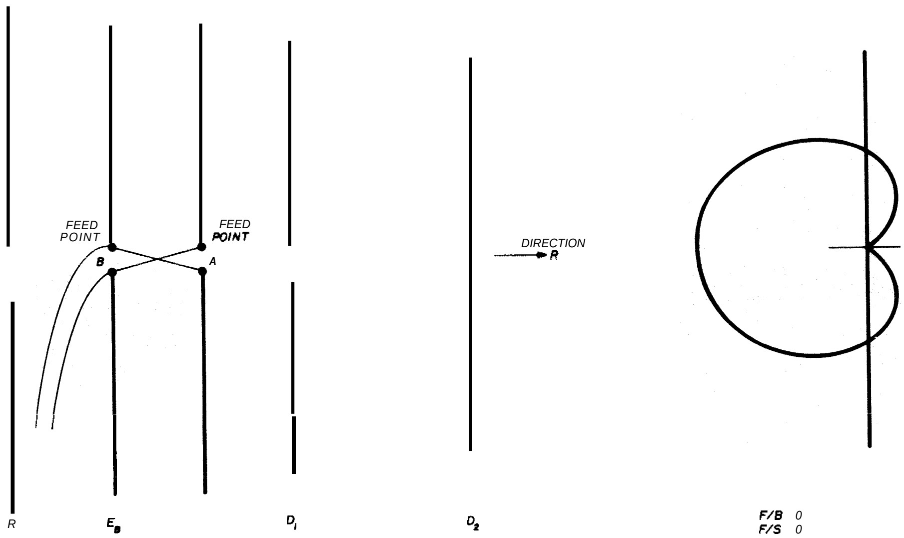


fig. 3. Reversing the A-B feed point changed the pattern dramatically, because the parasitic reflector element was overwhelmed.

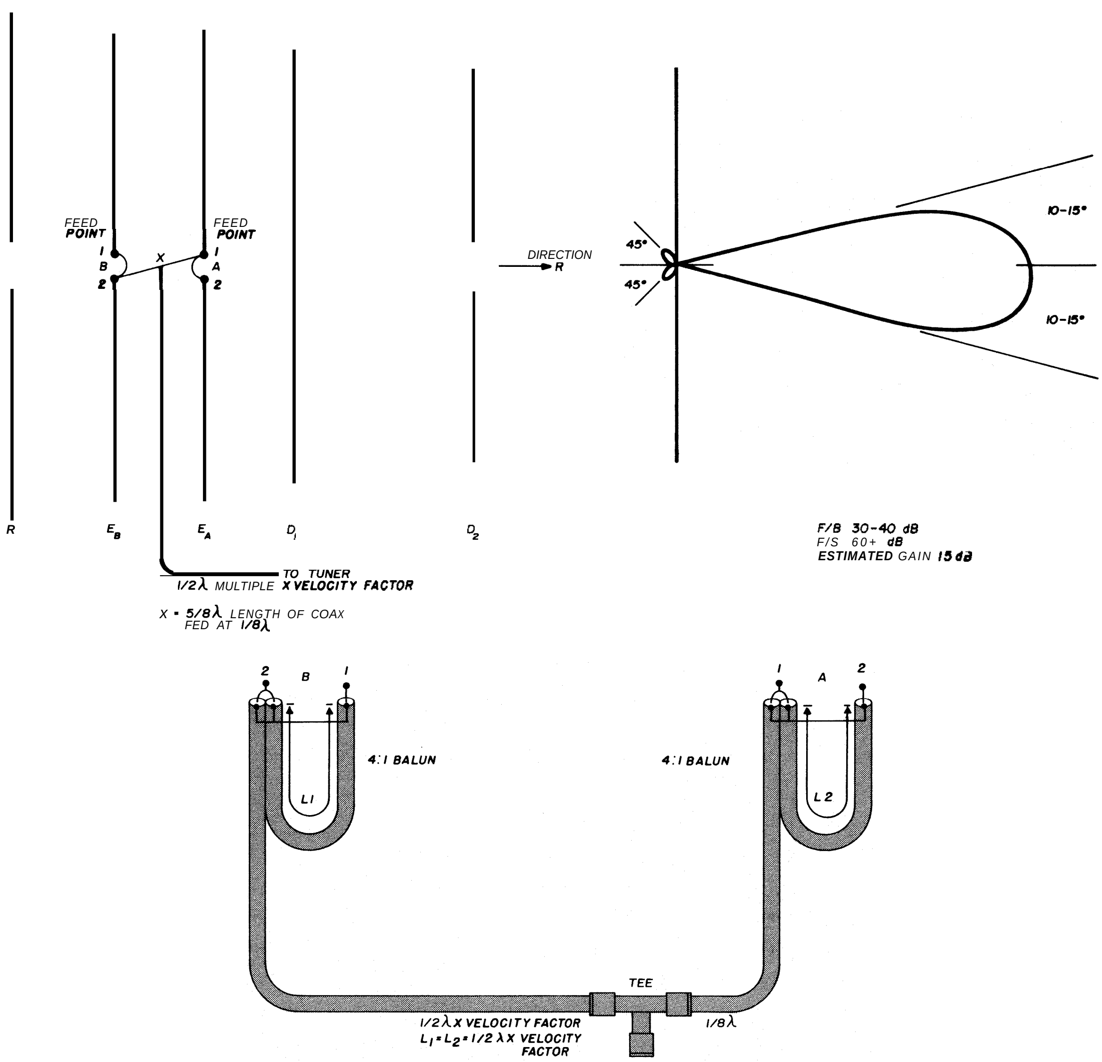
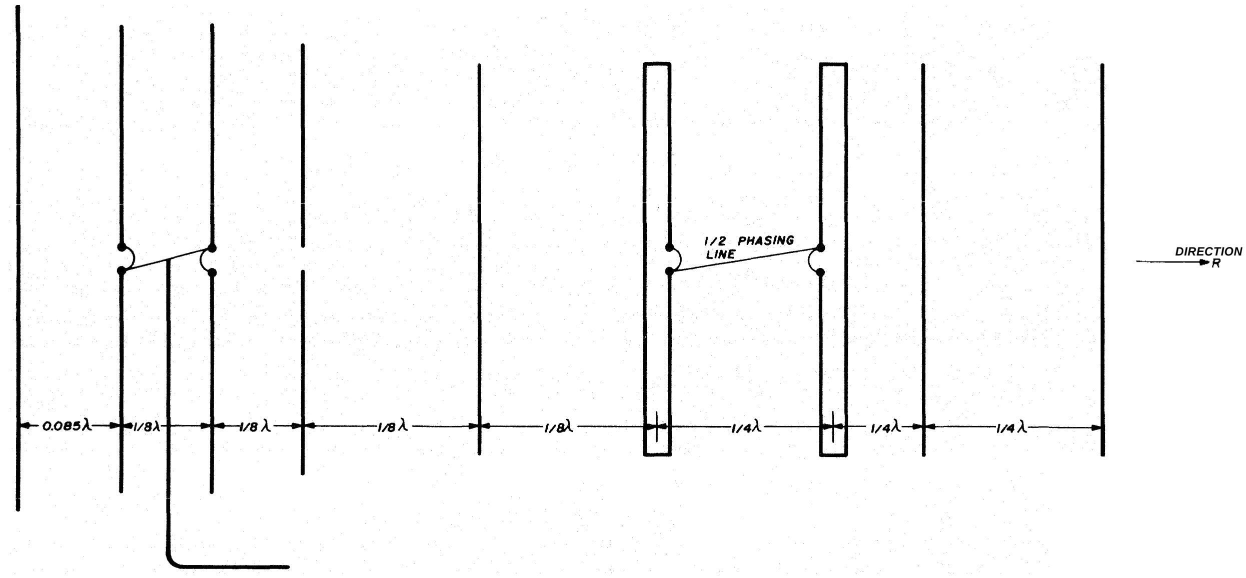


fig. 4. Adding a phasing harness between feedpoints B and A resulted in a large increase in response.



vide the correct phasing, so that waves emitted from E_B would coincide with those emitted from E_A and remain coincident with direction R. I added a 4:1 coaxial balun on each end of the phasing harness to maintain balance and permit polarity reversal (fig. 4).

Before installing the phasing harness I made a relative field strength reading using the arrangement in fig. 2. I then connected the phasing harness as in fig. 4. To my astonishment, the phasing harness produced a 1 - 1-1/2 S-unit increase in relative field strength, or an estimated 5-7 dB improvement in antenna gain. I tweaked the parasitic elements, and the pattern of fig. 4 was obtained. The length of reflector R is what would be expected in a normal Yagi. Directors D_1 and D_2 are shorter than the calculated value and are somewhat critical in adjustment.

On-the-air tests of the arrangement in fig. 4 suggested noticeable improvements — FIB was around 40 dB; F/S was 60 dB or more. Forward gain over a dipole, as suggested by the pattern, indicated 15 dB. Antenna bandwidth is approximately 500 kHz on 10 meters. The pattern is clean except for two small lobes off the back (when operated at ground level).

future experiments

Plans now are to modify a 20-meter beam by installing four halfwave elements in phase with parasitic elements and to explore the possibility of operating the parasitic directors as phased couplets, as shown in proposed test arrangement of fig. 5. Also, there seems to be no reason why the arrangement of fig. 4 can't be applied to quad antennas with the same results.

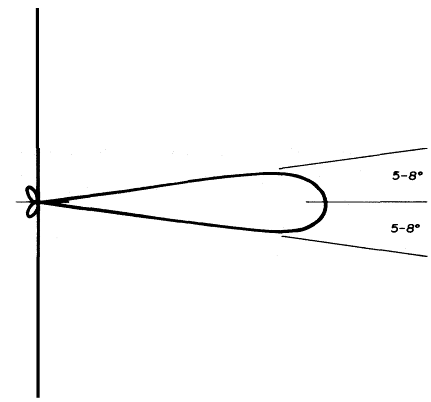


fig. 5. Future plans for four halfwave elements in phase with parasitic directors operating as phased couplets. Beamwidths shown at the half-power points.

acknowledgments

I wish to thank Dennis Timm and my daughter Lynnette, who helped with various measurements and adjustments during the antenna experiment. Also thanks to my wife, Shirley, who typed the manuscript.

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ham radio

10-meter ZL Special antenna for indoor use

Hampered by real-estate restrictions or a grouchy landlord? Try this full-size rotatable array that fits into an average room

To an Amateur, a huge antenna array high in the sky is beautiful. But more and more of us must live where outside antennas are frowned upon or forbidden. We're then faced with two choices: forego hf operation or use some sort of indoor antenna. I find the first choice unbearable. As for the second, antenna manufacturers, antenna handbooks, and Amateur-magazine articles offer little except loaded dipoles, loaded verticals, or random wires zigzagging like crazed snakes. Fortunately, there's no need to struggle with such inefficient antennas. Antenna theory applies equally well whether an antenna is located indoors or high on a mountain peak.

I live on the second (top) floor of a wood-frame, brick veneer apartment building. I operate from my bedroom using a full-size, rotatable, two-element, 10-meter, *indoor* ZL Special beam.[1] Just about anyone with a few simple hand tools and about $15 for materials can duplicate my antenna in a few hours.

why a ZL Special?

The ZL Special[1] is an all-driven array consisting of two unequal-length folded dipole elements spaced at 0.125 wavelength and fed 135 degrees out of phase. The array is unidirectional and has about 6 dBd forward gain and a good front-to-back ratio. The ZL Special is an old design and not currently in vogue. I suspect that its mechanical complexity, when constructed from aluminum tubing for outdoor use, and the fact that it's a single-band device have detracted from its popularity. My experimentation with it and several other antennas has uncovered a number of very good reasons for making the ZL Special the antenna of choice for indoor use..

The usual ways to squeeze an antenna with a 5-meter (16.5-foot) span into a 3.7 x 3.7 meter (12 × 12 foot) room, are a) to use loading devices, or b) bend the elements to fit the available space. Within reason, the ends of folded dipole elements may be bent without appreciable loss.

All antennas are adversely affected by unwanted coupling to nearby objects. Indoor antennas are generally close to house wiring, heating ducts, and other objects, and tend to couple energy to them. This problem is manifested by resonance shifts, drastic VSWR changes, and reduced gain and directivity. All-driven arrays, particularly those with folded dipole elements such as the ZL Special, are less troubled by unwanted coupling than are parasitic arrays. Quads are too large, even when reduced in size by loading, to be used indoors.

The ZL Special is a low-Q, hence wideband-response, array which is fairly tolerant of element length and spacing. Array impedance is about 70 ohms, compared with about 20 ohms for a parasitic Yagi. The ZL Special can be fed with 75-ohm coax using a simple sleeve-type bazooka balun.

A variation of the ZL array uses 0.375-wavelength elements, which reduces the antenna span by 25 per

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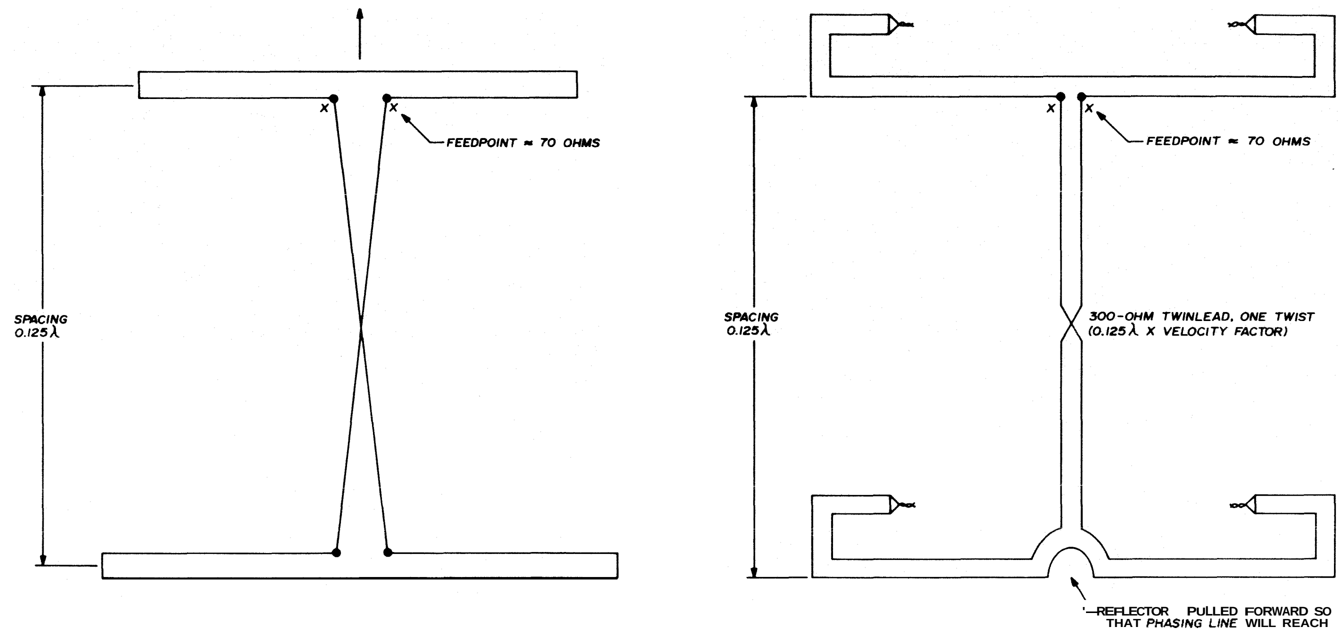


fig. 1. The ZL Special all-driven array is usually made of aluminum tubing and is cumbersome *(left)*. Use TV twinlead with the ends bent to fit your available space, and you have the indoor ZL Special *(right)*. The twinlead version has a phasing line shorter than the space between elements, and the rear element must be bowed forward slightly to connect the phasing line.

cent (described later). Spacing can be reduced to 0.1 wavelength. A 15-meter shortened ZL Special is about the same size as a standard 10-meter version and has only slightly less gain. A similarly reduced 20-meter array will fit a large 5.5 *x* 5.5 meter (18 *x* 18 foot) room or attic.

Attempts to use more than two elements, or to build multiband arrays indoors, are likely to be rewarded with less-than-hoped-for gain or with an array that performs poorly on several bands. I prefer one good antenna for my favorite band. I use temporary indoor dipoles for the other bands.

which design?

Dimensions for ZL Specials are many and varied.2 Element length depends on the type of construction. The vertical portions of the indoor ZL Special may act as capacitance hats and reduce the element length. These are dimensions I found to give the best results for a 10-meter array:

director length = *135/f meters (444/f feet)*
reflector length = *144/f meters (472/f feet)*
spacing = *37.2/f meters (121/f feet)*
phasing line = *spacing × 0.82 (velocity factor)*

where f is frequency (MHz).

Using a design frequency of 28.7 MHz gives a director length of 4.7 meters (15.5 feet), reflector length of 5 meters (16.5 feet), spacing of 1.3 meters (4.3 feet), and a phasing line of 1.1 meters (3.6 feet).

With no weather problems to contend with, the indoor ZL (fig. 1) uses lightweight wood for the boom and element supports. An earlier version used PVC pipe, which performed as well as the wood version, except that the PVC is flexible and it flopped when rotated and struck the walls and ceiling.

construction

Select unwarped, knot-free, pine 1 *x* 2 lumber and redwood furring strips. Cut to size (fig. 2) and give the wood pieces two coats of paint. (I chose to match the color of the walls and ceiling.) Hardware is steel angle brackets and mending (reinforcement) plates, which are inexpensive and available at any hardware store. The boom is attached to the TV mast with a U bolt.

I found it easier to build and make the initial beam assembly on a flat surface (living room floor), then reassemble it at the operating site. If you use machine screws to attach the hardware, you can assemble or disassemble the array in about ten minutes. (Quick disassembly avoids questions about the "clothesline" by guests – or the landlord!)

The idea is to have as much of the elements oriented horizontally as possible, so the framework dimensions are flexible. Figure on at least 7.5 cm (3 inches) wall-to-crossmember clearance. The boom length controls the spacing and should not be changed. The vertical end pieces can be reduced to as little as 30 cm (1 foot) to increase headroom, but the element length may have to be increased.

I built a simple, sturdy base from two 31 x 62 cm

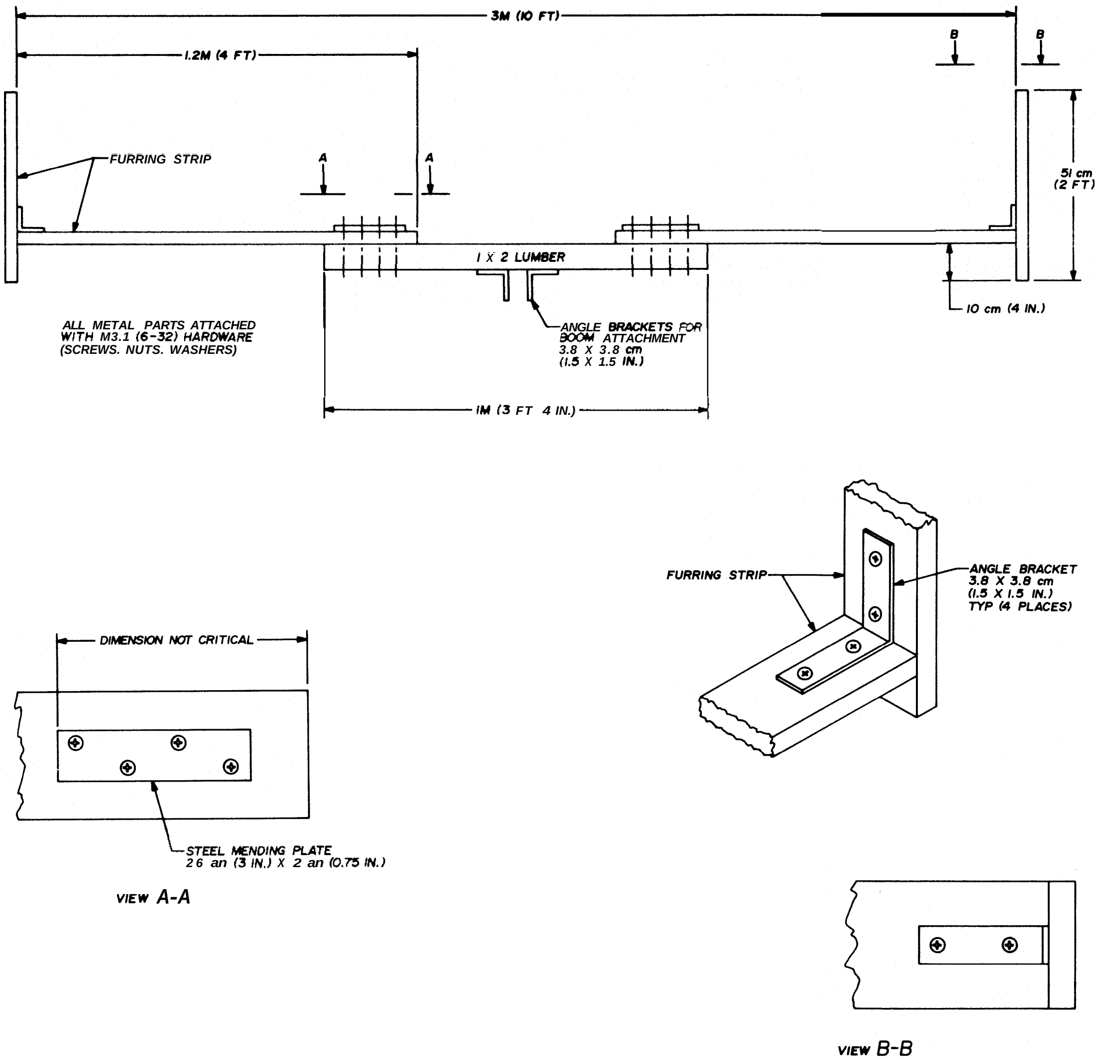


fig. 2. Crossmember construction. Two identical crossmembers are required. Pine 1 x 2 lumber and redwood furring strips provide a lightweight but strong structure. When the elements are in place, the structure is braced by the top guy wires and doesn't sag or flop around when rotated.

(1 × 2 foot) pieces of particle board (fig. 3). One board is flat on the floor and the other is vertical. A pair of TV mast clamps hold bearings (PVC pipe couplings) to allow mast rotation. A PVC pipe cap or an empty mayonnaise jar will serve as a bottom bearing. The base is steadied by clamping it to a solid object or by setting a couple of concrete blocks on the horizontal board.

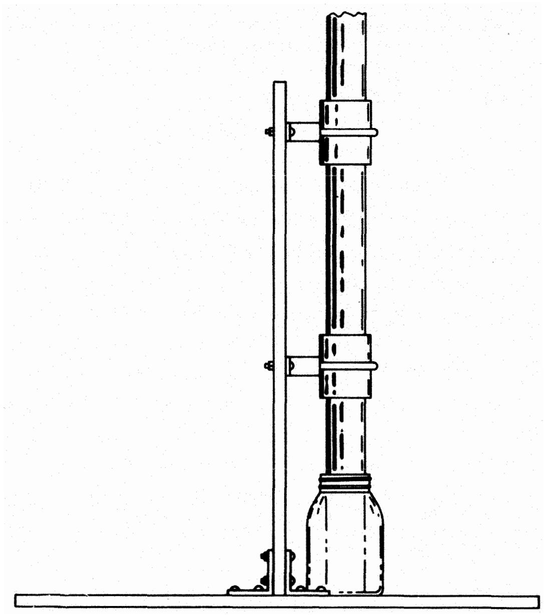
The elements are made from heavy-duty, foam-filled TV twinlead. The elements are looped around the vertical end pieces (taped in place) and strung across the top with wire or twine. Elements and top wires form a box at each end of the array. At moderate power, no additional insulation is required. PCB material with the foil removed makes excellent insulators. I used WA6TKT's method of bowing the reflector forward to fit the phasing line length.3

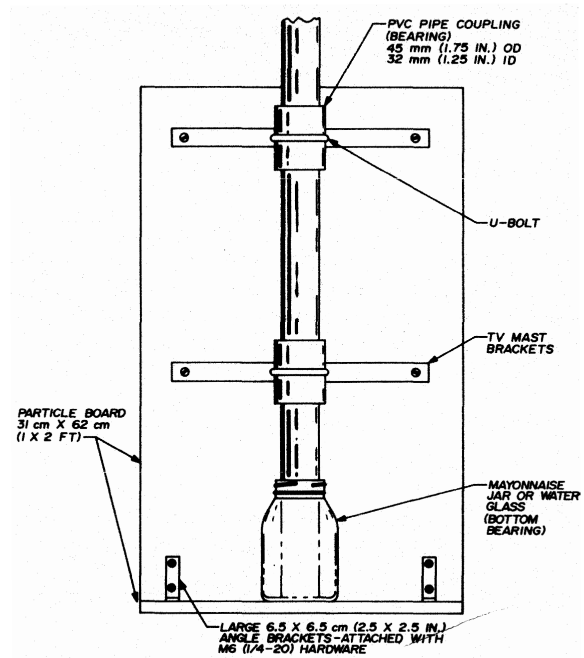
feeding the array

The feedline RG-59/U should be a multiple of 0.5 wavelength. A simple bazooka (sleeve balun) is formed by slipping a 0.25 wavelength piece of shield braid, removed from RG-81U or RG-11/U, over the antenna end of the feedline, fig. 4. The braid is soldered to the feedline shield at the 0.25-wavelength point, and no other connection is made to the braid. Wrap the braid with vinyl tape. The bazooka length is 2.6 meters (8.5 feet) for 10-meter operation.

performance

Performance data for my indoor ZL Special antenna appear in figs. 5 and 6. Gain, front-to-back ratio, and VSWR (fig. 5) are for a direction not affected by unwanted coupling. Proper performance of duplicate antennas is indicated by a VSWR minimum at the





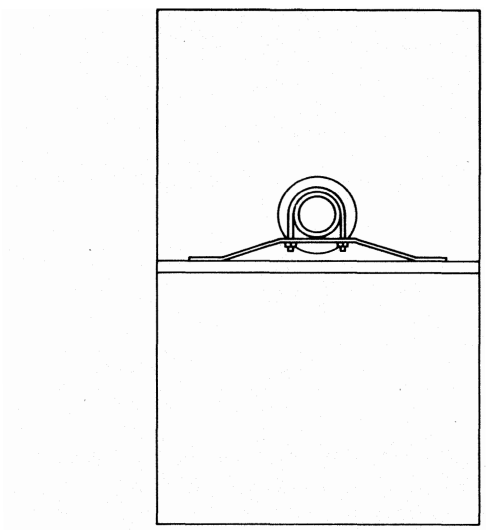


fig. 3. A simple but strudy base for the indoor array is made from two pieces of particle board or similar material and standard TV mast brackets. Total cost of the base is about $6.00 exclusive of the mayonnaise-jar bottom bearing.

design frequency (approximately 28.7 MHz). The front-to-back ratio is a function of phasing-line length, while gain is related to element spacing. Exhaustive pruning is probably not worth the effort once the array is working satisfactorily.

Fig. 6 shows VSWR variations of my array as it's rotated. The variation is caused by coupling to metallic objects in the room. A parasitic array (an early attempt) had a peak VSWR of 6:1 in the direction of the air-conditioning duct. Metal objects more than 0.25 wavelength distant don't cause resonance shifts or VSWR variations.

The difficulties of accurate measurement inherent to high-frequency arrays are compounded when the array is indoors. Repeated measurements using various techniques indicate a forward gain of about 6 dBd and a front-to-back ratio of 15-18 dB. The pattern is a cardioid and has a deep rear notch. VSWR is low over the entire 10-meter band and causes no problem for solid-state finals. I've checked the observed beam headings of hundreds of stations against a great-circle chart and found no significant errors. The only real difference between an indoor array and one outdoors is that the indoor array can be rotated in seconds rather than minutes.

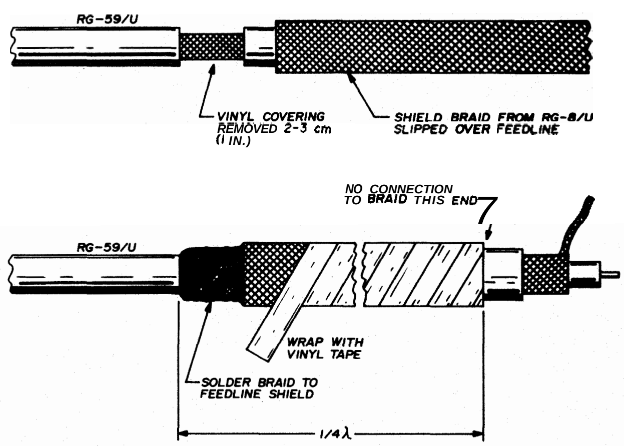


fig. 4. A simple bazooka balun: no impedance transformation, but it helps to eliminate feedline radiation. Aluminum foil could be used to wrap the coax, but the shield braid removed from larger coax makes a neater balun. Coax that's too lossy for a feedline is a good shield-braid source.

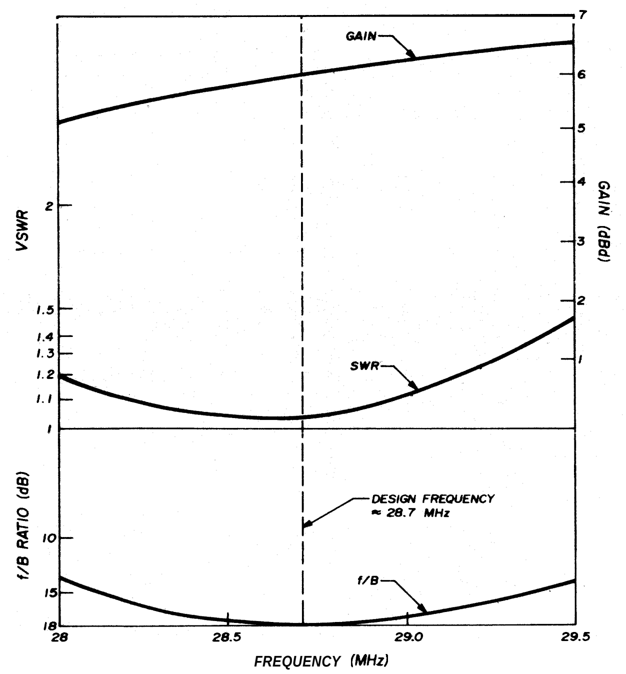


fig. 5. Indoor ZL Special performance data. The data are for a direction not affected by unwanted coupling. Design center is 28.7 MHz.

another version

A shortened version of the antenna is shown in **fig. 7.** It's a bit more complex, but the elements are 25 per cent shorter than in the design described above. It may be spaced 0.1 wavelength to decrease boom length but with some sacrifice in gain.

Element lengths are determined by the data given previously, multiplied by a factor of 0.75. For example, the director length for a design frequency of 28.7 MHz would be:

$$\textit{director length} = \frac{135}{28.7} \times 0.75$$
$$= 3.5 \textit{ meters (11.5 feet)}$$

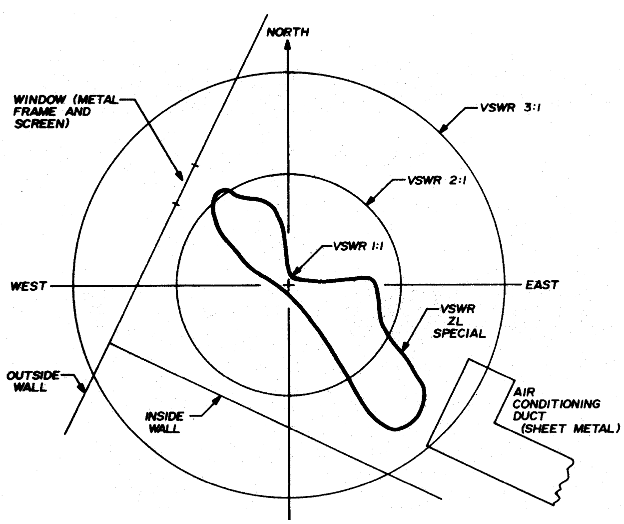


fig. 6. More performance data on the 28.7-MHz indoor array. Shown is VSWR variation as the array is rotated — the variation is caused by surrounding metallic objects.

The element ends may be bent up and around as shown in **fig. 1**B. My measurements of this shortened ZL Special were: gain approximately 5 dBd; front-to-back ratio about 12 dB.

concluding remarks

The indoor ZL Special is far and away the best indoor antenna i was abie to construct in two years of trying. With the indoor ZL, stateside contacts are 3 to 4 S-units better than my next best indoor antenna, phased groundplane verticals. During a recent 48-hour contest period, I worked 21 different countries on all continents using a barefoot Ten-Tec Argonaut, 3 watts PEP output!

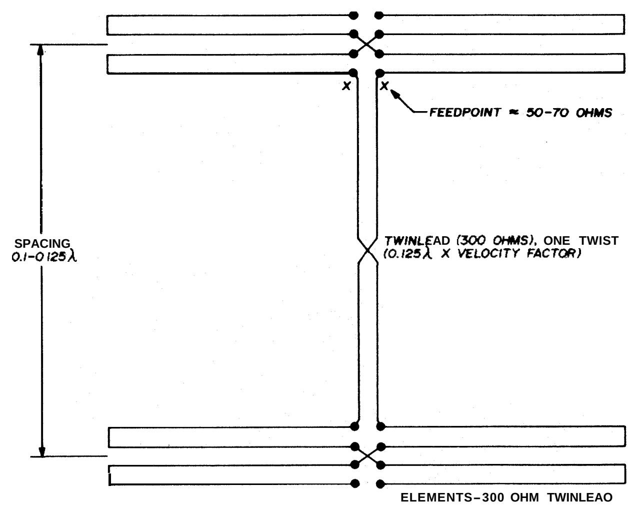


fig. 7. A shortened ZL Special. Spacing may be spaced 0.1 wavelength to decrease boom length. This small antenna provides about 4-5 dBd gain.

I hope nobody is naive enough to think that the performance of an antenna can be improved by placing it indoors; but neither should it be believed that indoor antennas cannot work. The multitude of variables make it impossible to predict how well an indoor array will perform at your location, but the indoor ZL Special is sure worth a try.

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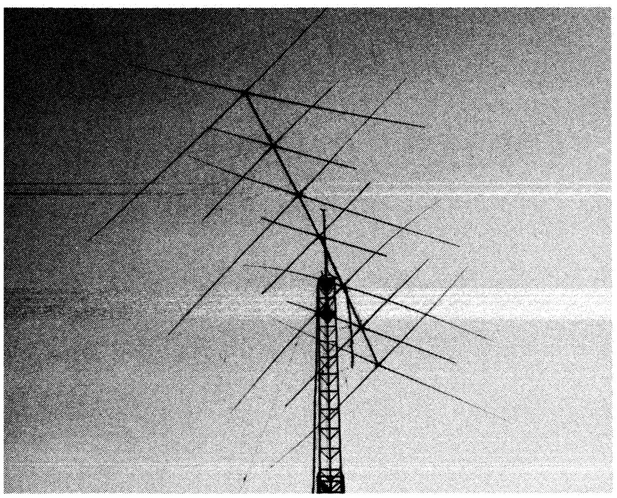

big quad — small yard

Here's how one Amateur erected a three-band quad antenna on a modest-size residential lot

Let me begin by saying that there are a few time-honored maxims about Amateur antennas. First, the result will be in direct proportion to the amount of time, effort, and money expended. Second, there's no such thing as too high or too big. Third, if the antenna stays up through the winter, it's not big enough!

With this philosophy in mind, I decided to build a big quad antenna with four elements on the Amateur 15- and 20-meter bands and seven elements on the 10-meter band, all on an 8.5-meter (28-foot) boom. The entire array was to be installed in a modest California backyard on a tilt-over tower.

The following is a photo essay on the project. I hope it will inspire others with a small yard to bite the bullet and put up a big antenna.

background

Last year I acquired a 21-meter (70-foot) guyed crank-up tower. With my mast it would allow the antenna boom to be nearly 23 meters (75 feet) up. Also it looked strong enough to handle something much larger than my old cubical quad on a 2.7-meter (9-foot) boom. I brooded, plotted, and schemed all winter, which, in southern California, is marked by windstorms up to 112 km (70 miles) per hour for a four-month period. I'd built a four-element quad for 15 and 20 meters the previous year and it worked great, but the place where the two boom lengths were joined together was weak and bent in the wind; thus the end of the four-element quad.

For six years I've been constructing and erecting quads in my back yard. And for six years, trees and bushes have been growing, cutting down the available space. Even so, I decided to build a quad with four elements on 15 and 20 meters and seven elements on 10 meters — all on an 8.5-meter (28-foot) boom.

This arrangement would give 0.13-wavelength spacing on 10 and 20 meters and 0.20-wavelength spacing on 15 meters. My experience indicated that these were acceptable spacings for the bandwidth and gain I was seeking. The antenna for each band would be fed directly through a gamma match, and all three coax cables would be run down to the station. I decided to use two reflector elements on 10 meters. (This should improve the front-to-back ratio

By George McCarthy, W6SUN, 2739 North Atherwood Avenue, Simi, California 93065

fig. 1. View from author's patio roof showing the work area.

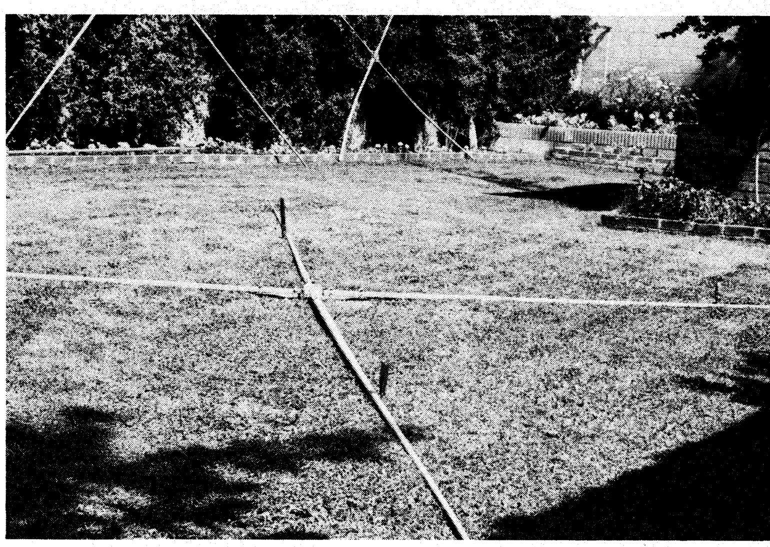

fig. 2. Method of assembling quad spreaders and elements.

fig. 3. Tower tilted over with mast and rotator installed and quad spreaders all over the place.

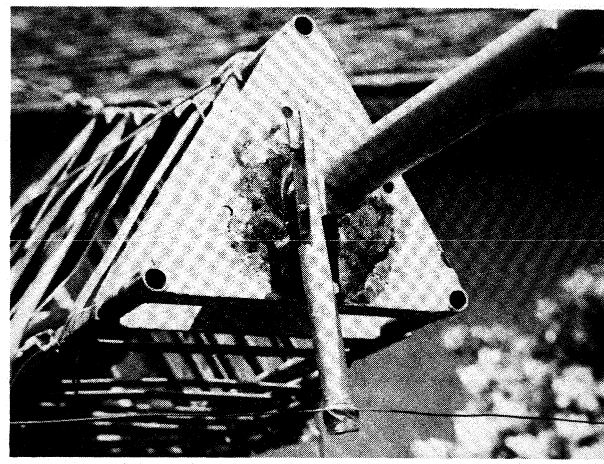

fig. 4. A length of PVC pipe, bolted to the mast with a muffler clamp. Ten-meter director wire is taped to the other end.

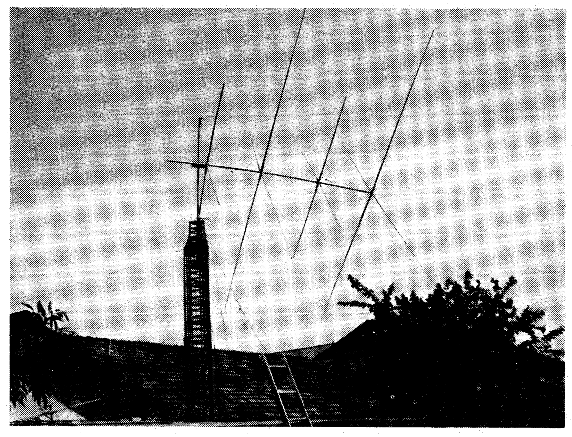

fig. 5. The "big quad — small yard" secret. Tower is raised and array is turned around.

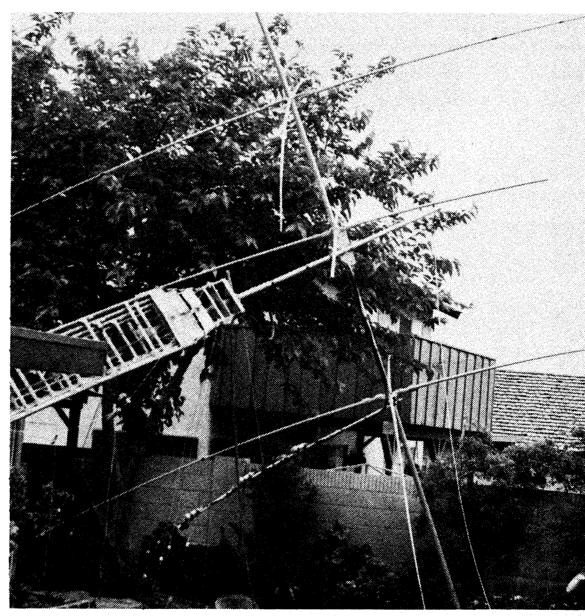

fig. 6. Assembly of the other end of the antenna. Driven element is in place; the 10-meter reflector is next. The second reflector is installed last.

a bit and allow me to string all three driven elements on the same spreaders, which was convenient.)

putting it up

The photos pretty much tell the story of how I managed to erect such a monster in a small area. I'll now relate some of the sidelights of the job.

I did all the work myself with one exception: the tower must be pulled over by someone else until gravity takes over and I can let it down with a winch. My thanks to Bob, W6TSH, and Joe, W6UL, who came over to help with this task.

A boat winch is the secret of my solo performance. The aircraft cable from the winch runs over the peak of the roof to the tower. That's right — I use the peak of the house as a gin pole! I figure that if anything goes it'll be the house. I raise and lower the tower from this position, which entails a lot of stopping and running around to make sure the wire isn't hung up on something.

From long experience, I no longer rely on predrilled holes in the fiberglass spreaders — they wear with wire movement. I stuck screwdrivers into the lawn to keep the spreader arms from distorting the perfect square. Then I ran the wires around the screwdrivers, moving the screwdrivers until I had a nice, tight loop that wouldn't pull the spreaders. I then taped the wire to the spreaders with PVC tape and applied PVC cement to keep the tape secure.

The short boom section connected to the boom-to-mast adapter is 1.4 meters (4-112 feet) long. It has a 4.8-mm (3116-inch) wall, which I turned down slightly for a distance of 152 mm (6 inches) at each end. The aluminum booms, which are 3.7 meters (12 feet) long and 51 cm (2 inches) OD, with 1.7-mm (0.065-inch) wall thickness, were slipped over the ends and bolted.

final comments

Putting up an array of this size in a small working area requires a lot of patience and the desire to succeed. My residential lot is only 20 meters (66 feet) wide by 31 meters (100 feet) long. Anyone who has attempted such a task will tell you that quad wires actually look for something in which to get tangled. Also, no matter how carefully you walk across a lawn full of strung spreaders you'll inevitably trip on some of the wires. Finally, my roof is made of thick shingles. No self-respecting wire would dare pass one up without trying to slide underneath.

I'm now in the process of tuning this monster, so I have no final words on performance. But it had better be great! Once tuned up, it'll be elevated to 22.3 meters (73 feet), where I hope to terrorize the ham bands — at least until the first big windstorm.

ham radio

More Details? CHECK—OFF Page 94

earth anchors

for guyed towers

Design data
for earth anchors
and how to make
a deadman anchor
for your tower guys

Often the design of a guyed tower fails to include the anchor design. This article describes construction and implementation of guy anchors.

holding power

The holding power of an anchor depends on anchor size and its depth in the ground. The weight of the earth above the anchor provides the holding force. Standard practice is to figure the holding force of earth in a cone shape above the anchor with a slope of 30 degrees relative to the vertical, as shown in **fig. 1.**

Note that, if the actual anchor were cone shaped, the total cone size would depend on depth, ***D.*** The cone volume, V, is:

$$V = 1/3\,\pi R^2 D \qquad (1)$$

Radius, ***R,*** is proportional to depth, ***D,*** where

$$R = D \tan 30 \text{ degrees} = D(0.577)$$

Therefore:

$$V = 1/3\,\pi (0.577D)^2 D \cong 0.35D^3$$

If the weight, W, of earth is assumed to be 100 pounds/foot^3 (45 kg/0.3 meter^3), and the holding force, F, = VW, then $F = 35D^3$. Holding force versus anchor depth is shown in **fig. 2.**

If you use 114-inch (6.5-mm) diameter steel cable for guy wire, the breaking strength is 5480 pounds (2488 kg). An anchor for this cable buried 5-112 feet (1.7 meters) would be appropriate.

table 1. Holding force at depths of 1-6 feet (0.3-1.8 meters).

depth		holding force	
feet	(meters)	pounds	(kg)
0		0	
1	(0.3)	35	(16)
2	(0.6)	280	(127)
3	(0.9)	945	(429)
4	(1.2)	2240	(1017)
5	(1.5)	4375	(1986)
6	(1.8)	7560	(3432)

By Ted Hart, W5QJR, Box 334, Melbourne, Florida 32901

deadman anchors

Screw anchors, available from lumber yards and marine-supply houses, have been used for antenna

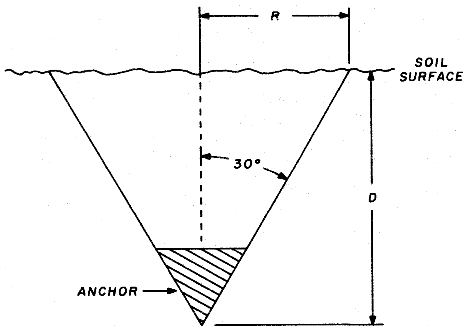

fig. 1. The holding power of an anchor depends on anchor size and depth in the ground. Weight of earth above the anchor provides the holding force. Standard practice for such anchors is to figure the holding force of the earth in a cone with a slope of 30 degrees.

guys but they're expensive. So I decided to make a deadman anchor from concrete. Here's how to do it:

1. Make a loop of your guy wire and tie it with a cable clamp.

2. Pass two pieces of concrete reinforcing rod, 5 inches (128 mm) long through the loop and tie the rods at right angles. Use heavy wire to make the connection.

3. Use a container, such as a plastic pail, to pot the assembly in concrete.

Holes for the deadman anchors can be made with a post-hole digger. Make sure the dirt is moist. Tamp

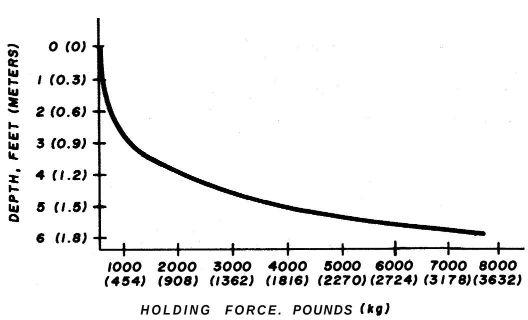

fig. 2. A plot of holding force as a function of depth for anchors buried in earth.

the dirt and wait a few days for the earth to settle. You'll have to tighten the guy wires several times during the first few days after installation while the cable aligns itself through the earth.

ham radio

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high-performance grounded-grid 220-MHz kilowatt linear

The Eimac 8877 is a high-mu ceramic-metal triode rated for use up to 250-MHz and several successful amplifier designs using this tube have been constructed for hf through vhf.[1,2,3] The 220-MHz amplifier described here has proven to operate very well during the last year, including several successful Earth-Moon-Earth (EME) contacts.

This 220-MHz 8877 linear amplifier is designed for the serious vhf DXer who demands reliable service combined with good linearity and efficiency. The amplifier requires no neutralization, is completely stable and free of parasitics, and is very easy to operate.

The amplifier is designed for continuous duty operation at the 1000-watt dc input level, and can develop 2000-watts PEP input for SSB operation with ample reserve. For operation at 2000-watts PEP the plate supply should be between 2500 and 3000 volts; under these conditions the amplifier will deliver 1230 watts output. With the higher plate-voltage supply, up to 14-dB gain can be obtained with an amplifier efficiency of 61 per cent; see **table 1**.

The 8877 triode has very good current division; that is, the grid current is quite low in comparison to the plate current. The grid current is typically about 15 per cent of the value of the plate current. the 8877 also has good gain and intermodulation distortion characteristics. The plate dissipation rating is 1500-watts. The cathode is indirectly heated; filament requirements are 5.0-volts at 10.5 amperes. The tube base mates with a standard septar socket.

the circuit

In the amplifier circuit shown in **fig. 1** the 8877 grid is operated at dc ground. The grid ring at the base of the tube provides a low-inductance path between the grid element and the chassis. The plate and grid currents are measured in the cathode return lead. A 12-volt, 50-watt zener diode in series with the negative return sets the desired value of idling current. Two additional diodes are shunted across the meter circuit to protect the instruments in case plate voltage arcs over to ground, or if there is an internal tube arc.

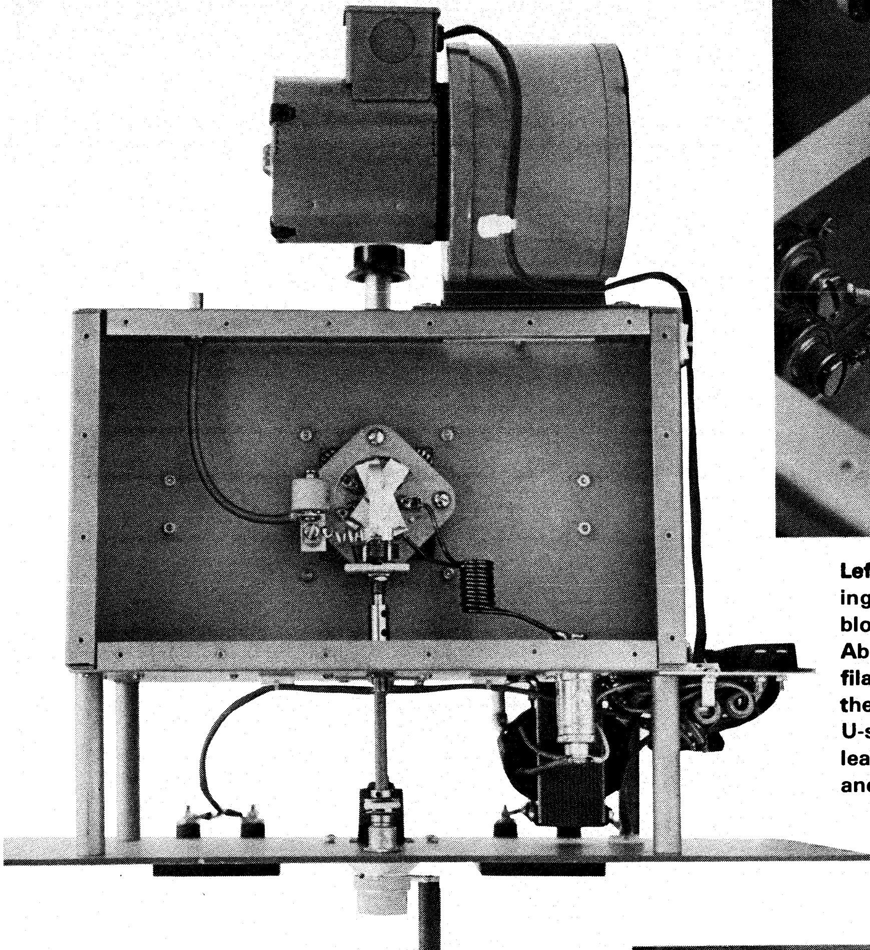
Standby plate current of the 8877 is reduced to a very low value by a 10,000-ohm cathode resistor. This resistor is shorted out in the transmit mode by the station control circuit. The resistor must be in the cathode circuit when receiving to eliminate the noise generated in the station receiver if electron flow is permitted within the 8877 tube.

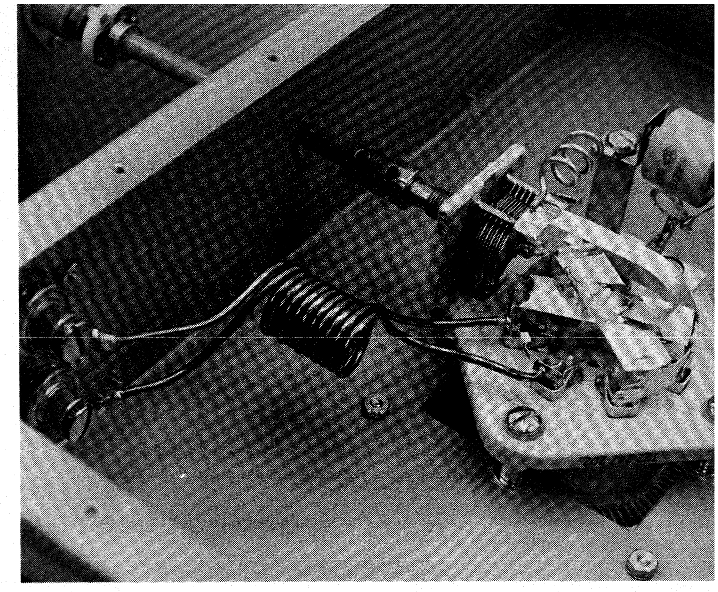
A 200-ohm safety resistor insures that the negative side of the power supply does not go below ground potential by an amount equal to the plate voltage if the positive side is accidentally grounded. A second safety resistor across the 1N3311 zener diode prevents the cathode potential from rising if the zener should accidentally burn open.

input circuit

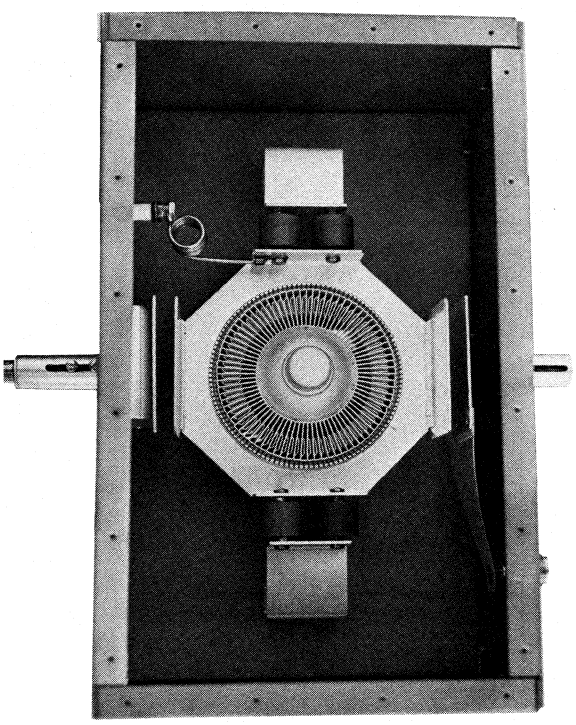
The cathode matching circuit is a T-network which transforms the input impedance of the tube (about 54 ohms in parallel with 40 pF) to 50 ohms at the coaxial input connector; the network consists of two series inductors and a shunt variable capacitor. The inductors are fixed and have a very low value of inductance; in fact, the rf return path through the chassis has about the same inductance value. To design the input circuit, many values of circuit Q were tried in the calculations. When the design equations yielded physically realizeable inductance values, then several combinations were tried in the actual amplifier. Since the stray inductances in the chassis and connecting leads in the socket were not included in the calculations, the final inductors were smaller in value than the calculated size. The actual inductors which resonated and provided a reasonable input match are specified in fig. 1 and are shown in some of the photographs. For those who build this amplifier I would expect that some minor variations in these coils might be required to attain an adequate input match.

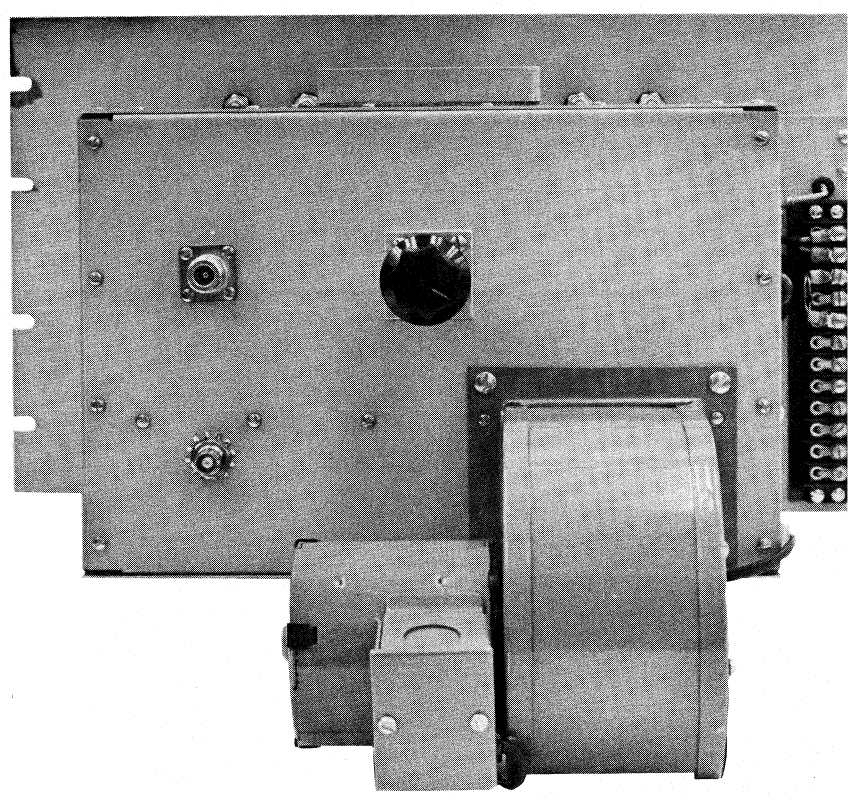
By Robert I. Sutherland, W6PO, EIMAC Division of Varian, 301 Industrial Way, San Carlos, California 94070





Left, underside view of the 220-MHz power amplifier showing the blower location as well as the input circuit. The blower is a Dayton 4C446.
Above, close-up view of the input circuit shows the bifilar filament choke L3 and L4 and the matching network. C2 is the 35-pF air variable mounted on the 8877 socket. L2 is the U-shaped strap connecting the capacitor to the cathode leads. L1 is the coil going between the variable capacitor and input line blocking capacitor C1.





Left, top view of the amplifier plate compartment. The 8877 tube is in the center with L5 and L6 to the left and right. The plate tuning capacitor C5 is at the bottom and the loading capacitor C6 is at the top.
Right, back view of the amplifier. The type-N connector is the rf power output; the BNC fitting is the connection for drive power. The knob is the loading adjustment. The terminal strip to the right is for the input voltage and control circuit connections. A **Millen** high-voltage connector is used for the plate voltage.

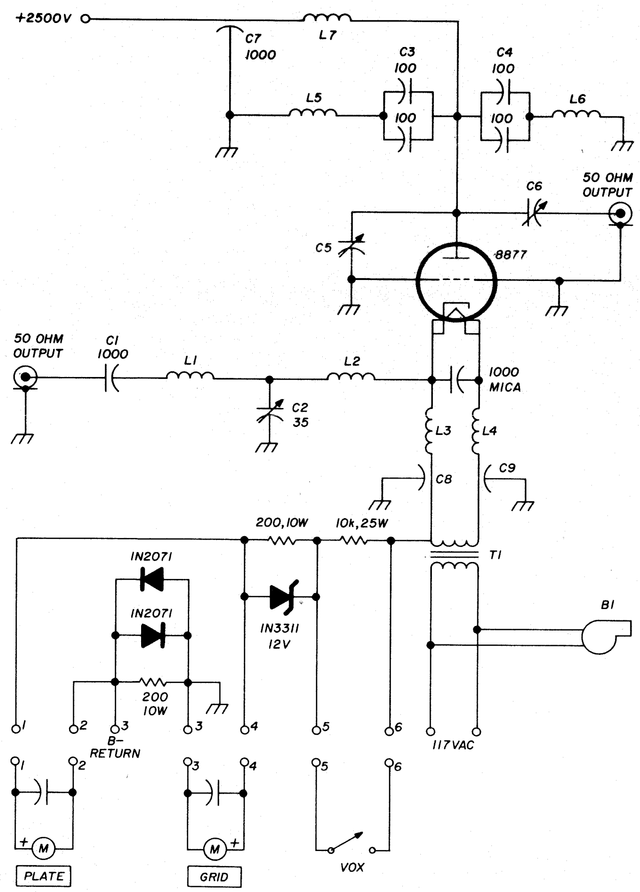


fig. 1. Schematic of the grounded-grid 220-MHz triode amplifier. Operating bias for the 8877 is supplied by a 12-volt zener diode in the cathode lead.

table 1. Performance of the 220-MHz grounded-grid 8877 rf power amplifier.

Plate voltage	3000 V	2500 V	2500 V
Plate current (single tone)	667 mA	800 mA	400 mA
Plate current (idling)	54 mA	44 mA	44 mA
Grid voltage	− 12 V	− 12 V	− 12 V
Grid current (single tone)	48 mA	50 mA	29 mA
Power input	2000 W	2000 W	1000 W
Power output	1230 W	1225 W	621 W
Efficiency (apparent)	61%	61%	62%
Drive power	48 W	69 W	20 W
Power gain	14 dB	12.4 dB	15 dB

C1	1000 pF ceramic transmitting type (Centralab 858S-1000)
C2	35 pF air variable (Hammarlund HF35 or Millen 22035)
C3,C4	Each consists of two parallel connected 100 pF, 5000 volt ceramic transmitting capacitors (Centralab 850S-100)
C5	Plate tuning capacitor (see **fig. 2**)
C6	Output loading capacitor (see **fig. 7**)
C7	1000 pF, 4000 volt feedthrough (Erie 2498)
C8,C9	0.1 uF, 600 volt feedthrough capacitor (Sprague 80P3)
L1	3 turns no. 14 (1.6 mm) wire, 1/4 inch (6.5 mm) inside diameter, 5/8 inch (16 mm) long
L2	Copper strap 1/4 inch (6.5 mm) wide, 2-1/2 inches (64 mm) long, bent into a U 5/8 inch (16 mm) wide
L3,L4	7 bifilar turns no. 12 (2 mm) enamelled wire, bifilar wound on 1/2 inch (12 mm) inside diameter
L5,L6	Plate resonators (see **fig. 5**)
L7	6 turns no. 14 (1.6 mm) wire, 1/2 inch (12 mm) diameter, 1 inch (25 mm) long
T1	Filament transforrmer rated at 5 volts, 10 amps (Stacor P-6433)

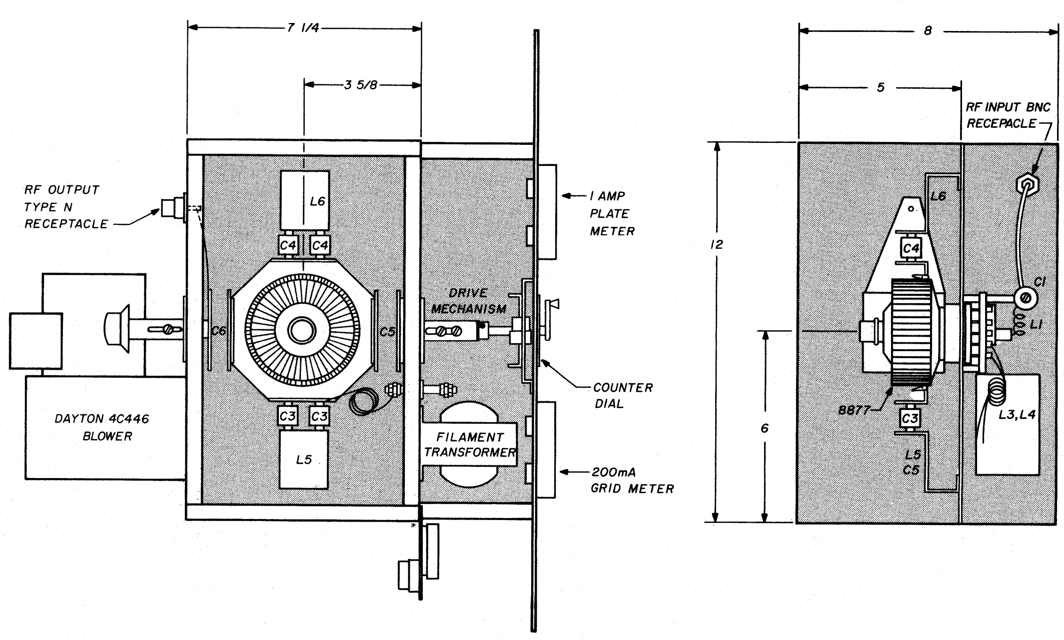


fig. 2. Structural details of the amplifier showing relative size and position of the various components. Assembly is made of aluminum panels.

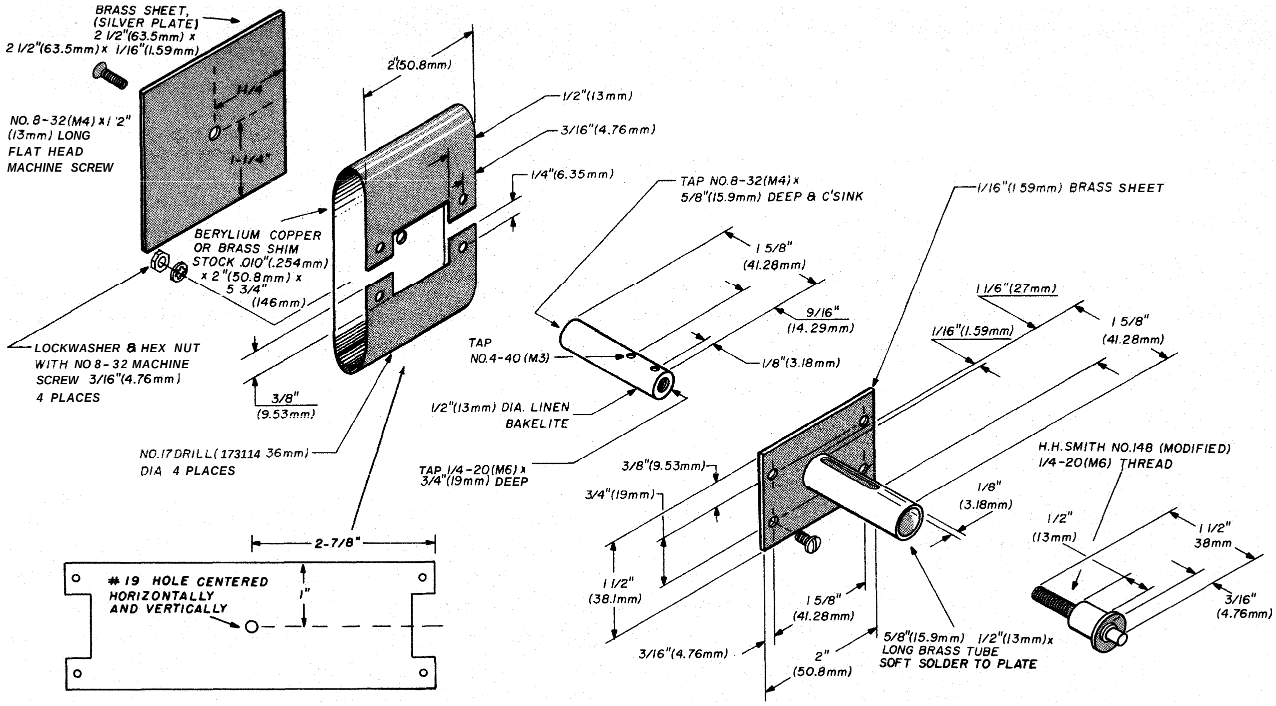


fig. 3. Variable plate portion of plate-tuning capacitor C5. Since there are no moving or sliding contacts which carry heavy rf current, this arrangement permits the capacitor to beadjusted under full power without erratic tuning.

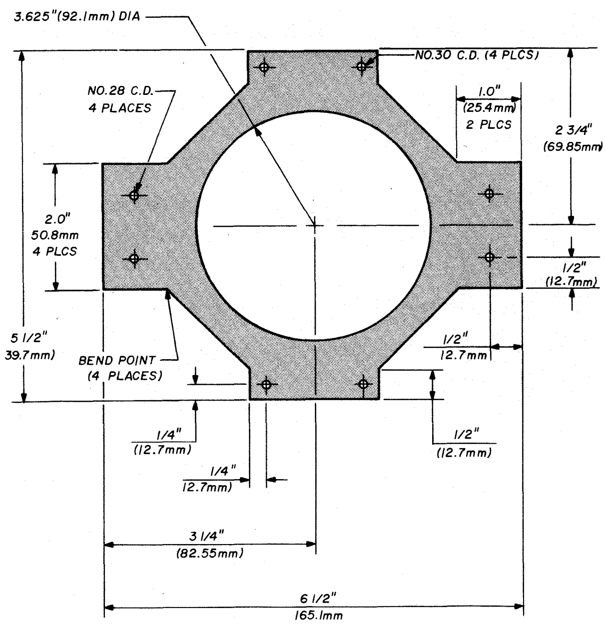
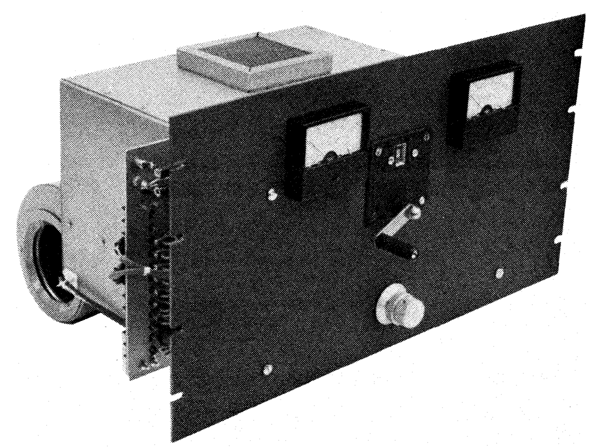


fig. 4. Anode collet and capacitor plate support pattern.

The underchassis layout of components is shown in the photographs. In the close-up view the bifilar wound coil in the foreground is the filament choke. The variable capacitor is C2, and L2 is the U-shaped strap connecting C2 with the cathode terminal. All the cathode leads and one filament lead are connected together with low inductance copper straps. Note that L2 is connected to the center point of all the cathode leads in an effort to equally balance rf drive to all sides of the cathode. At the frequency of 220-MHz, lead length and residual inductance are very important.

The inductor L1 connects capacitor C2 with the input blocking capacitor C1 at the top of an insulating piller. A section of RG-142B/U teflon-insulated coax connects the other side of C1 to the BNC coax input connector. It is difficult to see in the picture, but there is a 1000-pF chip ceramic capacitor connected from one heater pin to the other on the socket.

The socket for the 8877 is the Eimac SK-2210, the version with the grounded grid clips. The filament transformer is located between the aluminum enclosure and the panel. The filament voltage is fed



through the enclosure wall using 0.1 μF Sprague *Hy-Pass* feedthrough capacitors.

plate circuit

The plate circuit of the amplifier is a transmission-line type resonator. The line (L5 plus L6) is one half-wavelength long with the tube placed at the center. This type of circuit is actually two quarter-wavelength lines in parallel. One of the advantages is that each of the quarter-wavelength lines is physically longer than if only one is used. This is because only half of the tube output capacitance loads each quarter-wavelength section. Another advantage to this layout is a better distribution of rf currents around the tube seals.

The dc blocking capacitors are surplus Centralab 100-pF, 5000-volt ceramic capacitors. Two are used on each line to handle the rf current. The homemade variable capacitor C5 tunes the plate circuit. Note that this type of capacitor structure has no wiping contacts. All the rf currents flow through a fixed path which provides very smooth tuning with no jumping meter readings. The load capacitor C6 is constructed in a similar manner.

The plate choke L7 is visible in the photograph of the plate compartment. It is connected to the plate collet assembly with the Erie high voltage feed-through capacitor C7.

construction

The 220-MHz power amplifier is built in an enclosure measuring 8 x 12 x 7-1/4 inches (20 x 30 × 18 cm). The 8877 socket is centered on an aluminum deck 5 inches (12.7 cm) from the top of the enclosure. A centrifugal blower* forces cooling air into the under chassis area; the air escapes through the air-system socket, the teflon chimney (SK-2216), and then the tube. The warm air is exhausted through a "waveguide beyond cut-off" air outlet. This is an assembly which has expanded metal about 1/2 inch (12 mm) thick, mounted in a frame. A perforated aluminum cover may suffice in most cases, although restricts air flow slightly more and is not a very good rf shield at 220 MHz.

The plate tuning mechanism is shown in **fig.** 3. This simple apparatus will operate with any variable plate capacitor, providing a back-and-forth movement of about one-half inch. It is driven by a counter dial and provides a quick, inexpensive, and easy means of driving a vhf capacitor. The ground return path for the grounded capacitor plate is through a wide, low inductance beryllium-copper or brass shim stock which provides spring tension for the drive mechanism.

The variable output coupling capacitor is located at the side of the 8877 anode. The type-N coaxial

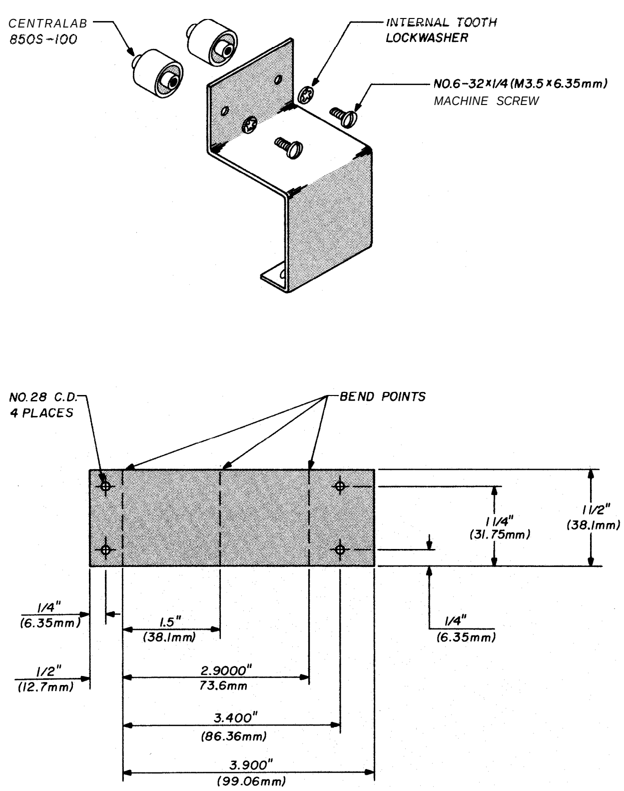


fig. 5. Plate line inductor pattern and bending layout for L5 and L6. Two assemblies are needed for the plate circuit.

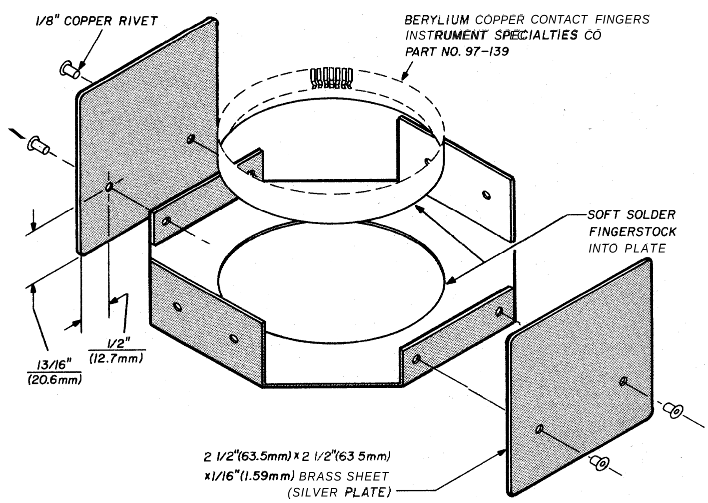


fig. 6. Anode collet and capacitor plate support assembly. The two fixed capacitor plates for C5 and C6 are mounted to the assembly using copper pop-rivets and then soldered. The two remaining bent-up edges are for mounting the blocking capacitors C3 and C4. The finger-stock is soft-soldered into the large hole in the center. A tight fitting aluminum disc helps to hold the finger stock in place while soft soldering with a hot plate.

*Recommended blower is the Dayton 4C446, a 115-Vac unit rated to deliver cooling air at 135 cubic feet per minute (3.8 cubic meters) with a static pressure equivalent to 0.2 inch (5 mm) of water.

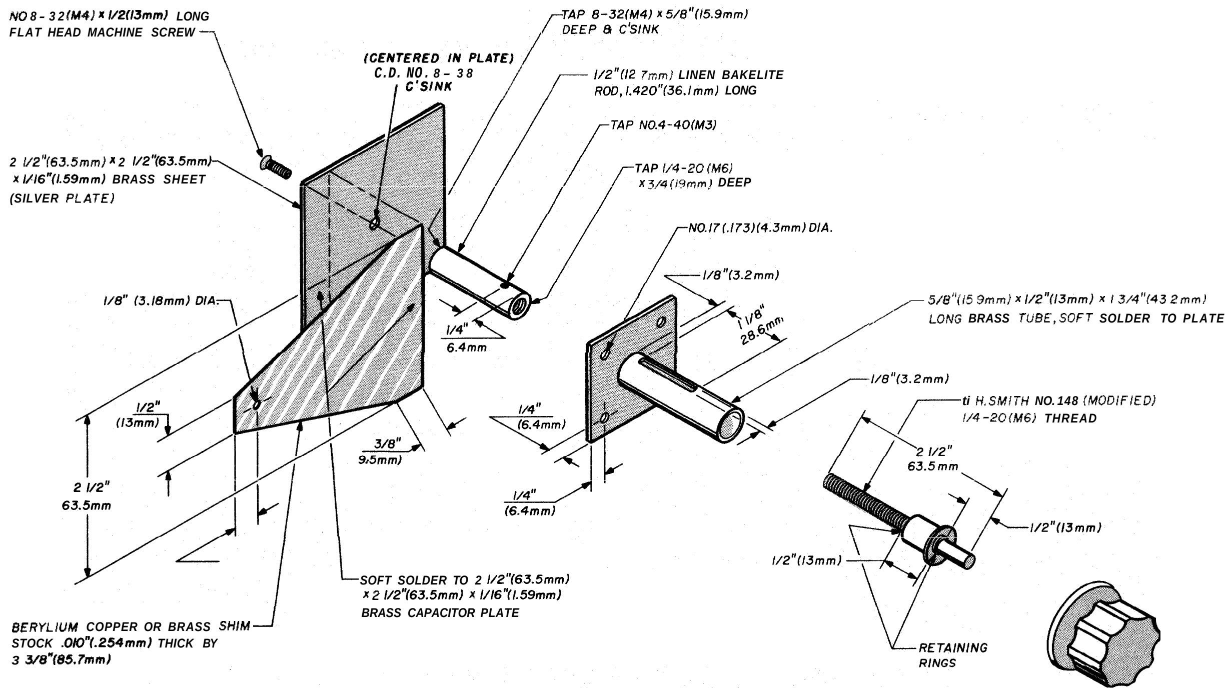


fig. 7. Variable plate portion of the loading capacitor C6. The beryllium-copper portion carries the rf current to the type-N coaxial connector as well as providing spring tension on the tuning mechanism. Because of the constant rf conducting path, the loading is very smooth with no jumpiness.

output connector is connected to the moveable capacitor plate by a wide beryllium-copper strap. The capacitor plate is driven in a manner similar to the tuning capacitor as shown in **fig. 7.**

The plate line is made up of two inductors L5 and L6 (see **fig. 5)** and the anode collet and capacitor assembly shown in **fig. 6.** With the inductor sizes given, the amplifier can be tuned from 220 to 222.5-MHz; no tests were run above 222.5-MHz.

The plate rf choke is mounted between the junction of the anode collet and a pair of the dual blocking capacitors. The high-voltage feedthrough capacitor is mounted on the front wall of the plate compartment. The blocking capacitors are rated for rf service, and inexpensive television-type capacitors are not recommended for this amplifier.

operation

Amplifier operation is completely stable with no parasitics. The unit tunes up exactly as if it were on the hf bands. As with all grounded-grip amplifiers, excitation should never be applied unless the plate voltage is on the amplifier.

The first step is to grid-dip the input and output circuits to near-resonance with the 8877 in the socket. An SWR meter should also be placed in series with the input line so the input network may be adjusted for lowest SWR.

Tuning and loading follows the same sequence as any standard grounded-grip amplifier. Connect an SWR indicator at the output and apply a small amount of rf drive. Quickly tune the plate circuit to resonance; the cathode circuit should now be resonated. The SWR between the exciter and the amplifier will not necessarily be optimum. Final adjustment of the cathode circuit for minimum SWR should be done at full power because the input impedance of a cathode-driven amplifier is a function of the plate current of the tube.

Increase the rf drive in small increments along with the output coupling until the desired power level is reached. By adjusting the drive and loading together it will be possible to attain the operating conditions given in the performance chart in **table 1.** Always tune for maximum plate efficiency: maximum output power combined with minimum input power. It is easy to load heavily and underdrive to get the desired power input but power output will be reduced if this is done.

references

1. R. Sutherland, WGUOV, "Two Kilowatt Linear Amplifier for Six Meters," *ham radio*, February, 1971, page 16.
2. R. Sutherland, WGUOV. "High Performance 144-MHz Power Amplifier," *ham radio*, August, 1971, page 22.
3. M. Partin, KGDC, "Custom Design and Construction Techniques for Linear Amplifiers," *QST*, September, 1971, page 24.

ham radio

Woodpecker noise blanker

The Russian over-the-horizon radar has been causing interference on the high-frequency bands — here's a noise blanker that helps

Anyone who operates regularly on the high-frequency Amateur bands has probably run into interference from the Russian over-the-horizon radar which operates between 10 and 30 MHz; because of its peculiar sound, it is popularly known as the "Russian Woodpecker." The noise-blanker circuit shown in **fig. 1** was designed especially by M. Martin of the Hahn-Meitner Institute in West Berlin to blank the Woodpecker noise pulses;[1] this unit is also suitable for blanking out the Loran pulses that plague long-distance communications on the Amateur 160-meter band.

Although the circuit of **fig. 1** was built for a 9-MHz i-f, it should be relatively easy to adapt the circuit to other i-f systems. The circuit requires only two integrated circuits and six transistors; it has a blanking range of about 80 dB and does not degrade the receiver's dynamic range.

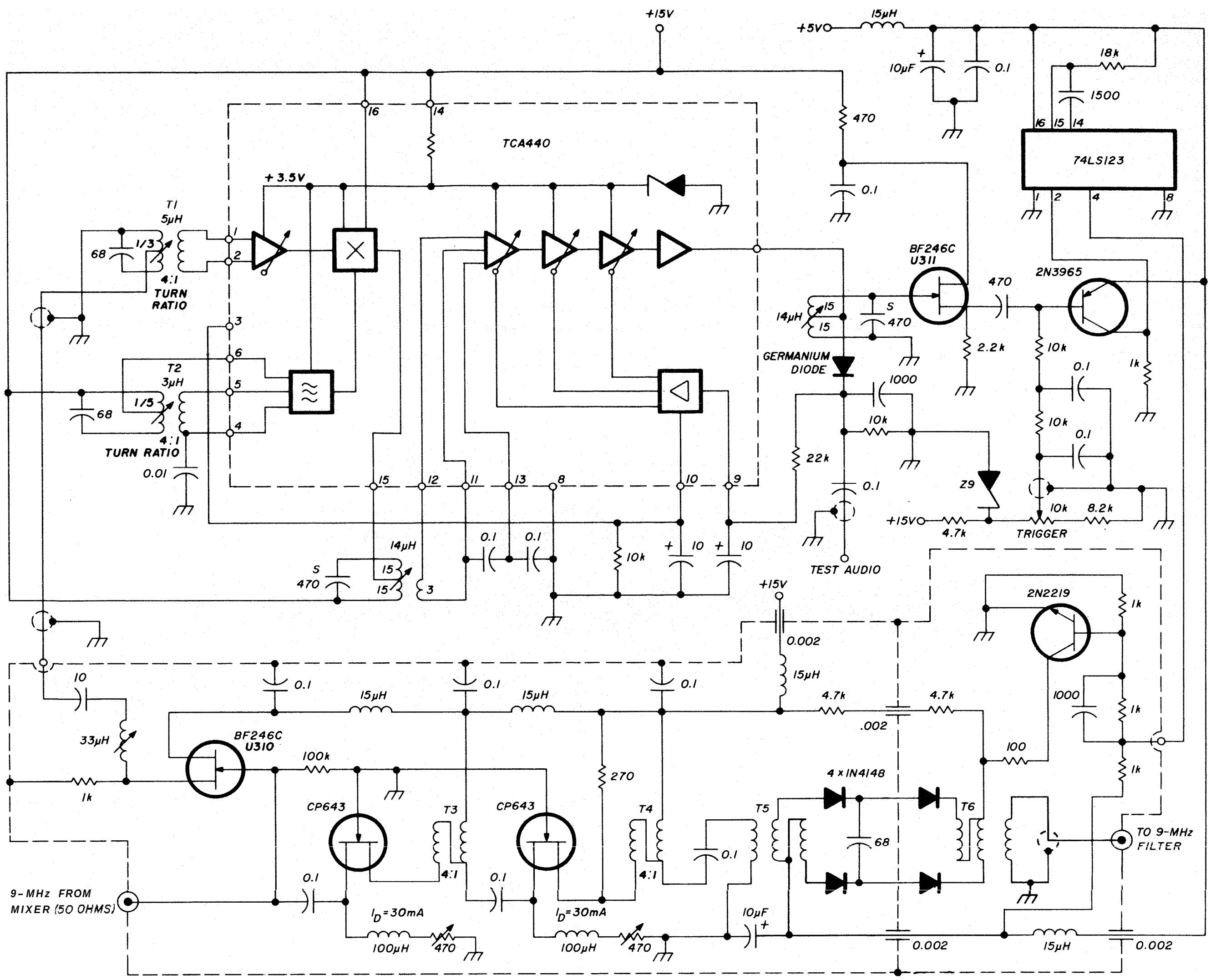
circuit description

The rf signal is picked up at the receiver's first mixer (9 MHz in this case), amplified by the CP643 fet amplifiers, and fed through the four diode gate, which is frequency compensated: the output is designed to drive a 9-MHz crystal filter. It should be possible to use this same basic circuit over the range from about 3 MHz to 70 MHz by changing the frequency tuned circuits.

A small fraction of the rf signal is coupled through the BF246C source follower and a tuned circuit to the Siemens TCA440 IC, which is actually a complete a-m receiver on a single chip;* this IC operates up to

*Circuit designers who are interested in developing the Woodpecker blanker for use in the Drake R4C, Collins 75S-3C, and other Amateur communications receivers please contact the editor.

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458



T1	Primary is 10 turns no. 28 (0.3 mm) on a FT37-61 ferrite core, tapped 3 turns from cold end; secondary is 2 turns no. 28 (0 3 mm)
T2	Primary is 7 turns no. 28 (0.3 mm) on a FT-37-61 ferrite core, tapped 2 turns from cold end; secondary is 1 turn
T3	Bifilar winding, 17 turns no. 28 (0 3 mm) wire on a FT50-61 ferrite toroid
T5, T6	Trifilar winding, 12 turns no. 30 (0.25 mm) wire on a FT37-61 ferrite toroid

fig. 1. Schematic of the noise blanker that can be added to most modern communications receivers for reducing Woodpecker interference between 10 and 30 MHz. This device is also suitable for blanking out Loran pulses on the Amateur 160-meter band.

40 MHz and is available in the United States. The TCA440 contains its own oscillator and converts the 9-MHz signal to a lower i-f (about 2 MHz) where it is amplified and detected. (The audio test output is for monitoring the AGC action of the TCA440 receiver section.) The BF246 source follower drives the 2N3965 amplifier which has an adjustable trigger threshold; this in turn drives the 74LS123 Schmitt trigger. The Schmitt trigger, through voltage-translator transistor 2N2219, activates the diode gate.

Designer Martin has shown that this arrangement has an intercept point of about 26 dBm and the switching gate has a depth of approximately 80 dB.

In practice, with this noise blanker, the Woodpecker noise pulses are completely nulled out, allowing the weakest high-frequency signals to be received successfully.

This circuit is relatively simple, easy to build, and not critical. Some care is required when building the switching gate, however, to eliminate rf signal leakage; good balance is required.

reference

1. Michael Martin, DJ7VY, "Moderner Storaustaster mid hoher Intermodulationsfestigkeit gegen den 'Specht' und andere Pulse," CQ DL (West Germany), July, 1978, page 300.

ham radio

automation for synthesized 2-meter mobile stations

Meet the *Auto-mate* — a design for improving operation of 2-meter radios using synthesizers

You say you've joined the crowd and have stopped buying crystals for your 2-meter rig? Now that you're into synthesizers and can dial up everything from the area's most valuable and used machine to the three-man operation **50** miles (80 km) away, no doubt you wish you could keep track of all the action. It gets rather scary when you try to manipulate all those dials in the darkness of your automobile.

This article may not solve all your problems, but it goes a long way toward making your mobile operation safer and more fun. It allows you to eavesdrop on the metropolitan chaos while keeping both hands on the wheel. I'll show you how to automate your synthesizer so that it "knows" exactly what you want when you dial in only the desired receiver frequency. I'll show you how to add scanning push-to-talk/push-to-receive controls to relieve you from "mobile thumb" derived from holding down the PTT button, and more. In the end you'll have 1) a radio setup that has a synthesizer up front with you and a trunk-mounted radio if you desire (sorry thieves!), 2) short microphone wires to avoid trash pickup in mobile operation, and 3) a unit you can run in complete darkness. I call it the ***Auto-mate.***

background

My project began with an article by Bob Fanning, K4VB, and Gary Grantland, WA4GJT.[1] This article showed how to build an 800-channel synthesizer from boards and parts supplied by the authors.

I had already fallen in love with the KLM 2700 synthesized radio I use for a base station on 2 meters, so synthesis had to be the way to go for mobile operation. I mounted a Heath HW-202 in the trunk of my Toyota. It became a case of running a huge wiring harness or settling for one channel (trunk chosen) and stopping every time I wanted to change channels. This doesn't make for the greatest operation when you travel around the country! In addition, I had alternator whine because I tried to run unshielded microphone lines to the trunk to simplify the wiring. Bob and Gary[1] gave me the solution to that one with their synthesizer, because you can modu-

By D. J. Brown, W9CGI, R.R. 5, Box 39, **Noblesville,** Indiana 46060

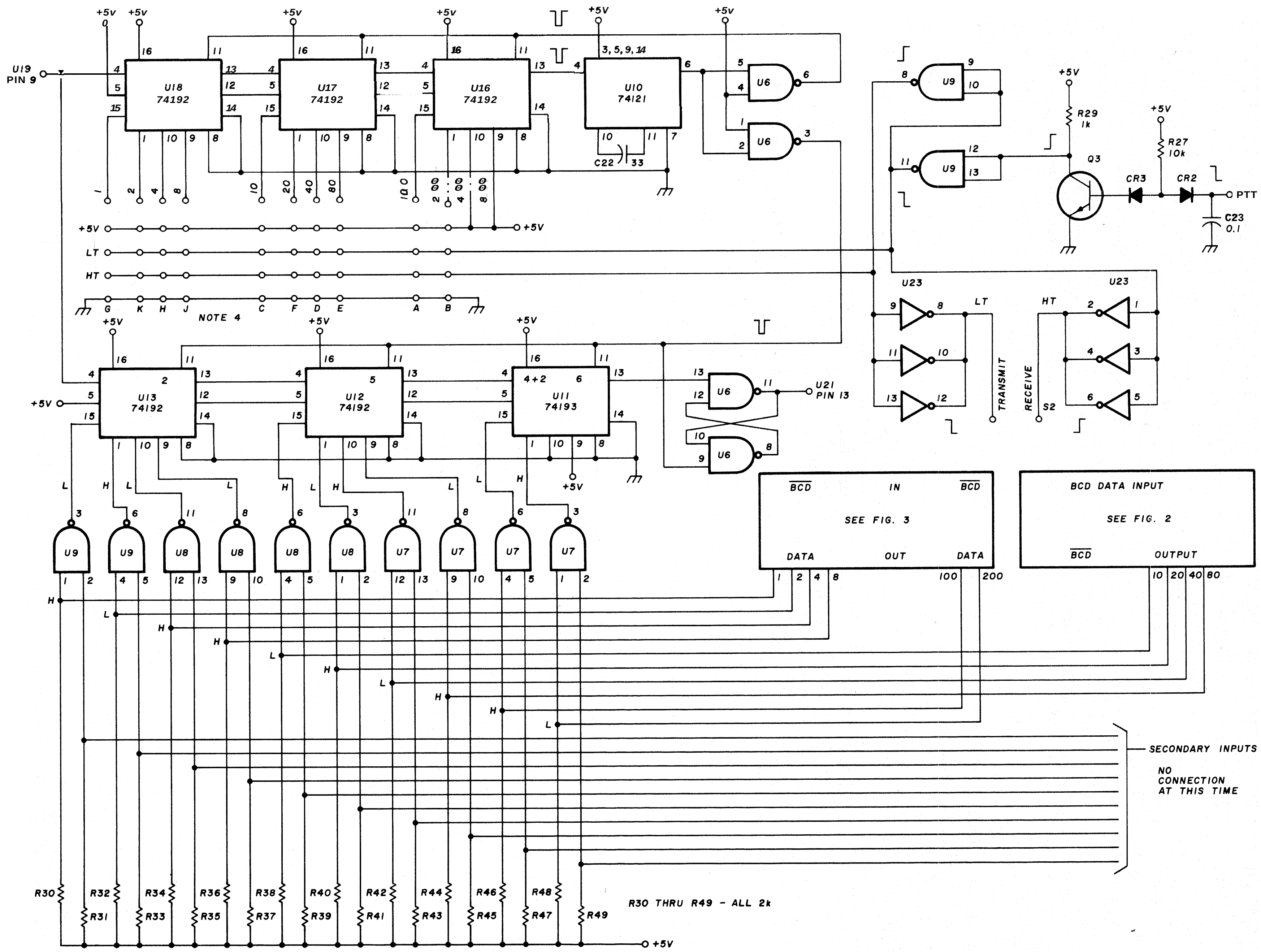


fig. 1. Counter portion of the synthesizer described by K4VB and WA4GJT shown in (reference 1). The thumbwheel switches are replaced by the circuits in this article.

late the sythesizer up front and run only two small rf coaxial cables to the trunk for full-channel control.

The synthesizer' used two BCD-encoded output switch sets to control channel selection and a separate switch to control transmit and receive modes. This can be quite confusing in a mobile, even in the daylight; at night it becomes a disaster! I can't give enough praise for the synthesizer, boards, parts, and most of all, the personal help by K4VB and WA4GJT. If you are about to go the synthesizer route on ***any*** rig, ***do*** read their article,[1] and look into this one further.

How would you like to dial up only the receive frequency on one BCD switch, set (three) switches and have a readout automatically tell you your switches are set correctly for both transmit and receive?

switching logic

If you're acquainted with BCD codes for the numbers 0-9 (you can learn them very quickly, I assure you), you can use inexpensive SPST switches for one BCD set and the second set can be eliminated altogether. If you arrange the BCD sets as two rows of four switches for the 100s and 10s of kHz, and a ninth switch to control the 146- or 147-MHz choice, you can do your setup in the dark.

The action is totally by feel. For example, you feel the front panel and locate the top row or switches. Flip up the right-hand three switches (A - B - C, from right to left – a decimal 7). Drop down to the second row. Flip the middle pair up (B - C, from right to left – a decimal 6). Make sure the MHz switch, placed by itself, is to the left (146 MHz, or lower segment), and you'll be on our local machine: 16/76. It's just that simple and you can do it blindfolded. If you're in Indiana, please, don't do this to prove a point while driving! It's really easier than the decimal-faced/BCD output switches and cheaper.

control circuit operation

Now you must be wondering what takes the place of the BCD set switches (at $10 per set), besides one set of SPST switches. Simple TTL gates! (And very few of them, thanks to our band plan setup.) Consult the tables included here and you'll see the very nice arrangement of our channels. Pay close attention to the numbers in bold type, as these are the only numbers on which my circuit operates. The 10s of kHz are fed in straight from the switches of that row (I suggest the bottom row). This is a 1 - 2 - 4 - 8 line combination in Bob and Gary's article.[1] The 1461147-MHz solution is by another single switch, rather than by a BCD deck. I simplified matters by wanting only the 2 MHz. The 100s of kHz (in bold face) are the only numbers processed on the gate board.

some examples

You feed the switch information into the gate board on A - B - C - D, and their respective outputs are marked 10 - 20 - 40 - 80, as in reference 1. The scheme will work on ***any*** synthesizer using the same encoding shown in fig. 1, which is from reference 1. The authors use TTL 7400 NAND gates at the switch inputs to allow for the dual switches. Some synthesizers require true BCD inputs directly to an up/down counter to set "jam" inputs. Just add inverters to the 10 - 20 - 40 - 80 lines for these models and wire the other switches for true data as well.

For my example frequency of 146.76 MHz, the 7 would be A - B - C low and D high, at the 10 - 20 - 40 - 80 lines that are my board outputs. If you must have true data, invert the 10 - 20 - 40 - 80 outputs – not the data from the switch feeding my gate board. My board has true data as inputs and it outputs inverted data as shown.

gating circuit

For the same example, the 7 is operated on my board (remember the A - B - C inputs are high and the D low). In receive, and in all simplex channels, the information is handled by gate U1 (fig. 2). The A - B - C high and D-low condition results in a low on U1 pins 3, 6, 8, and a high on pin 11. These are the outputs of my board and the inputs to the synthesizer 7400 gates. The gate on the synthesizer board[1] inverts the data and sets it for the jam inputs of the counters. Be sure to wire the 100s of kHz for true inputs to the gate board, and the 10s of kHz and MHz switches for the inverted data, as shown for the synthesizer board[1] since that's where they're connected. (See fig. 3 for details.)

144-147 MHz coverage

There's a good design scheme in the synthesizer1 that only allows the 4-MHz frequency spread from 144-148 MHz to be dialed in. A 4 is hardwired on the C jam line input (400). Only two wires come from the BCD deck switch that are used in the synthesizer for MHz. These two lines allow you to add a 0 - 1 - 2 - 3 to achieve 144 + 0 – 144 + 3 as a usable MHz figure. Thus, you get 144+ – 147+ MHz coverage, so I only have to enter a 2 or 3 on the two lines. The 2 (for 146 MHz) requires the A line to be low and the B line to be high. The 3 (for 147 MHz) requires the A and B lines to be 800-line high.

I ran the 200 line as a hardwire to ground as in the circuit of reference 1, causing the B input after the gate to always be high. Then I tied the 100 line to +5 volts through a 2200-ohm resistor. When the MHz switch is to the right in the 147-MHz position, it grounds the 100 line through the 2200-ohm resistor

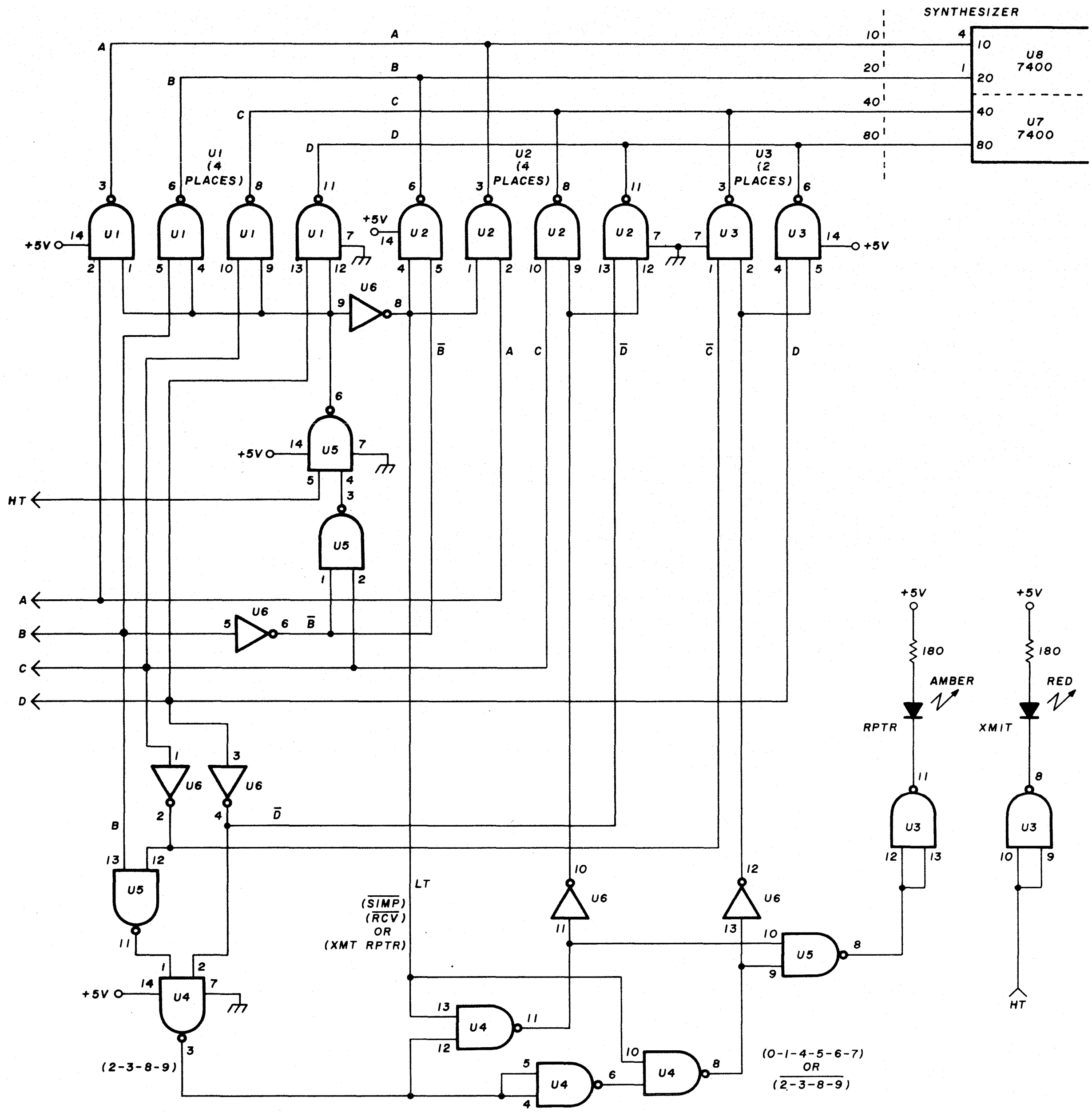


fig. 2. Schematic of the *Auto-mate* gate board. U1, U2, U3 are 7403s. U4 and U5 are 7400s. U6 is a 7404.

and causes the required high on the A line. For 146 MHz, the switch is merely an open circuit on the 100 line for a low on the A line.

The 10s-of-kHz-lines switches (1 - 2 - 4 - 8) are wired upside down from your normal true data switches **(fig. 4).** From my example 146.76 MHz), you want the switches to provide a low on the B and C (or 2 and 4) lines. The inversion to true data is handled by the gates on the synthesizer board.

All this allows switching in the ***receive*** frequency at all times. In my area, a 16/76 machine is referred to as 76; to hear you dial up, switch in, or tune in 146.76 MHz.

bandplan considerations

My board takes care of the required 146.16 MHz when, and only when, you want to transmit. You really don't care about the transmit frequency as long as it's a) correct for the bandplan (same for simplex, and split for repeaters), and b) in the legal band. I solved the first requirement by my circuit, which automatically senses the receive frequency dialed in as being either simplex or repeater and processes it accordingly. Bob and Gary[1] solved the second problem by limiting operation to 144-148 MHz on their synthesizer board. As long as these requirements are met why bother with dialing in the transmit frequency? For those who want to go upside down (i.e., 76/16 if the repeater is down), it's as simple as dialing in the *transmit* frequency. My board will still shift things correctly for the actual transmit cycle. If you dial in 146.16 MHz to receive, you'll automatically transmit 146.76 MHz. This proper shift holds true for *all* repeater pairs anywhere in the 146-147 MHz region.

gating-circuit operation

The simple gates are easy to follow, line-by-line, in **fig. 2.** I'm sure you want to know how the circuit does its tricks. For this, see the tabies. i'ii cover oniy the 146-148 MHz region **I use.**

All my board does in the repeater function is add or subtract the proper 600 kHz from the receiver frequency that you've input to the switches. The tables show you how the bandplan allows this function. In the 146-MHz region, the receive-frequency numbers dialed in, such as 6 - 7 - 8 - 9 for the 100s of kHz column, result in 0 - 1 - 2 - 3 respectively (i.e., 76 receive/16 transmit) **(table 3)**. For all these repeater pairs, my board gives a 600-kHz offset number no matter which one you dial in.

To set up a frequency (remember, choose the *receive* frequency), choose the MHz frequency by a switch totally independent of my board. Then choose the 100s of kHz going through my board by using the *receive* frequency. Then, choose the proper 10s of kHz and you're finished. When you press the PTT switch to transmit, my board will process the shift automatically whether you're in the 146- or 147-MHz region.

Gate U1 **(fig. 2)** handles all receive codes dialed in and all simplex transmit codes (the same as in the receive mode) and passes them to the synthesizer board. The left half of U2 operates on all repeater frequencies to pass line A (unaffected by the 600-kHz number shift), and an inverted line B (for all repeater shifts in transmit) to the synthesizer (lines 10 and 20). The right half of U2 and the adjoining half of U3 handle the C and D line inversions when required.

simplex operation

For all numbers 0-9, the B line is low and the C line high for only two numbers, 4 and 5, which detects the simplex frequencies you dial in. This is handled by U6 pins 5, 6 (inverting the B low to a high that can be gated in a TTL NAND gate), U5 and pins 1, 2 to gate the $\overline{B}$ and C together for a low at U5 pin 3, causing a high at U5 pin 6 and enabling all of U1 for simplex transmit. In receive, a high U5 pin 6 keeps U1 in use. This high is caused by a low at U5 pin **5** regardless of what occurs at U5 pin 4 and comes from the HT line (high on transmit; therefore low on

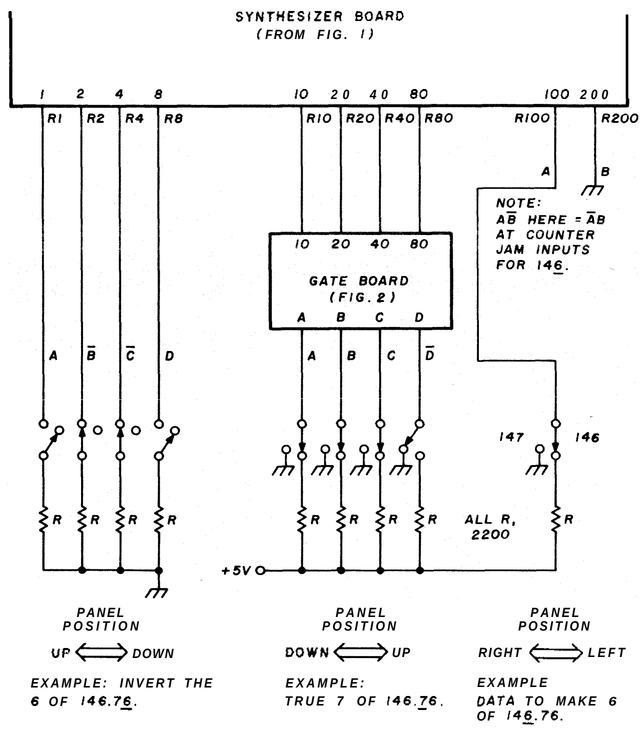


fig. 3. Panel-switch positions and relationship between the gate board and synthesizer board. The 100s of kHz should be wired for true inputs to the gate board. The 10s of kHz and MHz switches should be wired for inverted data (see text).

receive) of the synthesizer board connected to U5 pin 5.

Without going further into a line-by-line description, the other numbers of a repeater nature are detected by similar gating means and are used to control the function of the gates in the right half of U2 and left half of U3. All these outputs are paralleled so that only the correct one operates on the 10 - 20 - 40 - 80 output lines from the gate board. Control is maintained by lines such as the one line to all four gate inputs of U1 pins 1, 4, 9, 12. If this line goes low, regardless of the other inputs from A-B-C-D, all outputs will go high. In this case the gate is entirely out of the picture.

I'll be glad to answer any questions on the gate board upon receipt of a self-addressed stamped envelope. Questions on whether the synthesizer can be used on your radio should go to Bob and Gary.[1*] For questions on whether my scheme for automation will work between your switches and another synthesizer, send me a large copy of your schematic and I'll try to help you if I can. The Auto-mate should work on any synthesizer into which the count chain is fed as real frequency date, not as fancy codes!

Be sure to leave the leads a bit long between a) the

*G&F *Electronics*, *P.O.* Box 4151, *Huntsville*, **Alabama *35802.***

table 1. Amateur 2-meter bandplan for 146-MHz showing binary-coded decimal equivalents for input and output switching in the *Auto-mate.* Numbers in boldface type are those on which the circuit operates.

dial in receive frequency	type	desired transmit frequency	in code 146. x 1				outcode				relationship (I = Invert)			
			D	C	B	A	D	C	B	A	D	C	B	A
14601	I R	14661	0	0	0	0	0	1	1	0		I	I	
14604	I R	14664	0	0	0	0	0	1	1	0		I	I	
14607	I R	14667	0	0	0	0	0	1	1	0		I	I	
14610	I R	14670	0	0	0	1	0	1	1	1		I	I	
14613	I R	14673	0	0	0	1	0	1	1	1		I	I	
14616	I R	14676	0	0	0	1	0	1	1	1		I	I	
14619	I R	14679	0	0	0	1	0	1	1	1		I	I	
14622	I R	14682	0	0	1	0	1	0	0	0	I		I	
14625	I R	14685	0	0	1	0	1	0	0	0	I		I	
14628	I R	14688	0	0	1	0	1	0	0	0	I		I	
14631	I R	14691	0	0	1	1	1	0	0	1	I		I	
14634	I R	14694	0	0	1	1	1	0	0	1	I		I	
14637	I R	14697	0	0	1	1	1	0	0	1	I		I	
14640	S	14640	0	1	0	0	0	1	0	0				
14643	S	14643	0	1	0	0	0	1	0	0				
14646	S	14646	0	1	0	0	0	1	0	0				
14649	S	14649	0	1	0	0	0	1	0	0				
14652	S	14652	0	1	0	1	0	1	0	1				
14655	S	14655	0	1	0	1	0	1	0	1				
14658	S	14658	0	1	0	1	0	1	0	1				
14661	R	14601	0	1	1	0	0	0	0	0		I	I	
14664	R	14604	0	1	1	0	0	0	0	0		I	I	
14667	R	14607	0	1	1	0	0	0	0	0		I	I	
14670	R	14610	0	1	1	1	0	0	0	1		I	I	
14676	R	14616	0	1	1	1	0	0	0	1		I	I	
14679	R	14619	0	1	1	1	0	0	0	1		I	I	
14682	R	14622	1	0	0	0	0	0	1	0	I		I	
14685	R	14625	1	0	0	0	0	0	1	0	I		I	
14688	R	14628	1	0	0	0	0	0	1	0	I		I	
14691	R	14631	1	0	0	1	0	0	1	1	I		I	
14694	R	14634	1	0	0	1	0	0	1	1	I		I	
14697	R	14637	1	0	0	1	0	0	1	1	I		I	

switches and the gate board, b) the other switches and the synthesizer, and c) between the outputs of the gate board and the synthesizer. No high frequencies are on these leads so there'll be no radiation problem. Just don't dress the ieads down around the VCO area. If you leave the leads a bit long you can add scanning, push-to-talklpush-to-receive circuits, and more.

further automation: scanning and PTT/PTR

In this part of the article I describe another simple board requiring nine or fewer ICs, of which three are simple "less-than-25-cents" gates. The total IC cost, from a recent ad, is $3.56. This circuit may be added between the synthesizer input switches and the synthesizer board to provide scanning of a full MHz (or part), push-to-talk/push-to-receive (PTT/PTR) control to ease the "mobile thumb" problem, and full scan control from the PTT switch on the microphone.

It's a nice package in itself. Note that this board is connected between the ***receive*** encoder switches and the synthesizer board. You'll still transmit on whatever command is dialed into the transmitter switches. The small expense of building both boards makes fuli automation the way to *go.* Shouid your synthesizer need true BCD codes at the synthesizer board inputs I've provided information for the IC and wiring changes.

This part of the article is arranged into the following parts: Scanning counter/jam inputs (fig. **4**). PTT/PTR and scan/halt control circuits (fig. **5**). Input and output processing (fig. **6**), and what, where and why of the timing circuits (fig. ***7***).

scanning

Scanning is accomplished by feeding the binary outputs from a counter pair to the synthesizer gate inputs on the synthesizer board. These outputs change during scan and thus change the encoded input information choosing the channels. Which frequency band (MHz) that's to be scanned remains a

table 2. **Amateur 2-meter bandplan for** 147-MHz **showing binary-coded decimal equivalents for input and output switching in the *Auto-Mate.* Numbers in boldface type are those on which the circuit operates.**

dial in receive frequency	type	desired transmit frequency	in code D	C	B	A	out code D	C	B	A	relationship (I = Invert) D	C	B	A
14700	R	14760	0	0	0	0	0	1	1	0		I	I	
14703	R	14763	0	0	0	0	0	1	1	0		I	I	
14706	R	14766	0	0	0	0	0	1	1	0		I	I	
14709	R	14769	0	0	0	0	0	1	1	0		I	I	
14712	R	14772	0	0	0	1	0	1	1	1		I	I	
14715	R	14775	0	0	0	1	0	1	1	1		I	I	
14718	R	14778	0	0	0	1	0	1	1	1		I	I	
14721	R	14781	0	0	1	0	1	0	0	0	I		I	
14724	R	14784	0	0	1	0	1	0	0	0	I		I	
14727	R	14787	0	0	1	0	1	0	0	0	I		I	
14730	R	14790	0	0	1	1	1	0	0	1	I		I	
14733	R	14793	0	0	1	1	1	0	0	1	I		I	
14736	R	14796	0	0	1	1	1	0	0	1	I		I	
14739	R	14799	0	0	1	1	1	0	0	1	I		I	
14742	S	14742	0	1	0	0	0	1	0	0				
14745	S	14745	0	1	0	0	0	1	0	0				
14748	S	14748	0	1	0	0	0	1	0	0				
14751	S	14751	0	1	0	1	0	1	0	1				
14754	S	14754	0	1	0	1	0	1	0	1				
14757	S	14757	0	1	0	1	0	1	0	1				
14760	I R	14700	0	1	1	0	0	0	0	0		I	I	
14763	I R	14703	0	1	1	0	0	0	0	0		I	I	
14766	I R	14706	0	1	1	0	0	0	0	0		I	I	
14769	I R	14709	0	1	1	0	0	0	0	0		I	I	
14772	I R	14712	0	1	1	1	0	0	0	1		I	I	
14775	I R	14715	0	1	1	1	0	0	0	1		I	I	
14778	I R	14718	0	1	1	1	0	0	0	1		I	I	
14781	I R	14721	0	0	0	0	0	0	1	0	I		I	
14784	I R	14724	1	0	0	0	0	0	1	0	I		I	
14787	I R	14727	1	0	0	0	0	0	1	0	I		I	
14790	I R	14730	1	0	0	1	0	0	1	1	I		I	
14793	I R	14733	1	0	0	1	0	0	1	1	I		I	
14796	I R	14736	1	0	0	1	0	0	1	1	I		I	
14799	I R	14739	1	0	0	1	0	0	1	1	I		I	

function of the 1461147 MHz switch described previously and has no bearing whatever here. It's a manual, front-panel switch choice — no scanning involved.

Installation notes. To install the scanning circuitry, break the long leads described above and insert this in series. Trace the circuit from the switch line of 10s and kHz **A** line to the synthesizer gate input at R1 of **fig. 4.** You'll see two Xs on the line that were one at the same point before you open the lead, as previously connected. Each other pair per lead is the same, i.e., B to R2 — Xs were the same point, C to R4 — Xs, etc. Where you open the leads will depend on where you place your new board. I'd mount the new board, make the break, and reconnect two points of each lead, one lead at a time.

Counters. The counters chosen for scanning are binary (not BCD or decade) for a very good reason, even though each one must count only ten positions to cover the 100 possible frequencies of each megahertz. (Because of the original synthesizer input scheme, you could dial in every 10-kHz increment, even though the channels are every 30 kHz). (See **table 4.**) Note that in my synthesizer, a $\overline{\text{BCD}}$ (inverted) code is required at the synthesizer gate inputs. The easiest way to accomplish this and still scan in an up (increasing frequency) direction without a lot of inverters is as follows.

From **table 4,** you'll see that to obtain a decimal 0 at the synthesizer counter jam inputs you must feed the inverse BCD (i.e., $\overline{\text{BCD}}$) to the synthesizer board input gates. This happens to be a decimal 15. Or, for a BCD 0 (0000) you need a gate input of 15 (1111). For the next step of 1, you need a 14, for 2, a 13, and so on. Thus, you require a binary counter that sets to 15, counts downward to 6 (10 counts), detects the next count of 5 without changing from the 6 outputs (reset or load takes precedence over count), and

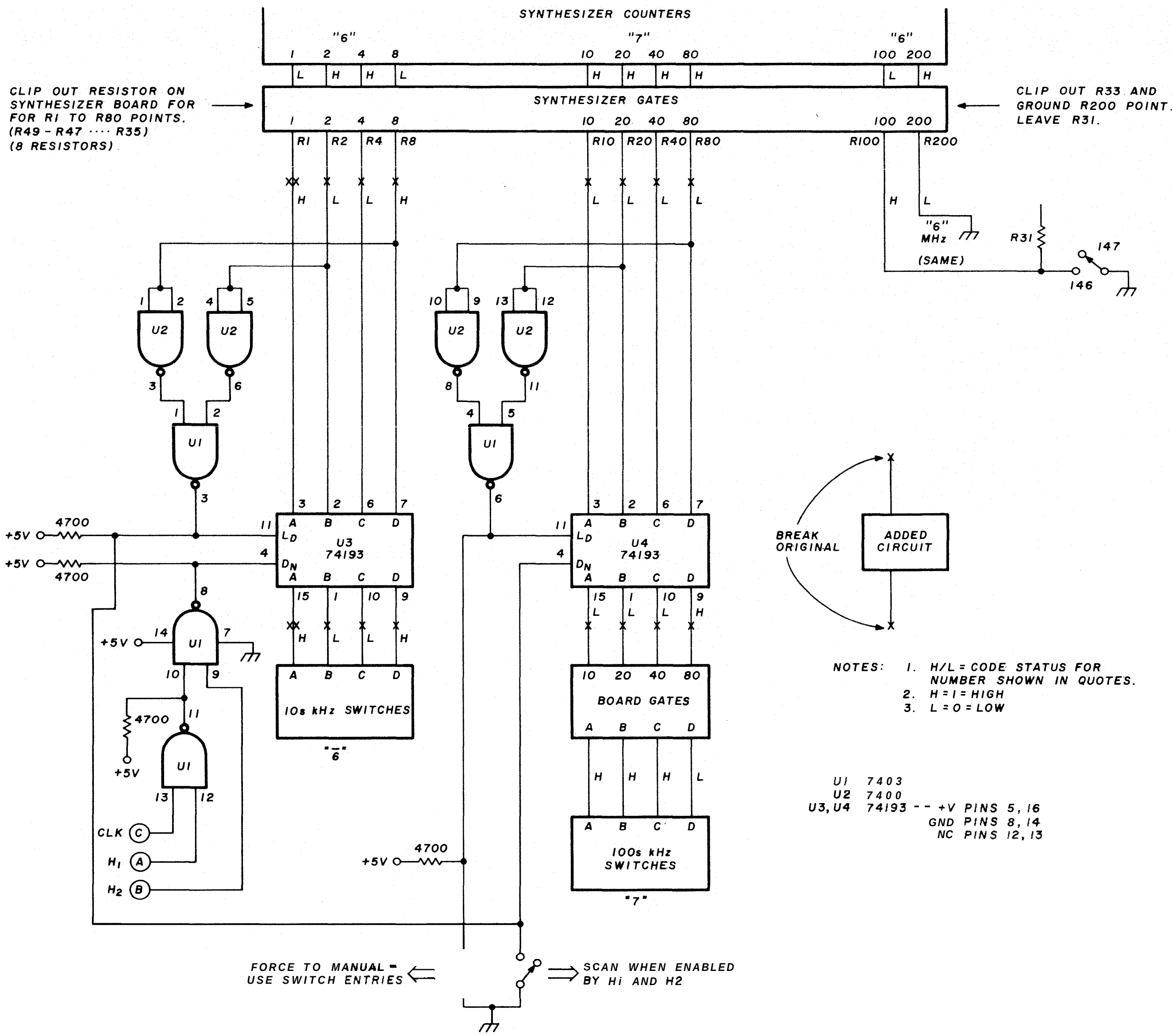


fig. 4. Scanning counter and jam inputs to the synthesizer counters. The 10s-of-kHz switch outputs and gate-board outputs are inverted BCD data, which can be fed directly into U3, U4, the counter ICs.

uses the detected 5 to set 15 again. This is the same as counting from 0 to 9 in inverted BCD — or from 15 to 6 in binary. It's the same if you want the same count sense (up) and inverted outputs.

The outputs of the 10s-of-kHz-switches and the gate board are inverted BCD outputs. They can be fed into the jam inputs of the new scanning counter set, U3, U4 (fig. 4). When the load line (U3, U4 pin 11) goes low, this information is passed directly to the outputs and to the synthesizer gate inputs — inverted and with the correct code ($\overline{\text{BCD}}$). This fact, (load line low to load the switch inputs) brings out some interesting sidelights and benefits.

First, for manual switch control and no scan, all you do is to force the load lines low with switch S1 (fig. 4). You're then in manual mode regardless of the states on H1 and H2 or whether the control (fig. 5) is in SCAN or HALT.

Second, the switches of 100s and 10s of kHz must be in their decimal zero position (all switches down) to feed the required 15 $\overline{\text{BCD}}$ to the counters (U3, U4, fig. 4) during reset, or load as it's called here. This is because, instead of clearing the counters (reset to 0 by a high on pin 14), you want to reset to 15. You do this by briefly pulsing the proper load line low with the outputs from U1 pin 6 or U1 pin 3 (fig. 4). If all switches are set to 00 (i.e., 146.000 MHz), the whole MHz segment will be scanned. Setting 00 puts the required 15-15 on $\overline{\text{BCD}}$ pulses on the synthesizer gate inputs.

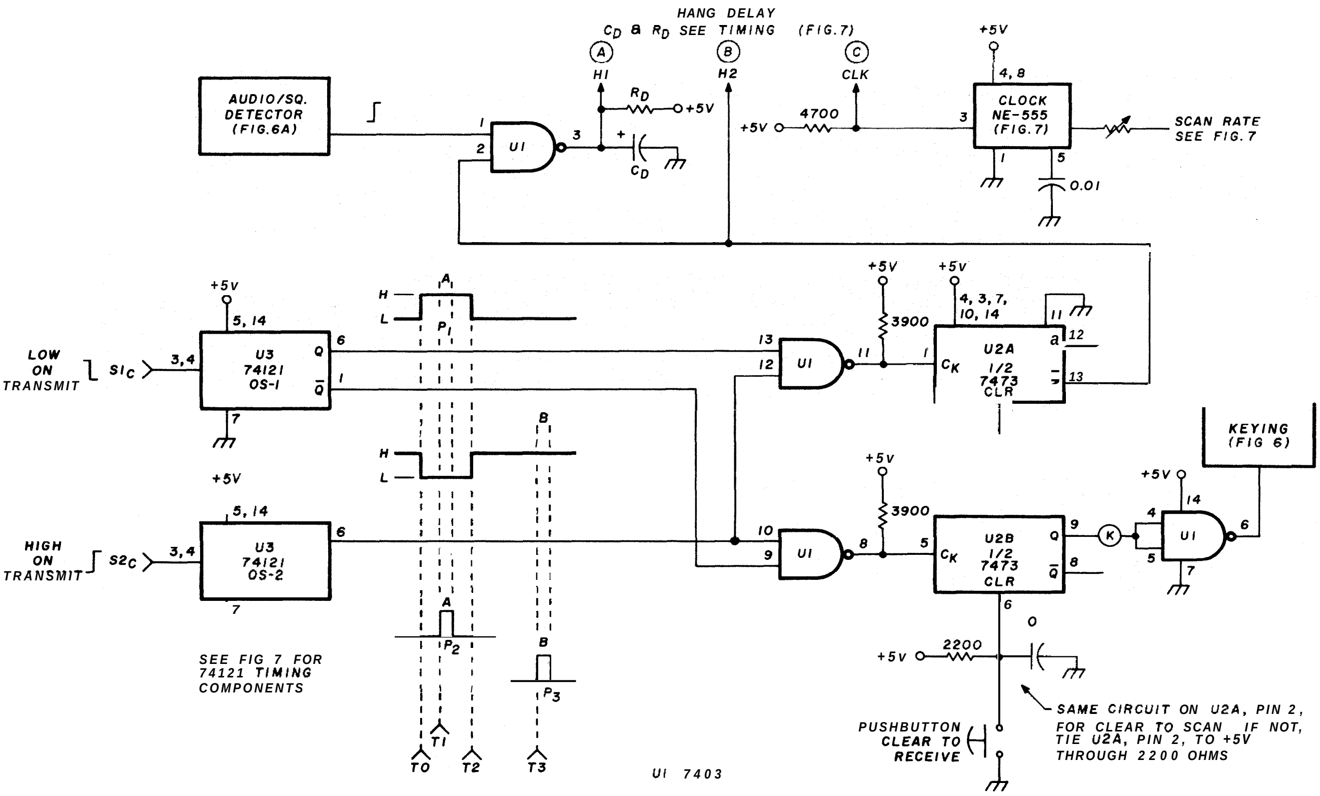


fig. 5. Push-to-talk and push-to-receive and scan/halt control circuits.

If you don't want the full MHz coverage, just set the 10s of kHz switch to 0 (all down — a must during all but manual mode), and the 100s of kHz switch to the ***lowest*** 100s of the kHz switch you wish to scan. Example: for the 146-MHz region, set the 100s of kHz switch to 6 (146.60) (B-C switches up), and you'll hear all repeater outputs. Set a 4 (146.40) (C switch up) and you'll get all simplex and repeater transmissions. This saves time by not scanning the repeater inputs.

If you like these ideas and have a synthesizer that requires BCD true data, you'll need a different counter scheme (fig. **8**). Suggestion: build the gate board and the scan counter set (fig. **4**) and add inverters at each of the upper Xs of fig. 4. If you then bring the outputs for true ***and*** inverted BCD to a plug, you can run the whole gadget on any synthesizer that requires direct frequency codes in BCD or $\overline{\text{BCD}}$, but not the models that require a special code.

PTT/PTR and scan/halt

Around my house there's just too much noise and unplanned interruptions to warrant VOX operations on any of the base station equipment. On the other hand, holding down a PTT microphone button for long periods during a 24-hour contest is no thrill either. Long ago I went to a push-to-talkipush-to-receive operation on all the base station radios; for mobile work it's even nicer. You can even go to a visor-mounted microphone and a steering-column or floor-mounted pushbutton for full hands-off control. No more hassles with a shift lever and the microphone cord!

how it works

As the PTT switch is closed, a pulse generated from OS-1, a 74121 one shot, (fig. **7**), is directed to two gate inputs. You can vary the width of this pulse to suit yourself as in fig. 7, but I've found that 1 second is a nice average number with which to start. When the PTT switch is released, another much shorter pulse is generated by OS-2. If the switch is released within the I-second timeframe of P_1 (fig. **5**), the high of P_1 and P_2 triggers U1 pin 11 low for a P_2 wide pulse (i.e., a short blip on the PTT switch).

This short pulse is stored as a change of state in one-half of a 7473 that controls the H2 line. A low on H2 turns off gate U1 (fig. **4**) and cuts off the clock pulses to the scan counter set — scanning stops. Another blip on the PTT and you're scanning again as the 7473 again changes state. This blip can be extremely brief. Unless you have a very quick transmitter, it won't be heard on the air. My circuit, used on a Heath HW202, is silent if I stab the PTT switch

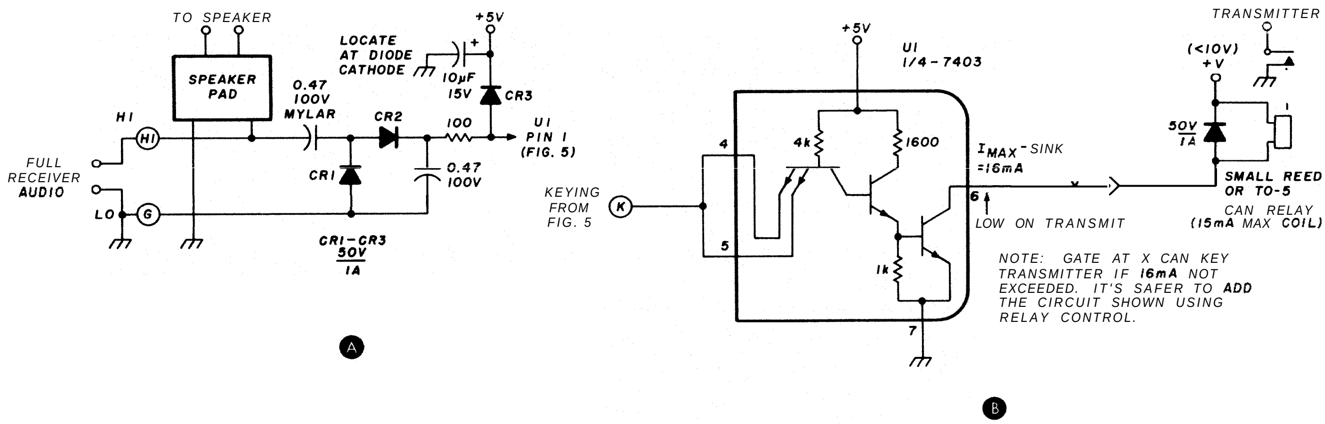
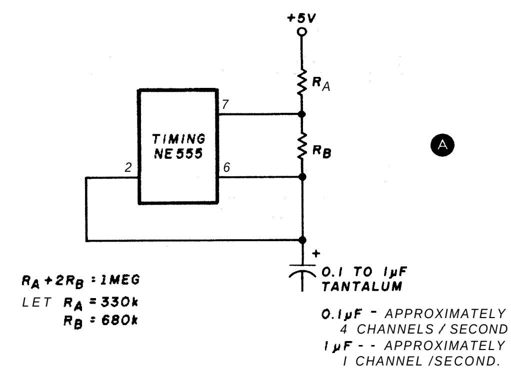
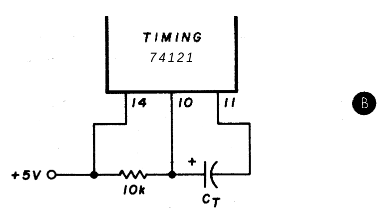


fig: 6. Input-output processing circuits. At (A) is a voltage doubler to handle the TTL gate input at U1 pin 5 (fig. 5). Component values have been chosen for maximum audio fidelity consistent with reliable halts during scanning. A keying system is shown in (B). U1, part of a 7403, will safely handle currents of 16 mA, but the relay circuit is recommended. CR4 is one of the 1N4000 family (approximately 50 volts at 1 ampere).





	C_T PULSE	PERIOD
OS-1	68 TO 100µF TANTALUM (≈1 SECOND)	T_0-TO-T_2
OS-2	6.8 TO 10µF TANTALUM (≈ 10mS)	P_2 OR P_3

NOTES:

1. HANG DELAY: R_D = 1k < R_D < 4700. C_D: ADJUST TO SUIT VOLUME OUTPUT OF RADIO. HANG MUST BE LONG ENOUGH TO HOLD DURING VOICE PAUSES AND DURING OPERATOR CHANGEOVER OR UNIT WILL RESUME SCAN.
2. THIS IS AUDIO - DERIVED HANG CONTROL AND WILL SKIP DEAD CARRIERS.
3. START WITH 4700 & 220µF/15V. REDUCE C TO SHORTEN HANG TIME.

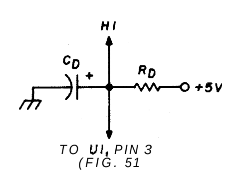


fig. 7. Timing circuitry; (A) shows the scanning clock, an NE-555 IC; (B) shows OS-1 and OS-2 timing-capacitor selection as a function of pulse length and period.

because the HW202 is relay-switched from receive to transmit.

Operation. To use the system, we'll start off in receive and unit scanning. Ah! There's Joe on the local machine. (How the scan stops to hear this in the first place is covered under **input/output processing,** but it does halt when it hears a station.) Then, to stay there and talk to Joe, blip the PTT microphone switch. Scan is now **Halt** through a H2 low. Joe finishes with Harry; now you want to talk to Joe. Firmly press the PTT for some period longer than you set up the P_1 pulse width. Release any time after that. Immediately when you *release* the PTT switch, you're on the air in transmit. As you near the end of your first go-around, again press the PTT switch firmly a few seconds before the end. When you *release* the PTT switch, you immediately return to receive — scanning is still unaffected and in **Halt.** Simple? Not much different, really, except the first release-to-transmit part!

For long-winded souls on quick-natured repeaters, you can even hook in an automatic timer to control the end of transmission. Several have appeared lately, so I won't go into any specifics here. Just wire the timer so that the act of going to transmit (U2A pin 13 low) triggers the timer on; the timer running out places a pulse low on U2B pin 6 (for a clear to receive command). See **fig. 5.** Wire so that a shorter conversation both resets to receive and resets the timer. Set the timer duration for about 10 seconds less than that of your local machines.

You don't even need a reset timer if the timer is of the 555 type. Just be sure to use the pulsed output to clear U2B and not toggle, as it does the PTT switch. If you've already returned to receive through the PTT

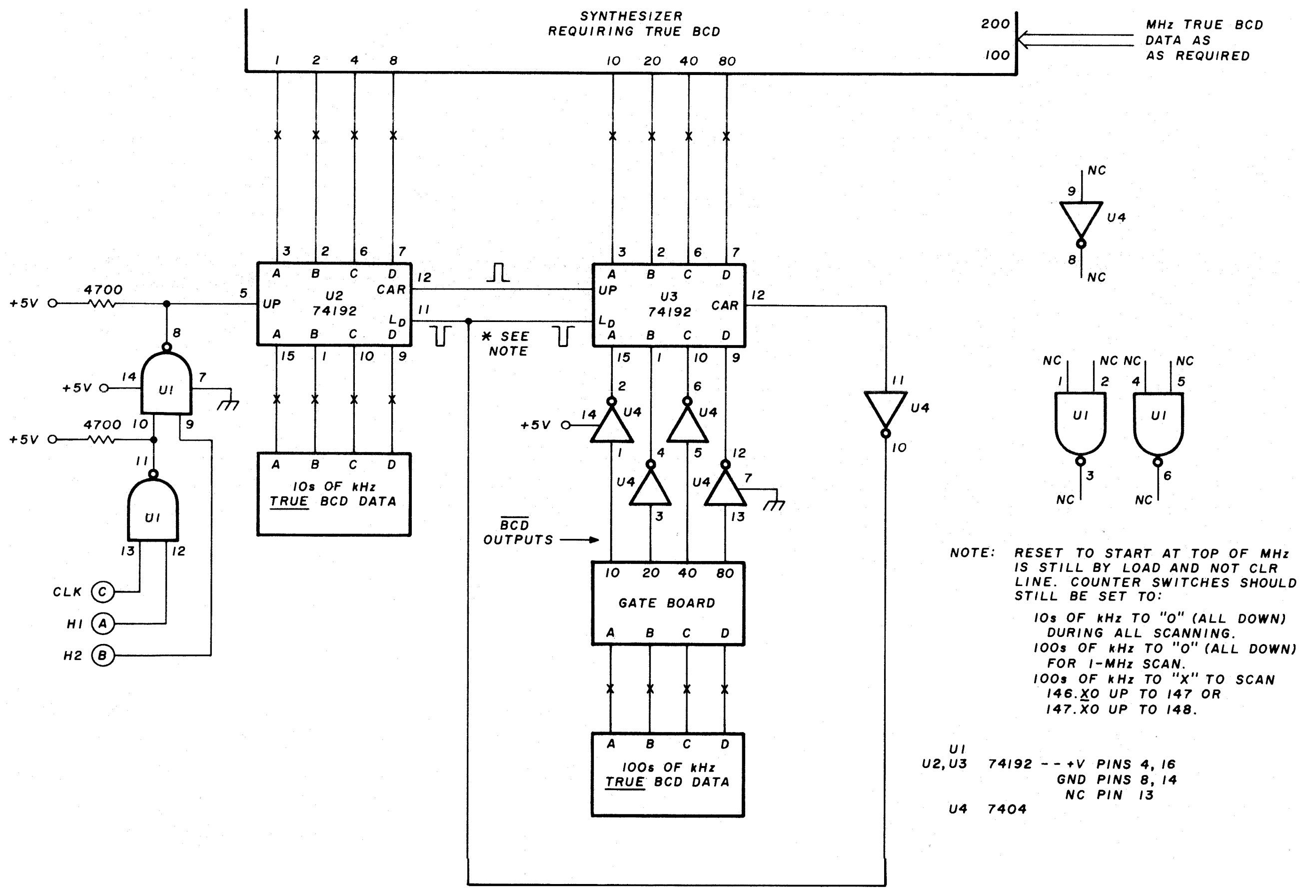


fig. 8. Schematic for synthesizers requiring true BCD input.

switch before the timer times out, this will pulse U2B to receive. If you return to receive ***and*** switch back to transmit before the first time period has run out (as often happens), the second return to transmit will again give you a full time period, as the 555 can be retriggered.

input/output processing

This is the easiest part of all. Output processing means whether or not to add the relay and/or additional transistor output stages to U1 pin 6 (fig. **5**) to handle key-line currents of greater than 15 mA or voltages higher than about +12 volts. If in doubt, use the relay and send a nice, firm relay ground connection back to the radio to key the transmitter.

As for input processing, the control lines into the control section of fig. 5 come from two points on the synthesizer I used. The control signals are TTL levels and a low is applied to OS-1 when the PTT switch is closed (keying the transmitter). A high is applied to OS-2 at that same time. If you don't have these controls, they should be easy to come up with. Just limit the high to about +5 volts. The low should be near ground to protect the inputs of OS-1 and OS-2.

I trunk-mounted my radio and wanted as few wires as possible back and forth, so I installed full volume audio to my synthesizer/control head and put a pad up front. This pad can be a low-impedance T pad if you have the room. With the radio volume control full clockwise or on, I put a resistor in series with the high-side speaker lead that reduced the volume to a comfortable level. The switch shorts out the resistor for the weak ones. I was cramped for space. With full volume coming forward the speaker is silent when full squelched and has plenty of audio available at the control head when a station comes on. Rectify this audio and you have a stop-scan signal, H1.

Looking at fig. 6A, a voltage doubler ensures that there's always enough voltage to handle the TTL gate input at U1 pin 1 (fig. **5**). Diode CR3 (fig. **6**) connected to +5 volts limits the input to U1 pin 1, fig. 5, to a TTL high level. The capacitor at the diode cathode ensures that no audio peaks over +5 volts will appear on the +5 volt line.

The 100-ohm resistor (fig. 6A) limits the gate input to +5 volts maximum without peak-limiting the audio peaks on the input side, which would distort the audio. You may have to decrease this value on some radios with low-volume output, but use a value as large as possible to still have reliable halts on all the stations that are on air (seen as lows on H1).

The RC network on H1 in figs. 5 and 7B puts a hang effect on the action of H1. Keep R_D within the limits shown and change C_D to keep the H1 line low between voice peaks or words. This is the alternative to running a wire from the radio to show a no-squelch condition. It also rejects any dead carriers with no modulation. If you use the wire, limit the voltage excursion to TTL levels and have a high for a station on frequency.

timing

Fig. 7 is self-explanatory as to the what and where, but here are a few of the whys. For OS-1 timing components, you're trying to create a pulse short enough that you don't have to hold the PTT down forever before releasing it to transmit. On the other hand, you don't want the pulse so short you could never use it for scan control. You can only blip your blipper so fast! I found the I-second pulse a good compromise. Blip controls scan reliably, and Joe won't mind waiting one more second to hear from you.

The return to receive is no problem, as you know

table 3. Processing scheme for 100s of kHz as a function of switch inputs for the gate board. Examples are shown for 146.76 MHz (receive) and 146.16 MHz (transmit).

100s OF kHz PROCESSING

DIAL IN * RECEIVE FREQUENCY	OUTPUT FOR TRANSMIT FREQUENCY *	SWITCH INPUT D	C	B	A	PROCESS (D LINE)	PROCESS (C LINE)	PROCESS (B LINE)	PROCESS (A LINE)
0	6	L	L	L	L	PASS	INVERT	INVERT AND PASS THROUGH U_B	PASS THROUGH U_B
1	7	L	L	L	H	PASS	INVERT		
2	8	L	L	H	L	INVERT	PASS		
3	9	L	L	H	H	INVERT	PASS		
4	4	L	H	L	L	PASS ALL THROUGH U_A			
5	5	L	H	L	H				
6	0	L	H	H	L	PASS	INVERT	INVERT AND PASS THROUGH U_B	PASS THROUGH U_B
7	1	L	H	H	H	PASS	INVERT		
8	2	H	L	L	L	INVERT	PASS		
9	3	H	L	L	H	INVERT	PASS		

EXAMPLE:

	D	C	B	A	
146.76 RECEIVE	L	H	H	H	SWITCH INPUT
NEEDS	L	L	L	H	OUTPUT ON TRANSMIT

A → PASS ALL A
B → INVERT AND PASS ALL B
C → INVERT AND PASS C
D → PASS D

	D	C	B	A	*	
SWITCH	L	H	H	H	7	RECEIVE FREQUENCY 146.76
PROCESS	L	L	L	H	1	
GATE BOARD OUT	H	H	H	L	14	(OR $\bar{1}$)
THROUGH SYNTHESIZER BOARD GATES AND AT COUNT JAM INPUTS	L	L	L	H	1	TRANSMIT FREQUENCY 146.16

table 4. Conversion from decimal to BCD code for up-scan (increasing frequency) at the synthesizer counter jam inputs.

UP SCAN IN DECIMAL NUMBER	NUMBER IN BCD CODE D	C	B	A	$\overline{BCD}$ REQUIRED BY SYNTHESIZER INPUT GATES D	C	B	A	THAT $\overline{BCD}$ IS BINARY FOR DECIMAL NUMBER = DEC
0	0	0	0	0	1	1	1	1	= 15
1	0	0	0	1	1	1	1	0	= 14
2	0	0	1	0	1	1	0	1	= 13
3	0	0	1	1	1	1	0	0	= 12
4	0	1	0	0	1	0	1	1	= 11
5	0	1	0	1	1	0	1	0	= 10
6	0	1	1	0	1	0	0	1	= 9
7	0	1	1	1	1	0	0	0	= 8
8	1	0	0	0	0	1	1	1	= 7
9	1	0	0	1	0	1	1	0	= 6

SCAN

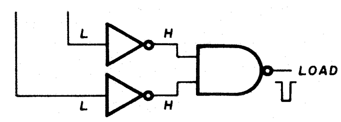
TO SCAN USING THE REQUIRED $\overline{BCO}$ CODE:

1. COUNTERS MUST DOWN COUNT.

2. RESET FOR FULL 1-MHz SCAN MUST BE TO THE NUMBER 15 (SWITCHES TO 14X.001.

3 RESET IS BY LOAD LINE AND OCCURS BY DETECTING THE NUMBER 5 ($\overline{10}$) AT OUTPUTS TO SYNTHESIZER BOARD.

10 N/A BUT $\overline{BCD}$ IS:
1 0 1 0 0 1 0 1 = 5



best when you're going to stop talking and turn it back over, Just press the PTT button a few seconds (over 1 second will do) before turning it over to receive. Return to receive is immediate upon PTT *release.* As for the P_2 or P_3 pulse (depending when you release the PTT), I wanted a pulse that was much shorter than P_1. You must take some time getting on and off the PTT for a scan blip, so that uses P_1 time. I figured a half second worst case, leaving half second if it's to be a P_2 scan-control pulse. Ten per cent of a half second (500 ms) is 50 ms, leaving a 90 per cent error margin, or P_1 safety zone. My capacitor happened to give me a 10-ms pulse that works just fine.

Just about any capacitor will give a pulse long enough. If your scan control PTT blips start putting you in transmit as well, the capacitor is too big, and P_2 is biting into the P_3 zone. Back off!

The C_D, R_D on H-1 depend on your radio. I've provided an R_D range and C_D starting point. Just use the advice under input processing to set things up.

scan clock

The scan counters (and even the switches) allow for increments of every 10 kHz, but there are stations only every 30 kHz. Therefore, the scan clock (fig. 7A) can run at a frequency three times per second as fast as you wish to scan the possible channels. With the components shown, you can make about 1.2-12 Hz, or less than one channel per second to four per second.

Start with a slow scan and another radio tuned to a channel with lots of activity if possible. Then start adjusting the scan clock capacitor, C, for faster rates (smaller C), after establishing R_D and C_D (fig. 7B) for the desired hang delay on channel. You could go up to the TTL counter limit of 32 MHz were it not for other limiting factors and the fact it's ridiculous anyway.

Some of the limits are found in a) the ability of the radio or, in my input circuit, to recognize a station and respond with a control stop only on H-1, b) the limit of the synthesizers to want data only so often (and settled data at that), and the fact that, as scan rates get high, trash is generated by switching that could cause havoc in the synthesizer itself.

Use the second radio to determine when your radio and scanner fail t~ stop on a station and remember, you're scanning only to remove the drudgery of always flipping switches.

stunt box fun

My scanner and synthesizer are such that I can still do a few cute tricks late at night when activity is low. Every time my synthesizer board wants new switch data input, the load line of the switch data input counters (pin 11 that controls the counters on the synthesizer board) goes low. The counter set (U3, U4, **fig. 4**) used for scanning happens to clock or advance on a low-to-high transition. With these two facts, I can cause some amusing things to happen! By making R_D and C_D provide a short pulse, like 10 ms, and using the synthesizer board load line as the clock input to the scan counter set, scan action really moves along! It only stops for 10 ms on each active channel and then moves on. By moving so fast, I can get 10-ms bursts of audio from each active channel – for a multiplex action. You wind up listening to more than one conversation at once. Scan control and PTT/PTR remain unaffected, but it would be a miracle if you could blip fast enough to stop on the channel you really wanted.

It's a novelty and demonstrates one of the upper limits of how fast is fast enough for scanning. All the commercial scanners are capable of scanning at many times the rates used, but you wouldn't be able to watch all the pretty LEDs go scanning along! The voices get pretty choppy in my speedy example above if you don't get the hang time just right. If several stations are on at once, it sounds like your local lodge meeting – everyone talking at once. It's just one of the cute things you can do with this miracle age of electronics. After all, now that you have a fully automatic station doing all the work, you must have something to do – it's called having fun!

reference

1. Bob Fanning, K4VB, and Gary Grantland, WA4GJT, "800-Channel 2-Meter Synthesizer," ***ham radio,*** January, 1979, pages 10-18.

ham radio

More Details? CHECK–OFF **Page 94**

Yagi antenna design:

more data on the performance of multi-element simplistic beams

Manipulating the boom length and element spacing of Yagi beams to maximize forward gain and front-to-back ratio

This is a continuation of last month's discussions of simplistic Yagi antennas. To provide continuity to the complete subject I shall continue the sequential numbering of tables and illustrations. Last month I presented the performance characteristics of 2, 3, and 4-element simplistic Yagi antennas over a range of useful boom lengths. Systematic detailed computations have also been made for simplistic Yagi antennas for 3, 6, and 7 elements. To illustrate the behavior of these larger and more complex antennas the characteristics of 6-element Yagi simplistic beams are shown in **fig. 13** where free-space antenna gain in dBi is plotted against frequency, F, for a range of boom lengths up to $1.5\lambda_o$. Numbered curves correspond to element lengths given in table 3 of reference 5.

The total range of results shows a number of characteristics of interest. First, the bandwidth over which gain is high is determined primarily by the frequency spread between the reflector and the director(~) Second, the shape of the gain curve is generally not flat in the region of interest; indeed it may be sloped and/or humped or dished. Usually the slope favors the higher frequencies. Third, the shape of the gain curve is more complex where the number of elements is large, and the shape of the F/B ratio varies enormously – much more than the shape of the associated gain curve. It may show more than one peak; moreover, the peak structure shifts very rapidly with boom length. The height of the peak does not necessarily seem to vary monotonically with the frequency separation of reflector and director(s). Very high F/B values (greater than 30 dB) are quite rare and when present are invariably very narrow banded. In addition, the frequency bandwidth of the

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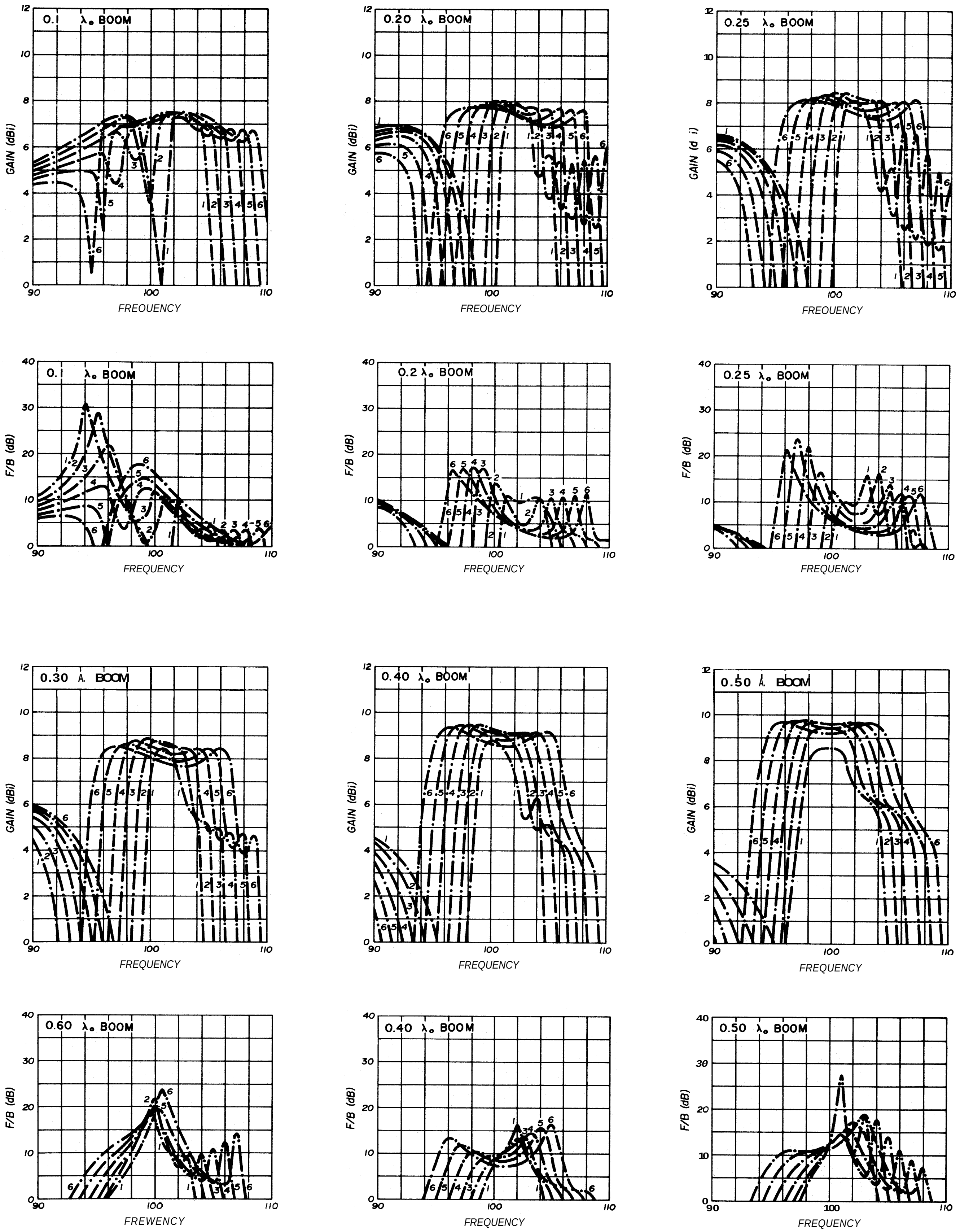
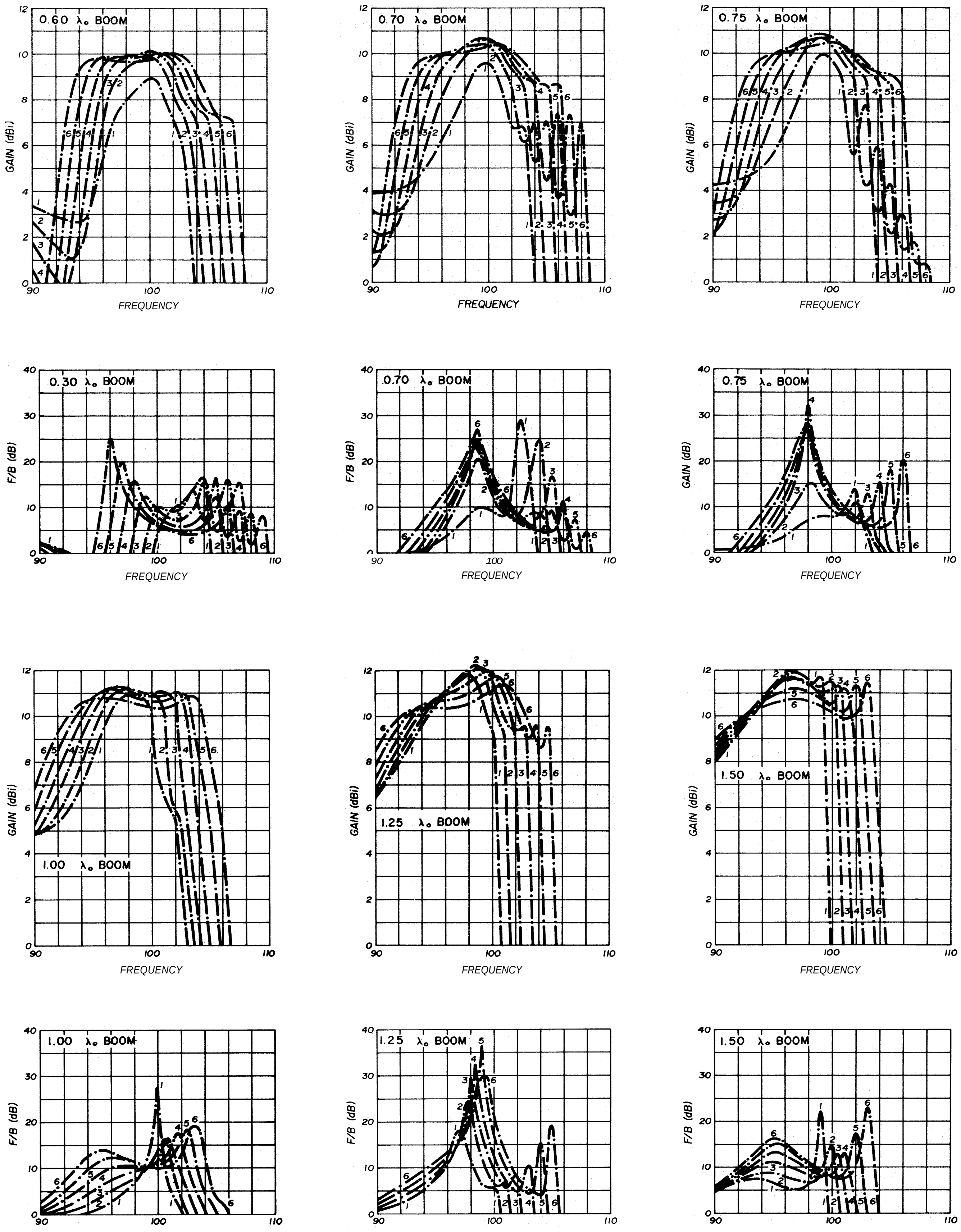


fig. 13. Gain and front-to-back (F/B) ratio in dB for six-element Yagi beams with boom lengths from $0.1\lambda_o$ to $0.5\lambda_o$, and changing reflector and director lengths (see table 3 reference 5 for complete data).



Gain and front-to-back (F/B) ratio in dB for six-element Yagi beams; boom lengths from $0.6\lambda_o$ to $1.5\lambda_o$.

table 4. Band-centered gain, dBi, and front-to-back (F/B) ratio, dB, vs boom length for various multi-element Yagi beams. Data for 3, 4 and 6 elements are plotted in figs. 11, 12, and 13.

boom length	3-elements		4-elements		5-elements		6-elements		7-elements	
(λ)	gain	FIB	gain	F/B	gain	F/B	gain	F/B	gain	FIB
0.10	6.980	12.21	7.216	12.14	7.481	8.36	7.493	7.12	7.486	6.65
0.20	7.960	14.10	7.901	5.49	7.877	7.90	7.759	5.56	7.590	4.34
0.25	8.233	19.88	8.466	8.06	8.262	10.06	8.260	7.83	7.923	5.12
0.30	9.072	12.32	8.797	7.66	8.640	9.47	8.600	8.00	8.353	6.62
0.35	9.254	11.37								
0.40	9.730	8.20	9.419	8.78	9.368	10.813	9.152	9.34	9.286	9.56
0.50	9.421	5.56	9.668	8.89	9.584	10.46	9.635	10.97	9.630	11.86
0.60	8.875	5.33	9.801	11.78	9.810	15.44	9.988	15.08	9.957	13.50
0.70	7.924	2.27	10.076	15.48	10.407	21.56	10.549	21.16	10.615	16.07
0.75			10.505	29.38	10.690	16.68	10.820	17.65	10.822	13.47
0.80			10.363	22.31	10.750	19.50	11.022	14.15	11.067	15.53
0.90			10.534	14.44	10.946	14.98	11.265	12.62	11.278	11.03
1.00			10.333	8.43	10.865	11.41	11.159	10.39	11.255	10.46
1.10			9.703	4.74	10.306	11.65	11.093	13.43	11.403	13.08
1.20					10.395	15.10	11.514	24.89	11.734	22.22
1.25					10.743	24.02	11.793	23.81	11.973	32.50
1.30					10.491	20.26	11.672	31.48	12.165	20.47
1.40					10.510	18.94	11.872	13.80	12.221	15.56
1.50					10.333	12.32	11.608	11.14	12.104	10.83

F/B parameter is undefinable because of the extreme variation in shape!

performance characteristics

If we look carefully at one of these plots, **e.g., fig. 11** (ref. 5) 3 elements, ***boom*** $= 0.25\lambda$, it becomes clear that it is quite difficult to simply characterize "the" gain and "the" F/B ratio. The ***maximum*** calculated gain at a single frequency is 8.9 dBi (curve 1) but a realistic gain at the center of a practical **4** per cent band (curve 3) is more like a 8.0 dBi! Even more difficult is the characterization of the F/B ratio. The *maximum* calculated F/B **(curve 5) is a whopping 38** dB, but this occurs ***only*** at a very specific frequency ($F =$ **1.00)** and for the situation where maximum gain is comprised (reduced to 7.3 dBi). How then can we characterize the results by a single gain figure and a meaningful F/B ratio?

Since gain is perhaps the most important parameter of antenna performance, and since a practical antenna must work effectively over a reasonable band, I have elected to specify the gain at the center of a **4** per cent band. For each case, **e.g.,** 3-elements, ***boom*** $= 0.25\lambda$, the band center is adjusted for each curve to give maximum gain performance over the entire **4** per cent band, and, finally, the specific curve is selected which yields best overall gain performance. I define "the" gain of this case as the gain at band center and "the" F/B ratio as the value at the ***same*** band center and the ***same*** selected curve! Note that the actual F/B may be significantly higher at some other frequency inside or outside the chosen band; we shall discuss this point shortly.

With this definition of band-center gain and band-center F/B, **table 4** has been constructed to show performance not only for 3-, **4-,** and 6-element beams but also for 5- and 7-element beams; **fig. 14** shows a plot of the gain information; this graph is remarkable in four respects. First of all, it demonstrates a practical upper limit to the gain achievable from a given boom length! Second, it demonstrates that this gain is almost independent of the number of elements distributed along its length as long as there are enough! Third, the achievable practical gain shows a ***slight*** preference for ***more*** rather than less elements on a boom. Finally, the "boom gain" — achievable gain from a given boom length — is not

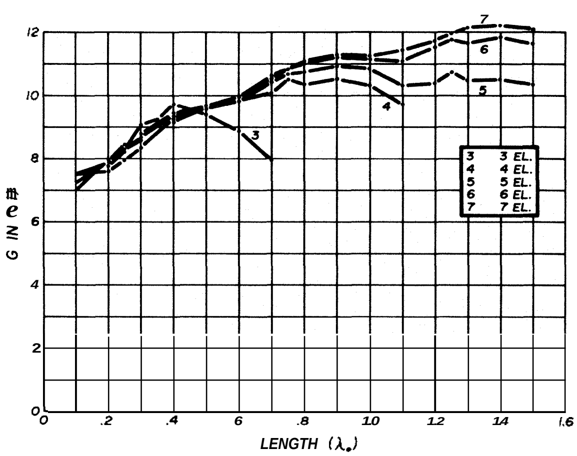


fig. 14. Yagi beam gain in dBi for 3, 4, 5, 6, and 7-element beams as a function of boom length in λ_o.

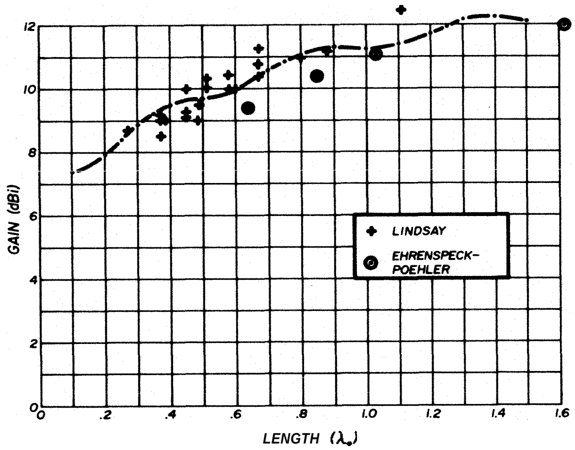


fig. 15. Theoretical Yagi beam gain envelope in dBi as a function of boom length in λ_o; comparison with experimental data of Lindsay (+) and Ehrenspeck-Poehler (o).

really a smooth function of boom length. Instead, it appears to exhibit "bumps" or oscillations with a fraction of a decibel amplitude and spacing at about a half wavelength!

This concept of boom gain, independent of the number of elements is not new; in fact, it was suggested by Ehrenspeck and Poehler[6] in a series of experiments using the automatic plotter built at the Air Force Cambridge Research Center. No claims by Ehrenspeck and Poehler were made as to the absolute accuracy of their results but they were able to demonstrate essential independence of gain on number of elements over rather wide limits for two long Yagi models (1.2λ and 6.0λ). If one accepts the idea of universal boom gain, it is instructive to compare the (upper envelope) curve of **fig.** 14 with Ehrenspeck and Poehler's experimental points, as well as the experimental results of Lindsay.' Lindsay made a number of models of varying boom length (but unstated element dimension schedules) and measured directivity at a design frequency of **440** MHz. All of these results are shown in **fig. 15** where the solid curve is the theoretical maximum gain (from **fig.** 14) and the keyed points are from Ehrenspeck-Poehler and from Lindsay. The Lindsay experiments provide remarkable confirmation of the universal boom gain curve! The Ehrenspeck-Poehler points all appear to lie slightly below the theoretical curve (by a fraction of a decibel). It is not clear that the slight discrepancy in absolute value is a real disagreement; it may be within the expected accuracy of the gain calibration technique used on the automatic plotter. It may also be due to lack of optimization; Ehrenspeck and Poehler used a fixed reflector reactance and it is hard to guess how much more gain they would have found with an optimized configuration.

Fig. 16 taken from **table 4** shows a plot of the center-band F/B ratio as a function of overall length. It is notable that there are three empirical values of overall length which seem to produce high values of ***F/B independent*** of the number of elements! These apparently favorable overall lengths are 0.25, 0.75, and $1.25\lambda_o$ – all odd multiple of a quarter wave. For the $0.25\lambda_o$ position only the 3-element beam shows a high value, but this is primarily caused by the definition of ***center-band*** F/B ratio.

element illumination

This remarkable phenomenon suggests that there might be a basic physical explanation covering all cases indeed; such a physical basis is not hard **to** find! Analogous to the physical optical illumination of an aperture by light, one can think of the Yagi boom length as illuminated by (electrical) excitation. Unlike the case of uniform illumination of an optical aperture, the Yagi illumination is ***not*** uniform but can be viewed as a series of discrete excitation points (elements) whose average envelope is quasi-uniform. Moreover, in the optical case the wave front is ordinarily plane (phase shift across the aperture is zero), whereas in the Yagi case the phase shift is purposely designed to cause the main diffracted "beam" to lie along the boom rather than broadside to the aperture as in the optical case.

The aperture produces a diffraction pattern (beam pattern) consisting of a "main beam" and several lobes; the ***number*** of lobes is determined basically by the size of the aperture in wavelengths; the ***amplitude*** of the lobes is determined by the way the aperture is illuminated (phase and amplitude).

An informative treatment of an end-fire array (aperture illuminated by a series of radiators having equal amplitudes) is given in Kraus[4] on pages 76-89. There are two interesting cases: the ***ordinary*** end-fire array in which the angular phase change between radiators is just equal to their spatial separation angle, and the ***increased directivity*** end-fire array first de-

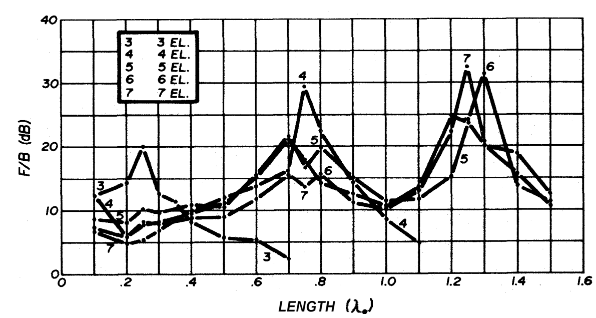


fig. 16. Band-centered front-to-back (F/B) ratio in dB for 3.4. 5.6. and 7-element Yagi beams as a function of boom length in λ_o.

table 5. Yagi null angles (fig. 17) as compared to ordinary end-fire arrays (OEF) and increased directivity end fire (IDEF).

boom length	freq	K = 1 null (degrees) Yagi	OEF	IDEF	K = 2 null (degrees) Yagi	OEF	IDEF	K = 3 null (degrees) Yagi	OEF	IDEF
0.25λ	0.972	—								
0.75λ	0.982	159								
0.25λ	0.992	120								
0.75λ	0.970	69	98.4	64.7	—	—	135.9			
0.75λ	0.980	66	97.7	64.3	171	—	134.5			
0.75λ	0.990	63	97.0	64.0	147	—	133.1			
1.25λ	0.976	51	71.5	48.8	102	111.5	91.4	—	—	135.0
1.25λ	0.986	51	71.1	48.6	99	110.6	90.8	171	—	133.7
1.25λ	0.996	48	70.7	48.3	93	109.8	90.2	150	—	132.3

rived by Hansen and Woodyard[8] where the radiator phase delay is larger than spatial separation by an angle π/n. The latter case is a good one with which to compare the Yagi antenna, because although the amplitude of the current(s) in the Yagi elements are not uniform, the phases are adjusted (by element reactance) to give highest directivity or gain. Kraus gives expected null angle directions for these two cases (null between lobes) as:

Ordinary end fire:

$$\theta_o = 2 \sin^{-1} \pm [K\lambda/(2nS)]^{1/2} \quad (1)$$

Increased directivity end fire:

$$\theta_o = 2 \sin^{-1} \pm [(2K-1)\lambda/(4nS)]^{1/2} \quad (2)$$

In our model the boom length ℓ_B is $(n-1)s$ so that we can rewrite these equations as:

Ordinary end fire (OEF):

$$\theta_o = 2 \sin^{-1} \pm [K(n-1)/(2\ \ell_B n)]^{1/2} \quad (3)$$

Increased-directivity (IDEF):

$$\theta_o = 2 \sin^{-1} \pm [(2K-1)(n-1)/(4\ \ell_B n)]^{1/2} \quad (4)$$

Note that where the number of elements, n, is large one would expect a high F/B ratio (a null at 180°) at particular values of boom length, ℓ_B, essentially independent of the number of elements!

Let's now examine the patterns of the cases where F/B is relatively high; I shall do this for the 6-element beam at three boom lengths: 0.25, 0.75, and 1.25 wavelengths long but will first find the precise frequency where the F/B ratio is maximum (presumably where the back radiation "null" occurs). It is instructive to also plot the pattern not only at this "best" frequency but also at frequencies just below (say −1 per cent and just above (say +1 per cent) of the best frequency. For all of these cases the reflector length was fixed at $0.50702\lambda_o$ $(FR = 0.95)$, and all director(s) length(s) were fixed at $0.45873\lambda_o$ (FR = 1.05).

Fig. 17 shows the H and E plane patterns of all of these cases. The H-plane pattern shown "nulls" between lobes which can move with frequency (equivalent to boom length in actual wavelengths). We can compare the angle at which these nulls occur to those of the end-fire arrays (**eqs.** 3 and 4); **table 5** lists these comparisons.

Note that these comparisons show qualitative agreement; also note that the computed Yagi results are in better agreement with the IDEF model (Hansen and Woodyard) than the OEF model. The more rapid shift of null angle(s) with frequency for the Yagi(s) compared to either end-fire model is not to be taken too seriously because, as we shift frequency, not only does the effective boom length in terms of actual wavelength change, but the element reactance(s) change significantly, i.e., the Yagi really becomes a different Yagi!

The details of the Yagi pattern depend on the particular way in which the boom is illuminated, i.e., on the details of element positions(s), element current magnitude(s), and element phase(s). The depth of the nulls depends on the degree of vectorial cancellation of back radiation; since the vectors themselves vary significantly with all Yagi parameters it is no wonder that complete cancellation is accidental and ordinarily impossible. The ***size*** of the lobes is determined primarily by the shape and phase delay of the Yagi ***illumination*** function. For the ***uniformly*** illuminated case the reader is referred to the uniform end-fire arrays (see Kraus[4], pages 79-88).

It will be noted that for non-uniformly illuminated

table 6. Element currents for a six-element Yagi beam with a boom length of 0.75λ; frequency F = 0.980 (assumes rf current of 1 ampere at 0° in the driven element).

element	current (amps)	phase (degrees)
Reflector	0.476	154.5
Driven element	1.000	0.0
Director 1	0.379	−121.3
Director 2	0.467	−169.7
Director 3	0.258	115.3
Director 4	0.458	32.6

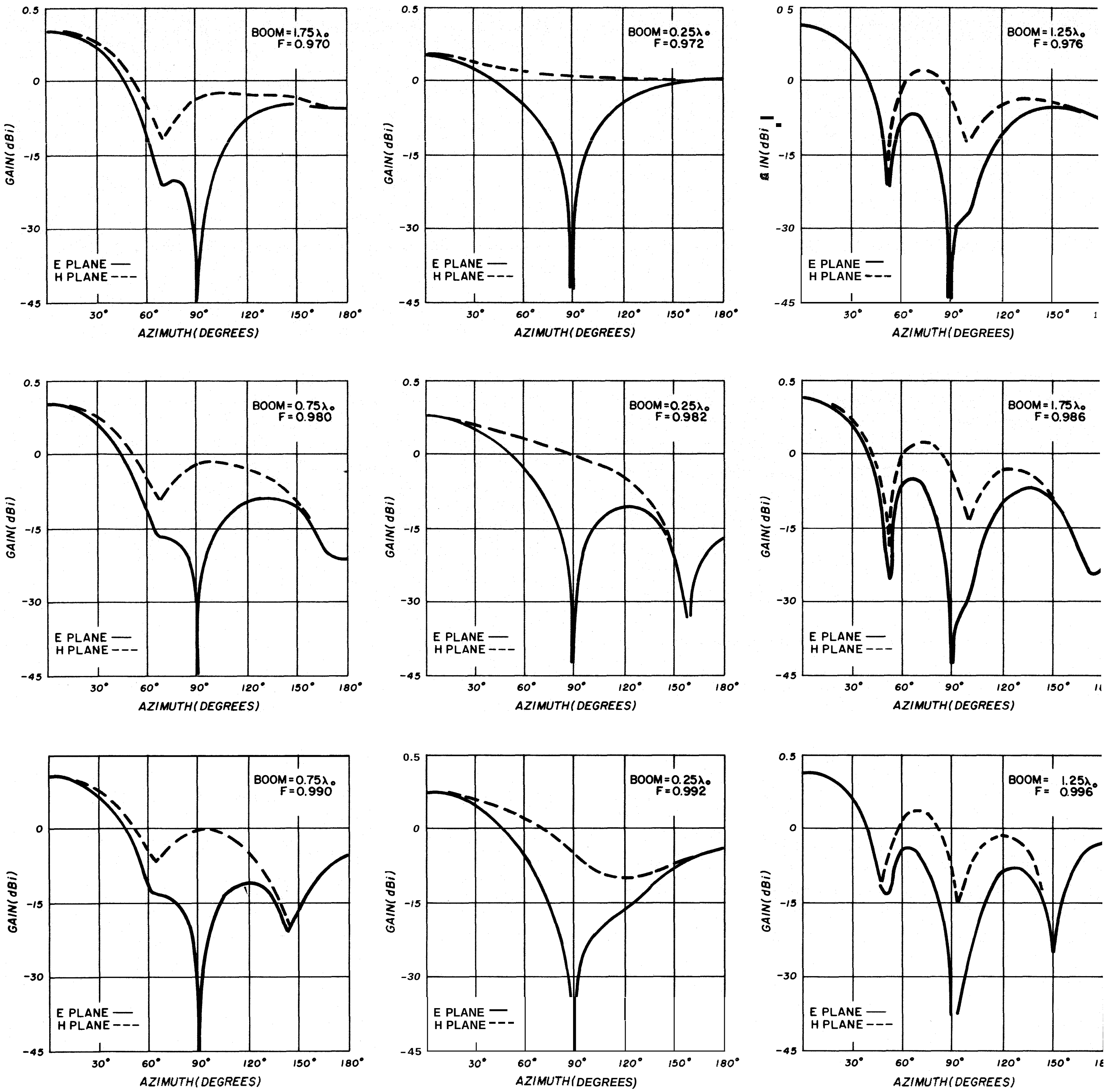


fig. 17. Yagi 6-element beam E and H-plane half patterns (0° to 180°) in dBi as a function of angle in degrees.

broadside structures (Kraus pages 93-121) the sidelobe level is highest for "edge" illumination, next for uniform illumination, and zero for illumination based on element amplitudes following the coefficients of a binomial series. One can expect the same kind of result for an end-fire array where edge illumination (2-element beam) produces high sidelobes, uniform illumination smaller sidelobes, and an illumination function falling at the extreme edges (reflector and end director currents smaller than in central elements) to produce still smaller sidelobes.

Unfortunately, for a given overall boom length the directivity or gain suffers somewhat as illumination is adjusted for smaller sidelobes. Moreover, in the case of a simplistic Yagi the illumination function is hard to adjust; the current amplitudes and phases are all

table 7. Computed F/B ratios for six-element Yagis with various boom lengths (see fig. 17).

length	freq.	F/B (180°)	E-plane F/B minimum (90°-180°)	H-plane F/B minimum (90°-180°)
0.25λ	0.982	25.7 dB	18.75 dB	7.5 dB
0.75λ	0.980	32.5 dB	19.65 dB	12.6 dB
1.25λ	9.986	35.0 dB	18.8 dB	14.9 dB

determined by element reactance(s) and position(s) and it is necessary to simply accept the result! To get an idea of the current(s) and phase(s) to be expected **table 6** shows the current(s) and phase(s) of the elements for the case of the 6-element beam, $boom = 0.75\lambda_o$, where the driver current is set at **1.0** ampere at 0° phase. Note that while the current amplitudes are not "uniform illumination," they seem to average out surprisingly alike! Incidentally, the current in the last director is always characteristically ***higher*** than that in the preceding director; this is due to the "end effect" (no mutual to an element ahead of it).

Thus we now have a consistent picture of the high F/B ratio Yagi design; the essence of correct design is to place the null between lobes exactly in the back direction! This will occur for simplistic Yagi antennas when the overall length is approximately an odd multiple of a quarter wavelength. The specific ***best*** design will involve optimizing the boom length and the boom ***illumination function*** (the particular element excitation currents and phases) to yield the best F/B ratio. Such optimization can be carried out around the best boom lengths; this will be the subject of a future article. For the present it is sufficient to note that really excellent F/B ratios are possible with these simplistic designs as long as one is willing to accept boom length(s) which are approximately odd multiples of a quarter wavelength.

As a final note on these simplistic ***best*** designs, not only is the back radiation low but the ***minimum*** (worst case) F/B ratio in the ***entire*** reverse direction (90° to 180°) is also surprisingly high! **Table 7** shows the result. These Yagi designs seem to be generally excellent!

With the exception of the 2-element beam case I have not yet commented on the driving point impedance of any of the Yagis shown. Remember that one can, by adjusting the length of the driver, always null out the driving point reactance at a designated frequency. The remaining resistance, however, just like that for the 2-element beam, varies enormously from case to case. It is very low for very short beams where there is very strong coupling between elements; moreover, in such cases it varies wildly with frequency as does the change in reactance with frequency. Thus to insure a reasonably reliable electrical feed system it is wise to keep element separation well above $0.05\lambda_o$; all such cases investigated have reasonably well behaved driving point impedances.

summary

Let me summarize the results for simplistic Yagi antennas:

1. 2 to 7-element beams with boom lengths to $1.5\lambda_o$ have been systematically explored.

2. Simplistic Yagis display a gain function where bandwidth is primarily a function of the resonant frequency separation between reflector and director(s). The bandwidth can easily be made several per cent of the central frequency.

3. The shape of the gain function is generally not flat in the region of interest. It is also more complex for beams which use a large number of elements.

4. Simplistic Yagis display a F/B ratio function with a shape that varies enormously from case to case. The shape may contain more than one peak and changes rapidly with boom length and/or frequency. It is so complicated that it is not possible to characterize its bandwidth.

5. High values of the F/B ratio (more than 30 dB) are quite rare; when they occur F/B is high only over a very narrow band of frequencies.

6. The spacing between elements should be generally greater than $0.05\lambda_o$ to realize a well-behaved feed.

7. The maximum practical gain of the simplistic Yagi is almost entirely determined by boom length. Maximum gain increases, but not steadily with boom length.

8. Best design for a high F/B ratio requires the approximate boom length to be an odd multiple of a quarter wavelength at the design frequency.

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ham radio

phased vertical antenna
for21 MHz

An economical approach
to beam antennas
for the
15-meter band

Most of the Amateur antenna handbooks don't describe the phased vertical beam. For economy and performance they should be used more! I built a phased vertical beam using the references indicated.[1-4] I hope the explanation given here will motivate others to try this antenna. There's probably an easier way to construct the beam. W7EL describes another method of matching in reference 4. However, I chose to use the more classical Wilkenson match. Perhaps there should be more experimenting and discussion on feeding this beam. W7EL's method would do away with the problem of locating 100-ohm noninductive resistors for termination and balance.

phased vertical array

The phased vertical beam has some good features. **Fig. 1** shows horizontal radiation patterns for a two-element phased vertical array using three different phase angles for feeding the two elements. If both elements are fed in phase (zero-degree phase angle), the radiation pattern will be bidirectional at right angles to the plane of the two elements. This is known as broadside radiation. If the antenna elements are fed 180 degrees out of phase, an end-fire pattern results, producing bidirectional coverage in the plane of the elements (end fire). If one element is fed 90 degrees out of phase with respect to the other, a unidirectional pattern is obtained. Thus, with two vertical elements, properly phased with relay switching, you can cover all four quadrants of the compass.

I was interested in only one direction, and it just happened that my brick fence was in a line where I wanted the antenna pattern to go. The rest was easy, because only holes had to be drilled with a concrete drill and lead slugs driven in place to mount the element insulators.

construction

Fig. 2 shows construction details for my two-element phased vertical beam. I obtained two pieces of 112-inch (12.5-mm) diameter aluminum tubing 12 feet (3.7 meters) long. I cut each tube at 11 feet (3.4 meters) for the 114-wavelength antenna elements. The formula is:

$$L = \frac{468/f}{2} \qquad (1)$$

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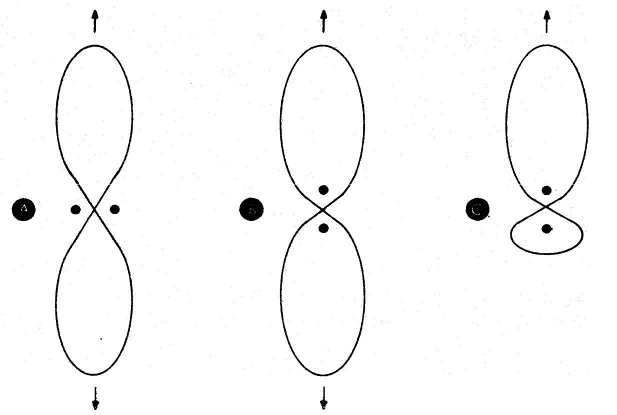

fig. 1. Radiation patterns for a two-element phased vertical array using three different phase angles for feeding the system. A shows the pattern when the elements are fed in phase. If the elements are fed 180 degrees out of phase, pattern *B* results. C shows the pattern when the elements are fed 90 degrees out of phase.

where L = length, feet

f = frequency, MHz

In metric terms **eq. 1** is:

$$L = \frac{145/f}{2}$$

Thus for my operating frequency (21.27 MHz) the element lengths were 11 feet or 3.4 meters. I mounted the two 1/4-wavelength elements on the fence, spaced 11 feet (3.4 meters) apart.

feed system

Any length of 52-ohm coax cable can be run to the antenna from the transmitter. The transmission line is connected to a coax T connector, as shown in **fig. 2**. I made two 1/4-wavelength sections of 72-ohm line (RG-11/U) and connected each line to the T connector. The other ends of the 1/4-wavelength lines were terminated with a 100-ohm noninductive resistor as shown.

The 1/4-wavelength coax line lengths are somewhat critical, as the slightest amount of variation at 21 MHz changed the frequency a great amount. These lengths were determined from:

$$L = \frac{246V}{f} \quad \textbf{(2)}$$

where V = cable velocity factor

f = frequency, MHz

Thus:

$$L = \frac{246\,(0.66)}{21.27} = 7.6\,feet$$

$$L = \frac{74\,(0.66)}{21.27} = 2.3\,meters$$

I checked the coax length with a grid-dip oscillator. Any loop in the cable will give a false reading of the length. K6DS came up with the idea of using an inch-wide (25.5 mm) piece of PC board with two holes and soldered up short. The wide surface provided enough pickup for the grid dipper. Putting the plugs on the ends lowered the frequency.

Each piece of coax I checked had a different vel-

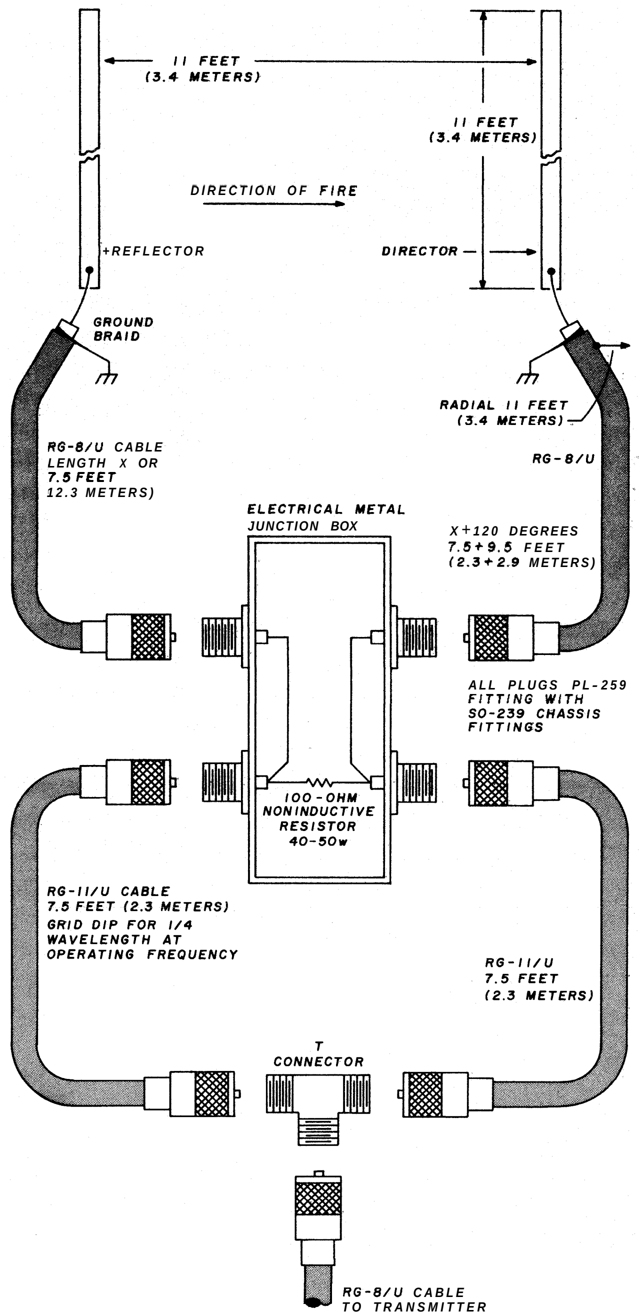

fig. 2. Construction details for the phased vertical beam for 21 MHz. Derivation of phasing-line lengths are discussed in the text.

ocity factor, V, varying from 0.6 to 0.68. You can check your coax by

$$V = \frac{fL}{246} \tag{3}$$

where V = velocity factor
L = length, feet

In metric terms **eq.** 3 is

$$V = \frac{fL}{75}$$

where length, L, is in meters.

final assembly

Fig. 2 shows the details of my array. The junction box, containing the 100-ohm noninductive resistor and the SO-239 coax chassis receptacles, was mounted on the brick wail. I coiled the phasing lines and attached them to the wall. The radial wires are 11 feet (3.4 meters) long. **Fig. 2** shows the method of attachment. For best results run as many radials as possible.*

A quarter-wavelength of RG-8/U cable, 7-1/2 feet (2.3 meters) long was connected between the junction box and reflector element. Another coax line (RG-8/U) 7-1/2 feet (2.3 meters) plus 120 degrees was connected to the director. (See **fig.** 3). This length was 17 feet (5 meters). This line isn't too critical because it mostly affects the antenna back lobe.

results

When the antenna was finished I walked around it with a field-strength meter. The readings concurred with the pattern in reference 3. It was surprising to see the signal fall off at 90 degrees to nothing on the back. I made the first on-the-air checks locally using four stations about 10 miles (16 km) away. Two of the stations were in front of the antenna and two were off about 45 degrees to the side. Stations off to the side noticed the same signal strength as my two-element Yagi. However, those in front couldn't hear the signal from my Yagi but reported S9 signals from the phased vertical. This would indicate that the phased vertical antenna has a lower radiation angle than that of the two-element Yagi mounted at 20 feet (6 meters) above ground.

Reports from stations in South America were varied. With some there was a noticeable difference between antennas; with most I received the same signal report. (One fellow said the phased array was louder than horseradish, a scale I'm not familiar with.) When signals from one antenna fade out, signals from the other will be maximum. It's interesting to switch between antennas.

"In a similar design for twenty meters5 the author used ground rods for each element as well as radial wires. As with any vertical antenna, the idea is to reduce ohmic losses in the system. Editor.

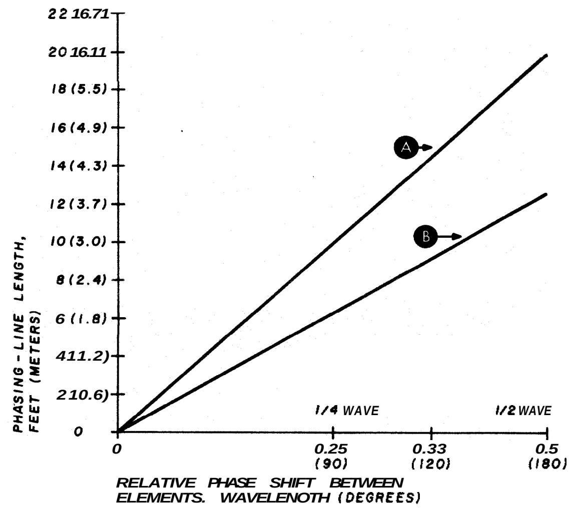

fig. 3. This curve may be used to determine the length of the 120-degree phasing line for the director element in fig. 2. A is the curve without velocity factor. B the length with velocity factor.

summary

To summarize my findings, I'd say that if I didn't have a two element Yagi beam, rotor and tower I wouldn't hesitate to use two of these phased vertical beams in place of it! The construction is far cheaper and I can find very little difference between the two. There's more of a null on the phased array off the back and sides, and the beam appears to be narrower than that of the two-element Yagi. The vertical antenna did not do what I had hoped: to get through a maze of power lines 100 feet (30 meters) away.

The most difficult part of the construction is locating 100 ohm noninductive resistors of 40-50 watts. The transmitter Looks at low SWR and loads up very nicely.

The ideal of W7EL might be a solution to feeding the antenna without a termination resistor and just using RG-8/U coax from the T connection. It will be interesting to hear the experiences of others and get this simple beam into more use. It is strange that it hasn't found much use before, because the Wilkenson match is well known in uhf work. Many of the 160-meter stations have been using it for a long time, anyone who has heard them will realize the advantage for low-frequency work where a beam other than vertical is almost impossible to erect.

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ham radio

antenna restrictions — another solution

How to install
a full-size
40-meter dipole
inside a mobile home

Antenna restrictions often make life difficult for hams who live in one of the new mobile-home parks. In fact, restrictions can make it almost impossible to get on the air without using a high degree of ingenuity. Articles have been written describing the use of fake (and not so fake) flagpoles as vertical antennas, but the problems of ground radials and low radiation angles limit the effectiveness of this kind of solution.

Lately many of the new double-wide modular homes are being built using conventional framing and roof construction with wood joists and rafters, wood sheathing and composition shingles. This leaves only the matter of space in the attic — how do you hide an 80-meter or even a 40-meter antenna inside if the long dimension is only 12-15 meters (40-50 feet)?

15-meter antenna

In my situation, our home is a 7.3-meter (24-foot) wide double unit in a park where restrictions prohibit outside antennas. As a retiree ham, off the air for about 40 years until 1978, I was able to solve my own antenna problem this way: For 15 meters I made up a simple dipole using four pieces of 1.8-meter (6-foot) long telescoping aluminum tubing, fed directly with 50-ohm coax and secured inside the attic close to the apex of the roof rafters. It works quite well, with low SWR and easy loading to one of the new "touchy"

By J. W. Bryant, N4AQD, 4736 Dauphine Boulevard, Tallahassee, Florida 32303

transceivers. As an afterthought, because I wanted to work primarily in one direction, I added a passive reflector mounted ***outside*** on the roof, down the slope far enough to provide the proper spacing. The reflector tubing was fastened to four small L brackets slipped under the shingle edges.

40-meter antenna

The 40-meter solution was a little more trouble. My mobile home is about 12 meters (40 feet) long, inside measure, so there was no way to insert a horizontal full-length halfwave dipole. I stretched the wire inside the limited attic space, then shoved the remaining lengths at each end (to make a total of 19.5 meters, or 64 feet) at 90 degree angles to the attic corners. This was done with 3.7-meter (12-foot) cane fishing poles, which I left in place to keep the wire extended.

To have some leeway to tune out any reactance, I improvised a tuner using series inductance and capacitance in each leg of the antenna at the feed point. I ended up with the arrangement shown in **fig. 1.** The center of the antenna and tuner were accessible through a hatch cut into the ceiling of the mobile home. Fortunately the location was right over the laundry alcove.

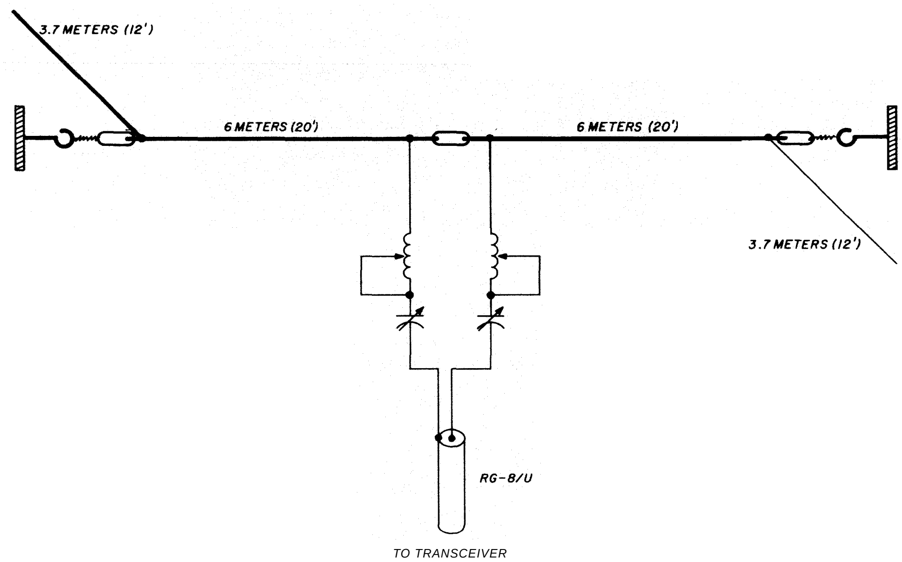

fig. 1. Author's inside 40-meter dipole. Horizontal sections occupy the length of the mobile-home attic. The end sections follow the corners of the attic to make a total radiator length of 19.5 meters (64 feet!. The tuner provides near unity SWR in the phone portion of the band.

tune up

For final tuning I used a dip meter coupled to a one-turn loop at the line-to-tuner feed point to determine the approximate resonance adjustment of the series reactances. Then I inserted an SWR meter at the antenna end of the coax and adjusted each side of the tuner for lowest SWR, which came out to be very near unity at the phone end of the band. I've not tried to tune the antenna to 75 meters, but I'm sure it can be done with added inductance in each tuner leg.

Contacts on both 15 and 40 meters have given good reports, even before I told them of the indoor nature of the antenna arrangement.

further reading

The bibliography provides other interesting approaches to the problem of erecting Amateur antennas in locations that restrict outside structures. Allen Ward, KA5N, describes a modified ZL Special that can be mounted in an apartment room. Spence Collins, N6SC, installed a 7-MHz Hertz antenna in his first-floor apartment – stretched approximately 8 meters (26 feet), with the highest point 2 meters (7 feet) from the floor. And for vhf buffs, Warren Hodges, WGDHX, and Bill Wise, WBGQEZ, tell how to camouflage a 2-meter antenna using a weathervane atop the house, which is in a restricted area.

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Ward, Allen **C.,** KA5N, **"10** *Meter* ZL *Special Antenna for Indoor Use," ham radio, May,* 1980, *page* ***50.***

ham radio

tone-encoder
for 2-meter autopatches

A design that produces clean sine waves tamed by AGC amplifiers for constant level — also featured is automatic PTT with delayed release

As the use of the Touch-Tone* system – both in telephones and in Amateur Radio – became widespread, a number of ICs were developed to generate the tones. Among the most popular are the Motorola MC14410 and the Mostek MK5087; there are several others on the market. These chips are generally superior to those in the original Western Electric *Touch Tone* (or *DTMF)* encoder, with its cup-core inductors. But they do have one disadvantage in common: they produce the tones by digital methods, so the outputs are not pure sine waves but are similar to the waveform shown in **fig. 1.** In telephone applications this doesn't matter, since the line itself acts as a lowpass filter, effectively restoring sine-wave purity. Also, being designed to work into such lines, the chips have greater output at higher frequencies to counteract the natural attenuation of the line.

When such a *Touch Tone* chip is used to activate a repeater autopatch through a 2-meter fm transmitter, these natural filter effects don't exist. Thus, in addition to harmonic distortion, the repeater autopatch must cope with deviation levels that vary according to the tones selected. This can cause problems in completing a call.

I designed a DTMF board that produces good, clean sine waves. In addition, it has agc amplifiers to keep the tones at constant levels. There's also provision for keying the transmitter automatically whenever any button of the *Touch Tone* pad is pressed and keeping it keyed until the dialing of the number is complete. The circuit will fit on a 3 x 5 inch (7.6 x 12.8 cm) board. After describing the circuit, I'll go in-

Touch Tone is the registered trademark of the American Telephone and Telegraph Company.

By **Chris Winter, WB0VSZ,** 610 South Clinton, Iowa City, Iowa 52240

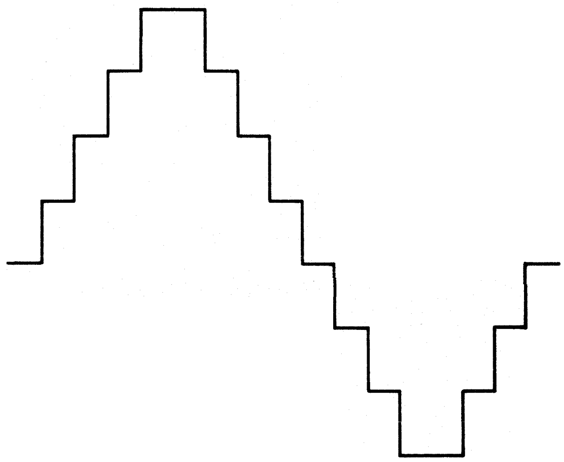


fig. 1. A sketch of the typical output of the MC14410 shows that it is not a true sine wave. This output is produced by a 16-stage walking-ring counter.

to the fine points of assembling and using the board — or of rolling your own, if that's your style.

circuit description

The tones are generated by a Motorola MC14410P. The advantage of this chip is that it has separate outputs for the low and high tone groups; this makes it easier to filter the tones. Since the MC14410 is widely used, there's no need to go too deeply into the circuit. The tone outputs are produced by walking-ring counters driving weighted-resistor networks. In addition to a dc level, each output waveform contains harmonics whose order is determined by the number of stages in the counter. A good source on this subject is reference 1.

The main schematic of the DTMF board is shown in **fig. 2.** The wiring of U1 is standard except for two things: capacitors C1-C8 are for RFI protection, and the output of pin 7 — a square-wave test signal, not normally used — drives the automatic keyer circuit.

filter considerations

The filters don't have to be extremely sharp or well-centered on the nominal center frequency, since they don't have to separate the tones. Bandpass filters are required to block both the dc levels and the high-order harmonics in the outputs from U1. I used a multiple-feedback configuration because it permits a bandpass filter with a single op amp. One stage per filter gives sufficient selectivity; this allows both filters and the two agc amplifiers to be built using a single quad op amp IC.

In reference 2, Moberg described design equations for the multiple-feedback filter where the two capacitors are of unequal values (see appendix). These equations give great flexibility in choosing the filter parameters and are easily turned into a program for a handheld calculator.

The design procedure starts with the values of the two capacitors and the three filter parameters: Q, gain A_0, and center frequency, f_0. The values of the three resistors are calculated (for filter-component designations see **fig.** 3). U1 will produce roughly 2 or 3 volts p-p. To avoid overdriving the agc amplifiers, a

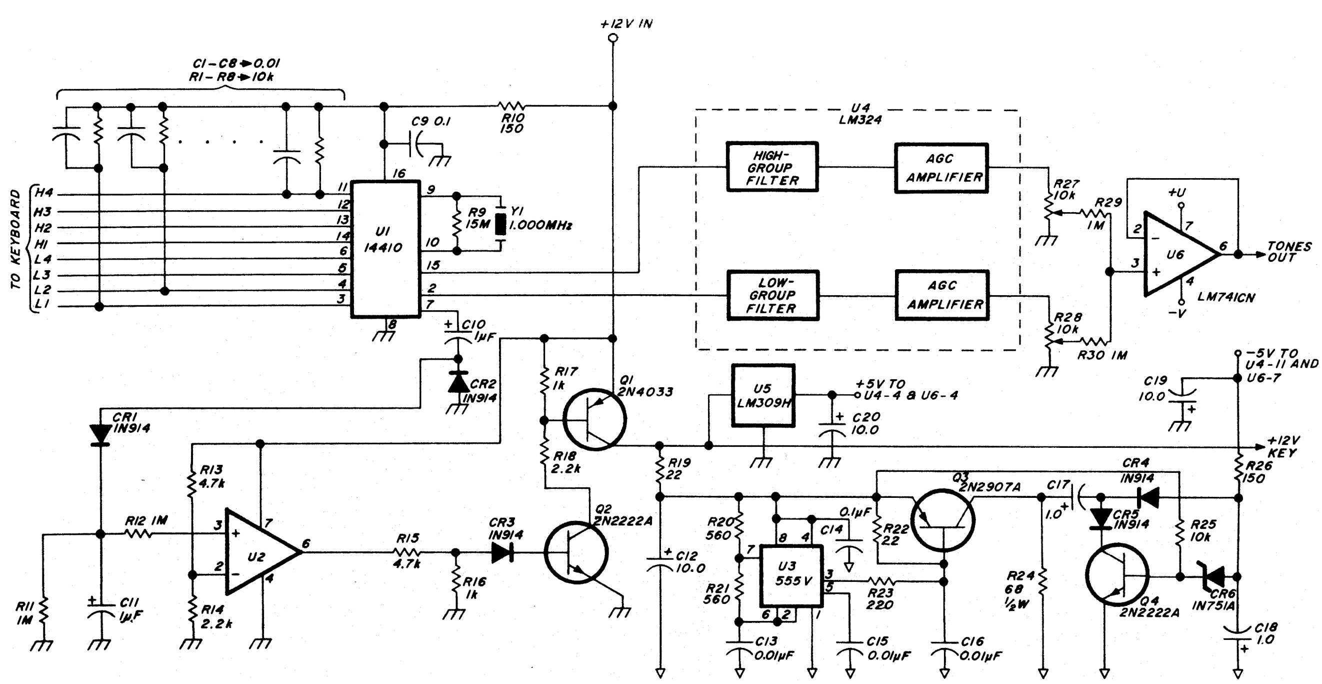


fig. 2. The main schematic of the DTMF board. Note the liberal use of bypassing and decoupling networks. To avoid duplication, filters and amplifiers are not shown in this illustration.

filter gain of unity or less should be used. Table 1 summarizes the results for the parameters I used in my design.

The components used for the filters are standard non-precision types. The response curve of **fig. 4** shows that the results with such components are quite close to the ideal. There are two reasons that I "broke up" R1; it made the layout come out better and allowed easy matching of the calculated value of R1. The 100-pF capacitor is another rf bypass.

keeping the outputs constant

The agc amplifier I use is based on one published in reference 3. It calls for a high-quality fet to give wide dynamic range and excellent linearity; here, there's no need for those characteristics, and substitutions can be made. My version of the circuit is identical to that of reference 3 except for the substitutions; it's shown in **fig. 5.** When the input level is below the threshold of compression (about 50 mV) Q1 gate is unbiased, and the full op amp gain of 70 is in effect. Larger amplitudes are rectified by Q2 and bias Q1 gate upward from its normal voltage of – 5V. QI's channel resistance drops, effectively pulling signal away from the op amp until equilibrium is reached. The attack time is a few milliseconds. R4 and C1 provide about one second of decay time. The compressed output is 1 volt p-p.

table 1. Summary of the design parameters used for the filters and the calculated resistor values.

	low group	high group
A_0	0.5	0.5
f_0	819.0	1,421.0
Q	3.357	3.351
C1(μF)	0.022	0.022
C2(μF)	0.068	0.068
R1(ohms)	19,195.0	11,040.0
R2(ohms)	665.6	384.2
R3(ohms)	39,241.0	22,582.0

The agc outputs amplifier go to a pair of pots so that the tone levels can be individually adjusted. Finally, U6, a voltage follower, combines the tones and provides a low-impedance output suitable for connection to the high-level audio input of any transmitter.

the split-supply problem

It would have been possible to power all the op amps from +12V and ground. Instead, to eliminate the need for coupling capacitors, I decided to run U4 and U6 from a split supply of ±5 volts. Then of course I was faced with the familiar problem of how to get a negative supply voltage in a vehicle.

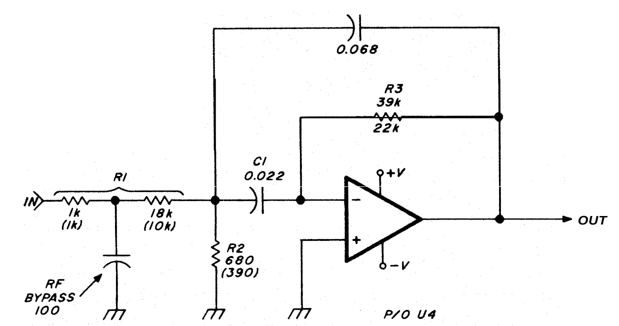


fig. 3. The two filter circuits differ only in resistor values. Those in parentheses are for the high-group filter.

I could have used a transistor radio battery, but it probably would have gone dead at a cruciai moment. Luckily, I found an inverter circuit which suited my needs perfectly.[4] It consists of a high-frequency pulse source, a bootstrap inverter, and a shunt regulator. I changed the oscillator to a 555 timer IC, U3, running in astable mode at about 100 kHz. U3 drives Q3, which supplies solid 12-volt pulses to R24 and C17. As the voltage at the junction of R24 and C17 falls from 12 volts to zero, the other side of C17 is forced to go negative; this series of spikes is then smoothed by C18. With a 12-volt input, you can get up to – 7 volts out. The shunt regulator (Q4, CR5, and CR6) holds the output constant within half a volt as the load varies from zero to 20 mA. In normal operation, the current drawn will be no more than 2 or 3 mA. With the zener rating shown, the inverter output should be close to – 5 volts. More about this later.

automatic PTT

One drawback of the MC14410 is that, unlike many DTMF chips, it lacks an output that can serve as an "any-key-pressed" indicator. But, as I mentioned, it

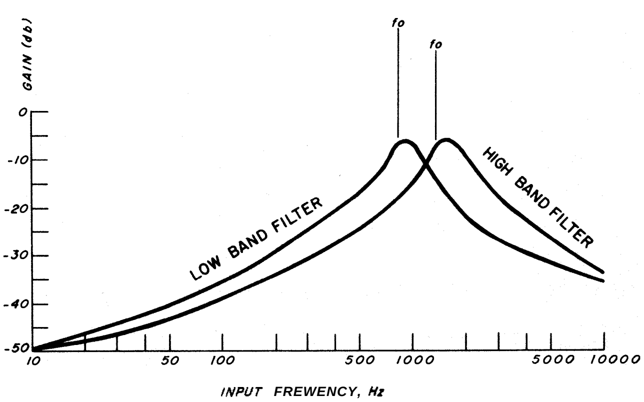


fig. 4. Filter response curves come close to the ideal; precision components are not required.

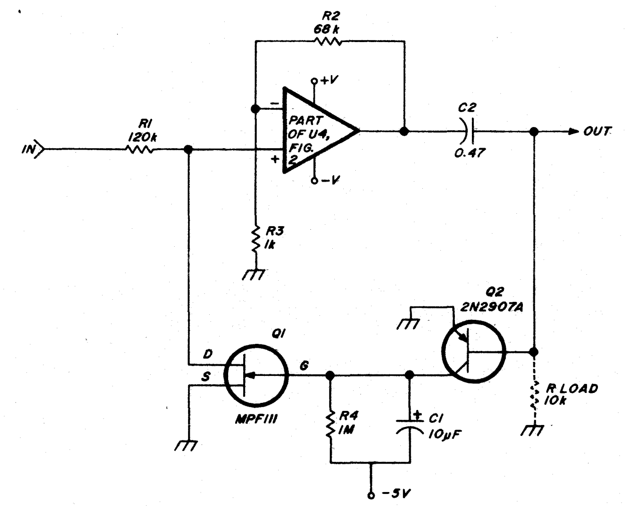


fig. 5. Simplicity and inexpensive components are the virtues of the amplifiers.

does produce a square wave on pin 7 when any key is pressed. This can be rectified and filtered, then used to key a transmitter.

In my circuit, U2 is wired as a comparator and drives Q1 and Q2. When any tone is activated, C11 will quickly charge to 8 volts. R13 and R14 set the threshold at 3.8 volts so that U2 output goes high (about 11 volts), and the full 12 volts appears on the keying line. This line can supply 300 mA with negtigible drop in voltage. Because of the RC time constant at U2 noninverting input, the keying line will remain high for three seconds after the key is released; thus, keys can be pressed at a reasonably slow pace without losing the carrier.

assembly and use of the module

For those who desire to build this DTMF module, I have a PC board available for $8.50 postpaid. The board is double sided, with plate-through holes, and plugs into a 15-pin edge connector (pin spacing 0.156 inch, or 0.4 cm). Including the connector pattern, the board dimensions are 3 x 5 inches (7.6 x 12.7 cm). Even if you don't go this route, I recommend using a PC board of some kind. It's important to minimize the switching noise radiated by the inverter, and a PC board does this better than point-to-point wiring. Conducted noise can also cause trouble; note the extensive use of decoupling networks and bypassing in my circuit. The bypassing helps with RFI but is not a cure. For that you need a tight metal enclosure.

component tolerances

While we're on the subject of the inverter, be warned that it's very finicky. That is, it's easily rendered inoperative by component tolerances. R24 is the key to the puzzle; if its value is too high, the inverter will not supply any current. The first unit I built worked well with 390 ohms; the second required less than 100 ohms. The zener can be another source of trouble. You may have to try several before you find the rating that gives you the output voltage you want. This, too, depends on R24.

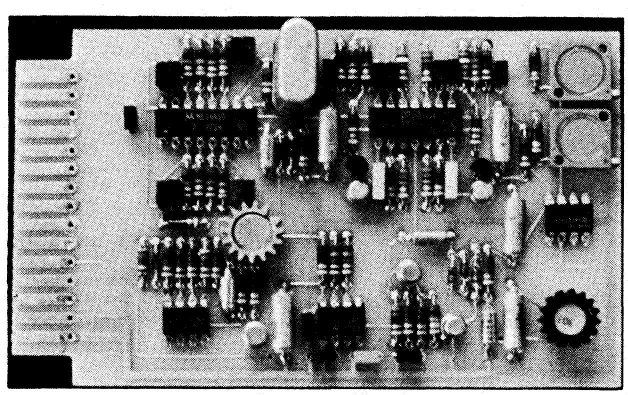
The following procedure works well: load the inverter output with a 1k resistor and select R24 by working down from 470 ohms. Go two steps lower still to give yourself a safety margin. Then make sure you have the right zener rating. Note, too, that for values between 50 and 100 ohms, R24 can get rather warm. You could probably get by with a quarter-watt resistor, but use a half-watt type and avoid the worry. It's a real hassle to test-select these components, but once it's done, the inverter will work reliably.

The components will tolerate normal supply voltage variations present in most vehicles, but it wouldn't hurt to limit the voltage to the 12-15 volt range. With a 12 volt supply and no external load on the keying line, the board will draw 25 mA in standby. When in use, the current will jump to 80 mA; the difference is due to the fact that only U1 and U2 are always powered up. The rest of the circuit is energized from the keying line.

interfacing

The DTMF module requires the following connections to the outside world: eight tone-select lines, the keying line, the tone output, and power and ground. All you should need in the way of off-board parts is a 12- or 16-button keyboard and something to interface the keying line to your rig's PTT line. This could be a relay or a power transistor.

If you want to modulate the rig through the mike input, you may need a fairly large series resistor to get



The completed prototype. Tho ICs, clockwise from the upper left are U1, 14410, and U4, LM324, both 14-pin DIP; U6, 741 op amp in an 8-pin DIP; U5, LM309H in a TO-5 package; and U3, 555 timer, and U2, 741 op amp, both in 8-pin DIPs.

the right signal and impedance levels. In other words, this board hooks up to the transceiver just like most *Touch Tone* pads. I intended it to be usable with a minimum of external parts but also to be part of a sophisticated access system. The system I have in mind includes an automatic dialer board and a third board that can automatically access one or several closed autopatches. These other two boards are currently under development and should soon be ready.

appendix

1. Filter design equations. Eqs. **1** through **3** are Moberg's (from reference 2). Eqs. **4** through **6** show how to design filter parameters A_0, Q, and f_0 from component values.

$$R_3 = \frac{Q(C1 + C2)}{2\pi f_0 \, C1C2} \qquad (1)$$

$$= R_3 \left[\frac{C1}{A_0(C1 + C2)}\right] \qquad (2)$$

$$R_2 = R_1 \left[\frac{A_0C2}{Q^2(C1 + C2) - A_0C2}\right] \qquad (3)$$

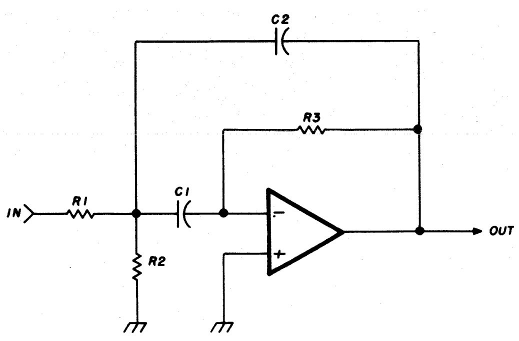
$$A_0 = \frac{R_3 \, C1}{R_1(C1 + C2)} \qquad (4)$$

$$Q = \sqrt{\frac{A_0 \, C2 \, (R_1 + R_2)}{R_2 \, (C1 + C2)}} \qquad (5)$$

$$f_0 = \frac{Q(C1 + C2)}{2\pi \, R_3 \, C1 \, C2} \qquad (5)$$

2. Calculator programs. The two programs following may be used with the HP-25C to calculate resistor values and the three filter parameters, A_0, Q, and f_0 for the multiple-feedback filter.

Program uses equations derived by Moberg.2 To calculate resistor values for multiple-feedback filter:



Input values required are $C1$, $C2$, A_0, f_0, Q

$$R3 = \frac{Q(C1 + C2)}{2\pi f_0 \, C1 \, C2} \qquad (1)$$

$$R1 = R3 \left[\frac{C1}{A_0(C1 + C2)}\right] \qquad (2)$$

$$R2 = R1 \left[\frac{A_0C2}{Q^2(C1 + C2) - A_0C2}\right] \qquad (3)$$

No warning needed. Simply store desired values in R1-R4, run program, and recall answers from R5-R7.

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	RCL 0	24 00	
	RCL 1	24 01	
	+	51	
	STO 3	23 03	
	RCL 7	24 07	
	RCL 5	24 05	
	÷	71	
	RCL 0	24 00	
	×	61	
010	RCL 3	24 03	
	÷	71	
	STO 2	23 02	Value of A_0
	RCL 5	24 05	
	RCL 6	24 06	
	+	51	
	RCL 1	24 01	
	RCL 2	24 02	
	×	61	
	×	61	
020	RCL 5	24 05	
	RCL 3	24 03	
	×	61	
	÷	71	
	f $\sqrt{x}$	14 02	
	STO 4	23 04	Value of Q
	RCL 3	24 03	
	×	61	
	2	02	
	g π	15 73	
030	×	61	
	÷	71	
	RCL 7	24 07	
	÷	71	
	RCL 0	24 00	
	RCL 1	24 01	
	×	61	
	÷	71	
	STO 3	23 03	Value of f_0
	GTO 00	13 00	
040			
050			

REGISTERS

0	1	2	3	4	5	6	7	8	9
C1	C2	A_0	f_0	Q	R1	R2	R3		

S0	S1	S2	S3	S4	S5	S6	S7	S8	S9

A	B	C	D	E	I

User Instructions

STEP	INSTRUCTIONS	INPUT DATA/UNITS	KEYS	OUTPUT DATA/UNITS
	Store the following values;	C1, F	STO 0	
		C2, F	STO 1	
		GAIN, A_0	STO 2	
		f_0, Hz	STO 3	
		Q	STO 4	
			R/S	
	Run program		RCL 5	R1, Ω
	Recall resistor values			
			RCL 6	R2, Ω
			RCL 7	R3, Ω

Program calculates the three filter paramaters A_0, f_0, and Q from resistor and capacitor values. Equations involved are:

$$A_0 = \frac{R3 \, C1}{R1 \, (C1 + C2)} \qquad (1)$$

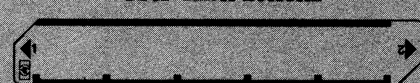
$$Q = \sqrt{\frac{A_0 \, C2 \, (R1 + R2)}{R1 \, (C1 + C2)}} \qquad (2)$$

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	RCL 4	24 04	
	RCL 0	24 00	
	RCL 1	24 01	
	+	51	
	×	61	
	RCL 3	24 03	
	2	02	
	×	61	
	gπ	15 73	
010	×	61	
	÷	71	
	RCL 0	24 00	
	RCL 1	24 01	
	×	61	
	÷	71	
	STO 7	23 07	Value of R3
	RCL 0	24 00	
	RCL 1	24 01	
	+	61	
020	RCL 2	24 02	
	×	61	
	÷	71	
	RCL 0	24 00	
	×	61	
	STO 5	23 05	Value of R1
	RCL 1	24 01	
	RCL 2	24 02	
	×	61	
	×	61	
030	RCL 4	24 04	
	gx^2	15 02	
	RCL 0	24 00	
	RCL 1	24 01	
	+	51	
	×	61	
	RCL 1	24 01	
	RCL 2	24 02	
	×	61	
	−	41	
040	÷	71	
	STO 6	23 06	Value of R2
	GTO 00	13 00	

REGISTERS

0	1	2	3	4	5	6	7
C1	C2	A_0	f_0	Q	R1	R2	R3

User Instructions



STEP	INSTRUCTIONS	INPUT DATA/UNITS	KEYS	OUTPUT DATA/UNITS
	STORE COMPONENT VALUES IN R0, R1, R5, R6, R7			
	RUN PROGRAM			
	RECALL ANSWERS FROM R2 - R4			

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ham radio

More Details? CHECK—OFF Page 94

solid-state T-R switch
for tube transmitters

Combine the advantages
of full break-in
CW operation
with the cost savings
of older equipment

A recent article in **QST** by Dave Shafer, W4AX, explained the advantages of full break-in QSK CW operation,' advantages that can't be provided in today's high-priced transceivers. The article brought to mind another by Stu Goodman, K2RPZ, which appeared a year earlier.* He was concerned with the economics of getting on the air and pointed out the possibilities of using older equipment. To quote Stu, "All that is needed is a T-R switch and an antenna . . ." The T-R switch described here uses solid-state components and provides the capability of full break-in operation with low-cost used transmitters with vacuum tubes.

T-R switches

The origin of the T-R switch stems from early radar days. An automatic device was required in the radar to prevent transmitted energy from reaching the receiver, but allowing the received energy to do so without appreciable loss. For fast, reliable break-in CW operation, using a single antenna for transmitting and receiving, the T-R switch is used today. However, with transceivers now dominating the Amateur-equipment market, T-R switches have all but faded from sight.

I checked several references in this regard with no results. Either they were too new and didn't mention T-R switches, or were so old that only vacuum-tube circuits were shown.

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Reference 3 provided some good design concepts. However, these designs were more applicable to solid-state transmitters. Between the extremes of too old and too new, I found a design by W4ETO that was described by W1ICP in the April, 1971, edition of QST (it later appeared in several editions of ***The*** Radio ***Amateurs*** Handbook.4) I modified the design slightly to improve some operating parameters. It should work well with any moderate-power, class-C vacuum-tube amplifier (such as a pair of 6146s with a plate supply of 750 Vdc).

theory of operation

A short expianation of how the T-R switch operates should be helpful. The principal purpose of the T-R switcn is to aiiow both the receiver and transmitter to be directly connected to the antenna. When the transmitter is keyed, the switch should reduce the amount of signal reaching the receiver to a safe level, but otherwise provide a near unity gain path between antenna and receiver. The noise figure of the device should also be low enough to prevent loss of receiver sensitivity on the'upper high-frequency Amateur bands. Some of the older T-R switches were designed to be placed at the transmitter output, but in some cases they caused RFI problems. While these circuits did protect the receiver, harmonics were generated in the T-R switch when the transmitter was keyed. To eliminate the harmonics associated with the circuit, the T-R switch is connected to the tube side of the transmitter pi-network; not to the antenna side. Any harmonics generated in the switch wiii be attenuated by the lowpass characteristic of the pi-network. The pi-network will also act as a preseiector, attenuating out-of-band signais.

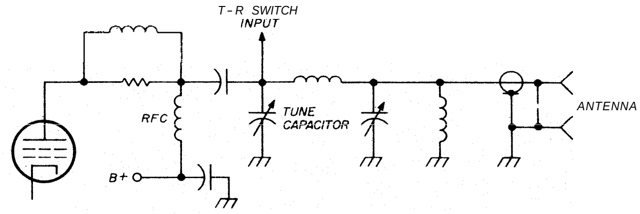

fig. 1. Typical vacuum-tube transmitter output circuit showing the pi network and connection point for the T-R switch.

Most transmitters use shunt-fed, class-C, final amplifier stages and pi-network impedance-matching networks, as shown in **fig. 1**. The input to the T-R switch is taken from the TUNE capacitor in the pi-network, rather than from the plate rf choke, so that high plate voltage (B+) won't have to be blocked in the switch. When the transmitter is keyed, the vacuum tube sends current pulses into the pi-network that are filtered before reaching the antenna. When the key is up, the final amplifier tube is cut off and has no effect on the received signal. Note that a T-R switch can be used only with class-C final amplifier stages. With linear class A or AB amplifiers, current flows in the final device even when no signal is being

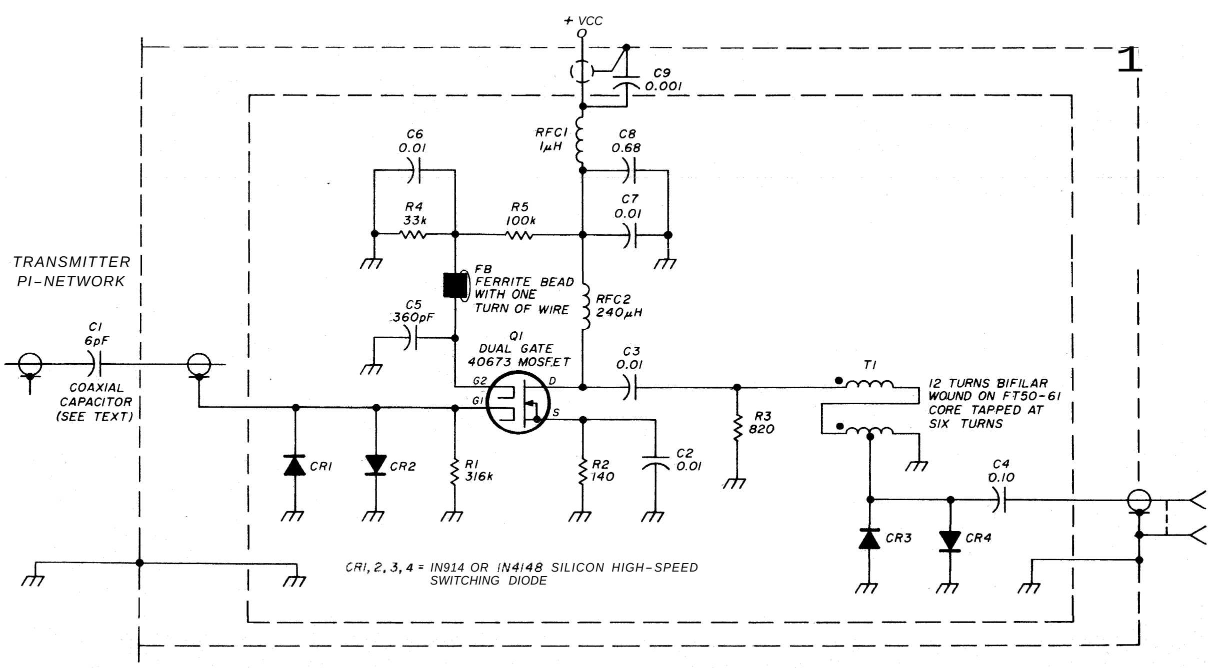

fig. 2. Schematic of a solid-state T-R switch that is suitable for use with moderate-power, vacuum-tube transmitters.

amplified. This action allows amplifier noise to pass directly into the receiver.

circuit details

The T-R switch circuit is shown in **fig. 2** and differs from W4ETO's original design in several ways. One is that a common-drain amplifier is used so that the overall gain can be set at unity. The pi-network transformation steps up the input voltage by the square root of the transformation ratio.

The switch input capacitor C1, and the parasitic capacitances of the two dioides and the mosfet form a voltage divider that reduces the signal level very close to the value it originally had at the antenna input to the pi-network. From the mosfet input at gate 1 to the output, the amplifier is designed for a nominal gain of unity. The mosfet g_m is typically 10 millisiemens, so that the 800-ohm load resistor sets the no-load gain at eight. The voltage stepdown in transformer T1 reduces the gain by a factor of four; adding a 50-ohm load reduces it by another factor of two. If the receiver input impedance is substantially higher than 50 ohms, the overall gain will be closer to two. A second set of limiting diodes, CR3 and CR4, is placed at the output to protect the receiver in the event of a T-R switch failure.

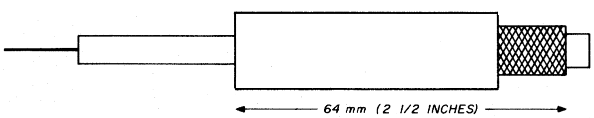

fig. 3. Construction details of the coaxial cable input capacitor (C1 in fig. 2).

Mosfet biasing. The mosfet is biased to operate with a drain current of 5 mA with a gate l-to-source quiescent voltage of – 0.7 Vdc. This allows the input signal voltage to switch to 1.4 volt p-p without appreciable gain compression. The diodes at the amplifier input will limit the signal to this same range so that they won't reduce the T-R switch dynamic range.

The voltage on gate 2 is set by a resistor divider at approximately 4 Vdc, setting the gate 2-to-source voltage to at least 3 volts. With these quiescent voltages and currents, the mosfet will have a transconductance of 10 millisiemens and a drain current swing of at least 10 mA p-p.

Bypass considerations. The mosfet is shunt fed to allow the drain-to-source voltage to remain above the minimum recommended by the manufacturer with the gate-2 voltage used. The value of the rf choke isn't critical. Values between 150 pH - 360 pH could be substituted. The decoupling of the mosfet drain supply is very conservative, using two capacitors in parallel (C7 and C8) to obtain several decades of effective bypassing. Gate 2 of the device is also decoupled with two capacitors and a ferrite bead to ensure the mosfet will remain stable through the uhf range. Again, the values of the decoupling capacitors aren't critical.

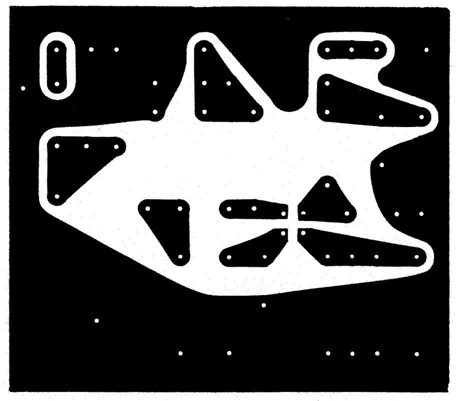

fig. 4. Printed-circuit board layout for the T-R switch, Component layout is shown in fig. 5.

Supply voltage. The supply voltage, $V_{cc}+$, for the T-R switch can range from +12 Vdc to +18 Vdc without much effect on performance. Additional decoupling of the input supply voltage is obtained with RFC1 and C9.

Input coupling. The only unusual component is C1, the input-coupling capacitor. Because the rf voltage levels are quite high and the required capacitance so low, a suitable commercial component would be difficult to find. An inexpensive substitute is a piece of coaxial cable. A typical piece of 50-ohm coax will have a capacitance of 1 pF/cm (30 pF/ft), so that a 6.3-cm (2.5-inch) center conductor-shield overlap will provide 6 picofarads of capacitance. A sketch of the coaxial capacitor is shown in **fig.** 3. The length of

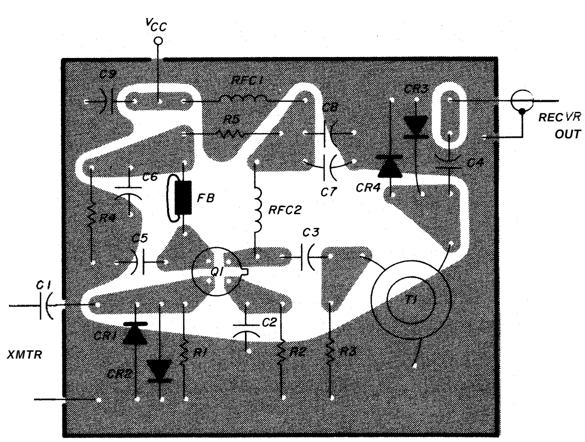

fig. 5. Component layout for tho T-R switch.

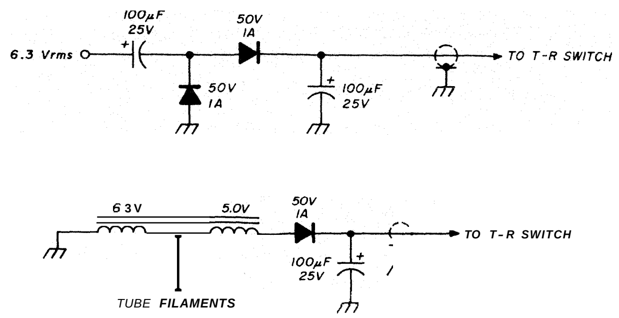

fig. 6. Two power supply options for the solid-state T-R switch. Supply voltage is not critical and may vary from +12 to +18 Vdc.

the un-overlapped portion can be as long as necesary to go from the pi-network to the T-R switch input. A piece of bare wire can be wrapped around the shield and soldered to complete the capacitor. For plate voltages less than +500 Vdc, RG-58/U can be used, with RG-59/U suitable for voltages up to +900 Vdc. With 70-ohm coaxial cable, the capacitance will be approximately 0.7 pF/cm (21 pF/ft), so an overlap length of 8.7 cm (3.4 inches) should be used.

construction

Circuit layout is shown in **fig. 4.** A 5.7 x 5 cm (2½ x 3¾ inch) single-sided copper-clad printed wiring board was used, and mounted in a 10 x 7.6 x 5 cm (4 x 3 x 2 inch) aluminum minibox. The board is mounted to the minibox using 2-cm (%-inch) aluminum angle stock. The whole assembly was mounted on the rear of the transmitter shielded pi-network cage. The only critical aspects of the mechanical assembly are to provide good rf returns between the pi-network ground and the T-R switch board ground, and between the output connector return and the board ground.

Only one board was constructed, so photo etching wasn't used to make the board. Copper tape was used on the back of the board to provide the component interconnections. For those who wish to use an etched board, a suggested layout is shown in **fig. 4;*** the component placement is shown in **fig. 5.**

If the transmitter doesn't have the necessary supply voltage available, **fig. 6** shows two possible solutions. Both use the power-transformer filament windings in the transmitter. If the 5-volt rms winding is used, be sure it isn't connected to the high-voltage supply through the rectifier tube. This winding also must be properly phased with the other filament winding to prevent the two voltages from bucking. The power supply can be mounted in any convenient location in the transmitter and connected to the T-R switch with a shielded wire. At the switch end, the shield should be connected to the minibox, with C9 between the supply voltage and the shield-box connection.

*A printed-circuit board and parts kit is available from Radiokit, Box 429, Hollis, New Hampshire 03049.

conclusion

The T-R switch has been in use for over a year and has worked well under various operating conditions. Remember that the T-R switch protects the receiver only; it will not prevent the receiver from overloading if its dynamic range or agc characteristics are not up to standard. In most cases, muting the receiver when the key is depressed is the best solution to this problem. Take the advice of W4AX and K2RPZ: Turn that bargain transmitter into an effective CW rig with the addition of a good T-R switch.

editor's note

The bibliography at the end of this article has been culled from ***ham radio*** for the benefit of CW enthusiasts interested in break-in control circuits.

The article by Al Brogdon, K3KMO, combines the advantages of electronic switching using a Johnson model 250-39 T-R switch and an antenna changeover relay.

Cal Sondergoth, W9ZTK, describes a solid-state system for use with separate receive and transmit antennas using low-power transmitters (under 100 watts). The article emphasizes receiver overload during transmit.

W.M. Mitchell, W8SYK, presents a single-transistor CW break-in circuit for stations with separate transmit and receive antennas. The design is for grid-block keying.

J.K. Boomer, W9KHC, shows a low-power, solid-state T-R switch using a PIN diode. The circuit handles power to 100 watts at any desired keying speed.

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2. Stu Goodman, K2RPZ, "A Bonanza Awaits You in the Ham-Ads," OST, December, 1977, page 11.
3. Wes Hayward and Doug DeMaw, Solid-State Design for the Radio Amateur, ARRL. Newington. Connecticut, 1977, pages 179-180.
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ham radio

understacking high-frequency Yagi antennas

A novel system for stacking beams on a tower to minimize mast damage in heavy weather

Installation of a single Yagi antenna on a tower, whether the tower is guyed or not, provides a clean looking system that will no doubt perform as predicted. If a commercially manufactured antenna is used, and there is no desire to make adjustments for optimum performance, mounting it in the clear away from all surrounding objects is the only way to do it. That, however, is not very cost effective. A second beam means a second tower, a third beam requires a third tower, and so on. I quit with three towers – but I want beams for 40 through 10 meters.

Stacking distances of 10 feet (3 meters) or more aren't generally suited to high-frequency beam installations because it's difficult to prevent the mast from bending in severe weather. Ultra-strong steel masts are expensive, heavy, and do not provide easy access to the top antenna. By understacking antennas, rather than overstacking them, a number of features develop which overcome these problems.

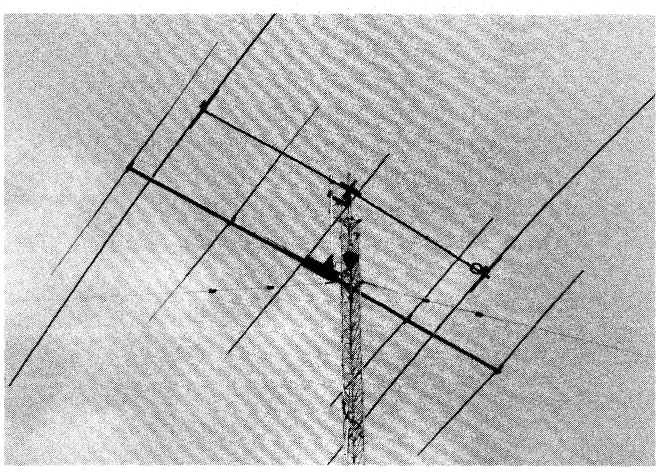
Understacking takes advantage of a unique relationship between the system weight and the effect of gravity. With conventional stacking arrangements, the weight of the top antenna ***adds*** to the bending moment at the base of the mast. Understacking ***subtracts*** the antenna weight from the bending moment because gravity aids in keeping the mast in a vertical position.

A number of other feature benefits come into play. First, installing, tuning, or repairing the stacked antenna is relatively easy because it is mounted within reach of the person climbing the tower. Properly designed tilting hardware makes it easy to reach any point on the boom of either antenna. Element repair or matching-network adjustments are relatively simple. Building the system shown in the photographs was a one-man project – no help was needed.

A primary objective for any antenna system is to have it remain intact during foul weather. One disasterous event which can destroy an antenna, especially one that is stacked above another, is a hurricane – or hurricane-force winds. Even with two days of notice (hurricanes are somewhat predictable!), an antenna stacked high on a mast offers little opportunity to take damage avoidance steps to weather the storm. With understacking, however, you can climb the tower and tie in the stacked antenna with heavy rope. When the boom of the stacked antenna is fastened securely to the tower face, the chance of mast damage is eliminated. Furthermore, the mast and boom of the stacked antenna are fastened at a point where the top set of guy wires have the greatest strength – right at the guy point. The tie-in procedure reduces the load above the top set of guy wires to one antenna surface instead of two.

Further security is built into the system by installing a torsion bar assembly at the very top of the tower; see the photographs. The benefit is obvious: during hurricane season (June through November), a ready-made set of guy wires is kept on hand. If they are needed, it is a simple task to install them (it can be done in less than an hour). With the extra guy wires in place, there is little chance of tower damage from high winds. Whether or not the antenna survives extremely high winds is a different matter. Keep in mind, though, it is far easier to repair a bent

By Robert M. Myers, W1XT, 221 Long Swamp Road, Wolcott, Connecticut 06716



The antenna installation at W1XT. Shown are a two-element, 40-meter Yagi and an understacked five-element, 10-meter Yagi.

antenna than it is to fix a bent tower!

There are a few minor benefits to understacking as well. The auxiliary mast is a torsion tube which reduces stress on both the tower legs and the rotor. The empty span between Yagis is ideally suited to the installation of a small vhf antenna.

mechanical components

Fabrication of the hardware is simple and can be done easily in the home workshop. There are three different pieces to the system: an II-foot (3.3-meter) long galvanized steel mast, two boom-tilt assemblies, and two angle brackets to support the auxiliary mast at the top of the main mast. Standard discount store automotive muffler clamps are used to hold everything in place. Plated clamps are usually found in the hardware department rather than in the automotive section.

The mast sections are made from 1%-inch galvanized waterpipe. Because waterpipe is specified by inside diameter (ID), not outside dimension (OD), the outer dimension for the 1%-inch pipe is slightly under 2 inches (50 mm). This size is ideally suited to 1-7/8 inch (48 mm) clamps although 2-inch (50 mm) clamps are satisfactory. Galvanized waterpipe is available from most plumbing supply dealers but be prepared for a stiff price. The material used here cost nearly $20!

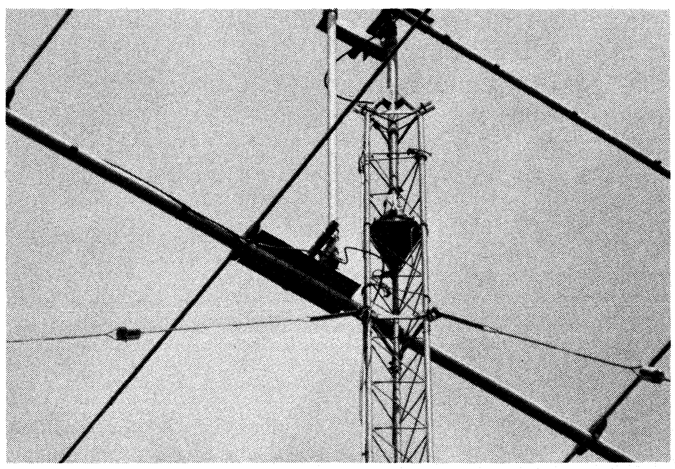
The two top mounted angle brackets are 12 inches (30 cm) long and provide adequate tower clearance for the auxiliary mast. The aluminum stock is 3 inches (8 cm) on a side and ¼-inch (6 mm) thick. The upper angle piece is equipped with six muffler clamps, three on each end; the lower one has two at each mast connection point.

Both boom-tilting assemblies are shown in the photographs. The %-inch (25 mm) thick plate (aluminum) has a horizontal pipe section held in place by a group of muffler clamps. Each Yagi has its boom-to-mast plate turned horizontal so that it may sit on the tilt assembly plate. Loose clamp hardware allows relatively free tilting of the boom to any position for the installation of the elements. The top boom-tilt assembly is fastened to the main mast and not the auxiliary one. This helps offset the leaning action caused by the lower antenna.

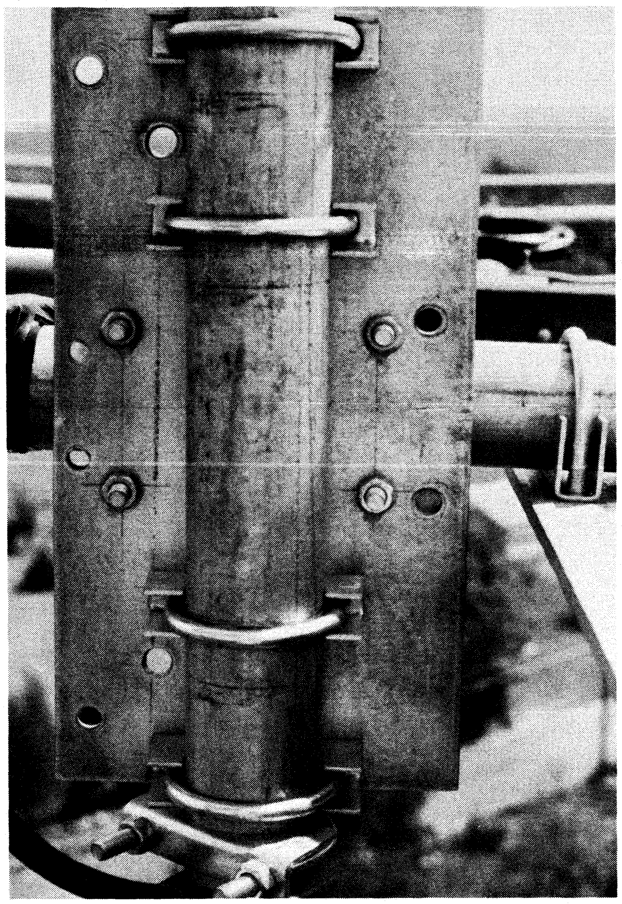
system description

A five-element, 24-foot (7.2 meter) boom 10-meter Yagi is stacked under a two-element electrically shortened 40-meter beam (Mosley S-402) which has 46-foot (13.8-meter) long elements. The Mosley antenna was selected because of its low wind surface profile as compared to a full sized array. The high Q of a loaded antenna makes it necessary to assure no detuning occurs as a result of other hardware being in close proximity to it.

A number of other factors were involved in selecting these antenna designs. The 10-meter beam has a boom length slightly longer than double the spacing between antennas which means that the top antenna interferes with rotating the lower boom vertical. Since the longest 10-meter element is shorter than the boom length for the 40-meter antenna (20 feet or 6 meters), it is a simple matter to turn one boom 90 degrees (horizontally) and tilt the 10-meter boom end and element up between the two 40-meter elements. The 24-foot (7.2 meter) boom length on the 10-meter beam was selected to provide element positions which would not be directly under the elements



The guy-wire torsion bracket is mounted just above the bottom rung of the tower top section. The 10-meter antenna rotates just above the top set of guys. An empty torsion assembly at the top of the tower is available for an additional set of guy cables if they should be necessary.



The 10-meter tilt assembly attaches to the auxiliary mast with four muffler clamps. Note the clamp at the very bottom of the mast. It is needed to keep the tilt assembly from slipping off the auxiliary mast and is an absolute requirement.

above. For one thing, this gives slightly more separation between elements than the vertical dimension indicates. More importantly, however, during a bad ice storm, the top elements won't droop and make contact with the lower antenna. Also, melting ice from the top elements will drop between the lower 10-meter ones. Falling ice can play havoc with the antennas beneath!

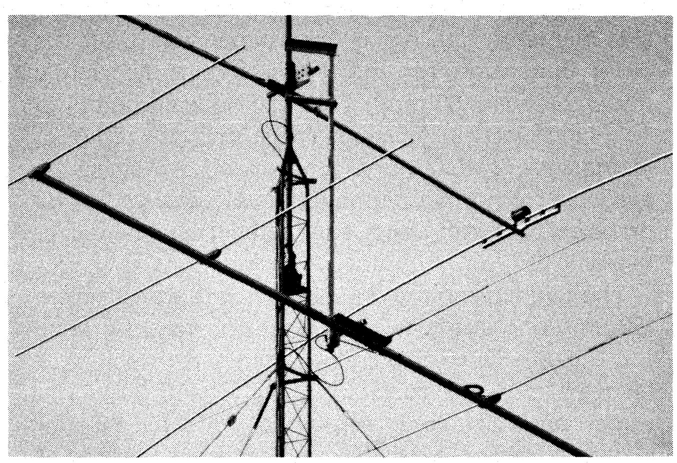
Interaction tests were required to assure no interaction between antennas. The Mosley antenna was installed first. An SWR curve was plotted, front-to-back measurements were taken, and the relative signal strength of a local broadcasting station were made (W1AW is 25 miles away and visible on a clear day). Next, the 10-meter Yagi was installed and similar measurements were performed. Rechecking the 40-meter tests showed no difference in the figures. Rotating the top antenna 90 degrees with respect to the lower one had no influence on the test results of either antenna. The conclusion is that stacking these two antennas 10 feet (3 meters) apart is sufficient to avoid detuning either one.

Mechanically, the system is stable and strong. The slightly off-center mounting of the auxiliary mast causes the main mast to lean to one side. This is counteracted to some extent by the top tilt assembly being mounted to the main mast with the horizontal pipe extending in the other direction. There is no binding in the tower top sleeve. The relatively long main mast to the rotor makes the misalignment insignificant. You should not attempt this type of understacking, however, if the rotor is mounted directly beneath the tower sleeve — or worse yet, if the rotor is mounted above the sleeve. Under these conditions, the lateral forces would destroy the rotor in short order.

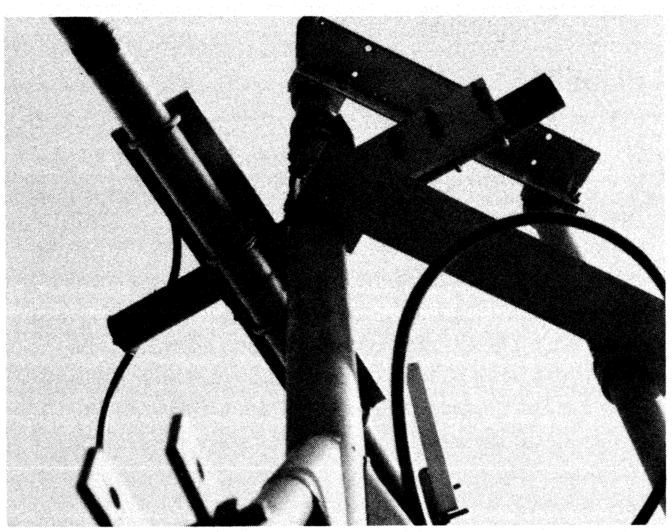
There have been two tests of the mechanical strength and reliability of the stacking procedure. In August, 1979, a severe storm whipped through central Connecticut, tearing roofs off buildings and uprooting trees. The local weather service measured wind velocities of 70 mph (110 km/h) for more than an hour. After the storm, the system inspection showed only a 40-meter element to be rotated around the boom; everything else was intact. During early September, Tropical Storm David generated wind gusts up to 70 mph (110 km/h); this time there was no damage. In neither case was the top set of guy wires installed (the extra set) or the lower boom tied into the tower face. A good deal of confidence was developed by these two events. After the first storm, it was a simple matter to tilt down the 40-meter boom and straighten the twisted reflector.

hardware installation

There are numerous ways an Amateur can approach this kind of a project. The testing requirements, however, dictated the order in which compo-



The top angle brackets have been offset to give a greater spacing between the tower legs and the 10-meter boom.



Two angle brackets are used to hold the auxiliary mast in place. One is attached above the 40-meter tilt assembly; the other is connected below.

nents were installed. It would be wise for anyone duplicating this system to perform tests similar to those mentioned earlier. If two antennas are put up and one doesn't operate correctly, it could be very difficult to determine the source of the problem.

First, install the tilt hardware to the main mast. Next, position the 40-meter boom on the horizontal tilt assembly pipe. With the appropriate boom end tilted down along side the tower, attach one element. It is necessary to tie a rope to the opposite boom end from where the first element is connected. It should be done before the boom leaves the ground. The rope is needed to pull the non-element end down after the first one is attached. With both elements connected, the antenna weight is balanced and vertical rotation around the tilt assembly is easy. Exercise extreme care when turning the boom up or down. Be sure the clamps can't slide off the horizontal pipe. The pulling rope must be strong and tied securely in place. If the rope breaks or slips half way through the tilting process, the heavy end will swing down with a vengeance.

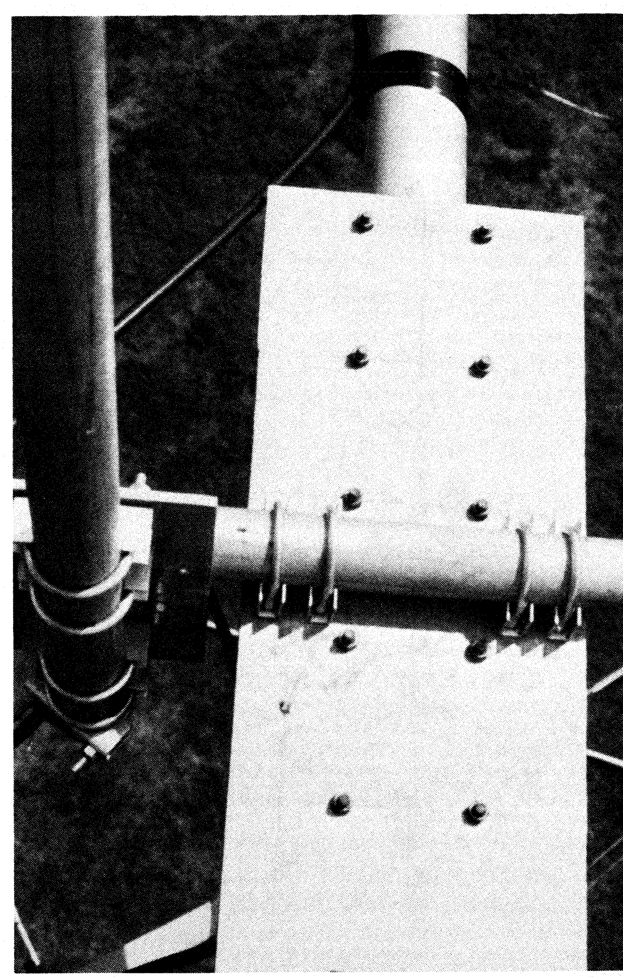
Once the 40-meter beam is installed and tested, the two auxiliary mast supports are clamped in place; one gooes above the 40-meter tilt assembly and the other mounts beneath it. Muffler clamps should be attached to the far end of the angle stock, ready to accept the auxiliary mast. Slip the mast pipe up through the clamps and secure all of the hardware. The 10-meter boom-tilt assembly can be attached to the auxiliary mast before it leaves the ground or after the mast is in place.

As with any antenna installation, safety is an absolute requirement. I have found the best procedure is to plan every step of the process in advance. Abbreviated notes are used to avoid mistakes.

boom-to-tower spacing adjustment

The auxiliary mast aligns on center with the tower when the support angle brackets are parallel to the 40-meter boom. The lower tilt assembly offsets the 10-meter beam sufficiently to clear the tower during rotation. Sway in the auxiliary mast caused by wind may allow the 10-meter boom to occasionally bump into the tower leg at some headings. Twenty turns of polypropylene rope are wrapped around the 10-meter boom where it comes close to the tower. The rope acts as a bumper pad during very high winds.

To increase the spacing between the boom and the tower legs, change the position of the angle supports toward perpendicular with the 40-meter boom. The lower beam will need to be repositioned slightly for a corrected heading; an 8-inch (20 cm) clearance



The homemade 10-meter beam mounts to the tilt assembly with four muffler clamps. The coaxial cable loop must be positioned so that rotation of the system doesn't crimp or cut it.

is adequate. Note the offset of the angle brackets in some photographs.

pitfalls

It is possible to forget some of the basics of good engineering practice. For instance, the auxiliary mast is indeed a 10-foot (3-meter) long lever arm and will flex the main mast pipe. An antenna of much larger dimensions than described here would likely cause one of the masts to bend if extremely severe weather were encountered. For use with bigger systems, hard steel tubing is recommended in place of waterpipe.

Another important consideration is tower loading. The tower shown here is 100 feet (30 meters) of Rohn 25 guyed at 33, 66, and 91 feet (10, 20, and 27 meters). The unsupported top section has a torsion guy assembly mounted just below the bottom rung which keeps lateral forces off the tower leg bolts. The load rating for Rohn 25 tower is 6 square feet (0.55 $meter^2$) of antenna; many Amateurs exceed that with large six-element Tribander. The surface area rating for the Mosley S-402 is about 3.8 square feet (0.35 $meter^2$). The lower five-element Yagi adds another 2.5 square feet (0.23 $meter^2$) of surface area. This tower is sufficiently loaded for maximum safe operation (note that tower load ratings assume a rotor, mast, and cables and should not be included in the calculations).

other combinations

Interaction between antennas is always a possibility when more than one antenna is placed on a tower. Many Amateurs have experienced difficulty in operation when 15- and 40-meter antennas are mounted together on the same tower. The end result is usually poor front-to-back ratio with the 15-meter system. The high *Q* of a Tribander accentuates the problem. For this reason, you should be cautious about installing a Tribander and a loaded 40-meter beam on the same support. If interaction does result, the simple solution is to turn one of the antennas 90 degrees with respect to the other. Double dial calibration would then be required.

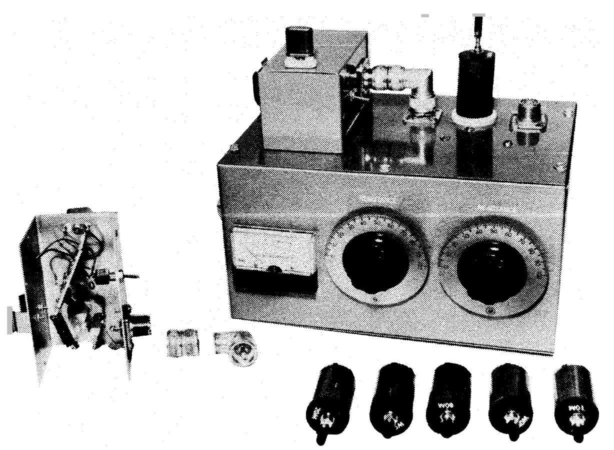
A combination of antennas ideally suited to understacking is a small "Christmas Tree" of monobanders for 20, 15, and 10 meters. The largest antenna should go at the top and the smallest in the middle of the auxiliary mast. In this manner, the heaviest Yagi is mounted just above the top tower sleeve, and the next largest antenna will have its weight at the bottom of the auxiliary mast to counteract the bending moment.

No matter what combination of antennas is selected, be sure not to overload the tower. Hardware falling from the sky is hazardous to your neighbors!

ham radio

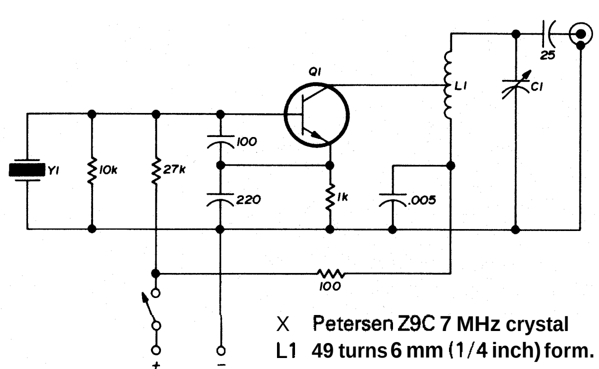
More Details? CHECK — OFF Page 94

the Macromatcher: increasing versatility



Several years ago I built the Macromatcher and the pickup **coils**[1] and used **it** with a grid-dip meter for antenna matching. This combination is rather cumbersome if used at the antenna terminals when the antenna is at the top of the tower. An alternative method is to use the above setup on the ground with one-half wavelength of coax to give the same characteristics as at the antenna.

I built a crystal-controlled transistorized signal source (powered by a 9-volt battery and mounted on the Macromatcher) for 20 meters **(fig. 1)** and 40 and 75 meters **(fig. 2).** The photo shows the complete unit. The 20-meter oscillator is on the left, and the 40- or 75-meter oscillator is mounted on the Macromatcher. A Bud minibox CU-2103-B, 102 x 58 x 58 mm (4 x 2-1/4 x 2-1/4 inches), was used. The part of the minibox with the lips is fastened to the Macromatcher as shown.



X Petersen Z9C 7 MHz crystal
L1 49 turns 6 mm (1/4 inch) form. Tap at 25th turn

fig. 1. Twenty-meter oscillator.

Some thought and planning must be done to ensure proper positioning so that the oscillator output (coax receptacle) mates with the Macromatcher input receptacle.

Connections between the Amphenol fittings are as follows: an 83-877 (double male) from the 83-1R (SO-239) on the oscillator to a 83-1AP (angle) to the 83-1R (SO-239) on the Macromatcher (see photo). The oscillator was built in the other portion of the minibox, the part without the lips.

When the unit is to be used on a different band, all that's necessary is to remove the four screws holding the minibox together, unscrew the coax connector on the Macromatcher, change the battery, secure the other oscillator to the Macromatcher, and the unit is ready to use.

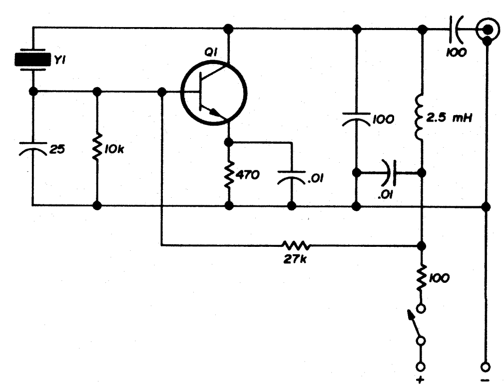


fig. 2. Forty- and 75-meter oscillator.

With a slightly larger minibox a VFO with a bandswitching arrangement could be built, which would have more versatility. You'll note that the dial skirts of my unit haven't been reversed and remarked for resistance and reactance values.

acknowledgment

A special thanks to Bob Henry of Satterfield Electronics for the photographs.

reference

1. **The** ***Radio Amateur's Handbook,*** 1979 **edition, chapter** 16, **page** ***25.***

ham radio

By Arnie Bachmann, K9DCJ, Route 1, Blue Mounds, Wisconsin 53517

the ham notebook

two for one

If the title sounds like poor odds on a tout sheet at Pimlico and not applicable to ham radio, have patience and read on. There have been a number of times in more than thirty-three years as an Amateur when I wished to use more than one receiver on a single antenna system at the same time. Whether corrected directly or through decoupling amplifiers, sensitivity was lost or undesired interaction in the form of rf oscillations resulted.

For use on 2-meter fm I built a quarter-wave ground plane antenna

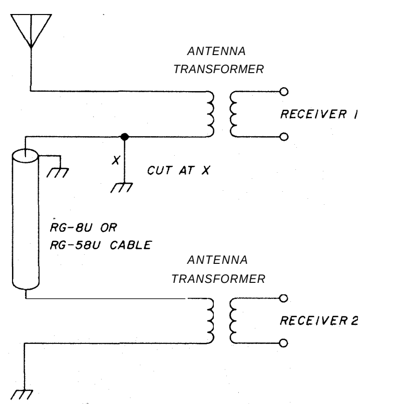


fig. 1. Circuit for combining one or more receivers to a single antenna.

for base-station use. Some months later I obtained a tube-type aeronautical monitor receiver (I'm involved with flying and wanted to listen to the aeronautical service). In the aeronautical communications with which I work, a maximum of three vhf receivers are connected so that the primary side of all the antenna transformers are series connected; the bottom end of the antenna coil in the last receiver is grounded. These are double-conversion, fixed-frequency receivers with a minimum of 1-MHz frequency separation between the three series-connected receivers.

Without a relative signal-strength meter in either the ham or the aeronautical receivers, I not only maintain my Amateur operation but also copy aeronautical ground station transmitters that I know are putting 10 watts into over 46 meters (150 feet) of RG-8/17, combined. Thus, with about 3 dB power loss in the coaxial cables, I operate both receivers simultaneously. I have more than 20 MHz frequency separation between the receivers. I'm about 32 km (20 miles) airline from the airport.

The only modification to any receiver with a grounded end on the antenna transformer is the addition of a chassis-mounted rf connector such as an SO-239 or a BNC, or N type connector (fig. 1). The ground connection is lifted and wired to the connector. Then, for single-receiver operation, this point is again restored to ground by a jumper plug of the appropriate type. For dual- or triple-receiver operation, these former ground points are series connected with suitable lengths of 50-ohm coaxial cable. The ungrounded end of the last receiver is grounded once again to re-establish the antenna-circuit continuity. The length of the coax jumpers is somewhat critical for the uhf and vhf ranges for optimum operation but will be less so in the hf range.

Jack Struthers, W2OZY

TS-820 filter switching modification

The addition of the YG-88C 500-Hz crystal filter to the TS-820 is a worthwhile operating aid. When this filter is installed according to the instructions furnished, the filter will be automatically selected when the TS-820 MODE switch is in either the CW or TUN position. It's convenient to be able to select the wider 2400-Hz standard filter during tune up or while scanning the CW portion of the band in use. Here's a scheme for selectable filter switching using no new hardware or new holes.

Fig. 2 shows the back of the TS-820 METER switch. It's a double-pole, five-throw (DP5T) switch with only one-half being used; i.e., single-pole five-throw, or SP5T. Thus half

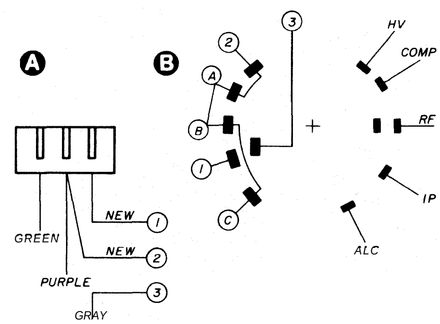


fig. 2. Modifications for selectable filter switching for the TS-820. The meter switch (S-1) is shown at (A) viewed from rear of panel. New connections to connector IF2 are shown in (B).

of this switch is available for other uses.

The speech processor is inoperative during CW and TUN modes, so the COMP position of the METER switch is ideal for activating and selecting the 500-Hz filter. During tune up, as the METER switch is switched through ALC, IP, and RF, the 2400-Hz filter is activated, allowing higher meter readings and finer adjustments.

For CW you can choose filters by switching to COMP to activate the 500-Hz filter or use any of the other METER switch positions for 2400-Hz filter operation.

Modification of the TS-820 is straightforward and should take less than an hour:

1. Remove cabinet top and bottom covers.

2. Locate and remove connector IF2 from the bottom of IF Board X48-1150-00.

3. Refer to the color-coded wiring of connector IF2 as shown in Figure 25, page 34, of the TS-820 Operating Manual.

4. Remove gray wire from IF2. Place the blade of a small screwdriver into the slot above the wire and gently pull on the gray lead.

5. Unsolder the gray wire from the connector tip. Solder a new wire approximately 46 cm (18 inches) long to the connector tip and reinsert the tip into the connector IF2.

6. Route the new wire along the large wire bundle and up through the chassis to the back of the METER switch. Cut to length, strip, tin, and solder to the switch at terminal 1 as shown in **fig. 2.**

7. Splice a second new wire approximately 46 cm (18 inches) long to the free end of the gray wire, route it to the switch, and solder at 3.

8. Remove the purple wire from IF2 using the blade of a small screwdriver as in step **4** above.

9. Solder one end of a third new wire approximately 46 cm (18 inches) long to the connector tip along with the purple wire. (There will now be two wires soldered to the single connector tip.) Reinsert the connector tip back into connector IF2.

10. Route this last new wire to the switch and solder at 2.

11. Connect and solder a short piece of uninsulated, tinned wire between terminals 2, A, B, and C. Be sure that the uninsulated wire does not touch terminal 1.

12. Neatly dress all leads (tie to existing wire bundles if desired) and replace the top and bottom cabinet covers.

This completes the modification. During tune up, as the meter is cycled through ALC, IP, and RF the 2400-Hz filter will be operative, as it will be during **SSB.** Additionally, the COMP position of the METER switch will activate the 500-Hz filter only while in the CW or TUN mode. Thus, the ability to read compression level during **SSB** operation with the speech processor is retained.

Don Jacobson, K7OAK

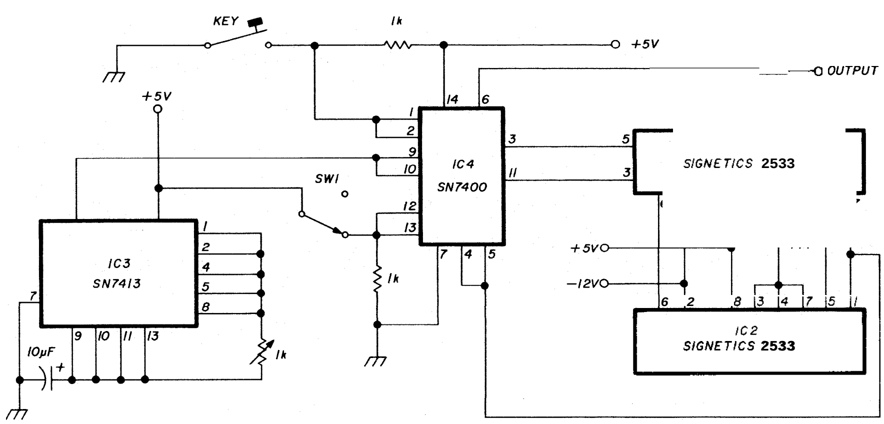
2048-bit memory keyer

I do quite a lot of moonbounce operating on 144 MHz and this can sometimes result in an hour or so of continuously sending callsigns. I designed the keyer in **fig.** 3 to make operating easier. All I have to do is program the callsigns, and the keyer does the rest.

The circuit is simple. The key is an ordinary Morse hand key. SW1 is the information in-out switch. The 1-k variable resistor is adjusted so that information going into the keyer just fills the memory. Decoupling capacitors are left out for the sake of clarity, but at least 200 μF should be used from −12 volts to ground and at least 500 μF from +5 volts to ground.

The output will drive a solid-state keying switch or sensitive reed relay directly. The keyer is indispensable for contest work; other uses include beacon and repeater identification. Memory capacity is 2048 bits.

Dave Price, GW4CQT



IC1, IC2	Signetics 2533
IC3	SN7413
IC4	SN7400

fig. 3. Schematic of the memory keyer by GW4CQT, which takes the drudgery out of contest work for sending ID, CQ, and callsigns. Memory capacity is 2048 bits.

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contents

on top of the news

by Joe Schroeder

It is sometimes fashionable to say that the spirit of Amateur Radio has gone, arid today's Amateur is an uninformed uninvolved appliance operator. What took place Friday evening, May 23rd, on 75 meters was a stinging rebuttal to that gloomy assessment. Every night that week AMSAT members and other interested Amateurs had been meeting on 3850 at 02002 for a progress report on the launch of AMSAT's Phase-III satellite. After one of the French Ariane launch vehicle's four rocket motors failed during launch that morning and put Phase III (with a lot of other expensive space projects) into the Atlantic Ocean near Devil's Island, the group met once again to commiserate with each other over the disaster.

What began as a wake quickly turned into an almost unprecedented outpouring of support. With AMSAT's President, Tom Clark, W3IWI, as net control, most of the active North American participants in Phase III's development were joined by perhaps one hundred other check-ins from across the United States, Canada, and even Cuba. There were eulogies for the Phase-III bird and the loss of its unique capabilities, to be sure, but the predominant message over the following several hours was, "Let's keep moving ahead!" And this message came, significantly, not only from the already involved AMSAT membership but from bystanders — many of whom checked in to say: "I haven't been on OSCAR yet but always admired what you guys were doing. With your loss today it's time I became involved, so my check for membership plus a contribution is in the mail. How else can I help?" Needless to say, such support provided a priceless boost for those who'd heard their efforts of the past several years splash down in the ocean just twelve hours earlier . . . and they reflect that real Amateur Radio spirit that has too often been passed off as dead.

What does the loss of Phase III mean to AMSAT, and along with AMSAT, to the Amateur community? It means the loss of years of very hard work by a relatively small group of Amateurs in a half dozen countries. It means the loss of the $150,000 in hard cash that AMSAT invested in Phase III. It means the loss of more than a year, and possibly several years, before a new free-world Amateur satellite can be put up to replace OSCAR 7 (still operational well beyond its designed lifetime but showing its age) and OSCAR **8.**

What's needed to keep our space program going? First and foremost, money, and plenty of it. Space efforts cost money, and the kind of sophistication that makes our "amateur" satellites suitable traveling companions for the best efforts of the pros cannot be accomplished on a shoestring. The Phase III investment brought AMSAT's treasury to a dangerously low point, and it's going to need rapid infusions of new money if we are not going to lose momentum. The second need is participation, people to volunteer for all kinds of tasks from bookkeeping and basic administration to state-of-the-art design work and computer programming. Finally, AMSAT needs members, for, by joining, an Amateur becomes not only a contributor but an ***involved*** contributor.

The response to AMSAT's needs was almost instantaneous after the news of Phase III's loss. Following the pledges on the Friday-night net, AMSAT's mail box has been bulging with new member applications and contributions. Within a few days, Amateur Electronic Supply in Milwaukee, Ham Radio Center in St. Louis, and the Ham Radio Publishing Group had all pledged $1,000 each to AMSAT, and many more industry contributions are expected.

What can you do? Join AMSAT. Annual dues are only $10 a year before July 1, $20 thereafter; life memberships are now $100, going up to $200 after July 1. Contributions to AMSAT are tax deductible. AMSAT, Box 27, Washington, D.C. 20044 is the place to send your check. Do it today and help demonstrate that Amateur Radio spirit is as real now as it ever was!

Joe Schroeder, W9JUV
associate editor

comments

Factory service

Dear HR:

The editorial in the January, 1980, ssue of *ham radio* is interesting, but somewhat misleading. Even though t's stated that some manufacturers 'equire their dealers to provide warranty service, the general impression, after reading the editorial, is that all Amateur dealers apparently do not have the capability to provide adequate after-sales service.

I agree it is a bad situation when a manufacturer requires its dealers to provide the warranty service, and then does not supply the dealer with up-to-date service information or parts.

However, Trio-Kenwood goes to great lengths to avoid this situation. First, every authorized Kenwood dealer is required to have the facilities to perform after-sales service. The customer may send his equipment either to the dealer or to us for warranty service. We are constantly mailing up-to-date service information to our dealers, and they are equipped with all of our service manuals and other technical material.

When you bring a rig to an authorized Kenwood dealer for repair, you are bringing it to a factory-trained technician. We conduct service seminars each year for the dealers, providing them with several days of intensive instruction. Our dealers are reimbursed for parts and labor, and are adequately stocked with the most commonly needed parts. Therefore, in most cases, the local Kenwood dealer can repair a rig as quickly and as thoroughly as we do, saving the customer several days of \shipping and avoiding possible shipping damage or loss.

The "Kenwood Users' Report" in the January, 1980, issue of *Ham Radio Horizons* shows that more rigs are serviced by the dealers than by us, and that the majority of owners are *satisfied* with the service; and local service is certainly much more convenient for the customer.

We support the local dealer by training him to perform after-sales service. We *must* support the local dealer, because Amateurs depend on him for advice on the selection and use of equipment, as well as fast and competent service. A *responsible* Amateur Radio manufacturer, besides offering full factory-backed warranties, also keeps his dealers fully prepared to offer after-sales service. I wish your editorial had recommended contacting these dealers first, for service.

Kenneth M. Bourne, W6HK
Manager, Marketing Services
Trio-Kenwood
Communications, Inc.

We did not mean to imply that customers who require service should not seek help from their dealers first; we simply wanted to point out that dealer service is only as good as the support those dealers receive from the manufacturers, and not all manufacturers have the extensive dealer support program sponsored by Trio-Kenwood. Too many Radio Amateurs take good customer service for granted, when it is not always available — regardless of the reputation or integrity of the dealer.

Editor

ni-cad battery charging

Dear HR:

I enjoyed WA6TBC's article on the any-state ni-cad charger in the December issue of *ham radio.* Ni-cad batteries are something of a problem . . . they are, to a point, difficult to maintain. They either run down entirely and reverse polarity in one or more cells, or else are damaged by mere overcharging. My own experience has shown, however, that batteries are better kept on constant charge (a trickle); that minimizes annoyances all around. For example, I have a Vivitar electronic flash unit which has been on charge all the time for more than a year; test shows the unit to be operating very well. I installed a 100-ohm series resistor to reduce the charging current; this was placed right at the charger jack inside the flash unit. My impression is that ni-cad batteries tolerate all-time trickle charging. At any rate, this trickle charging insures that I have an operative unit always ready for use.

I might mention that I run all my calculators on chargers, and to date I have never had any problem with inoperative batteries due to "overcharging." Acquaintances tell me that they run their Hewlett-Packard calculators all the time off their chargers. I even run my Non-Linear Systems Miniscope on its charger all the time; the system has an automatic current limiting system built in.

At any rate, ni-cad batteries and gel batteries are a mixed blessing. Convenience dictates that each calculator, electronic flash unit, or oscilloscope be ready and running on a second's notice, without regard to state of battery charge. Hence the need for constant trickle charging.

Robert H. Weitbrecht, W6NRM
Belmont, California

(Continued on page 12)

comments

(Continued from page 6)

frequency synthesizer

Dear HR:

Tom Cornell's article on the CMOS 2-meter synthesizer in the December issue is an example of a well-designed circuit, along with a down-to-earth description that will probably warm the soldering irons of many interested readers. It is for these reasons that I would like to suggest that any would-be builders examine the owner's manual of their radio to determine the method of frequency modulation.

In the article, Tom mentions that he was using the synthesizer with a Regency HR-2B. Undoubtedly, the synthesizer will work fine with that radio because it uses a reactance-type phase modulation technique. It has been my experience, however, that most present day crystal-type rigs and transmitter strips use varactor diode crystal rubbering techniques to deviate the carrier. Thus, using the synthesizer with this type transmitter will produce a clean carrier, but will lack any modulation. Therefore, would-be builders would be well advised to check their rigs before buying parts.

Don Cwynar, WA3AXS
Reading, Pennsylvania

You bring up a good point. Older fixed-frequency and crystal controlled fm rigs such as the Regency HR-2B used by author K9LHA were based on a phased modulator which work fine with the CMOS synthesizer — newer equipment often has a varactor across the crystal to deviate the carrier, For owners of these newer rigs K9LHA is designing a phase modulator which will allow the use of his CMOS synthesizer; it will be published later this year in ham radio.

Editor

EI2W six-meter report

Dear HR:

EI2W commenced operations on the six-meter band on 20 October, 1979, when VE1AVX was worked at 1423 GMT. This report is for the period from 20 October to 20 December, 1979 (inclusive). VE1AVX with 1000 watts of power and an 11-element beam on a 30-foot (10 meter) boom has been the outstanding signal on the band.

EI2W has been using low power; a FT620B transceiver, kindly loaned by South Midlands Communications Limited, of Southampton, England; output about 10 watts. The beam used is a 3-element spaced 0.2 wavelength and a folded dipole driven element. During the two months' operation 1552 QSOs were made with approximately 600 different stations in all W/K call areas plus VE1, VE2, VE3, VE4, XE, KP4, and VP2 (Virgin Island).

Activity has been much greater during this cycle than during the International Geophysical Year (IGY) 1957158. This has been due in some part to the use of SSB as against a-m during the IGY period.

The best day's operation was on the 18th of November when 106 stations were worked in all W call areas and VE1-2-3-4 and VO; very little fading was noticed on that day. The highest recorded MUF at this station during the period under review was on December 15th when it rose to 62.75.

On December 11th KØSFH was worked in Kansas, the USA station only using 3 watts; his signal was R5, S7/8 in Dublin.

A total of forty-three states have now been worked including California, Nevada, and the Dakotas. Tests with a tilted antenna will be carried out during the middle of January with VE1AVX; this arrangement met with great sccess during the IGY. Information on propagation and F2 Layer working is being collected and a full report on the present cycle will be made later in 1980.

H.L. Wilson, EI2W
9 Haddington Lawn
Glenageary, County Dublin
Republic of Ireland

rotary-dial mechanism

for digitally tuned transceivers

An ingenious application of up/down counters and photo-detectors for continuous tuning using a single knob

More and more ham equipment now uses digital synthesizers for frequency control. The usual method for programming the synthesizer is with rotary or thumbwheel switches. Although this works well for channelized communications, such as that on the 2-meter fm band, it's cumbersome for continuous tuning, such as on the high-frequency bands. For continuous tuning it's necessary to program the synthesizer indirectly, through an up/down counter.

Tuning can then be done, for example, by using two pushbutton-operated pulse generators, one of which sends a slow series of pulses to make the counter count up, and the other to make it count down. Fast and slow pushbuttons can be added, or alternatively a joystick and variable-rate pulse generator can be used to vary the tuning speed. Most of us, though, are used to tuning with a knob, and this is still about the most flexible method. In building a frequency-synthesized high-frequency transceiver, I couldn't find a readily available digital dial knob mechanism suitable for use with an up/down counter, so I built my own.

The mechanism I built is an improved version of that described by Earnshaw.[1] His readout consisted of a metal disk with holes punched in it through which two phototransistors viewed a small light bulb. By suitable placement of the detectors and with the appropriate logic circuitry, it was possible to generate separate pulses as the shaft was rotated: one for each direction of rotation (how this works is discussed in detail later).

Since I needed at least a hundred pulses per shaft revolution, and didn't want to spend time drilling holes, I used photographic methods to generate "holes" in the disk. The new disk was made by contact-printing the computer-generated pattern shown in **fig. 1** onto heavy graphic arts film.* Also, additional CMOS logic circuitry was added to provide twice as many pulses per revolution as there are marks on the disk, as well as circuits to drive several types of counters.

Instead of the phototransistors, photointerruptors specifically designed for this type of application were used. Given the mask, only simple tools and readily available parts are needed to duplicate this unit. It can be wired to give 50, 100, or 200 pulses per revolution.

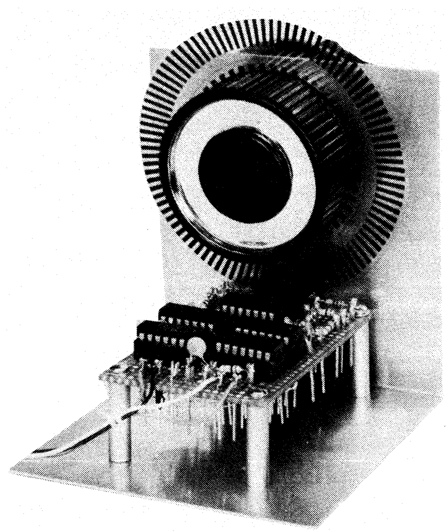
A single photodetector can be used to indicate the shaft rotation rate, but not the direction. To tell the direction of rotation it's necessary to use a second detector spaced one-fourth the mark-to-mark distance (or an odd multiple thereof) from the first detector. To see how this works, consult **fig. 2.** In the position shown, both detectors sense a dark region (mark), which means that their outputs are at logic 0.

Suppose the disk is moved to the left. As the mark goes past detector B, its output goes to 1. We will indicate this as $\bar{A} \cdot B\uparrow = L1$; that is, if **B** changes from 0 to 1 while **A** is 0, we generate a left-rotation pulse. As the shaft continues to rotate, **A** will go to 1 while B remains 1; we denote this $A\downarrow \cdot B = L2$. Continuing, **A** will remain as 1 as B goes back to 0, denoted by $A \cdot B\downarrow = L3$; and finally we get $\downarrow A \cdot \bar{B} = L4$. We are now back to a position with both detectors over a mark, the way we started. Now consider moving the mask to the right. The sequence of transitions will be $A\uparrow \cdot \bar{B} = R1$, $A \cdot B\uparrow = R2$, $A\downarrow \cdot B = R3$, and $A \cdot B\downarrow = R4$.

Although there are eight possible combinations of transitions from the center of one mark to the center of the next, it's not good practice to use both the turn-on and the turn-off transitions at the same posi-

***Photo disks are available from the author for $1.00 each. Please send a self-addressed, stamped envelope.**

By Chet B. **Opal, K3CU,** 5414 Old Branch Avenue, Camp Springs, Maryland 20031



The assembled prototype mounted on an aluminum angle bracket for checkout and alignment.

tion; the encoder might then be sensitive to vibration and electrical noise when the detector is at the very edge of a mark. Consequently, in the circuit described in this article, only four of the eight transitions are used: thus A t - is~used but $A\downarrow\cdot\bar{B}$ is not. As a result, there is a 114-mark backlash in the dial, but in practice this is not noticeable.

The schematic of the circuit necessary to process the detector signals is shown in **fig.** 3. I used GE type H21A5 interrupters, but TI type TIXL45 and Monsanto type MCA8 should work as well (all these types use a photo-darlington detector transistor to avoid the need for an output amplifier). Because the opto-couplers have a gain of about 1/20, the load resistor on the output transistor should be about twenty times the value of the current-limiting resistor to the LED emitter. The values shown for these resistors (R1-R4) are for 5-volt operation; they should be increased in proportion if the circuit operating voltage is made higher (the circuit will work over the 3 to 15 volt range).

The outputs of the detectors are squared up by Schmitt triggers U1A and C, and the complements of these signals are generated in U1B and U1D. Thus A, $\bar{A}$, B, and $\bar{B}$ signals are available from this gate. Type D flip-flops (U2 and U3) are used to sense the transitions from light to dark. U2A has $\bar{B}$ applied to its data (D) input and A applied to its clock input; it therefore clocks (in the notation used above) on $A\uparrow\cdot\bar{B}=R1$. (The flip-flop is reset later through the R input when B returns to 1.)

As the flip-flop sets, Q goes to 0 and produces a negative-going pulse (with duration determined by C1 and R5) at pin 1 of NAND gate U4A. Similarly, U2B generates pulses at pin 2 of this gate on the $A\downarrow\cdot B=R3$ transition. Normally, both inputs of this gate are at 1, so its output is at 0. When either of its input pulses goes to 0, its output briefly goes to 1. This pulse is buffered and inverted by U3A to provide a negative-going TTL-compatible output pulse (discussed further below). In similar fashion, pulses for the other direction are available at the U5B output. I've labeled these pulses CW and CCW, although depending on the mechanical arrangement, CW may actually produce pulses for counterclockwise rotation and vice versa.

Not all up/down counters use separate up and down clock pulses; some use a "direction" (U/$\bar{D}$) control signal and a single clock pulse. To generate signals compatible with these requirements, a flip-flop consisting of U4C and U4D is used. A negative-going pulse at pin 8 (the CCW pulse signal) sets this flip-flop, while a CW pulse resets it through pin 13. Both polarities of the direction signal are available at the outputs of this flip-flop ("CCW direction" and "$\overline{\text{CCW}}$ direction"). The two direction pulses are first combined by U5C, delayed by C5 and R9 to allow the direction control signal time to stabilize, and then buffered by U5D to produce a negative-going "$\overline{\text{either}}$" pulse.

construction

A photo of the prototype digital dial is shown; it

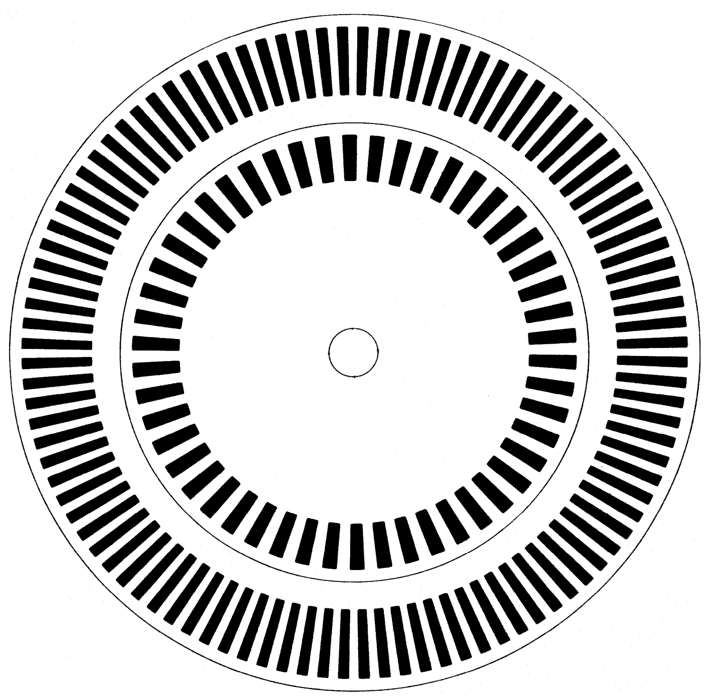


fig. 1. Optical mask used to encode shaft rotation (shown actual size). inner set of marks generate 50 pulses per revolution; outer set 100 pulses per revolution.

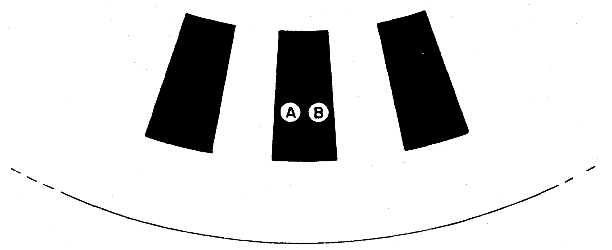


fig. 2. Fields of view of the two optical detectors, A and B, spaced one-quarter the mark period, if the mask is rotated leftward, detector B will sense light before detector A; conversely, if the mask is moved to the right, detector A will change first. The appropriate transitions are electronically processed to generate pulses indicating the extent and direction of rotation.

was assembled on an aluminum angle bracket for checkout and alignment. The whole assembly was then mounted in the final enclosure.

A method must be found to make the disk rigid and to mount it to the shaft. I resolved both problems by cementing the disk with epoxy to a large knob, slightly smaller in diameter than the mark pattern. For good adhesion, it helps to roughen the region on the disk and the knob where the cement will be with fine sandpaper (protect the rest of the disk with adhesive tape). The knob is then attached to a panel bearing, such as the Millen 10061; no doubt a satisfactory bearing could be salvaged from an old tuning drive, ten-turn pot, or something similar.

The electronic circuitry and detectors are mounted on a small section of perforated fiberglass board. The detector mounting holes should be slotted to allow adjustment of their separation. The gap between the two interrupters should be about 1/16 inch (1.6 mm) for the 50-mark pattern and about 1/64 inch (0.4 mm) for the 100-mark pattern. Point-to-point wire-wrap techniques were used for wiring. The board was mounted on the aluminum bracket with 1-inch (2.5-cm) aluminum stand-offs.

alignment

Alignment is not critical. Shaft height and knob position should be adjusted so that the disk always clears the interrupters but so that sector marks pass in front of the detectors at all shaft angles.

Apply power and check with a scope or voltmeter that the **A** and B signals oscillate as the shaft is rotated. Check that the "direction" signal changes when the direction of rotation is reversed; if necessary, adjust the interrupter spacing until this signal is not erratic. Final fine tuning (which probably is not really necessary) is done by observing the "either" pulses with an oscilloscope while the knob is spun.

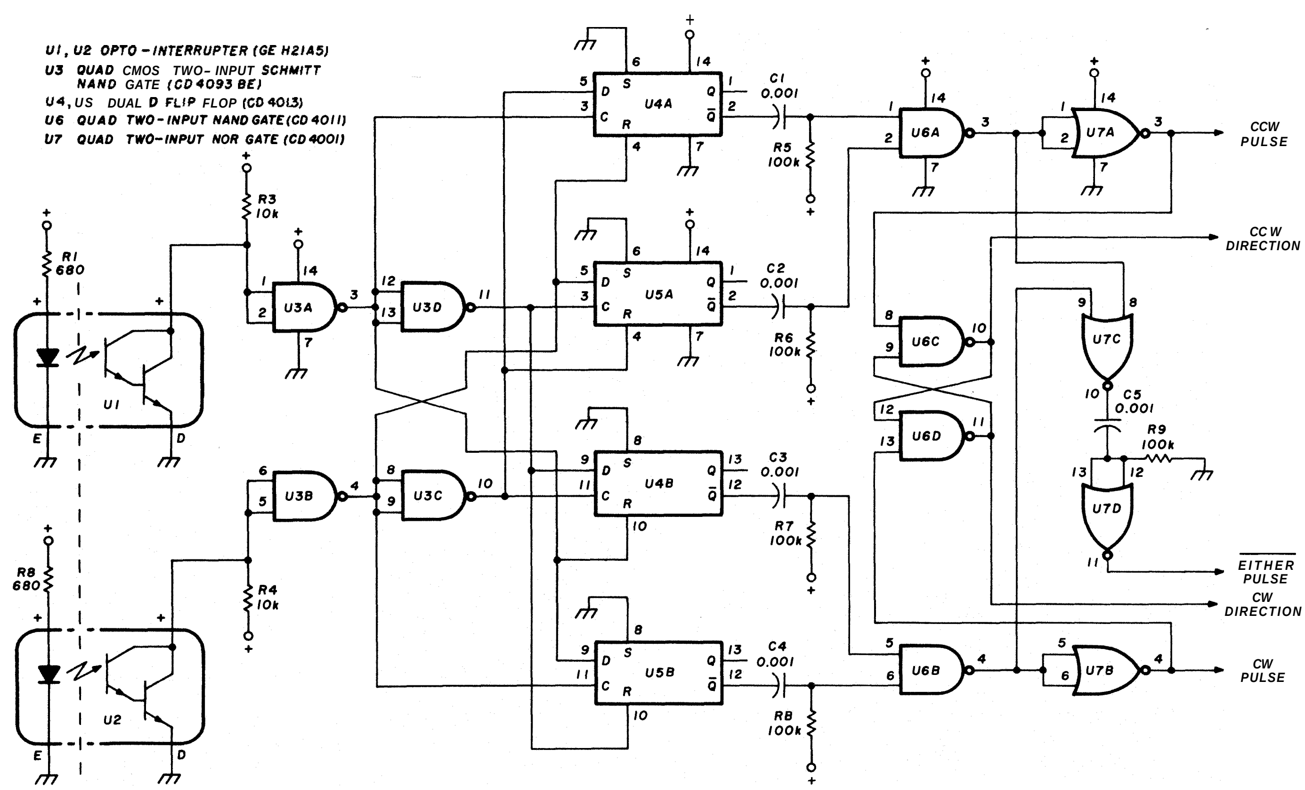


fig. 3. Schematic of the electronic processing circuit. GE H21A5 interrupters are used, but TI type TIXL45 and Monsanto type MCA8 should work as well. Output transistor load resistors, R1-R4, are for 5-volt operation. They should be increased proportionally if the circuit operating voltage is made higher.

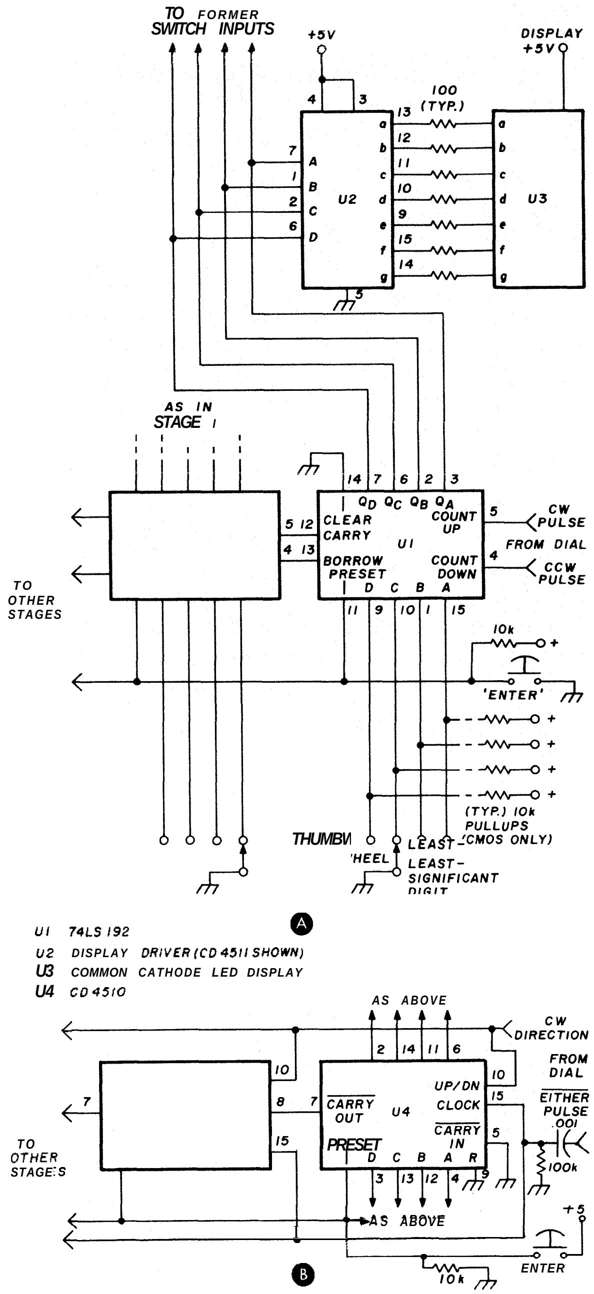


fig. 4. Examples of upldown counters that can be driven by the encoder. (A) shows an upldown counter with preset entry and an optional LED readout. The 74192 devices are highly recommended because of pinout compatibility with TTL and CMOS units. In (B) an upldown counter using a type CD 4510BE is shown, which is also compatible with the encoder. Standard IC power connections not shown.

The interrupter spacing should be tweaked until these pulses are evenly spaced.

applications

The simplest use of this dial system is as a supplement to existing pushbutton-tuned systems that already have an upldown counter. In this case it will be necessary only to merge the signals from the internal pulse generator with those from the dial mechanism. It may be possible to do this by simply adding a couple of diodes, although I have no direct experience with such a modification.

up/down counters

Most applications2 will require the addition of an up/down counter. The original thumbwheel or rotary switch wires are routed to the counter, and a pushbutton is added to initialize the counter to the settings on the switch dials. After that, tuning is with the shaft encoder.

Fig. 4A shows a suitable up/down counter with preset entry and an optional LED readout, if the transceiver does not have one. The circuit uses 74192 upldown counters, which are highly recommended because they are available as pin-compatible TTL (74192), low-power TTL (74L192) low-power Shottky TTL (74LS192), and CMOS (74C192). The logic type used should be the same as that in the synthesizer. The output of the dial electronics is compatible with all types.

Some other counters, such as the CMOS 4510 and TTL 74168, use the direction and clock control signals as shown in **fig. 4B.** This type of counter often presents multiple control signal and/or clock loads to the encoder, so buffers will be needed when using the encoder with non-CMOS units.

additional remarks

I've been using the prototype encoder to tune a 5-5.5 MHz frequency-lock loop in an experimental 80 and 20 meter receiver. I've tried 100-, 40-, 20-, and 10-Hz tuning increments. To me, 100-Hz steps are intolerably coarse for CW work, 40-Hz steps are noticeable, but 20-Hz steps seem smooth enough. With 10-Hz increments, the 100-mark disk gives a leisurely 2-kHz/revolution tuning rate. Although the knob is easily spun for fast tuning (the only drag is in the panel bearing), pushbuttons for faster tuning would be nice. Those readers who use 100-Hz tuning increments should use the 50-mark disk; if this gives tuning which is too fast, disable half the pulses by disconnecting C2 and C4.

I'd be pleased to correspond with anyone considering use of the encoder and would appreciate reports on successful applications. I'd be most grateful for an SASE with all correspondence.

references

1. *Lester Earnshaw, "Basics of the Digital VFO — A Tunable Synthesizer," ham radio, November, 1978, page 18.*
2. *Examples of equipment that could use the encoder are described by Earnshaw (op. cit.) and by Raymond C. Pettit, W7GHM, "Frequency Synthesized Local-Oscillator System for the High-Frequency Amateur Bands," ham radio, October, 1978, pages 60-65.*

ham radio

Yagi antenna design: optimizing performance

Considerations for optimizing element length and position for maximum forward gain and front-to-back ratio

Over the past two months I have explored the properties of ***simplistic*** Yagi-Uda antennas, i.e., antennas of a given boom length but having uniformly spaced elements one of which is a reflector, one of which a driver, and with director parasites all of uniform length.[1,2] In real life, however, one is not restricted to the simplistic design; it is therefore interesting to examine a number of departures from the simplistic design to see if there are ways to further improve performance; in this and succeeding articles I shall attempt to explore a few ideas in a systematic way. It will soon be apparent that some of the departures from the simplistic design produce only subtle changes in performance, while others are major departures which produce significant changes in performance. It is fortunate that the accuracy inherent in computation can show very subtle changes; these changes, though small, can usually be trusted although the absolute accuracy of the model on which computation is based may not be better than a few per cent.

Departures from the simplistic design may be accomplished in a number of ways, but primarily by allowing the ***lengths*** of the directors (hence their resonant frequencies) to vary and by allowing the ***placement*** of the element(s) on the boom to change. Additionally, for a given boom and a given total number of elements, the number of reflectors (and hence directors) can be changed. This is a much more drastic change, and produces more pronounced performance variations. These changes will be analyzed in this and subsequent articles. Only free-space performance will be investigated at this time.

I shall start with a "good" simplistic design (6-element Yagi on a boom 0.75 λ_o long), but will first change the lengths of all parasites to bring the center frequency of the desired **4** per cent band of maximum gain to $F = 1.0$. **Fig. 1** shows the main properties of this test antenna and the required element lengths. Note that the position of maximum F/B ratio is somewhat lower ($F = 0.988$) than the gain band center ($F = 1.0$); **fig. 2** shows the pattern of this antenna at band center ($F = 1.0$).

There are two visible nulls, the first one ($K = 1$) occurs at 87° and the second ($K = 2$) at 144°. The second null can be identified with the peak in the F/B ratio occurring at $F = 0.988$; at this lower frequency the null ($K = 2$) moves out to 180°. This antenna can be operated at best gain over the band ($F = 1.0$) and compromise F/B (17 dB at $F = 1.0$), or operated at the frequency of best F/B ($F = 0.988$, F/B = 38 dB) and somewhat compromise the gain available. In either case this Yagi-Uda design seems to be a good one and I shall use it as a test case around which certain departures from simplistic design can take place.

Since we are interested in ***very*** subtle changes I show in **table 1** the detailed performance in the region of chief interest, accurately calculated for this antenna over the frequency range from $F = 0.970$ to $F = 1.030$. The driving point reactance at a given frequency is somewhat arbitrary; it can be easily shifted or offset by changing the length of the driven element. The length of the driven element, however, remains fixed (free space resonance = 1.0) throughout this series of explorations so that reactance changes can be properly sensed.

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table 1. Computed performance characteristics of the 6-element Yagi over the frequency range from $F = 0.970$ to $F = 1.030$.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.058	12.928	24.677	-38.229
0.972	10.145	14.043	24.454	-35.512
0.974	10.225	15.279	24.190	-32.775
0.976	10.299	16.676	23.889	-30.006
0.978	10,369	18.298	23.559	-27.199
C.980	10.434	20.245	23.206	-24.347
0.982	10.495	22.700	22.838	-21.443
0.984	10.554	26.040	22.463	-18.483
0.986	10.608	31.184	22.088	-15.464
0.988	10.659	38.034	21.723	-12.381
0.990	10.706	31.810	21.375	-9.232
0.992	10.748	26.459	21.055	-6.016
0.994	10.785	23.044	20.770	-2.732
0.996	10.816	20.570	20.531	0.623
0.998	10.841	18.634	20.346	4.050
1.000	10.857	17.045	20.227	7.548
1.002	10.864	15.693	20.186	11.131
1.004	10.861	14.520	20.235	14.789
1.006	10.847	13.485	20.390	18.525
1.008	10.821	12.562	20.668	22.339
1.010	10.782	11.731	21.090	26.232
1.012	10.730	10.979	21.679	30.206
1.014	10.665	10.296	22.468	34.262
1.016	10.586	9.675	23.496	38.398
1.018	10.495	9.112	24.814	42.610
1.020	10.393	8.603	26.489	46.889
1.022	10.281	8.148	28.611	51.214
1.024	10.162	7.747	31.301	55.542
1.026	10.038	7.401	34.722	59.793
1.028	9.913	7.113	39.097	63.809
1.030	9.790	6.889	44.709	67.290

If we keep the average value of the length(s) of the directors constant we can explore "linear" length tapers and "parabolic" tapers. I will define a linear taper as a uniform linear free-space resonant frequency progression from D1 to *D4* keeping the ***average*** resonant frequency constant. In other words, the director resonant frequencies are linearly related to director position measured from the center of the director assembly (to keep the average value constant). Similarly, a "parabolic" taper is one in which the directors' free-space resonant frequencies are proportional to the square of the distance from the center of the director assembly, with the further condition that the average vaiue of director resonance is held constant. If we define the change in resonant frequency of *D1* as A, then with a total of four directors all other director resonances will change:

Element	Linear Taper	Parabolic Taper
Director 1	+A	+A
Director 2	$+\Delta/3$	- A
Director 3	$-\Delta/3$	- A
Director 4	- A	+A

The degree or magnitude of taper is fixed by the size of A; moreover, A can be chosen either as an increase or decrease in resonant frequency.

I have selected and investigated six taper schedules which are delineated in **table 2;** as indicated, antenna performance in the critical region of interest are shown in **tables** 3 to **8.** These results are to be compared with the simplistic design shown in **table 1.**

Table 3 shows the results for a " – 2 per cent" linear taper and it is obvious that all performance parameters are virtually unchanged! **Table 4** shows the results for a " – 4 per cent" *linear* taper, and even in this rather extreme case performance is almost totally unchanged in the central frequency region! Remember that for this case the free-space resonant frequency of *D1* has dropped from 1.06 to 1.02 and it is easy to see that performance deteriorates at higher frequencies (*D1* no longer behaves like a director). Nevertheless, it is remarkable that the linear taper — even of this magnitude — has virtually ***no*** effect on the performance of the Yagi-Uda antenna!

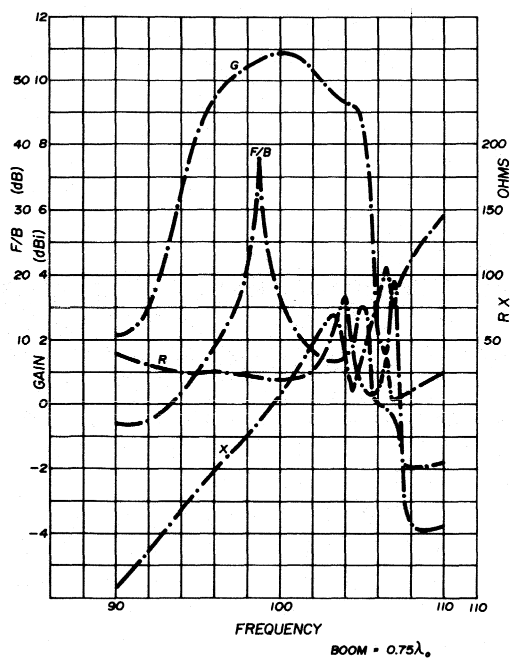
Table 5 shows results for a – 2 per cent parabolic taper; **tables 6, 7,** and **8** show results for parabolic tapers of – 1, + 1, and + 2 per cent, respectively. In comparing these tables with the standard simplistic Yagi-Uda antenna we again see that truly minimal changes are made to the chief performance indices. The variations in maximum F/B ratios are not of great significance; I shall come back to this point later.

Analogous to the ***length*** (and corresponding resonant frequency) taper variations I have also investigated element ***placement*** variations along the boom. Again it is possible to vary the space intervals between elements linearly and pseudo parabolically. **Table 9** shows element positions for six schedules I have investigated and the individual results are shown in **tables 10** through **15.**

Table 10 shows the results where elements are crowded towards *D4*. Note that truly large changes in placement have been made! Similarly, **table 13** shows the results where elements are severely crowded towards the reflector. **Tables 11** and **12** are intermediate schedules; **tables 14** and **15** show results where end spaces are (relatively) increased.

table 2. Schedule of director lengths (λ_o) for the six 6-element Yagi-Uda performance characteristics listed in table 3 through table 8.

table	taper type	Dir 1 (λ_o)	Dir 2 (λ_o)	Dir 3 (λ_o)	Dir 4 (λ_o)	
3	Linear	0.46341	0.45719	0.45263	0.44524	-2%
4	Linear	0.47306	0.46028	0.44817	0.43667	-4%
5	Parabolic	0.46341	0.44524	0.44524	0.46341	-2%
6	Parabolic	0.45873	0.44964	0.44964	0.45873	-1%
7	Parabolic	0.44964	0.45873	0.45873	0.44964	+1%
8	Parabolic	0.44524	0.46341	0.46341	0.44524	+2%



element	length (λ_o)	free space resonance
Reflector	0.50195	0.96
Driven Element	0.48167	1.00
Director 1	0.45414	1.06
Director 2	0.45414	1.06
Director 3	0.45414	1.06
Director 4	0.45414	1.06

fig. 1. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o. Note that the position of maximum F/B ratio (F = 0.988) is somewhat lower than the gain band center ($F = 1.0$). The pattern of this antenna at the band center is plotted in fig. 2.

These tables all show that these placement variations have only a ***very*** minor effect on directivity or gain, and while the maximum ***F/B*** ratio is somewhat affected (generally adversely), we shall soon see that it may not be very significant.

Up to this point we have looked at taper schedules which are ***linear*** and ***parabolic*** and which also involve director ***length,*** or resonant frequency, and element ***placement*** along the boom. It is truly remarkable that ***all*** of these schedules produce minimal changes in antenna performance; it is therefore plausible that ***combinations*** of these schedules will also produce minimal performance variations. This leads to the conclusion that the original simplistic design (dimensions listed in **fig. 1**) is just about as good as any. No real improvement on gain can be expected by any new tricky design; as far as ***F/B*** ratio is concerned, it will soon be apparent that you can "tune up" the maximum ***F/B*** ratio starting with almost any of these schedules.

A summary of raw performance of all of these cases is shown in **table 16,** where, in addition to data already shown in the previous tables, information on pattern (not explicitly shown here) has been added. This table shows that all cases produce about the same gain; the very small variations are due to the effective "illumination" pattern of the boom aperture.[2] The ***F/B*** ratio at central gain frequency ($F = 1.0$) varies somewhat, but a very slight change in operating frequency would easily make them all comparable. The frequency position of maximum F/B ratio and the angle of the second null at $F = 1.0$ are related. ***Lower*** frequencies of maximum ***F/B should*** correspond (at central frequency) to *longer* effective boom illuminated apertures, thus corresponding to ***lower*** null angles at $F = 1.0$ and somewhat ***higher*** gain. An examination of **table** 16 shows all of these quantities to be well correlated; it appears therefore that all results are understood and self-consistent.

From all of this information it is reasonable to draw a general conclusion that the simplistic Yagi design gives about as much gain as any other design off the

table 3. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director lengths tapered linearly at – 2 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.074	12.997	23.856	–39.110
0.972	10.160	14.083	23.636	– 36.425
0.974	10.238	15.280	23.379	– 33.725
0.976	10.309	16.622	23.089	– 31.003
0.978	10.375	18.164	22.771	– 28.250
0.980	10.435	19.993	22.433	– 25.462
0.982	10.492	22.262	22.082	– 22.634
0.984	10.544	25.284	21.723	– 19.762
0.986	10.592	29.878	21.365	– 16.843
0.988	10.635	39.987	21.016	– 13.876
0.990	10.674	38.240	20.684	– 10.859
0.992	10.708	29.358	20.375	– 7.794
0.994	10.736	25.064	20.099	– 4.679
0.996	10.758	22.209	19.862	– 1.516
0.998	10.774	20.068	19.673	1.694
1.000	10.781	18.356	19.542	4.948
1.002	10.780	16.927	19.478	8.251
1.004	10.770	15.706	19.491	11.594
1.006	10.750	14.643	19.593	14.972
1.008	10.720	13.705	19.796	18.381
1.010	10.680	12.872	20.115	21.812
1.012	10.629	12.128	20.567	25.259
1.014	10.568	11.462	21.173	28.707
1.016	10.497	10.869	21.960	32.141
1.018	10.418	10.343	22.957	35.535
1.020	10.331	9.884	24.203	38.851
1.022	10.240	9.492	25.742	42.033
1.024	10.145	9.171	27.623	44.991
1.026	10.049	8.925	29.894	47.585
1.028	9.954	8.764	32.572	49.596
1.030	9.861	8.701	35.588	50.696

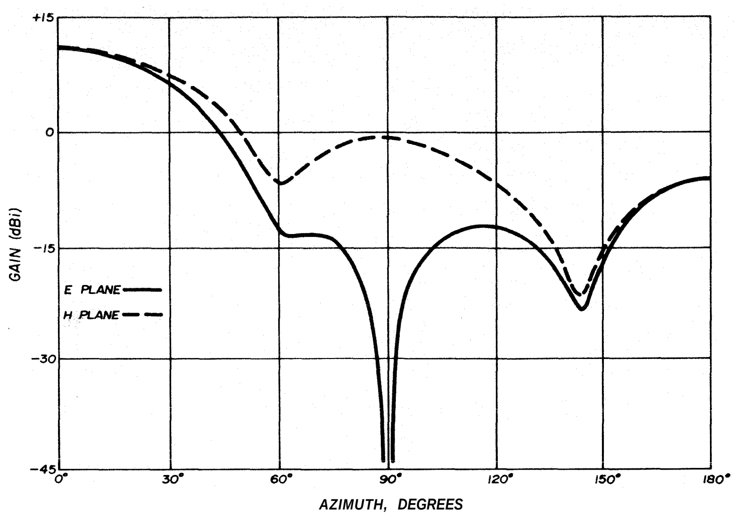


fig. 2. Radiation pattern of a 6-element Yagi beam on a $0.75\ \lambda_o$ boom (only 180 degrees are shown). The pattern nulls occur at 87 and 144 degrees; the second null is correlated with the peak of F/B ratio at $F = 0.988$.

same boom length. Tapering element lengths or element position intervals along the boom is of no apparent value. The characteristic of the directors which *is* important is the ***average*** length (or average free-space resonant frequency). But this conclusion has been demonstrated only for a boom length of $0.75\ \lambda_o$; we must be careful not to generalize too much. Recall that the NBS data (see **fig. 7** of the NBS report[3]) suggested that for booms longer than one wavelength some improvement in gain over simplistic Yagi-Uda performance could be obtained with particular director length schedules. My calculations support the NBS result; nevertheless, for boom lengths shorter than one wavelength the simplistic design is as good as any!

It can be seen from **table *16*** that the best "null" ($K = 2$) positions give quite different values of maximum F/B. Indeed, ***good*** nulls or correspondingly ***high*** values of ***F/B*** must be viewed as accidental vectorial cancellations in the reverse or back direction; such good cancellations will not generally occur with any arbitrary boom illumination. Note that the various cases shown in **table *16*** display maximum ***F/B*** ratios ranging from 22 to 40 dB. It is an interesting exercise to see if there is some way to significantly enhance the maximum ***F/B*** ratio by some variational procedure.

Let us start with the simplistic Yagi-Uda design (***fig.* 1**) and vary the ***position*** of, say $D3$, along the boom. We now know that small variations in position will not significantly affect gain, but vectorial cancellation effects in the back direction can be expected to be significant. If we can find a position for $D3$ which maximizes the ***F/B*** ratio, its vectorial contribution in the back direction should be approximately out of phase with the residue from all other elements. At this point some other element (say $D1$) can be positioned for a new (still higher) maximum ***F/B*** ratio; after this is done $D3$ can be readjusted again for a new maximum ***F/B*** ratio, etc. By iterating the two adjustments it should be possible to continuously improve ***F/B*** ratios, presumably to as high a value as desired! With such an iteration procedure it is desirable to start with a fairly good value of F/R so that only small variations in element position can have a significant effect.

I have carried out such an iteration (using $D3$ and $D1$) for the simplistic Yagi-Uda design and have arrived at the following positions:

element	$X(\lambda_o)$
Reflector	0.000
Driven Element	0.150
Director 1	0.28967
Director 2	0.450
Director 3	0.58945
Director 4	0.750

table 4. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director lengths tapered linearly at – 4 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.064	13.741	20.930	– 40.263
0.972	10.152	14.915	20.598	– 37.496
0.974	10.232	16.225	20.240	– 34.709
0.976	10.305	17.715	19.860	– 31.898
0.978	10.371	19.458	19.464	– 29.058
0.980	10.432	21.568	19.058	– 26.185
0.982	10.486	24.235	18.648	– 23.276
0.984	10.535	27.758	18.241	– 20.331
0.986	10.577	31.908	17.841	– 17.348
0.988	10.613	31.851	17.456	– 14.328
0.990	10.642	27.723	17.092	– 11.271
0.992	10.663	24.260	16.755	–8.181
0.994	10.676	21.650	16.450	– 5.058
0.996	10.679	19.600	16.183	– 1.907
0.998	10.672	17.921	15.959	1.268
1.000	10.654	16.504	15.785	4.463
1.002	10.624	15.276	15.666	7.677
1.004	10.580	14.197	15.607	10.897
1.006	10.523	13.237	15.612	14.115
1.008	10.452	12.374	15.687	17.320
1.010	10.365	11.594	15.835	20.498
1.012	10.264	10.886	16.058	23.631
1.014	10.147	10.242	16.356	26.698
1.016	10.013	9.653	16.721	29.669
1.018	9.861	9.115	17.140	32.511
1.020	9.689	8.621	17.578	35.181
1.022	9.489	8.159	17.977	37.637
1.024	9.252	7.713	18.235	39.849
1.026	8.955	7.252	18.205	41.835
1.028	8.562	6.722	17.715	43.715
1.030	8.012	6.025	16.654	45.750

table 5. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director lengths tapered parabolically at – 2 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.013	16.381	19.732	– 42.299
0.972	70.112	18.110	19.184	– 39.363
0.974	10.207	20.196	18.625	– 36.379
0.976	10.298	22.808	18.063	– 33.345
0.978	10.386	26.155	17.504	– 30.258
0.980	10.471	29.690	16.954	–27.117
0.92	10.553	29.326	16.421	– 23.921
0.984	10.632	25.681	15.910	– 20.669
0.986	10.706	22.446	15.428	– 17.361
0.988	10.776	19.930	14.980	– 13.998
0.990	10.840	17.920	14.572	– 10.578
0.992	10.897	16.257	14.211	–7.103
0.994	10.946	14.841	13.903	– 3.573
0.996	10.986	13.610	13.653	0.013
0.998	11.014	12.522	13.469	3.655
1.000	11.030	11.547	13.359	7.354
· 1.002	11.033	10.662	13.332	11.124
1.004	11.021	9.856	13.398	14.956
1.006	10.993	9.120	13.570	18.854
1.008	10.950	8.444	13.864	22.822
1.010	10.892	7.824	14.299	26.865
1.012	10.820	7.256	14.903	30.991
1.014	10.735	6.737	15.709	35.208
1.016	10.641	6.267	16.768	39.527
1.018	10.540	5.846	18.148	43.959
1.020	10.436	5.475	19.953	48.511
1.022	10.336	5.157	22.337	53.182
1.024	10.243	4.897	25.544	57.936
1.026	10.163	4.702	29.962	62.653
1.028	10.104	4.583	36.221	66.971
1.030	10.070	4.555	45.262	69.868

This design, optimized for maximum F/B ratio (at $F = 0.990$) produces the performance displayed in table **17.** A careful comparison of this table with the original simplistic model shows virtually identical performance in all respects ***except*** that the F/B maximum has gone up from an excellent 38 dB to an astounding 98 dB! Even this high value is not a real limit; it is limited only by the number of iterations which were made.

Notice that these astronomical values of F/B are of no practical significance. It occurs at essentially a single frequency and its effective bandwidth becomes vanishingly small. Moreover, extremely small variations in the Yagi-Uda dimensions will upset the cancellation; in practice you could not likely construct a mechanically satisfactory, fully optimized Yagi-Uda antenna. Nevertheless, the mathematical iteration shows that it is possible, in principle, to obtain (at a single frequency) an arbitrarily high F/B ratio. It is likely that there are a large number of potential solutions involving iterations with other elements. Furthermore, we now know that the variations in F/B maxima shown in table **16** result from the particular illumination chosen, and it is very likely that minor element placement variations could make an arbitrarily high F/B ratio design starting from any of the cases shown.

To understand this iteration procedure it is helpful to show the vectorial contributions of each element to the forward and back waves. The (current) contribution from a given element will be a vector whose magnitude is the magnitude of element current and whose phase consists of two parts. The first part is the actual (time) phase of the element current referred to some time origin (say the driver current) and the second is the (space) phase change due to the element position referred to some space origin (say at X = 0 along the boom). Note that this second part changes sign in going from a forward wave to a reverse wave! **Fig.** 3 shows these (current) vectorial contributions at $F = 0.988$ for the original simplistic Yagi-Uda design **(fig. 1)** to both forward and reverse waves.

table 6. Performance characteristics of a 6-element Yagi beam with a boom length of 0 75 λ_o, director lengths tapered parabolically at – 1 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.040	14.338	22.533	– 40.493
0.972	10.132	15.667	22.133	– 37.696
0.974	10.219	17.186	21.704	– 34.863
0.976	10.300	18.973	21.253	– 31.986
0.978	10.378	21.151	20.787	– 29.060
0.980	10.453	23.931	20.313	– 26.082
0.982	10.524	27.615	19.839	– 23.048
0.984	10.592	31.590	19.372	– 19.956
0.986	10 656	30.418	18.920	– 16.804
0.988	10.717	26.254	18.490	– 13.590
0.990	10.773	22.942	18.089	– 10.314
0.992	10.824	20.432	17.725	– 6.975
0.994	10.870	18.445	17.406	– 3.572
0.996	10.908	16.807	17.138	– 0.105
0.998	10.938	15.417	16.932	3.427
1.000	10.959	14.209	16.794	7.024
1.002	10.970	13.140	16.736	10.700
1.004	10.969	12.183	16.770	14.446
1.006	10.956	11.321	16.909	18.265
1.008	10.930	10.538	17.168	22.160
1.010	10.890	9.825	17.568	26.133
1.012	10.838	9.173	18.134	30.191
1.014	10.772	8.578	18.896	34.338
1.016	10.694	8.035	19.899	38.580
1.018	10.607	7.543	21.198	42.922
1.020	10.512	7.101	22.875	47.367
1.022	10.413	6.710	25.041	51.906
1.024	10.312	6.372	27.863	56.516
1.026	10.215	6.089	31.585	61.122
1.028	10.126	5.869	36.575	65.547
1.030	10.051	5.720	43.373	69.362

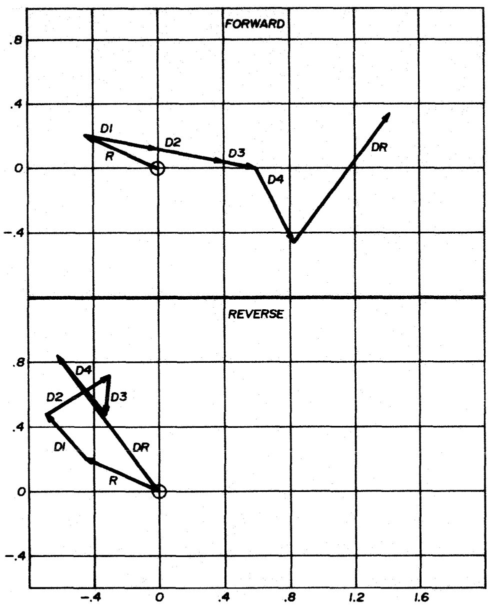


fig. 3. Current vectorial contributions at F = 0.988 for 6-element Yagi (boom = 0.75 λ_o), forward and reverse waves.

Note that for the forward wave the individual element contributions do not all fully reinforce the forward wave; in fact, the contribution from the reflector is even ***negative*** with respect to the final total (current) vector! This curious result is typical of all Yagi-Uda arrays. Note that the contributions to the back or reverse wave, in total, nearly cancel out, leaving only a small residue which accounts for the 38 dB F/B ratio. Now it is easy to see conceptually what happens in the iterative procedure to reduce the backwave residual.

If you look at the reverse wave vector plot, it is easy to imagine that as $D3$ is moved along the boom, the $D3$ vector rotates around ***its*** origin. The backwave residual will then be changed along an axis at right angles to the $D3$ vector and can be minimized by the $D3$ position. After this is done ***another*** element, say $D1$, can be moved along the boom; its vector contribution is at a different angle and can therefore reduce the residual still further. Thus, in principle, iterative motions of two elements whose backwave vectors contribute at different angles can ultimately reduce the backwave residual to as low a value as desired.

The iterative convergence will be most rapid if the two element vectors are orthogonal; nevertheless, it can converge adequately for many element combinations. Of course, this conceptual picture is oversimplified; as any element is moved on the boom not only does its vector rotate, but ***all*** element currents and phases readjust somewhat. However, these readjustments are usually minor and in practice cause little difficulty as long as you start with a reasonably small residual as shown in **fig.** 3.

Fig. 4 shows the vectorial contributions for the optimized Yagi-Uda beam. Note that the element contributions are only slightly modified in the optimization procedure.

At this point I must issue a warning. Recall that the mathematical model being used in these computations involves certain approximations. These approximations make relatively little difference in the calculations for forward gain, but they become crucial in calculations involving vectorial cancellation or closure for back radiation. Thus the explicitly calculated ***positions*** for and ***magnitude*** of a very high F/B ratio are not to be trusted. Nevertheless, the ***general*** behavior is still valid. The real Yagi-Uda can still be made to have a high F/B ratio, just as our mathematical model shows, but it may occur at a slightly different frequency and it may require slightly different positions for $D3$ and $D1$ in the final optimization.

table 7. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director lengths tapered parabolically at +1 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.066	11.872	26.168	−35.821
0.972	10.150	12.855	26.121	−33.131
0.974	10.225	13.922	26.025	−30.437
0.976	10.293	15.096	25.884	−27.728
0.978	10.355	16.414	25.702	−24.993
0.980	10.412	17.927	25.485	−22.222
0.982	10.465	19.720	25.240	−19.407
0.984	10.514	21.939	24.975	−16.542
0.986	10.558	24.879	24.699	−13.619
0.988	10.599	29.294	24.419	−10.635
0.990	10.635	38.438	24.147	−7.584
0.992	10.667	38.763	23.891	−4.463
0.994	10.693	29.441	23.663	−1.269
0.996	10.714	25.008	23.472	2.000
0.998	10.727	22.079	23.330	5.345
1.000	10.732	19.886	23.248	8.768
1.002	10.728	18.127	23.244	12.280
1.004	10.714	16.660	23.328	15.873
1.006	10,688	15,401	23.519	19.547
1.008	10.650	14.301	23.833	23.302
1.010	10.597	13.325	24.293	27.136
1.012	10.530	12.450	24.923	31.048
1.014	10.448	11.661	25.755	35.033
1.016	10.350	10.946	26.823	39.083
1.018	10.237	10.296	28.174	43.188
1.020	10.108	9.708	29.862	47.325
1.022	9.965	9.177	31.956	51.463
1.024	9.808	8.701	34.542	55.548
1.026	9.640	8.280	37.723	59.490
1.028	9.462	7.916	41.618	63.148
1.030	9.277	7.613	46.355	66.289

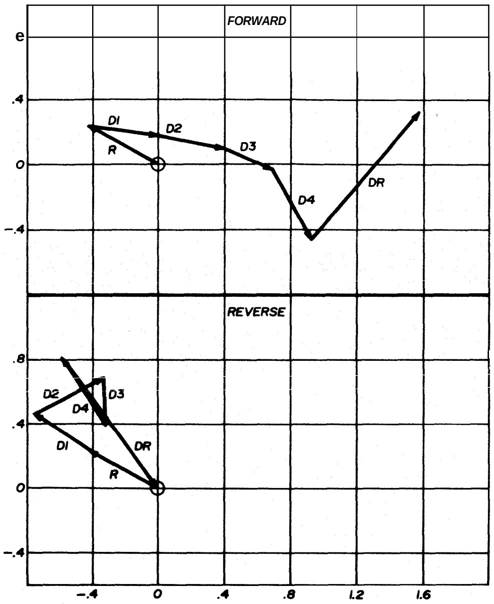


fig. 4. Current vectorial contributions for the optimized 6-element Yagi at $F = 0.990$.

optimum design

We now have the necessary tools with which to design truly excellent Yagi-Uda antennas. We start first from a knowledge that the boom length should be approximately an odd number of quarter wavelengths for an initial simplistic design; we have seen that such a boom length promotes an inherently high F/B ratio at a frequency near the center of the best gain band; boom length also determines ultimate gain. After the boom length is chosen a resonant frequency schedule is chosen (see appropriate figures from the simplistic Yagi-Uda articles[1,2]) for reflector and director(s); a preliminary calculation is then made to accurately determine the frequency of maximum F/B ratio which will not necessarily correspond with the frequency at the center of the best gain portion. The *useful* band, however, is now to be centered around the F/B point; it is necessary to insure that there is enough gain bandwidth left for the intended purpose.

Remember that the overall gain bandwidth is basically controlled by the resonant frequency schedule of the parasites. This bandwidth should not be larger than necessary, because gain is compromised somewhat as the bandwidth increases.

Now translate the frequency ($F1$) of best F/B to F = 1.0 by multiplying all parasite lengths by $F1$; a new preliminary calculation, possibly iterated once more, will insure that the best F/B ratio is exactly F = 1.0. Next, alternately vary the X position of $D3$

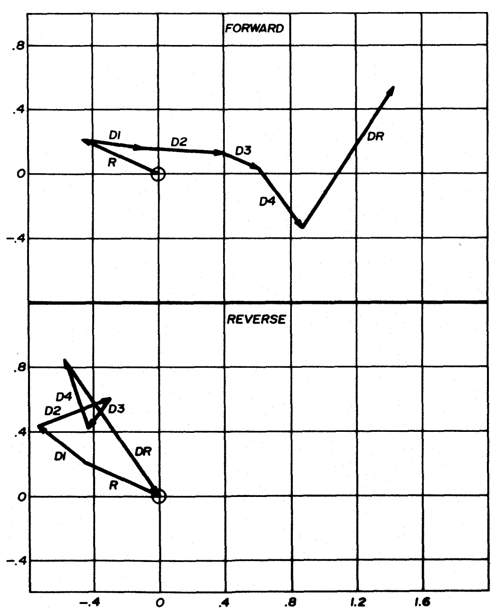


fig. 5. Current vectorial contributions of the *03-01* optimized Yagi; forward wave, above, and reverse wave, below.

table 8. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director lengths tapered parabolically at +2 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.065	11.048	27.104	−33.539
0.972	10.145	11.945	27.202	−30.843
0.974	10.215	12.904	27.249	−28.160
0.976	10.277	13.942	27.243	−25.474
0.978	10.332	15.081	27.187	−22.775
0.980	10.382	16.353	27.086	−20.051
0.982	10.426	17.805	26.944	−17.289
0.984	10.465	19.507	26.770	−14.482
0.986	10.500	21.575	26.572	−11.621
0.988	10.530	24.216	26.358	−8.697
0.990	10.555	27.812	26.138	−5.704
0.992	10.575	32.685	25.924	−2.637
0.994	10.588	33.646	25.725	0.508
0.996	10.594	28.730	25.555	3.736
0.998	10.592	24.844	25.424	7.050
1.000	10.581	22.026	25.348	10.453
1.002	10.559	19.844	25.342	13.954
1.004	10.525	18.071	25.421	17.548
1.006	10.476	16.576	25.602	21.236
1.008	10.412	15.281	25.902	25.017
1.010	10.330	14.139	26.345	28.891
1.012	10.228	13.116	26.954	32.855
1.014	10.106	12.189	27.756	36.904
1.016	9.962	11.341	28.786	41.032
1.018	9.794	10.559	30.080	45.228
1.020	9.602	9.835	31.684	49.473
1.022	9.384	9.161	33.649	53.741
1.024	9.140	8.530	36.034	57.993
1.026	8.870	7.938	38.907	62.169
1.028	8.570	7.378	42.337	66.185
1.030	8.241	6.846	46.394	69.921

table 9. Schedule of director placement on boom (λ_o) for the six 6-element Yagi-Uda performance characteristics listed in table 10 through table 15.

table	taper type	element position on boom (λ_o) Refl	DR	D1	D2	D3	D4
10	Linear	0	0.200	0.3750	0.5250	0.650	0.750
11	Linear	0	0.175	0.3375	0.4875	0.625	0.750
12	Linear	0	0.125	0.2625	0.4125	0.575	0.750
13	Linear	0	0.100	0.2250	0.3750	0.500	0.750
14	Parabolic	0	0.200	0.3167	0.4333	0.550	0.750
15	Parabolic	0	0.175	0.3083	0.4417	0.575	0.750

and then $D1$ to get larger and larger values of F/B at $F = 1.0$ until the value is sufficiently high.

design example

An example will illustrate this design procedure. For example, let's choose a boom length of 0.780 λ_o. From an inspection of the results of our test Yagi-Uda simplistic design (**fig. 1**) we can probably use the same parasite resonant frequency schedule and still obtain an adequate ultimate gain bandwidth performance. Listed below are the initial element positions along the boom:

element	X (λ_o)	initial length (λ_o)	intermediate length (λ_o)
Reflector	0.000	0.50195	0.49343
Driven El	0.156	0.48167	0.48167
Director 1	0.312	0.45414	0.44643
Director 2	0.468	0.45414	0.44643
Director 3	0.624	0.45414	0.44643
Director 4	0.780	0.45414	0.44643

table 10. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director spacing tapered linearly, large positive interval, directors crowded toward $D4$.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.872	11.767	43.699	−38.627
0.972	9.912	12.501	42.699	−36.912
0.974	9.952	13.285	41.580	−35.082
0.976	9.992	14.127	40.362	−33.122
0.978	10.034	15.036	39.064	−31.025
0.980	10.078	16.025	37.706	−28.783
0.982	10.123	17.099	36.306	−26.392
0.984	10.171	18.259	34.883	−23.851
0.986	10.221	19.481	33.453	−21.161
0.988	10.273	20.686	32.032	−18.323
0.990	10.326	21.689	30.634	−15.341
0.992	10.381	22.191	29.272	−12.219
0.994	10.435	21.951	27.960	−8.960
0.996	10.489	21.034	26.709	−5.571
0.998	10.540	19.734	25.528	−2.055
1.000	10.587	18.319	24.429	1.581
1.002	10.628	16.928	23.427	5.334
1.004	10.661	15.619	22.524	9.196
1.006	10.683	14.407	21.732	13.164
1.008	10.691	13.289	21.060	17.233
1.010	10.682	12.256	20.518	21.398
1.012	10.654	11.300	20.116	25.656
1.014	10.604	10.412	19.866	30.001
1.016	10.529	9.586	19.780	34.432
1.018	10.427	8.816	19.875	38.944
1.020	10.299	8.098	20.168	43.535
1.022	10.143	7.428	20.680	48.203
1.024	9.961	6.805	21.438	52.946
1.026	9.756	6.226	22.478	57.762
1.028	9.530	5.693	23.843	62.649
1.030	9.289	5.205	25.595	67.602

table 11. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director spacing tapered linearly as shown in table 9 (mild positive linear interval).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.039	12.781	33.045	−37.524
0.972	10.096	13.710	32.479	−35.253
0.974	10.151	14.723	31.839	−32.918
0.976	10.204	15.841	31.136	−30.510
0.978	10.256	17.094	30.381	−28.020
0.980	10.308	18.519	29.588	−25.443
0.982	10.359	20.164	28.767	−22.772
0.984	10.409	22.078	27.931	−20.005
0.986	10.459	24.247	27.093	−17.140
0.988	10.509	26.340	26.262	−14.177
0.990	10.557	27.212	25.450	−11.116
0.992	10.603	25.986	24.667	−7.956
0.994	10.646	23.746	23.923	−4.700
0.996	10.685	21.521	23.228	−1.350
0.998	10.719	19.567	22.592	2.094
1.000	10.747	17.882	22.023	5.629
1.002	10.767	16.414	21.536	9.264
1.004	10.777	15.123	21.136	12.988
1.006	10.775	13.973	20.836	16.799
1.008	10.760	12.938	20.648	20.697
1.010	10.729	11.998	20.584	24.681
1.012	10.682	11.140	20.659	28.749
1.014	10.617	10.353	20.893	32.903
1.016	10.534	9.629	21.306	37.142
1.018	10.432	8.961	21.925	41.466
1.020	10.312	8.347	22.785	45.875
1.022	10.174	7.783	23.929	50.368
1.024	10.022	7.268	25.415	54.941
1.026	9.857	6.802	27.321	59.584
1.028	9.684	6.386	29.750	64.273
1.030	9.505	6.023	32.855	68.965

Initial performance of this Yagi-Uda model is shown in table 18; the frequency for maximum *F/B* is $F = 0.984$. Shortening all elements by approximately this frequency factor yields the intermediate design also shown above. Performance for this intermediate design is shown in table **19.** Note that since all lengths were not scaled (boom not scaled), this intermediate Yagi-Uda is not really quite the same as our starting model; the maximum *F/B* ratio has, in fact, fallen to 27 dB. However, this is of no concern; it is now time to iteratively vary *D3* and *D1* positions to "tune up" the *F/B* ratio. Alternatively, if our concept of optimization is correct, iterative variations of *D3* and *DR* could also tune up the *F/B* ratio. I have carried out both iterations and the resulting optimized Yagi parameters are as shown:

element	length (λ_o)	*D3-D1* opt X (λ_o)	*D3-DR* opt X(λ_o)
Reflector	0.49343	0.000	0.000
Driven El	0.48167	0.156	0.175595
Director 1	0.44643	0.291564	0.312
Director 2	0.44643	0.468	0.468
Director 3	0.44643	0.64075	0.6328873
Director 4	0.44643	0.780	0.780

Performance of this *D3-D1* optimized antenna is shown in table 20; it is nearly the same as that of the intermediate design (table **19)** except that the *F/B* ratio at $F = 1.0$ has gone up from 27 dB to an astronomical 120 dB! Similarly, the performance for the *D3-DR* optimized antenna is shown in table **21.** Again an astounding *F/B* ratio figure is achieved; moreover, the newer optimized beam performance is essentially identical with that of the first optimized model!

It is instructive to examine the final vector contributions to forward and reverse waves; **fig.** 5 shows such a plot for the *D3-D1* optimized Yagi-Uda and **fig.** 6 a similar plot for the *D3-DR* optimized model. Note that they look similar, differing only in minute details. Incidentally, it is noteworthy that the reverse plots show vectorial contributions going around the

table 12. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director spacing tapered linearly as shown in table 9 (mild negative linear interval).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.877	11.719	17.840	-41.204
0.972	10.024	12.962	17.848	-38.079
0.974	10.155	14.333	17.848	-34.966
0.976	10.271	15.879	17.838	-31.856
0.978	10.375	17.669	17.821	-28.741
0.980	10.468	19.823	17.800	-25.615
0.982	10.552	22.556	17.777	-22.470
0.984	10.627	26.310	17.758	-19.300
0.986	10.694	31.942	17.750	-16.100
0.988	10.753	34.618	17.757	-12.865
0.990	10.805	28.634	17.788	-9.590
0.992	10.849	24.386	17.851	-6.272
0.994	10.885	21.483	17.956	-2.905
0.996	10.913	19.314	18.111	0.512
0.998	10.932	17.592	18.328	3.984
1.000	10.942	16.170	18.622	7.513
1.002	10.943	14.958	19.011	11.116
1.004	10.934	13.909	19.511	14.783
1.006	10.914	12.989	20.145	18.516
1.008	10.884	12.173	20.939	22.317
1.010	10.843	11.445	21.930	26.185
1.012	10.793	10.795	23.158	30.117
1.014	10.733	10.213	24.682	34.104
1.016	10.664	9.694	26.572	38.128
1.018	10.588	9.236	28.926	42.152
1.020	10.507	8.836	31.868	46.112
1.022	10.422	8.495	35.562	49.886
1.024	10.336	8.216	40.211	53.254
1.026	10.252	8.001	46.034	55.818
1.028	10.171	7.857	53.178	56.886
1.030	10.098	7.795	61.476	55.329

table 13. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director spacing tapered linearly with the elements crowded toward the reflector (large negative interval taper).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.271	8.830	11.990	-47.108
0.972	9.569	10.190	12.078	-43.541
0.974	9.823	11.648	12.199	-40.009
0.976	10.040	13.240	12.349	-36.506
0.978	10.225	15.016	12.523	-33.027
0.980	10.383	17.045	12.720	-29.565
0.982	10.517	19.420	12.940	-26.112
0.984	10.631	22.216	13.185	-22.662
0.986	10.727	25.165	13.458	-19.209
0.988	10.807	26.553	13.764	-15.746
0.990	10.873	24.901	14.109	-12.267
0.992	10.925	22.296	14.501	-8.776
0.994	10.964	19.998	14.950	-5.237
0.996	10.992	18.123	15.468	-1.675
0.998	11.009	16.585	16.071	1.924
1.000	11.014	15.299	16.776	5.563
1.002	11.010	14.210	17.609	9.241
1.004	10.995	13.272	18.595	12.963
1.006	10.971	12.455	19.768	16.728
1.008	10.939	11.741	21.173	20.528
1.010	10.899	11.114	22.864	24.352
1.012	10.852	10.564	24.916	28.172
1.014	10.800	10.085	27.420	31.940
1.016	10.744	9.674	30.495	35.570
1.018	10.686	9.326	34.284	38.911
1.020	10.627	9.044	38.945	41.704
1.022	10.571	8.829	44.607	43.505
1.024	10.518	8.685	51.240	43.611
1.026	10.472	8.620	58.368	41.024
1.028	10.435	8.643	64.591	34.749
1.030	10.408	8.770	67.340	24.792

clock ***twice*** corresponding to the $K = 2$ null which we have constructed.

There is one final point worth mentioning. An examination of **tables 1, 20,** and **21** reveals that the frequency for the best F/B ratio is not generally quite the same as the frequency center of the gain bandwidth. It is offset by an amount which depends only on the boom length. This offset is of small importance as long as the gain bandwidth is large enough; it is nevertheless possible to empirically measure the offset frequency as a function of boom length. Let us fix the frequency of best F/B ratio as $F = 1.0$, and designate the frequency of (central) best gain (4 per cent BW) as F_G (Offset $frequency = F_G - 1.0$); empirical results are shown in **fig.** 7.

Note that if the boom length is $0.63\lambda_o$ the offset disappears. For booms shorter than this value the offset is negative, and for booms longer than $0.63\lambda_o$ the offset is positive. But it is clearly possible to design a satisfactory Yagi over a considerable range of boom lengths without incurring an offset which is comparable to the bandwidth itself; it is only neces-

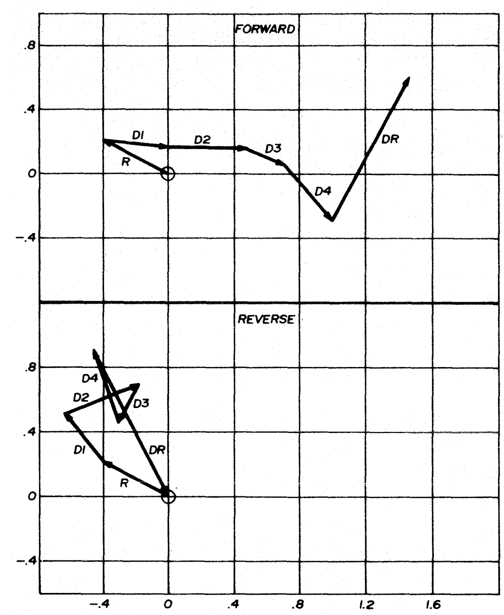


fig. 6. Current vector contributions for the $D3$-DR optimized Yagi beam.

table 14. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director spacing tapered pseudo parabolically according to the schedule of table 9 (large positive interval).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.248	15.471	33.711	-34.346
0.972	10.301	16.499	33.381	-31.789
0.974	10.350	17.639	33.016	-29.188
0.976	10.398	18.923	32.623	-26.535
0.978	10.442	20.389	32.209	-23.924
0.980	10.485	22.087	31.780	-21.049
0.982	10.526	24.059	31.344	-18.204
0.984	10.564	26.253	30.909	-15.284
0.986	10.600	28.196	30.485	-12.285
0.988	10.634	28.611	30.079	-9.203
0.990	10.664	27.043	29.703	-6.034
0.992	10.691	24.764	29.366	-2.774
0.994	10.714	22.595	29.079	0.580
0.996	10.731	20.705	28.855	4.033
0.998	10.743	19.071	28.707	7.588
1.000	10.749	17.645	28.651	11.250
1.002	10.747	16.388	28.706	15.021
1.004	10.736	15.262	28.891	18.909
1.006	10.716	14.243	29.228	22.918
1.008	10.685	13.312	29.746	27.056
1.010	10.643	12.457	30.482	31.330
1.012	10.588	11.667	31.479	35.747
1.014	10.522	10.935	32.794	40.313
1.016	10.442	10.254	34.504	45.035
1.018	10.350	9.621	36.705	49.912
1.020	10.247	9.033	39.535	54.935
1.022	10.132	8.488	43.177	60.072
1.024	10.009	7.986	47.893	65.246
1.026	9.879	7.526	54.045	70.288
1.028	9.745	7.109	62.141	74.841
1.030	9.611	6.739	72.838	78.159

table 15. Performance characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, director spacing tapered pseudo parabolically according to the schedule of table 9 (mild positive interval).

frequency (F)	gain (dBi)	FIB ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.195	13.965	30.953	-35.247
0.972	10.257	14.993	30.705	-32.716
0.974	10.315	16.131	30.412	-30.152
0.976	10.369	17.412	30.079	-27.546
0.978	10.420	18.888	29.713	-24.890
0.980	10.469	20.636	29.322	-22.177
0.982	10.515	22.792	28.915	-19.400
0.984	10.558	25.606	28.500	-16.556
0.986	10.599	29.598	28.086	-13.639
0.988	10.637	35.381	27.683	-10.646
0.990	10.673	34.817	27.302	-7.573
0.992	10.704	29.112	26.952	-4.419
0.994	10.732	25.234	26.645	-1.182
0.996	10.755	22.471	26.391	2.142
0.998	10.772	20.342	26.203	5.553
1.000	10.782	18.610	26.096	9.053
1.002	10.785	17.148	26.086	12.658
1.004	10.780	15.884	26.190	16.356
1.006	10.765	14.770	26.428	20.151
1.008	10.740	13.777	26.823	24.044
1.010	10.704	12.882	27.405	28.036
1.012	10.655	12.069	28.209	32.129
1.014	10.595	11.328	29.282	36.321
1.016	10.522	10652	30.680	40.609
1.018	10.437	10.033	32.481	44.981
1.020	10.341	9.470	34.781	49.413
1.022	10.236	8959	37.715	53.861
1.024	10.123	8.501	41.468	58.239
1.026	10.004	8.096	46.283	62.379
1.028	9.882	7.747	52.474	65.967
1.030	9.761	7.458	60.408	68.405

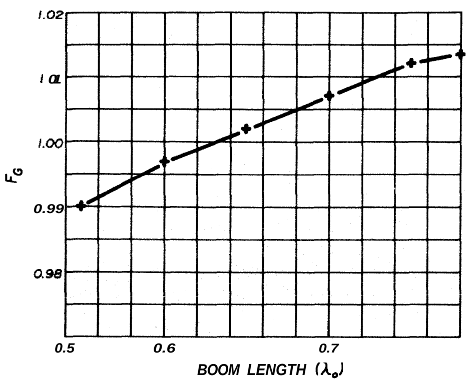


fig. 7. Plot illustrating frequency ratio for best central gain to best *F/B*. Note that frequency offset disappears for a boom length of 0.63 λ_o.

sary to take this offset into account in fixing the original bandwidth over which gain must be high. From a gain consideration alone the longer booms are best; that is why the example I used for illustrative purposes had a boom of 0.78 λ.

number of reflectors

It is interesting to consider a major change in possible Yagi-Uda antenna design: to explore the effect of changing the number of reflectors in a Yagi-Uda array. Up to this point we have assumed only a single reflector with a variable number of directors. It is tempting to consider increasing the number of reflectors in the hope of a significant improvement in the average F/B ratio over the entire bandwidth to be used. This question is now easily explored. I shall assume that the simplistic test Yagi-Uda of **fig. 1** will be our standard. To keep conditions other than the number of reflectors as constant as possible I shall keep the total boom length constant at 0.75 λ and the total number of parasites constant at five. We shall compare the cases where the number of reflectors is zero, on (our test standard), two, and three. **Fig. 8** shows frequency-swept gain curves for all four cases; the curves are keyed to the legend on the diagram. Severe resonance effects are noticed near the free-space resonances of the reflector ($FR = 0.96$) and the directors ($FR = 1.06$); these resonances, however, were purposely spread far enough to allow the **4** per cent band of interest to display a good gain figure,

The highest curve (curve 1) displays gain for the standardsimplistic Yagi-Uda (same as **fig. 1**) and it is clearly the best performer. The zero reflector case (curve 0) yields substantially less gain in the region of interest; it also contains no resonance effect at the reflector frequency, because there is no reflector. The two- and three-reflector cases (curves 2 and 3) show progressive loss of gain over the original standard; the reason is to be found in the much lower currents induced in the additional reflectors.

Shown below are the reflector currents when the driver is excited by one ampere at the central frequency ($F = 1.0$):

number of reflectors	magnitude of reflector current (amps) 1	2	3
1	0.477		
2	0.043	0.538	
3	0.031	0.046	0.626

Note that the reflector next to the driver has substantial current while all other drivers are hardly excited at all. Thus where there are multiple reflectors, the ef-

ble 16. Performance comparison of 6-element Yagi beams with ırying director lengths and element positions along the boom ıows little gain variation.

ɔm ble	gain (dBi)	F=1.0 F/B (dB)	max F/B	at freq	resist (ohms)	angle K=2 null
1	10.857	17.04	38.03	0.988	20.23	144
3	10.781	18.36	39.99	0.988	19.54	144
4	10.654	16.50	31.91	0.986	15.79	138
5	11.030	11.55	29.69	0.980	13.36	138
6	10.959	14.21	31.59	0.984	16.79	141
7	10.732	19.89	38.70	0.992	23.25	150
8	10.581	22.03	33.64	0.994	25.35	153
0	10.587	18.32	22.19	0.992	24.43	150
1	10.747	17.88	27.21	0.990	22.02	147
2	10.942	16.17	34.62	0.988	18.62	141
3	11.014	15.30	26.55	0.988	16.78	138
4	10.749	17.65	28.61	0.988	28.65	144
5	10.782	18.61	35.38	0.988	26.10	147

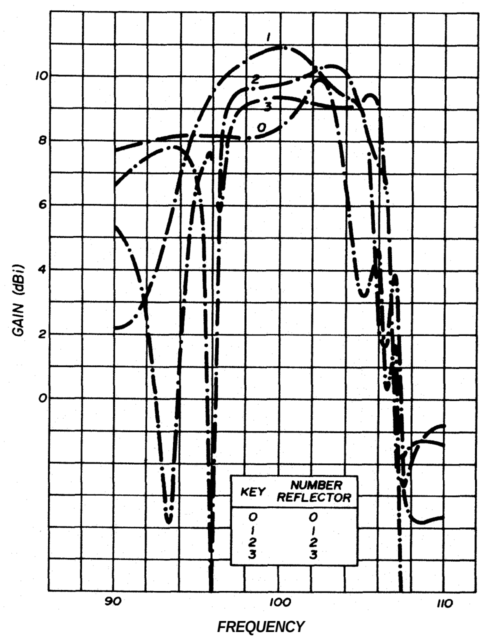


fig. 8. Gain of a 6-element Yagi beam vs. number of reflectors, overall number of elements held constant.

table 17. Performance vs frequency characteristics of a 6-element Yagi beam with a boom length of 0.75 λ_o, element positions optimized for maximum F/B ratio (at $F = 0.990$).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.960	9.482	8.066	26.753	–50.678
0.962	9.631	8.848	26.922	–47.892
0.964	9.762	9.661	27.057	–45.159
0.966	9.877	10.512	27.150	–42.467
0.968	9.980	11.410	27.195	–39.803
0.970	10.072	12.366	27.189	–37.154
0.972	10.154	13.397	27.133	–34.507
0.974	10.229	14.523	27.029	–31.849
0.976	10.298	15.773	26.881	–29.169
0.978	10.361	17.191	26.695	–26.458
0.980	10.420	18.843	26.476	–23.705
0.982	10.476	20.839	26.235	–20.903
0.984	10.528	23.386	25.980	–18.044
0.986	10.576	26.947	25.720	–15.124
0.988	10.621	33.001	25.467	–12.137
0.990	10.663	98.800	25.230	–9.078
0.992	10.701	33.035	25.021	–5.946
0.994	10.734	27.029	24.852	–2.737
0.996	10.762	23.517	24.735	0.551
0.998	10.785	21.024	24.685	3.920
1.000	10.801	19.090	24.716	7.372
1.002	10.810	17.505	24.845	10.919
1.004	10.810	16.165	25.094	14.554
1.006	10.800	15.008	25.486	18.276
1.008	10.781	13.990	26.049	22.087
1.010	10.750	13.086	26.817	25.985
1.012	10.707	12.277	27.833	29.965
1.014	10.653	11.549	29.153	34.021
1.016	10.587	10.893	30.847	38.136
1.018	10.509	10.303	33.009	42.280
1.020	10.421	9.775	35.761	46.397
1.022	10.324	9.307	39.263	50.385
1.024	10.219	8.900	43.721	54.062
1.026	10.109	8.555	49.385	57.097
1.028	9.996	8.276	56.488	58.895
1.030	9.884	8.069	65.091	58.439
1.032	9.776	7.943	74.616	54.170
1.034	9.675	7.912	82.965	44.366
1.036	9.585	7.994	85.869	28.942
1.038	9.507	8.216	79.043	12.212
1.040	9.442	8.618	63.459	1.471

fective boom length is shortened and we therefore should expect the gain to fall appreciably. **Fig.** 9 shows the F/B ratio for these same four cases. Clearly the standard Yagi-Uda antenna (curve 1) is superior to the zero reflector case (curve 0). In the two-reflector case (curve 2) the peak of maximum F/B (corresponding to the $K = 2$ null) has moved significantly higher in frequency. We have already learned that this occurs when the effective boom length is reduced (in this case by the relatively ineffective first reflector). This effect is exaggerated in the three-reflector case (curve 3) where the effective boom is still shorter due to the first two relatively ineffective reflectors.

Thus we now see that there is a very good reason why a Yagi-Uda should contain one and only one reflector in the linear boom array; one is definitely needed to improve the gain and F/B. More than one reflector reduces the effective boom length and therefore gain; also, because of the relatively small currents induced in the extra reflectors, they do very little to the basic Yagi-Uda F/B ratio potential.

missing parasites

A common observation among Amateurs who have had large Yagi-Uda antennas in operation over a period of time is that when a parasitic element is broken or even entirely missing the Yagi continues to perform surprisingly well. We may now examine quantitatively just what occurs; for comparison I shall use the same 6-element simplistic Yagi-Uda design of **fig. 1.**

When a parasite is missing, the individual element currents all readjust to new values; such a readjustment changes the effective boom illumination function and therefore must cause a change in Yagi-Uda antenna performance. Starting with the standard 6-

table 18. initial performance characteristics of the 6-element Yagi discussed in the text (boom length $= 0.78\lambda_o$).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.198	14.828	23.551	–35.790
0.972	10.292	16.185	23.337	–32.860
0.974	10.379	17.740	23.107	–29.906
0.976	10.460	19.570	22.866	–26.920
0.978	10.535	21.802	22.618	–23.898
0.980	10.606	24.640	22.369	–20.835
0.982	10.671	28.327	22.126	–17.725
0.984	10.732	31.831	21.895	–14.566
0.986	10.787	30.198	21.685	–11.353
0.988	10.837	26.307	21.501	–8.085
0.990	10.881	23.196	21.354	–4.758
0.992	10.918	20.814	21.252	–1.371
0.994	10.947	18.917	21.204	2.078
0.996	10.969	17.352	21.221	5.591
0.998	10.981	16.024	21.314	9.170
1.000	10.983	14.873	21.498	12.815
1.002	10.975	13.860	21.786	16.528
1.004	10.955	12.959	22.197	20.309
1.006	10.924	12.149	22.751	24.159
1.008	10.881	11.419	23.473	28.076
1.010	10.825	10.756	24.395	32.058
1.012	10.757	10.156	25.553	36.100
1.014	10.679	9.612	26.997	40.193
1.016	10.590	9.121	28.787	44.320
1.018	10.492	8.681	31.003	48.451
1.020	10.387	8.293	33.747	52.534
1.022	10.278	7.956	37.152	56.479
1.024	10.167	7.674	41.383	60.131
1.026	10.056	7.449	46.637	63.216
1.028	9.949	7.288	53.114	65.264
1.030	9.850	7.200	60.879	65.480

table 19. Performance characteristics of the intermediate design 6-element Yagi described in the text; maximum F/B at $F = 1.0$.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.077	7.614	24.286	– 40.497
0.972	9.271	8.371	24.167	– 37.504
0.974	9.447	9.162	24.048	– 34.535
0.976	9.609	9.994	23.925	– 31.583
0.978	9.757	10.874	23.794	– 28.645
0.980	9.892	11.812	23.652	–25.715
0.982	10.016	12.821	23.498	– 22.786
0.984	10.130	13.916	23.332	– i9.852
0.986	10.236	15.120	23.154	– 16.907
0.988	10.333	16.461	22.966	– 13.946
0.990	10.424	17.978	22.772	– 10.962
0.992	10.509	19.718	22.574	– 7.950
0.994	10.587	21.730	22.377	– 4.906
0.996	10.660	24.001	22.187	– 1.825
0.998	10.727	26.186	22.007	1.297
1.000	10.789	27.120	21.846	4.464
1.002	10.845	25.947	21.709	7.677
1.004	10.894	23.789	21.603	10.941
1.006	10.938	21.658	21.537	14.257
1.008	10.974	19.805	21.518	17.627
1.010	11.002	18.221	21.556	21.054
1.012	11.022	16.859	21.662	24.539
1.014	11.033	15.672	21.846	28.083
1.016	11.034	14.627	22.123	31.687
1.018	11.025	13.695	22.506	35.352
1.020	11.005	12.860	23.015	39.077
1.022	10.974	12.106	23.669	42.863
1.024	10.931	11.422	24.493	46.706
1.026	10.877	10.800	25.518	50.604
1.028	10.813	10.234	26.784	54.552
1.030	10.738	9.721	28.337	58.533

element simplistic design, I have made calculations of performance when one parasite is missing. Frequency swept plots of gain and F/B ratio are shown in **figs. 10** and **11;** the individual curves are keyed to the iegend in the diagram. Note that there is still significant gain displayed for any of the cases.

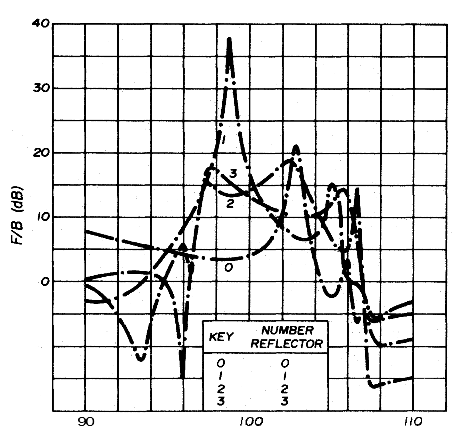


fig. 9. Front-to-back ratio vs. the number of reflector elements for a 6-element Yagi beam.

The greatest loss in performance occurs when the reflector, R, is missing; this, of course, is analogous to the previously discussed zero reflector case but now with a shorter (residual) effective boom. The most surprising aspect of **fig. 10** is the small but real *increase* in gain occasioned by the ioss of $D3$. This can only be understood if the readjustment element currents constitute an effective boom illumination function slightly *longer* than that for the fully populated beam; in this event we would expect the frequency for maximum F/B to be *lower* than that for the standard case. **Fig. 11** shows this to be true. For all other cases of missing parasites the frequency of maximum F/B is *increased,* indicating a *shortened* effective boom length and hence lowered gain. The lowered gain is verified in **fig. 10.**

Thus a missing parasite is not always disastrous. However, if you look at the performance at the frequency of best F/B, the original fully populated Yagi-Uda is best.

table 20. Performance of the 6-element Yagi where the positions of directors $D1$ and $D3$ have been varied to "tune up" the F/B ratio.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.064	7.989	27.181	– 43.099
0.972	9.235	8.747	27.090	– 40.279
0.974	9.390	9.541	26.981	– 37.483
0.976	9.532	10.378	26.850	– 34.704
0.978	9.662	11.266	26.694	– 31.935
0.980	9.781	12.217	26.511	–29.167
0.982	9.890	13.246	26.302	– 26.391
0.984	9.992	14.373	26.069	– 23.601
0.986	10.087	15.629	25.813	– 20.788
0.988	10.177	17.054	25.539	– 17.947
0.990	10.261	18.715	25.252	– 15.070
0.992	10.342	20.723	24.957	– 12.153
0.994	10.418	23.284	24.659	– 9.190
0.996	10.491	26.860	24.366	–6.178
0.998	10.561	32.929	24.083	–3.112
1.000	10.627	119.848	23.819	0.010
1.002	10.690	33.007	23.580	3.192
1.004	10.749	27.018	23.375	6.437
1.006	10.804	23.524	23.212	9.746
1.008	10.853	21.049	23.100	13.123
1.010	10.898	19.133	23.050	16.570
1.012	10.935	17.569	23.072	20.089
1.014	10.966	16.249	23.181	23.684
1.016	10.989	15.109	23.386	27.358
1.018	11.003	14.107	23.707	31.113
1.020	11.008	13.214	24.166	34.952
1.022	11.002	12.412	24.787	38.876
1.024	10.986	11.686	25.602	42.889
1.026	10.958	11.026	26.645	46.990
1.028	10.919	10.425	27.969	51.180
1.030	10.868	9.876	29.634	55.447

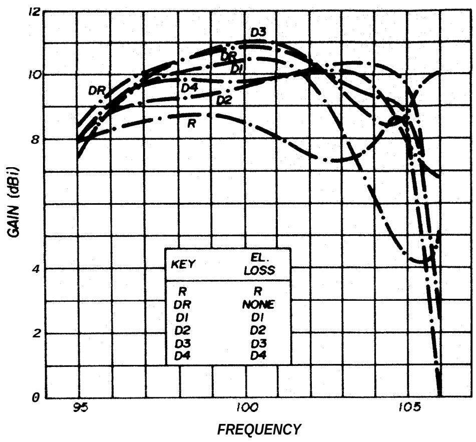


fig. 10. Forward gain of a 6-element Yagi showing performance when one element is missing. F/B under similar conditions is plotted in fig. 11.

It is now apparent that a Yagi-Uda antenna really "wants to work." Even major changes, such as a missing inner director, due to automatically readjusted element currents, works surprisingly well. It is now perfectly obvious why the Yagi-Uda antenna is so popular: it will provide ***reasonable*** performance no matter how it is constructed. It will provide top performance, especially in the F/B ratio, only if carefully made in accordance with the design rules presented in this article.

summary

In this article I have explored the effects of departures from the simplistic design previously given. The results show:

1. Director length taper schedules have no apparent beneficial effect on gain or F/B for boom lengths smaller than one wavelength. The important design parameter is the ***average*** director length – not the taper schedule.

2. Element placement schedules on the boom also have a marginal effect on gain or F/B for boom lengths less than one wavelength.

3. The simplistic design is as good as any design for boom lengths less than one wavelength.

4. A Yagi-Uda linear array on a given boom is best when it involves one and only one reflector element.

5. The F/B ratio at a given design frequency can, in principle, be increased without limit by iterative design procedure.

6. Very high values of F/B will be available only over very narrow bandwidths.

7. The Yagi-Uda antenna is basically very tolerant of major faults. Even missing parasitic elements cause surprisingly little deterioration in gain.

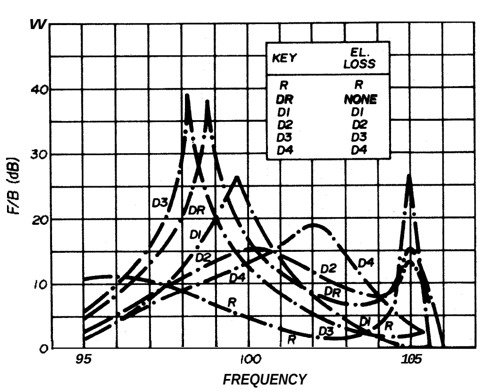


fig. 11. Front-to-back ratio of a 6-element Yagi showing the effect of a missing element. Forward gain under similar conditions is shown in fig. 10.

table 21. Performance of the 6-element Yagi where the positions of the driven element (DR) and director ($D3$) have been optimized through computer iteration.

frequency (F)	gain (dBi)	FIB ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.432	9.225	31.638	– 40.673
0.972	9.554	9.938	31.476	– 38.105
0.974	9.667	10.688	31.276	– 35.547
0.976	9.770	11.483	31.037	– 32.990
0.978	9.866	12.331	30.758	– 30.427
0.980	9.955	13.243	30.441	– 27.848
0.982	10.039	14.234	30.089	– 25.246
0.984	10.118	15.325	29.704	– 22.613
0.986	10.193	16.545	29.294	– 19.943
0.988	10.265	17.936	28.862	– 17.229
0.990	10.334	19.565	28.414	– 14.466
0.992	10.401	21.540	27.955	– 11.651
0.994	10.465	24.069	27.493	– 8.779
0.996	10.528	27.615	27.035	– 5.846
0.998	10.588	33.654	26.587	– 2.852
1.000	10.646	150.334	26.157	0.208
1.002	10.701	33.673	25.751	3.334
1.004	10.754	27.655	25.379	6.528
1.006	10.803	24.132	25.048	9.792
1.008	10.848	21.629	24.766	13.126
1.010	10.888	19.684	24.543	16.533
1.012	10.922	18.092	24.388	20.014
1.014	10.950	16.744	24.312	23.569
1.016	10.970	15.574	24.327	27.201
1.018	10.982	14.542	24.448	30.910
1.020	10.984	13.620	24.690	34.698
1.022	10.976	12.788	25.072	38.566
1.024	10.957	12.033	25.618	42.515
1.026	10.927	11.342	26.355	46.545
1.028	10.885	10.710	27.317	50.658
1.030	10.831	10.131	28.544	54.849

references

1. J.L. Lawson, W2PV, "Yagi Antenna Design: Performance of Multi-Element Simplistic Beams," ***hamradio,*** May, 1980, page 18.
2. J.L. Lawson, W2PV, "Yagi Antenna Design: Gain and Front-to-Back Ratio of Multi-Element Beams," ***ham radio,*** June. 1980, page 33.
3. P. Viezbicke, ***Yagi Antenna Design,*** NBS Technical Note 688 U.S. Department of Commerce, Washington, D.C., December, 1976.

ham radio

checking transmission lines with time-domain reflectometry

Most hams **are** familiar with the frequency-domain reflectometer, commonly known as the SWR bridge. In its various forms, this device can report the status of a transmission line under operating conditions at a single frequency (that of the transmitter – one frequency at a time). It shows the reflection coefficient, or SWR, depending on the scales employed. But if something is amiss (high SWR), the operator can't tell either the exact nature of the problem or its location. Both shorted and open lines will give the same reading regardless of length. Time-consuming tests may be required to localize the fault to the antenna, the transmission line itself, or the connectors. For as simple a test as continuity/absence of shorts, climbing the tower (in winter, yet) may even be required to disconnect the antenna and gain access to the distant terminals of the line.

An alternative approach, time-domain reflectometry, permits all measurements to be made in armchair comfort at the station end of the line. It can reveal the presence of open or short circuits or excessive resistance in the line. Additionally, the location of the problem can be determined, often within a foot or two, without going outside. The idea is not new but is not widely published in the Amateur literature.

time domain reflectometry

Here's the principle. A step of voltage is applied to the end of the line by a square-wave generator or pulse generator. The pulse has a very fast rise time. Iritially the generator sees only the characteristic impedance, Z_0, of the line, which determines the current according to Ohm's law.

When the pulse reaches a discontinuity in the line (such as a break, high resistance, short circuit, or the end), it is reflected, or absorbed, or a combination of the two. The amount and phase of reflection are determined by the impedance of the discontinuity. Any reflected current travels back to the generator where it combines with the outgoing current to produce a resultant. The generator, if mismatched, will cause a second reflection, and so on, until the pulse dies away due to line attenuation.

Now if we connect an oscilloscope across the generator terminals (in the station) we'll see (if the sweep rate is right) the initial voltage step (a in fig. **1A**), the resultant voltage caused by the sum of reflected and incident voltages (b), and so on for each stage in the reflection process. We'll see these reflections in real time as they occur (hence the name of the technique).

The time required for the first reflected pulse to return (t_1 in the figure) is twice the travel time for signals in the line. Thus:

$$t_1 = \frac{(2\ell_1/V)}{C} \qquad (1)$$

where ℓ_1 is the line length, V is the velocity factor,[*] and C is the speed of light, $3 \cdot 10^8$ meters/sec.). Of course if the line is resistively terminated in Z_0, no reflection occurs, and the generator waveform will be unmodified by returning pulses (fig. **1G**).

Multiple reflections can be seen under favorable circumstances, such as when two types of line are connected in series, or a lossy connector is used. The nature of a resonant antenna can even be determined when the resonant frequency is within the oscilloscope bandpass.

*The apparent velocity factors seen by this technique seem to be lower than those published. Thus foam dielectric cables typically showed a V = 0.6-0.7 vs the 0 8 found in the handbooks. This may be partly due to inaccuracy in the scope time-base generator, or partly a real effect – perhaps V is a function of frequency.

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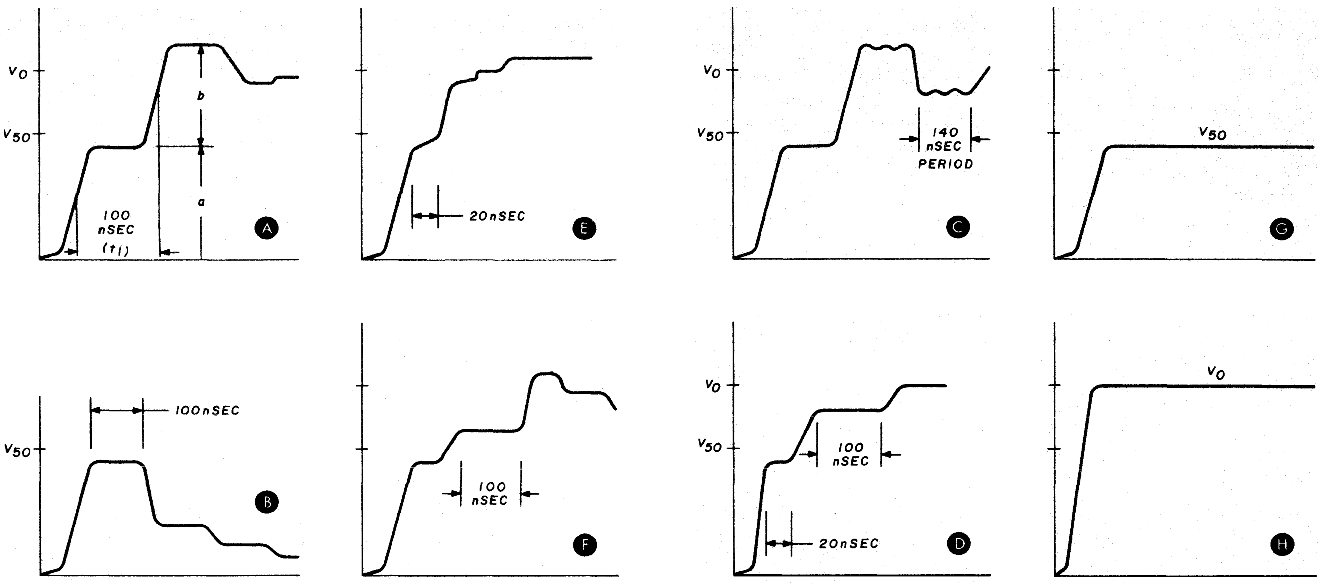

fig. 1. Time-domain reflectograms using the test setup in *fig.* 2 for various line lengths and termination impedances (see *table 1*). A through C show the effect of termination on a single length of line. D through E show the effect of a series resistance at an intermediate point. F shows what happens when two lines of different characteristic impedance are connected in series. G shows a properly terminated line of any length. H shows the effect of no load (open-circuit generator output).

instrumentation

The apparatus for a practical setup is shown in **fig. 2.** To get sharp reflections the scope must have a wide bandwidth and the pulse or square-wave generator must have a fast rise time. My 15-MHz scope gives quite good results with lines of about 10 feet (3 meters) or more in length. TTL logic oscillators are quite good as generators if buffered or padded to 50 ohms. The scope calibrator can sometimes be used. Repetition rate (square-wave frequency) is slow enough to let the reflections die away after each pulse, but fast enough to give a good bright trace on the scope (typically 100 kHz-1 MHz). When testing real antennas, the voltage level should be kept to a minimum to avoid QRM. The minimum will be determined by the scope sensitivity.

A transformer is needed when using lines other than 50 ohms. For 300-ohm line (TV twinlead), a simple transformer on a FT-82-43 toroid worked well (primary, 50 ohms, 28 turns, secondary, 300 ohms, 43 turns). This transformer passes the frequencies involved (100 kHz-15 MHz) quite well, appearing transparent to the pulses. However, other baluns, matching devices, or transformers are not usually designed for this frequency range, so they will usually appear to be near short or open circuits, depending on their dc characteristics.

test patterns

A number of typical patterns are shown in **fig. 1** using the test setup of **fig. 2. Figs. 1A-C** show the effect of termination on a single length of line. Note

table 1. Data for determining line length and resistance parameters for responses shown in fig. 1.

figure	$_1$ length feet	(meters)	$_2$ length feet	(meters)	Z_o' (ohms)	R_{series} (ohms)	Z_1 (ohms)	Z_2 (ohms)
1A	34.0	(10.5)	0		—	—	∞	—
1B	34.0	(10.5)	0		—	—	0	—
1C	34.0	(10.5)	0		—	—	(note 1)	—
1D	6.6	(2.0)	34	(10.5)	50	100	∞	∞
1E	6.6	(2.0)	34	(10.5)	50	100	∞	50
1F	6.6	(2.0)	34	(10.5)	75	0	∞	∞
1G	(note 2)		0		—	0	50	—
1H	0		0		—	—	m	—

Notes:

1. Termination was a 10-25-20-meter trap dipole. Note superposition of 14 and 21 MHz oscillations giving − 140 ns period (*7 MHz* = 27-74)
2. Any length. 2m and 10.5m were tried, using a dummy load.

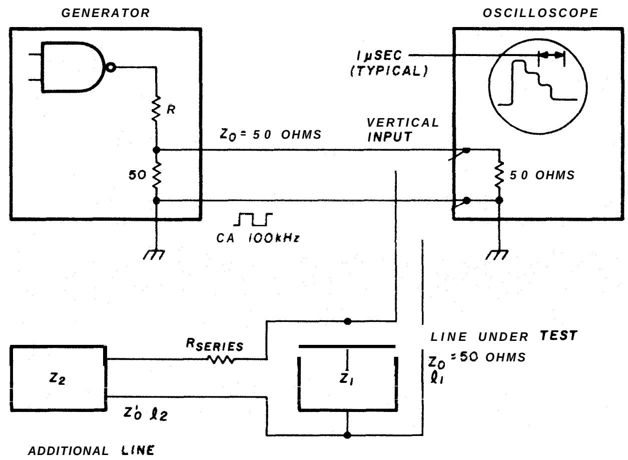

fig. 2. Test setup for line measurements. Oscilloscope must have a wide bandwidth and generator output must have a fast rise time. A 15-MHz scope gives good results with lines of moderate length (10 feet or 3 meters). R is adjustable to vary the generator output level. Notation corresponds to *table 1*.

that the trap dipole (no balun) is nearly an open circuit. In all three cases, the same length is obtained, within the accuracy of the scope time base. **Table 1** provides line lengths and impedance terminations for the time-domain reflectograms in **fig. 1.**

Figs. 1D-E show the effect of a series resistance at an intermediate point (such as a corroded connector). Note that in E, the properly terminated line looks like a pure resistance. Thus the effective length is ℓ_1 and the effective termination is **150** ohms.

Fig. 1F shows what happens when two lines of differing Z_0 are connected in series. The impedance bump shows clearly. And finally, **fig. 1G** shows a properly terminated line of any length. **Fig. 1H** is included to show that the voltages in the other cases are reduced from the generator open-circuit output, since the 50 ohms of the transmission line/load are in parallel with the generator output impedance.

closing comments

Since the technique is based on a step function, it covers the bandwidth from dc to ***1/(rise time of scope or generator)***. Thus it's not suited for critical vhf applications except to show the location of a gross defect, such as a short circuit in the line. Neither will it tell much about the steady-state characteristics of an antenna at a fixed frequency. For this you need an impedance bridge or SWR meter. But when something goes wrong, it's sure nice to know exactly where — and time-domain reflectometry gives the answer to that.

bibliography

Allen, David M., "A Practical Experimenter's Approach to Time-Domain Reflectometry," *ham radio,* May, 1971, pages 22-27.

ham radio

More Details? CHECK — OFF Page 94

open quad antenna

An interesting approach to quad antenna design using phased radiators

In this article I describe a novel approach to the classic two-element quad antenna. I call it the "open quad." The designs described are for the Amateur 144- and 432-MHz bands, but they can be scaled for lower-frequency bands.

design approach

In this design I've used the regular driven element and reflector of the quad and have added director elements in the form of V and inverted V elements in a quasi-Yagi configuration. **Fig. 1** shows the general idea, and **fig. 2** shows the physical arrangement for a 10-element array for the 432-MHz band. Note that the parasitic elements are aligned with the two points of maximum current. Note also that the ends of the directors have been bent horizontally for a short distance. (This modification might be more useful for a larger antenna.)

advantages

The advantages of the open quad over the classic quad or Yagi derive not from any revolutionary concept, but rather from an attempt to combine the advantages of both designs:

1. Easier adjustment of the quad reflector for better front-to-back ratio.

2. Easier excitation of the driven element because of relatively higher impedance at the feed point.

3. Absence of feeders allows the exact phase of excitation, for top and bottom parasitic element.

4. High Q of director elements will increase gain.

5. Double V configuration makes for a collinear effect, which lowers the vertical radiation angle.

6. Possibility of eventually inserting a smaller antenna for operation on more than one band.

open quad for 432 MHz

In this design I attempted to duplicate the 13-element two-meter antenna described in reference 1. It is reproduced here in proportional scale. (I lacked a boom long enough for 13 elements.) Its characteristics include the following points:

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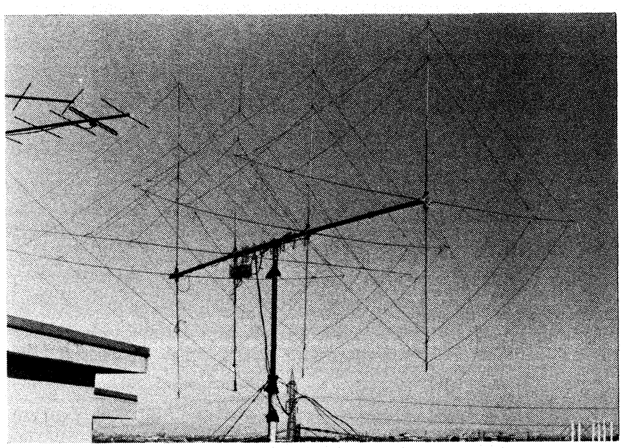


Open quad for the high-frequency bands based on the principles discussed in the text.

1. Forward gain: 17 dBd
2. Front-to-back ratio: 28 dB
3. Front-to-side ratio: 46 dB
4. Feedpoint impedance: 100 ohms

construction dimensions

I made the element lengths as follows for 432 MHz:

radiator: 27.9 inches (70.8 cm)
reflector: 29.7 inches (75.4 cm)
directors: 13.4 inches (34.0 cm) per element (two required for each director)

Element spacing for the 432-MHz quad was as follows:

reflector-to-radiator:	3.7 inches (9.5 cm)
radiator-to-director 1:	2 inches (5.5 cm)
director 1 to 2:	2.3 inches (6 cm)
director 2 to 3:	2.3 inches (6 cm)
director 3 to 4:	5.3 inches (13.5 cm)

The remaining directors were spaced 10 inches (26 cm).

The boom is about 6.4 feet (2 meters) long and made of fiberglass; the quad is fed with a 50-ohm coax cable and a 75-ohm quarter-wave matching section.

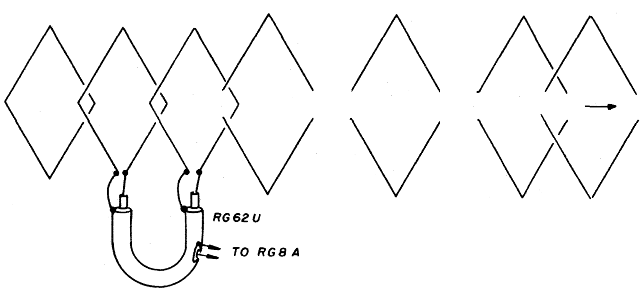


fig. 1. Design approach for the open quad antenna using two phased driven elements, with directors in the form of Vs and inverted Vs in a quasi-Yagi configuration.

open quad for 144 MHz

In this design I tried to achieve the desirable effect of two phased radiators spaced one-quarter wavelength apart. It's a well-known fact that maximum energy transfer between two antennas is a function of their spacing. The exact spacing follows a sequence of minimum and maximum current occurring at one-quarter wavelength between the two antennas. Tests have shown that an open quad with two in-line radiators, properly phased, will produce a signal not less than 0.7 times the maximum obtainable with the radiators spaced at the ideal distance.

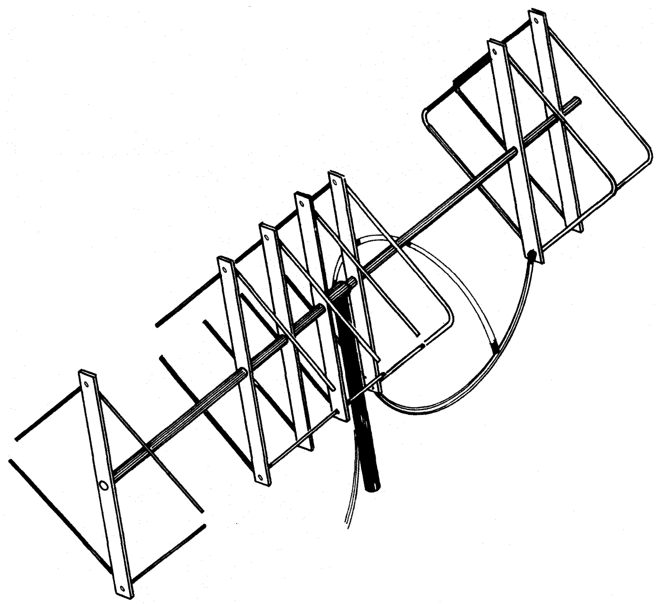


fig. 2. Physical arrangement of the open quad designed for the 432-MHz band.

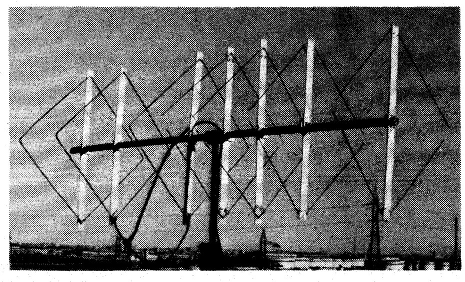
Characteristics of the 144-MHz open quad are:

Forward gain:	15 dB/d
Front-to-back ratio:	24 dB
Front-to-side ratio:	40 dB
Input impedance:	50 ohms

construction

I made the elements from aluminum tubing 0.2 inch (5 mm) in diameter. The boom was of fiberglass, as for the 432-MHz antenna. (The fiberglass boom was necessary to arrange one or more antennas across the boom of my high-frequency antenna.)

The matching section for the two radiators used lengths of 93-ohm coax cable (RG-62/U). The phasing line to the first (rearmost) radiator element was 6 inches (15 cm): that to the second radiator was 22



Author's seven-element open quad for the 2-meter band.

inches (56 cm). This matching section provided the proper phase shift between each radiator element for the energy transfer described above. The junction of these lines was soldered to a 50-ohm feedline.

Element lengths for 144 MHz were:

reflector:	6.98 feet (2.13 meters)
radiators:	7.40 feet (2.25 meters)
directors:	3.35 feet (1.02 meters) each part

Element spacing for the 144-MHz antenna was:

reflector to radiator 1:	11 inches (27.9 cm)
radiator 1 to radiator 2:	20 inches (51 cm)
radiator 2 to director 1:	8.2 inches (21 cm)
director 1 to director 2:	8.2 inches (21 cm)
director 3 to director 4:	15.75 inches (40 cm)

I tested a new, improved version using ten elements on a 13-foot 14-meter) boom with results better than those of the big commercial antennas used in Europe.

open quad for 10, 15, and 20 meters

It should be an easy matter to add one or two "open" directors to an existing high-frequency quad. I recommend the following spacing:

reflector to radiator:	0.14λ
radiator to director 1:	0.12λ

on-the-air tests

I tested the high-frequency open quad shown in the photo with a station 20 miles (32 km) away with the following results:

Direct line to station:	S9 + 10 dB
180 degrees from station:	S3
90 degrees from station:	signal audible but not readable

concluding remarks

The open quad can surely be improved, both from a mechanical and electrical standpoint, by someone with more sophisticated and precise instrumentation. I shall be glad to correspond with anyone wishing more information.

reference

1 William Orr, W6SAI, The *Radio* Handbook, 18th edition, Howard Sams, Indianapolis, 1969

ham radio

More Details? CHECK—OFF Page **94**

microwave-frequency converter for vhf counters

Theory and construction of a frequency scaler that can be used to convert 500-2500 MHz to frequencies acceptable to a 500-MHz counter

Measuring frequency accurately to 500 MHz is inexpensive and easy today. Several good 500-MHz counters are available for about $100.00. These are of little value to the microwave experimenter, however, whose world is just beginning at 500 MHz. For lower-frequency microwave measurements, up to 2500 MHz or so, a good counter will set you back $1000 or more.

This article describes a heterodyne frequency converter that will convert frequencies from 500-2500 MHz to below 500 MHz, enabling you to use your 500-MHz counter for microwave measurements. Depending on your resourcefulness at pleading with your local manufacturers' representatives to supply you with a few key parts, your parts cost will be anywhere from nothing to $100.00. Construction techniques should be followed closely unless you're familiar with uhf construction practices.

theory

The heart of this project is a double-balanced mixer, the Engelmann Microwave MLP-101 (see **fig. 6**). I recommend it because of its low cost — $35.00. Neglecting third- and higher-order products, a balanced mixer output spectrum contains two frequencies — $F_{RF} + R_{LO}$ and $F_{RF} - F_{LO}$ as in **fig. 1**.

For our purposes, $F_{RF} - F_{LO}$ will be the spectral component of interest since it translates an unknown radio frequency to a 0-500 MHz i-f by a known LO frequency: in this case 1000 or 2000 MHz. This makes your present counter direct reading in the lower two Amateur microwave bands, 1215 - 1300 MHz and 2300 - 2450 MHz, by simply mentally adding a 1 or a 2 in front of the displayed frequency, depending on the LO frequency you choose. To gain an overview of how the local-oscillator signal is generated, see **fig. 2.**

Local oscillator. The local oscillator string consists of a 1-MHz crystal reference frequency oscillator, a phase-locked-loop, X500 frequency multiplier stage, and two frequency-doubler stages. The 1-MHz reference oscillator is a conventional CMOS inverter using a parallel-resonant crystal. The loop phase detector is a 4046 CMOS IC using the sample-and-hold phase detector. All other stages are fairly conventional with a

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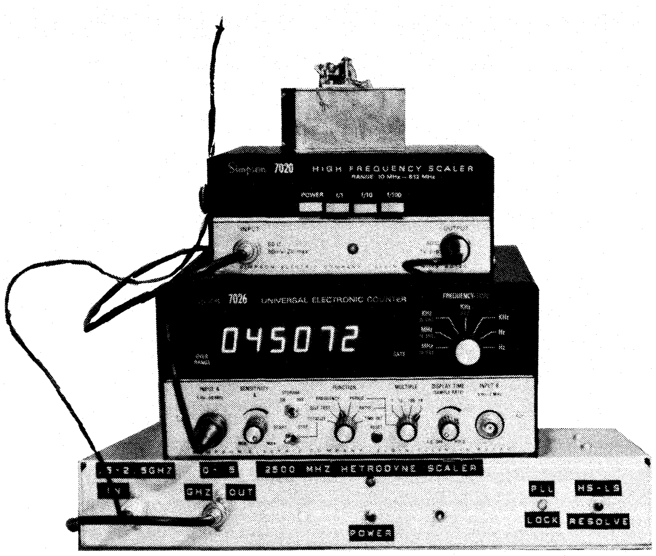


The heterodyne scaler in operation. Oscillator frequency measured 2450.72 MHz.

few exceptions. The loop filter is a two-pole design as opposed to the usual single-pole variety. The second pole resides in the "Spurrie filter" (see **fig. 2**). Its purpose is to attenuate unwanted phase-detector outputs ($N \times 1$ MHz), which would otherwise frequency modulate the VCO, producing sidebands at 500 ± 1, 2, 3 . . . MHz. As it stands, all such reference frequency spurs are at least 50 dB down from the 500-MHz fundamental.

Frequency multipliers. The frequency doublers may seem a bit unusual to some at first glance. This is because uhf transistor frequency multipliers don't operate in the same way as their high-frequency counterparts. High-frequency multipliers operate on the principle of reduced conduction angle (60°-120°) collector-current pulses containing large quantities of harmonics, which may be filtered by a high-Q tank circuit to yield the desired harmonic output. Alas, very few transistors can switch fast enough to provide low conduction angles at 1000 MHz. Rise storage and fall times preclude operation as a conduction-angle-based frequency multiplier above a few hundred MHz or so.

Fortunately, a thoughtful electron god has included other nonlinearities into transistors that make them useful frequency multipliers. The most important of these is the varactor effect in the collector-base (C-B) junction. (See **fig.** 3 for a general explanation of how vacactor multipliers work.)

Transistor-varactor-effect frequency multipliers must do three things simultaneously: 1) they must amplify the input frequency, 2) they must apply this amplified F_{IN} efficiently to their own C-B junction for frequency multiplication, and 3) they must extract and filter the desired harmonic from the C-B junction.

analysis

The circuit of **fig. 4** shows how to accomplish the objectives outlined above. To be a good F_{IN} amplifier, the transistor's conjugate input impedance must be matched; in this case to a 50-ohm input. From **Z**-parameter data, the transistor's input is $Z = 13 + j14$ at 1 GHz. This translates into a parallel-equivalent conductance of about 0.04-0.04B. Since Mho $= 1/ohm$, the transistor looks like a 25-ohm resistor in parallel with a 25-ohm inductor. My input-matching strategy is to cancel the 25-ohm inductance with a parallel 25-ohm capacitive reactance (LN2), leaving only the 25-ohm resistance. A quarter-wave impedance transformer, LN1, of $Z_0 = 35$ ohms converts 25 ohms resistive into 50 ohms resistive impedance.

$$Z_0 = \sqrt{Z_{IN}Z_{OUT}} = \sqrt{50 \times 25} = 35\,ohms \qquad \textbf{(1)}$$

The 25-ohm capacitive reactance was especially chosen, however. It consists of an eighth wavelength, at F_0IN, of $Z_0 = 25$-ohm line open circuited at the far end. At $2F_0IN$, the desired output frequency, it is a quarter-wave open-circuited line and therefore an effective $2F_0IN$ short circuit.

This line shorts the transistor base to ground at $2F_0IN$, meaning that $2F_0IN$ energy can flow from the transistor C-B junction to the $2F_0IN$ filter without dissipating in the resistive B-E junction. This $2F_0IN$ trap greatly improves frequency-doubler efficiency.

On the collector side of the frequency doubler, L1C1 form an "F_0IN idler." This tank circuit is roughly analogous to the sine-wave current source in **fig.** 3. Note that when properly adjusted, L1C1 are not series resonant at F_0IN. (Series resonance would short-circuit the F_0IN source, leaving no energy for the current source to drive F_0IN through the varactor to generate harmonics.) Instead, the series combination of C1 and C_{CB} (average value) and L1 are parallel resonant to provide a high circulating F_0IN current

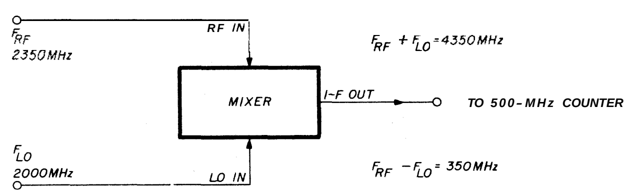
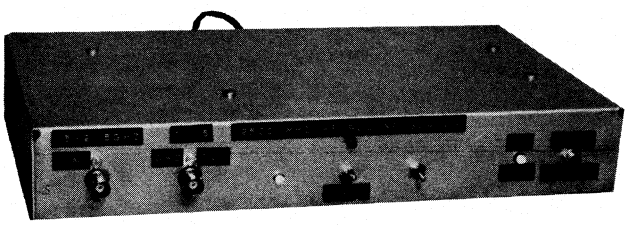


fig. 1. The heart of the heterodyne frequency scaler is a double-balanced mixer. The spectral component of interest is $F_{RF} - F_{LO}$, which makes your frequency counter direct reading in the lower two Amateur microwave bands, 1215-1300 MHz and 2300-2450 MHz.



Front panel of the heterodyne scaler.

through C_{EB}, where harmonic generation occurs.

LN3, in conjunction with C2, forms a half-wave-length line, high-Q output filter tuned to $2F_0IN$. The ratio of C2 to C3 allows for input and output matching.

The broadband amplifier (Q_{AMP} in **fig. 6**) makes up for the 7-dB loss in the mixer, giving the frequency converter as good or better sensitivity than the counter with which it's used.

The only part you may have difficulty obtaining inexpensively (or for free) is the 2500-MHz double-balanced mixer. The most inexpensive suitable unit I've found is the MLP101 (or MLF101) made by Engelmann Microwave; it cost me $35.00.

The I-MHz crystal should be of time-base quality, as its frequency is multiplied by 2000 on the converter high-MHz range. If your counter has a 1-MHz time base, you can omit the crystal.

The 0.032-inch (0.8-mm) Teflon/glass double-sided board specified for this project may be obtained from Oak Laminates. Cincinnati Millacron polyester glass board (Milliclad[R]) is a very good, inexpensive substitute for the TFE glass but is only available in 0.062-inch (1.6-mm) double-sided board. Its dielectric constant is a function of frequency, and some experimentation will be necessary to make it work, except as indicated in **fig. 5.**

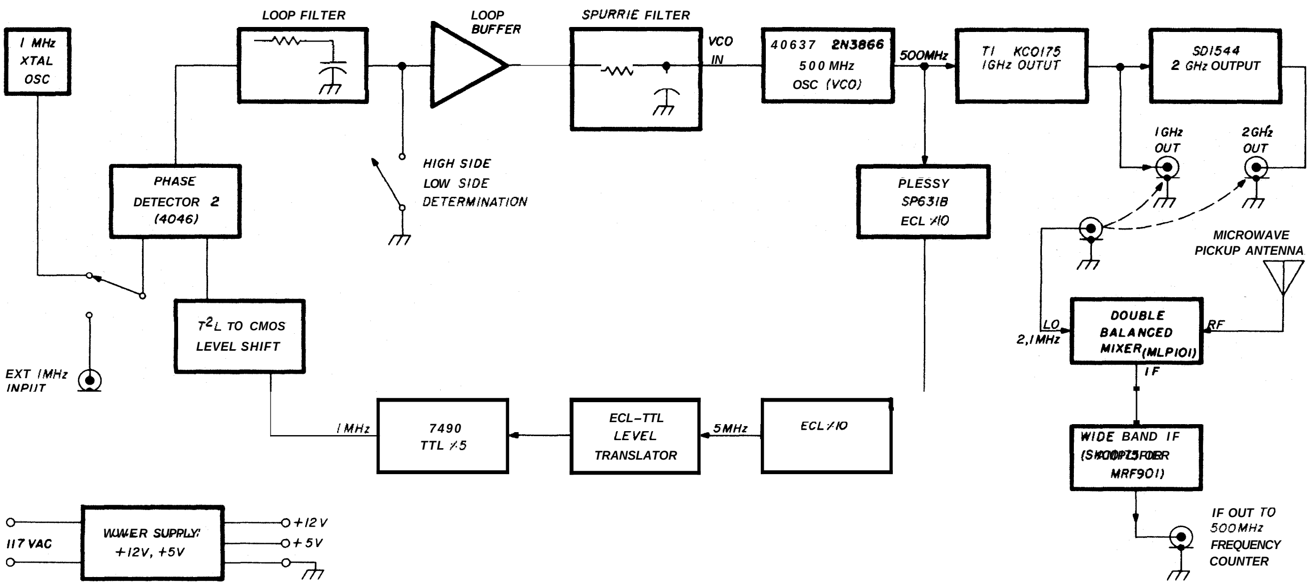
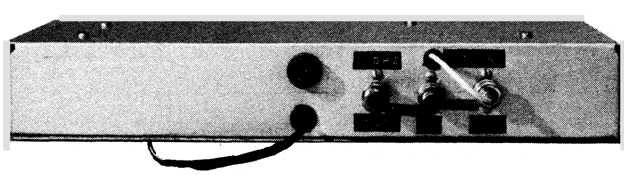


fig. 2. Heterodyne frequency scaler block diagram. Local-oscillator string consists of a 1-MHz crystal reference frequency oscillator, a phase-locked x 500 frequency multiplier, and two frequency doublers. The varacter effect in the frequency-multiplier transistors is used for efficient frequency multiplication in the Gigahertz region.

Amplifier gain is 10 dB ± 1.5 dB between 5-500 MHz. Try to obtain an SKC0175 for this stage, as any other device probably won't work too well without extensive circuit mods. The broadband amplifier can be omitted at some loss (about 10 dB) of scaler sensitivity.

obtaining ı

The only nonstandard parts used i the converter (**fig. 6**) are transistors QD1, QD2, the Plessy SP631B, and of course the MLP101 double-balanced mixer IC. Possible substitutes for the Texas Instruments SKC0175 (QD1) are the 2N5770 and the 2N3866. Possible substitutes for QD2, a Solid State Microwave SD1544, are the 2N5108 and the MRF8009. The Plessy SP631B may be replaced by any ECL 500-MHz divide-by-ten prescaler chip.



Rear panel of heterodyne scaler showing L.O. changing jacks.

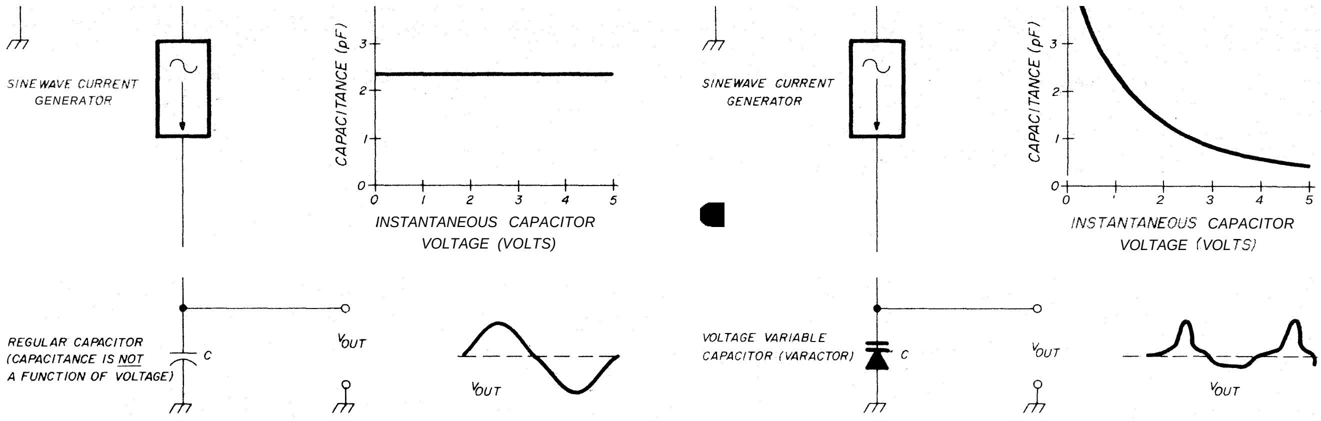


fig. 3. Comparison of a sine-wave current generator terminated by a regular capacitor, A, and one using a varactor, B. In A, V_{OUT} is an undistorted. phase-lagging sine wave (i.e., a cosine wave). In the varactor application, B, as V_{OUT}, increases in amplitude, C decreases, making V_{OUT} rise ever faster. V_{OUT} will then become badly distorted, creating harmonics that may be filtered out to yield the desired output signal.

construction

Construction of the 2500-MHz heterodyne scaler may be divided into two parts: 1) power supply, reference oscillator, phase detector, loop filters and buffer; and 2) the VCO, divider string, doublers, mixer, and broadband i-f amplifier.

Part 1 may be built using any construction practices you desire, as layout and grounding aren't critical. Part 2, however, should be built using plain unetched 0.064-inch (1.6-mm) or thicker copper-clad board (material uncritical) as a goundplane.

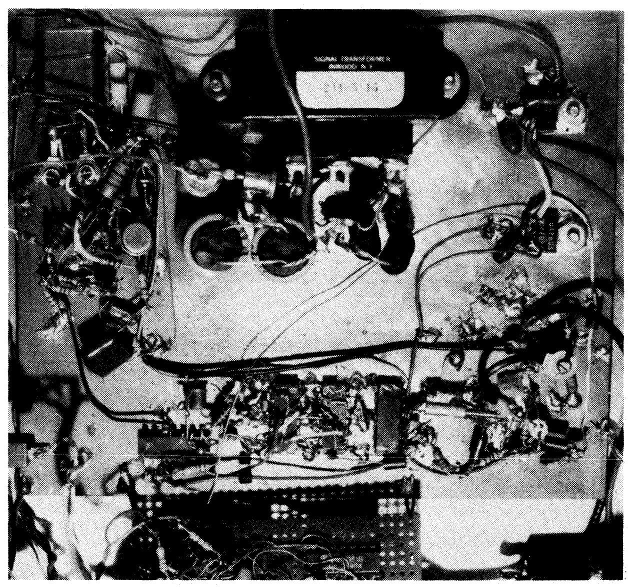
Mount the divider ECL and TTL ICs by turning them upside down and bending all their ground pins to meet the copper-clad board, then solder them in place. All ECL point-to-point wiring can then be made using very ***short*** lengths of solid hookup wire. By proper forethought as to layout, it's possible to eliminate any inter-IC lead length greater than 5/8 inch (16 mm). If for some reason you must make a longer run in the divider chain, use shielded 50-ohm cable. Run all bypass capacitors from their IC pins to the groundplane with lead length as short as possible. possible.

The same basic construction practices outlined above apply to the VCO and broadband amplifier, except that here you deal with transistors instead of ICs. The doublers, as they use stripline circuitry, deserve special mention. Stripline, as done in industry, is a photoetching process that leaves dual-sided Teflon copper-clad board with all its copper intact on one side (the groundplane) and etched strip-like conductors on the other side. The strip conductors work with the groundplane to form a transmission line whose impedance is a function of strip width, height above groundplane, and dielectric constant of the Teflon glass board.

For our purposes, however, we can use a different technique, which is much more inexpensive, easier, and alterable too. It involves cutting the desired lines out of Teflon PC board stock, and then cementing them to a far-less-expensive, unetched G10 or phenolic-base circuit board.

To duplicate this technique, obtain a 6 x 6 inch (152 x 152 mm) piece of single-sided, unetched, copper-clad board (material not critical). Use steel wool to polish the copper side until it shines. Then refer to **fig. 5** to obtain stripline dimensions for each line used.

Using your stock of Teflon-glass board, cut out lines to these exact dimensions. Polish both sides of



Internal view of scaler showing power supply and L.O. frequency synthesizer. This board outputs 500.000 MHz.

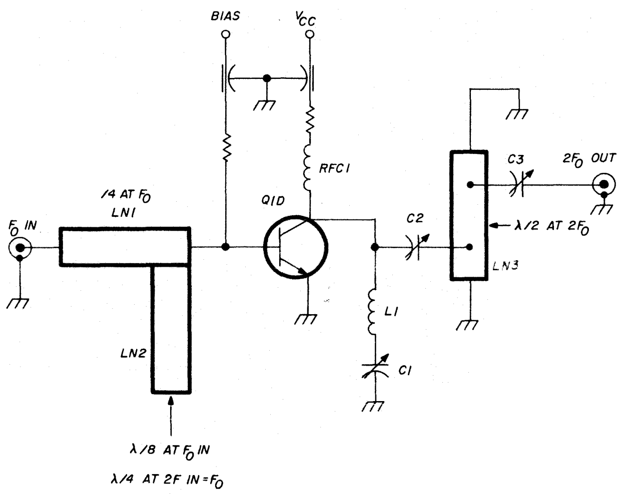


fig. 4. How to accomplish the objectives of a transistor-varactor-effect frequency multiplier. Stripline techniques are used for impedance transformation.

each line with steel wool. Using all doubler, mixer, and i-f amplifier parts, come up with a suitable layout for this assembly. When you're happy with the layout, drill holes for all feedthrough capacitors and transistors in the 6 x 6 inch (152 x 152 mm) circuit board chassis. Use ***Super Glue*** to mount all polished striplines to the ***copper side*** of this board to form instant, repairable striplines. Finally, mount both transistors and other parts with leads as short as possible. Mount the MLP101 mixer by soldering it to the circuit board in two or three spots, being very careful not to overheat the device.

A well-shielded box is recommended to house the heterodyne scaler, as it radiates on many harmonics of 1 MHz. It's a good idea to use feedthrough capacitors on the power-line cord to prevent radiation leakage.

tune up

Tune up begins by locating a volt-ohmmeter, oscilloscope *(bandwidth ≥ 5 MHz)*, a frequency counter, and a set of nonmetallic alignment tools. Make yourself comfortable and expect several hours of entertainment (or frustration).

1. Break the connection between the 4700-μF filter cap and the 7812 regulator. Insert a milliammeter here to measure total supply current. If you read 250-350 mA, all is probably well.

2. Connect a voltmeter to the +12 and +5 volt sources, just to make sure the regulators are wired correctly. The collector of Q_{D1} should read 7 volts ±1 volt, that of Q_{D2} **8 volts** ±1 volt, and that of Q_{AMP} about 5 volts ±1 volt. If these voltages are other than specified, adjust stage bias accordingly; that is, if the collector voltage is too low increase the bias resistor value and ***vice versa.***

3. Ready your frequency counter, and, with the reference frequency switch set to INTERNAL, measure the 1-MHz reference frequency at the 4046, pin 14. Set this frequency to 1.0000 MHz.

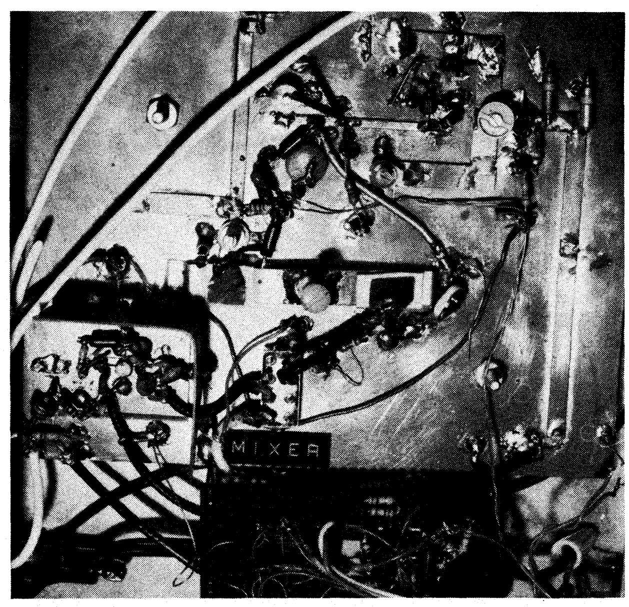
4. Loosely couple your counter to the VCO emitter through a 1-pF capacitor, after closing C_{TUNE} completely and disconnecting the 500-MHz OUT terminal. A frequency between 350 and 500 MHz should be observed here.

5. Observe the waveform at TP1. It should be a clean, jitter-free, TTL-level square wave of 700 kHz to 1 MHz. If not, adjust the 0.8-6 pF piston trimmer on the VCO emitter until the proper signal is obtained. You'll want 1-2 turns more capacitance (from the trimmer) than the minimum required to get a clean waveform at TP1.

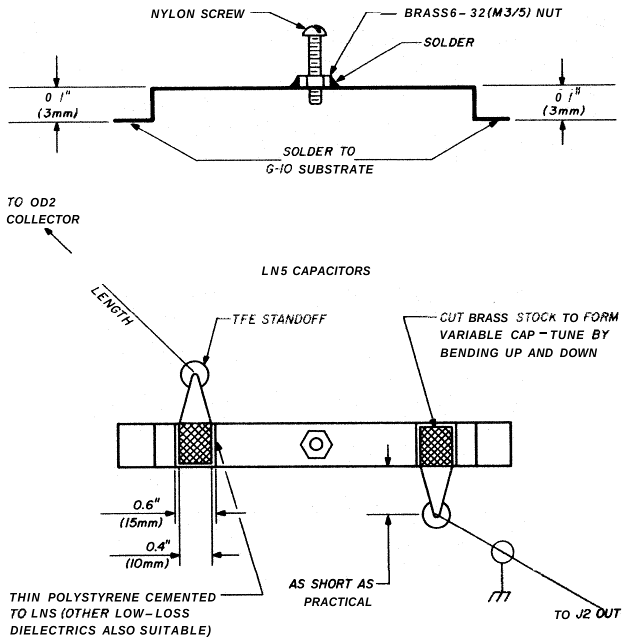
6. Slowly reduce the capacitance of C_{TUNE} until the loop locks. Lock is established when the OUT-OF-LOCK pilot goes out and the measured frequency of the VCO is exactly 500.0 MHz.

7. Connect a 47-ohm, 114-watt resistor, with ***short leads,*** between 500-MHz OUT and ground. The VCO will probably lose lock at 450 MHz or so.

8. Tune C_{TUNE} and C_{LOAD} for maximum power ***and*** a locked loop by measuring the dc voltage at TP2. You should measure at least 1 volt, but don't tune for more than 1.5 volts.



Internal view of scaler showing the two frequency multipliers and the broadband i-f amplifier.



LN1	3.9 x 0.2 x 0.032 inch (99 x 5 x 0.8 mm) glass Teflon
LN2	3 x 0.2 x 0.064 inch (76 x 6 x 1.6 mm) Cincinnati Millacron *Milliclad* or 3.5 x 0.2 x 0.032 inch (89 x 4 x 0.8 mm) glass Teflon
LN3	2 x 0.1 x 0.032 inch (52 x 3 x 0.8 mm) glass Teflon
LN4	0.975 x 0.2 x 0.032 inch (25 x 5.5 x 0.8 mm) glass Teflon
LN5	2.7 inches (70 mm) of 0.39-inch (10-mm) brass with hole drilled in exact center and a 6-32 (M 3/5) brass nut soldered squarely over hole, then bent as shown
L1	1.4 inches (35 mm) no. 12 (2.1-mml bare copper wire bent around a 0.5-inch (13-mm) coil form. A wooden dowel is suitable
L2	3 turns no. 20 (0.8-mm) enamelled wire wound on a 0.1-inch (3-mm) coil form, closely spaced to start with
L3	20 turns no. 30 (0.25-mm) enamelled wire wound around a 0.5-watt resistor lead. Form the coil to occupy 0.4 inch (9 mm)

fig. 5. Construction details for the microwave striplines and inductors used in the frequency scaler.

9. Disconnect the 47-ohm resistor and reconnect *LN1* to 500-MHz OUT.

10. Move the voltmeter to TP3 and terminate BNC connector $J1_{OUT}$ with about 100 feet (30.5 meters) of RG-58/U or RG-174/U cable (far-end termination is not critical).

11. Tune both doubler 1 variable caps until you obtain maximum otuput at TP3. You should obtain 1.5-2.5 volts. If not, stretch *L2* turns for maximum output.

12. Verify that the loop OUT-OF-LOCK pilot light is still out. If not, slightly adjust the VCO C_{TUNE} control until loop lock is re-established.

13. Connect the lossy coax load to $J2_{OUT}$. Jumper $J1_{IN}$ and $J2_{OUT}$ with a coax cable.

14. Adjust the 1-GHz idler and the nylon screw in *LN5* and its two flap caps **(fig. 5)** for maximum output at TP4. If the signal suddenly disappears, try a little more 1-GHz idler capacitance. Several iterations will be required to get the maximum voltage level at TP4. About 2 volts should be obtainable if a quality Schottky detector or point-contact mixer diode is used at TP4. if your output is a little low here, don't worry. Accurate voltage measurements using crude peak detectors are a joke at 2 GHz anyway!

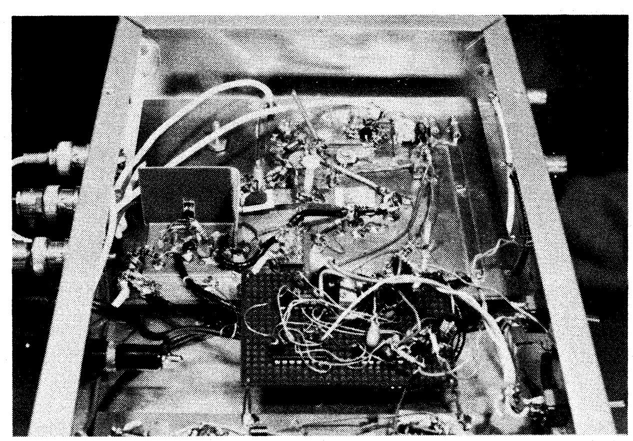
15. Remove the coax dummy load from J2 and insert the *LO* cable into J2. The reading at TP4 should be little different than with the dummy load.

further hints

If you build this unit, I assume you have some pet microwave project (or at least an oscillator) that you'd like to test and improve. Now is the time to do it. If you think your unit operates between 1.5 and 2.5 GHz, leave the heterodyne scaler set up the way it is. If you think your unit's output is between 500 and 1500 MHz, connect the LO_{IN} jack to $J1_{OUT}$. Connect your frequency counter to 0-500-MHz OUT. A suitable pickup antenna for this frequency is a half-wave coaxial dipole with a 6-inch (152-mm) overall length for the low band and a 2-112 to 3 inch (64 to 77 mm) length for the high band.

If you have any problems with scaler sensitivity — especially on the high band — and have followed my instructions, try one more thing.

Move the peak detector associated with TP4 to the actual *LO* terminal of the mixer. If there is much less voltage here than at the original location, some cable and connector work is in order. On the same subject,



In the foreground is a late scaler addition, an auto-ranging circuit. This allows the scaler to choose the correct L.O. frequency automatically or allow manual selection by a front panel switch.

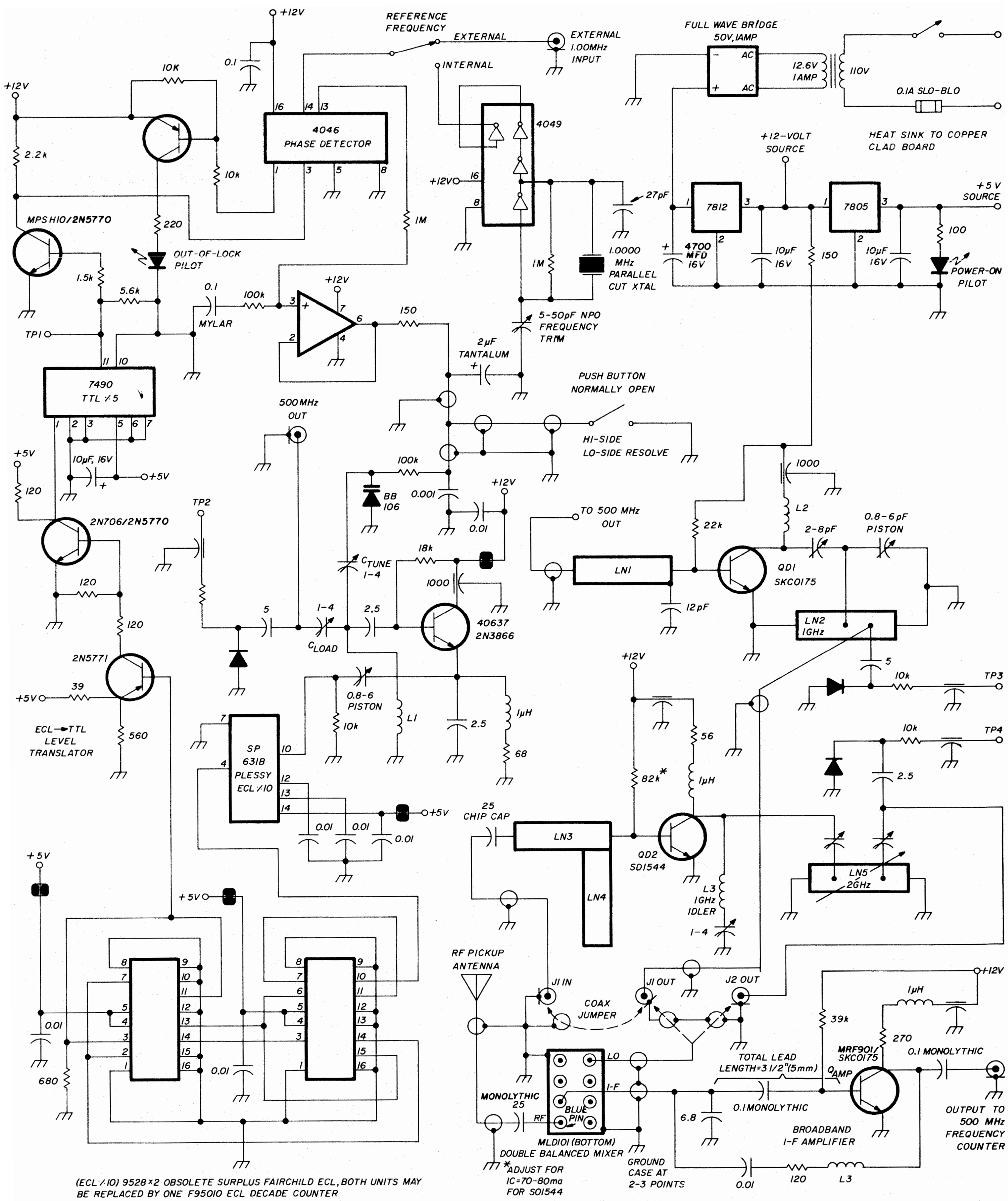


fig. 6. Schematic diagram of the 2500-MHz heterodyne frequency scaler.

if you're interested in only one band, forget about the back-panel jacks and handwire the connections.

operation

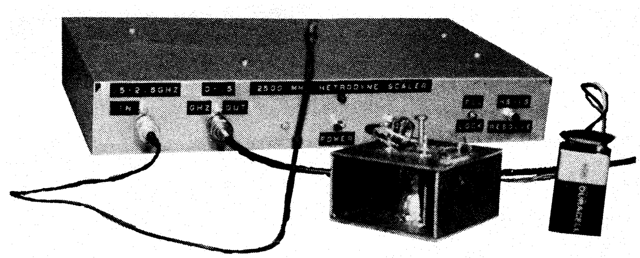
Operation of the heterodyne scaler is nearly as simple as using your counter with one or two exceptions. The first is the need to depress the HIGH SIDE-LOW SIDE RESOLVE switch when making a measurement. If, when this switch is depressed, the indicated frequency increases, your microwave signal is on the high side of the 1- or 2-GHz local-oscillator frequency, and your actual frequency is the *LO* frequency ***plus*** the counter displayed frequency. Conversely, if the measured frequency decreases when this switch is depressed, your microwave signal is on the low side of the *LO* frequency, and its actual frequency is the *LO* frequency in use ***minus*** the frequency your counter is measuring.

The second exception is overload protection. Your counter probably has fairly good overload anti-burnout circuitry; this scaler does not. The usual back-to-back hot-carrier diodes that provide protection at low frequencies are simply too reactive to do much good at 2.5 GHz. So be careful, and this scaler will serve you well. Don't do anything rash like trying to measure the frequency of your microwave oven by inserting the pickup antenna directly into the oven cavity!

Caution must also be used when trying to measure frequencies closer than 2 MHz away from the 1- or 2-GHz *LO* frequency. The double-balanced mixer as well as the broadband amplifier frequencies fall off very rapidly in this region; and with high input signal levels, i-f harmonics may well be stronger than the i-f fundamental. Besides, to measure closer than 10 MHz to the *LO*, the *dv/dt*-sensitive ECL prescaler in your counter will have to be bypassed. Some ECL devices become very confused with slowly rising and falling wavefronts.

other uses

The heterodyne scaler has other uses besides accurate frequency measurement in the lower microwave bands. Most notable among these is its use as a receiving converter for 1296 or 2304 MHz (or anywhere in between). This project was designed as a high-level mixer, so I made no attempt to characterize *LO* noise skirts or system noise figure. Because *LO* noise drops as you move farther away from the *LO* frequency, I would guess receiver noise figure would be best between 600 and 900, 1100 and 1400, 1650 and 1850, and 2150 and 2350 MHz. In a nutshell, my advice to anyone using this circuit as a receiving converter is to use a high gain antenna, a 3 dB (or better) noise figure preamp, mast-mounted if possible, and a feedline such as 314-inch (19-mm) hardline.



Close-up of scaler, scaler pickup antenna, and a microwave oscillator under test. Note polarization of osciliator.

Another possible use of the heterodyne scaler is a stable 2.000-GHz *LO* source for a transmitting converter. For this you need a 432-MHz transmitter, mixer, and a preamplifier-power amplifier chain.

If you have any problems, questions, or comments concerning the scaler, or live around Chicago and would like to attempt communications at or above 2300 MHz, please get in touch with me.

addresses of electronic parts manufacturers

Texas Instruments
Semiconductor Components Division
P.O. Box 5012
Dallas, Texas 75222

Engelman Microwave
Skyline Drive
Montville, New Jersey 07045

Solid-State Microwave
Montgomeryville, Pennsylvania 18936

Plessey Semiconductor Products
1674 McGraw Avenue
Santa Ana, California 92705

Fairchild Camera and Instrument Corporation
464 Ellis Street
Mountain View, California 94042

addresses of rf PC-board manufacturers

Oak Materials Group
Laminates Division
174 North Main Street
Franklin, New Hampshire 03235

Cincinnati Millaron
Molded Plastics Division
Blanchester, Ohio 45107

ham radio

variable-inductance variable frequency oscillators

A comprehensive discussion of VFO circuits including a variometer VFO, an iron-vane VFO, and copper-vane VFO

This article deals with a number of variable-frequency oscillators, which I developed over seven or eight years, mainly for portable use.

Inductance-tuned VFOs are not a new idea. Long before the days of solid-state VFOs some manufacturers were using them in commercial ham gear. The Collins PTO (permeability-tuned oscillator) is an example. The original PTO was a precision-built factory product, but the inductance-tuned VFOs described here are simple enough to be built by any construction-minded Amateur. This might no longer be the case, however, if it becomes impossible to obtain good dial drive mechanisms through retail sales outlets.

The importance of simple VFO design may seem questionable at a time when the industry appears to be rushing into synthesizers. The answer is evident when one notes that data in an advertisement by a leading manufacturer indicated that its latest transceiver drew over an ampere from a 12-volt source in "receiving standby" position. In contrast, the receivers I've designed for portable use have drawn less than 50 milliamperes; something of the order of 10 milliamperes at 9 volts is possible with direct-conversion designs. Thus there continues to be a need to develop VFOs that are smaller, more economical, and more stable.

The reasons for the variable-inductance approach for both portable- and fixed-station use are these:

1. An inexpensive variable capacitor may cause noise and frequency-jitter when tuned.

2. Even a good variable capacitor, if at all available, may be large and heavy by modern standards, besides being expensive.

3. A variable-inductance tuning system can be low cost.

4. Variable-inductance tuning lends itself best to bandspread tuning, as in the Amateur bands.

5. Vernier dial drives that ensure backlash-free tuning are available from Japan and also from England (Jackson Bros.).

design considerations

The basic circuits used in my experiments are the series Colpitts (originally known as the Clapp oscillator), **fig. 1,** and, in one instance, the Hartley oscillator, fig. **2.** The Seiler circuit, not used, is a modification of the series Colpitts. It is, of course, desirable to make the oscillator frequency as stable as possible consistent with cost and size.

One source of frequency instability is the active element, the transistor, which is much more unstable with current and temperature changes than a vacuum tube. At a given frequency, the transistor equivalent internal capacitances change with power-supply voltage and biasing changes, thus causing frequency shift. Temperature changes have the same effect. One source of temperature change is the dc operating currents and the other is any change in the ambient (i.e. surrounding) temperature.

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Oscillator stability. An important factor in ensuring oscillator stability is to make the frequency much dependent on the LC circuit and little dependent on the transistor parameters. The usual way to do this is to start with a high-Q LC resonant circuit and couple it loosely to a very low impedance circuit – actually to two low impedance circuits, one for actuating the transistor, and the other at 180 degrees phase reversal for feedback from the transistor.

I've usually worked with fets because, like vacuum tubes, they have high input and output impedances, whose changes with temperature and voltage variation will have a minimum effect on the low impedances with which they are made to interface. The fets I've used have usually been N-channel, dual-gate, gate-protected mosfets such as RCA's 40673 or 3N211, because the impedances and mutual conductances are higher in mosfets than in jfets. No information in the Motorola HEP literature is given as to whether their mosfets are gate-protected. Older fets such as RCA's 3N128 and 3N140 are not gate protected and thus are very hard to use in this application.

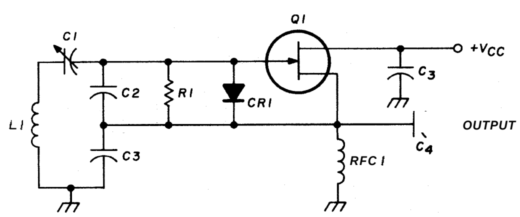


fig. 1. Series-Colpitts oscillator. C1 for tuning. Feedback ratio is determined by C2 and C3. C4 is an rf bypass capacitor and C5 a coupling capacitor.

Power source. To further minimize the effects of transistor instability in small portable equipment, I believe it's most economical of space and battery drain to power each oscillator with its own 9-volt battery rather than use elaborate voltage regulators with a single, common power source. Note that a battery has internal resistance, so that attempts to power another circuit from the same battery will result in frequency shifts whenever the latter circuit's current drain is changed. This effect results from the additional voltage drop in the battery's internal resistance, caused by the current to the second circuit. Besides rf shielding can be made very effective when the battery is kept in the shield box.

Battery current. If the battery current, which is normal for oscillator operation, is too high, oscillator frequency will drift after the voltage is applied, while the transistor is stabilizing to a slightly higher internal temperature. In experiments with mosfets a number of years ago I found that, in high-frequency oscillators, the drain current should not exceed about 2 milliamperes. Much lower values than this can be used to advantage, except that a too-low value can produce a noisy circuit or a hard-starting oscillator.

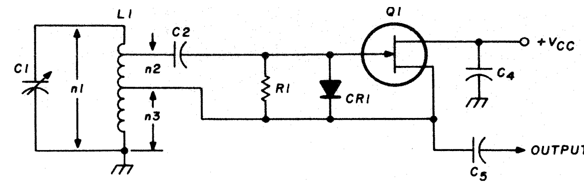


fig. 2. Hartley oscillator. C1 is for for tuning. Feedback ratio is determined by n2 and n3. C2 and C4 are coupling capacitors and C4 is an rf bypass capacitor.

The current can be decreased by decreasing the voltage on the agc gate of a dual-gate mosfet (G2 in the 40673). It can also be controlled by changing the feedback ratio, which is C2/C3 in **fig. 1**, or n2/n3 in **fig. 2.**

The relationship of feedback-ratio-to-performance is not simple, but fortunately a 1:1 feedback ratio is frequently best. If the resonant-circuit Q is high enough, isolation can be improved by making C1 relatively small compared with C2 and C3 **(fig. 1)** or by placing the high n2 tap considerably below the top of L1 in **fig. 2.** Carrying this procedure too far produces high drain current, noise, parasitic oscillations, or no oscillation.

Oscillator keying. If you wish to key a transmitter in the oscillator circuit, a low-current design is desirable to minimize chirp caused by transistor heating when the key is depressed. Conditions for low current frequently mean closer coupling of the transistor to the LC circuit, so that there is a limit to the effectiveness of this way of reducing chirp (e.g., making C3 in **fig. 1** too small).

On the other hand some designs with good isolation diminish chirp but introduce slow drift because of high current (e.g., making C3 in **fig. 1** too large). However, another factor makes direct keying undesirable. A key-click filter usually entails RC circuits with slow time constants. Charging and discharging of the capacitors produces transients in the dc voltages across the transistor terminals, causing frequency shift and, hence, chirp, as the voltage rises and falls.

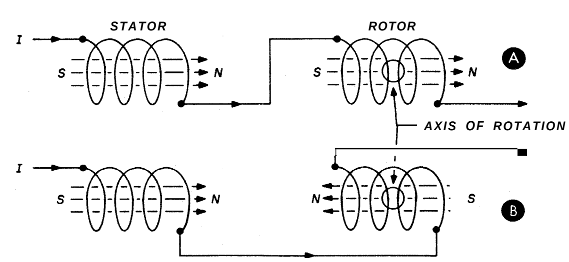


fig. 3. Principle of variometer. (A) fields aiding; (B) fields opposing.

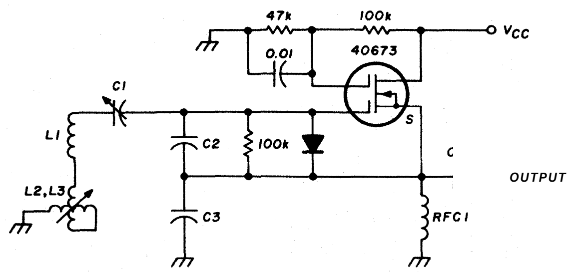


fig. 4. Variometer-tuned series-Colpitts VFO. **Tuning is by L2L3. C1 is for calibration. Feedback ratio is determined by C2 and C3.**

Temperature effects on inductors. In the matter of ambient temperature effects, it has been pointed out that even powdered-iron toroid inductors have a temperature coefficient too great for many purposes.[1] However, I've used these inductors for compactness in the VFOs of portable QRP equipment and have found that they did not drift too rapidly when the equipment was used under a canopy on a ship in the tropics, even with the sun setting.* Outdoor operation in a simple metal box in the sunlight is more difficult.

Air-core toroids. Ferrites have much worse temperature characteristics than powdered iron, so almost the only alternative is an air-core coil. However, there's still the issue of what kind of air-core coil to use. A number of years ago I evolved a simple construction procedure for making air-core toroids. These are much larger than their powdered-iron equivalents, but they still can be confined to a small shield box; since a toroid's magnetic field is closely contained, the shield box has little effect on inductance and *Q*, producing mainly an increase in capacitance.

One of my early air-core toroids had a *Q* of about 300 and was used in a signal generator tuning from 12 to 23 MHz. With a 200-milliwatt final amplifier preceded by a buffer, it was possible to key the oscillator and transmit on the fundamental frequency at 14 and 21 MHz without very bad chirp. But key clicks were bad, so that direct keying became undesirable, as discussed above.

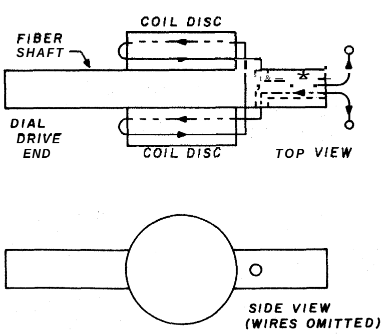


fig. 5. Split-rotor coil construction.

Temperature stability of air-core toroids built for a recent series of tests was disappointing, as described further on. Perhaps the toroidal shape itself produces a high temperature coefficient when subjected to the expansion and contraction of copper and plastic.

Capacitors. Capacitors are temperature sensitve too. In frequency-determining circuits, experimenters have found the small polystyrene types to be the best. Both DeMaw[1] and Eaton[3] correctly note that overall frequency drift can be compensated for by experimenting with capacitors of different temperature coefficients, but that this is a difficult process, the results of which might be hard to duplicate. And enclosing the oscillator in a constant-temperature oven, an obvious alternative, would not be feasible for small, portable equipment.

In addition to being temperature sensitive, certain capacitor types are highly sensitive to humidity. Mica compression trimmers in rf circuits have been known for many years to have some of the properties of barometers. It used to be considered a cute trick to place a sensitive back-biased vacuum-tube voltmeter across the output of an i-f stage and watch the effect of blowing one's moist breath through a glass tube onto a tuning capacitor.

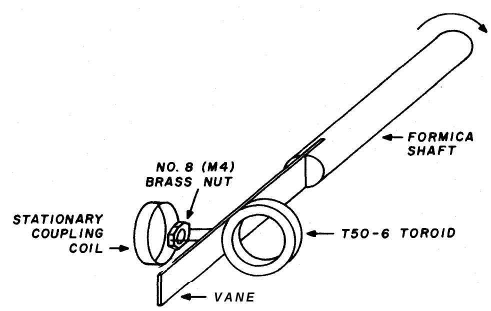
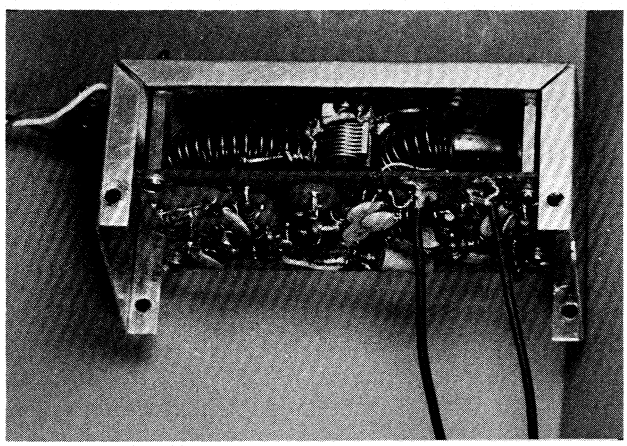


fig. 6. Tuning couplet. Vane is of epoxy circuit-board material. Leave two small copper spots for the supporting wires. Alternatively, support nut with insulating spacer block. Hold T50-6 toroid to vane with cyanoacrylic cement.

Even ceramic trimmers leave a lot to be desired, which makes it appear that adjustable air-dielectric trimmers might be the best for stable oscillators. Some very small air trimmers, with very close plate spacing, are available. Having once worked with installations on yachts, I wonder if the spacing in some capacitors is small enough that precipitation, which sometimes forms when hot, moist air cools on a metal surface, might not short-circuit capacitor plates; i.e., is a "dew-drop shunt" a possibility, especially if the dew drop happens to be salty?

Shielding. In small, portable equipment, a limited amount of frequency drift can be corrected by resetting the oscillator to a reference calibration point, as in zeroing the VFO to a frequency marker. The important thing here is to be sure that the VFO does not drift appreciably during a contact. To obtain improved results with a small increase in size, the temperature of the oscillator can be made to change slowly under conditions where the oscillator might have drifted rapidly, as when moving equipment outdoors into the sun or into a chilly breeze. This can be done by using both an internal and external shield box. Such an arrangement produces double rf shielding, which not only decreases rf interference locally, but also is one way of reducing feedback from the higher stages of the transmitter to the VFO, which may be a source of instability. For best rf shielding, as in rf screened rooms, grounding the boxes to each other should be at *one* point, with precautions in treating rf cables and power leads the same as in the screened-room case.

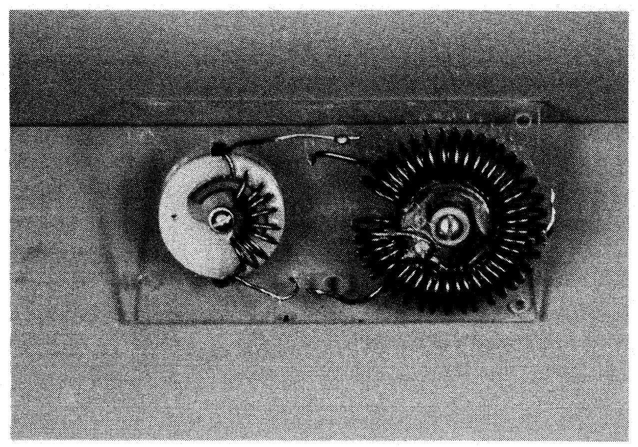


Coil board for VFO H-2. Air-core toroid and iron-vane tuning unit are shown in position.

the variometer VFO

In the early 1920s the variometer was used to tune stages of the then-prevalent trf (tuned radio-frequency) broadcast receivers. It consisted of one stationary coil, inside of which was one rotatable coil. **Fig.** 3 shows the principle, with the rotor shown beside the stator for illustrative purposes. Drawing **A** shows the rotor in the "fields-aiding" position. The magnetic fields of the two coils are mutually aiding, which results in a larger value of inductance than the sum of the two. In the "fields-opposing" position, B, the circuit inductance becomes small.

Commerical variometers were made with split stators and rotors, which were shaped to ensure the maximum coupling of electromagnetic field lines, and thus the maximum range of inductance values,

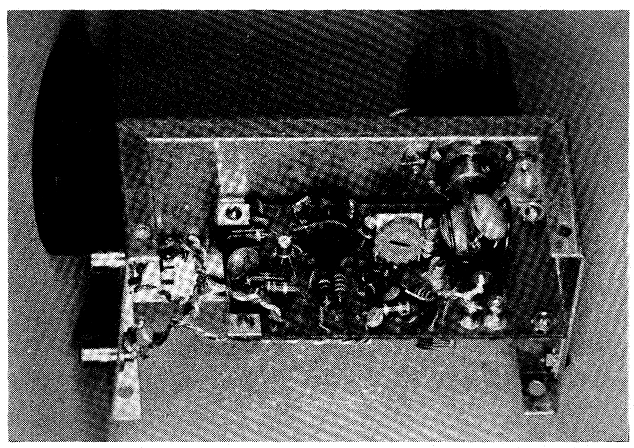


VFO H-2 assembly. Air-core toroid and copper-vane tuning unit are visible under circuit board.

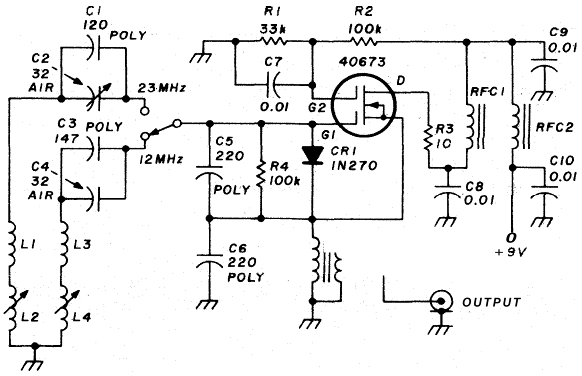
as the rotor was turned 180 degrees inside the stator. The main problem with the variometer is that most of the circuit resistance remains unchanged for all rotor positions, so that the Q becomes very small at one end of the tuning range. This results in broad tuning and low gain at the high-frequency end of the dial in trf receivers. However, for bandspread tuning, which is required for covering only a few hundred kHz in an oscillator, a small variometer can be placed in series with the main inductor, so that the worst-Q position of the rotor will have only a small effect on the circuit Q.

Fig. 4 shows a series Colpitts circuit tuned by variometer L2L3 in series with L1. Capacitor C1 can be used for calibration.

Construction. Fig. 55 shows how the split rotor in the experimental model was built. The tuning shaft is a 1/4-inch (6.35-mm) phenolic rod. Cemented to its sides are two circular disks of insulating material,



Variorneter VFO. Variometer tuning unit is in corner attached to Jackson dial drive behind tuning knob.



C1	120-pF polystyrene
C2, C4	32-pF air trimmer
C3	147-pF polystyrene
C5, C6	220-pF polystyrene
C7-C10	0.01μF
L1	9-1/2 turns no. 18 (1-mm) wire on T50-6 core
L2	1-1/2 turns no. 16 (1.3-mm) wire close to couplet
L3	21 turns no. 20 (0.8-mm) wire on T50-6 core
L4	4-112 turns no 18 (1-mm) wire close to couplet
T1	primary 19-112 turns no. 30 (0.25-mm); secondary 4 turns no. 24 (0.5-mm) wire on FT37-63 form or equivalent
RFC1,	23-112 turns no. 32 (0.2-mm) wire
RFC2	on FT-37-63 form or equivalent

fig. 7. Couplet-tuned VFO for 15- and 20-meter receiver with 9-MHz i-f.

each carrying a one-turn coil section. Leads from the rotor go through a drilled portion of the shaft to the rear, and come out to stationary terminals in the rear. These leads are twisted, plastic-covered, stranded no. 24 (0.5-mm) wire. They emerge at terminals close behind the shaft to minimize the adverse effect of continual flexing. The dial drive is at the front of the shaft. The spring effect of the twisted wires was enough to rotate the shaft against dial friction. (This could have been eliminated by a friction bearing as described later.)

The disks of insulating material in the rotor could be replaced by powdered-iron cores to some small advantage. The split stator consists of two self-supporting, one-turn, coils connected in series with the rotor, and into which the rotor meshes.

The variometer-VFO design illustrates the principle. The oscillator was never used, because better approaches became evident. However, it did serve as a stimulus for thinking about inductance tuning. In other respects, this particular oscillator might be called the "nostalgia special."

Performance. In all-inductance tuning, as you might guess from simple considerations, for a fixed ratio of inductance change to total circuit inductance, the frequency coverage or bandspread achieved decreases directly as frequency decreases. This is accurate when the bandspread is a small fraction of the main frequency; it becomes less accurate as the fraction increases. Design for the lower high frequencies suffers from this frequency effect, especially at 3.5-4.0 MHz, where a large bandspread may be desired.

a rotating-couplet VFO

This design allowed the most compact construction of all that were built. The heart of the unit is a "rotating-couplet," whose design I evolved from recollection of an aligning tool of the 1930s. This magic wand had a powdered-iron slug at one end and a brass slug at the other end. Inserting the iron slug into an rf solenoid coil increased the coil inductance. If this act tended to increase the receiver output at a given frequency, the rf coil needed more inductance; a decrease meant it needed less. The brass slug worked in the opposite manner, since induced eddy currents in the brass reduced the coil inductance.

Construction. The couplet shown in **fig. 6** consists of a powdered-iron toroidal core (I used T50-6) placed opposite a brass slug and in the same plane. The whole couplet is rotated by a formica shaft. At one extreme of a 180-degree rotation, the toroid is coupled closely to a stationary coil, increasing its inductance a maximum amount. Rotation away from the maximum position decreases the effect of the iron. At the 90-degree point, the coil inductance should be the same as if the coil were alone. Past the 90-degree point, the brass slug rotates into the coil, reducing its inductance. An 8-32 (M4) brass nut was used as the slug. The core and the slug are mounted

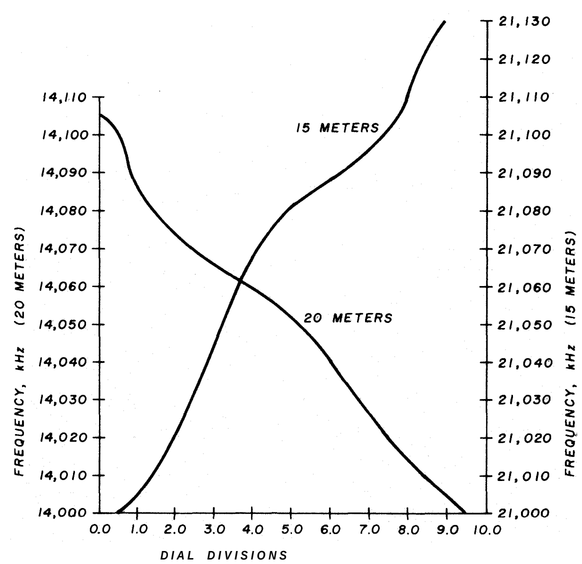
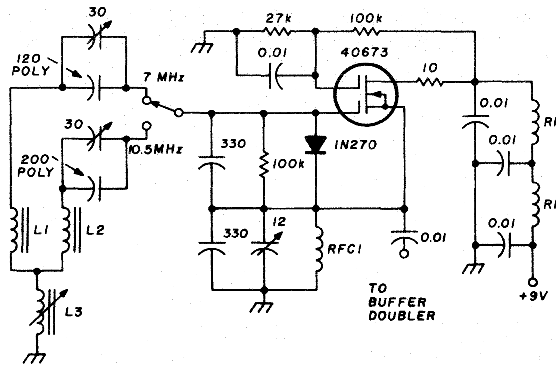


fig. 8. Frequency calibration of receiver using couplet-tuned local oscillator.



L1	33-1/2 turns no. 26 (0.3-mm) wire on T50-6 form
L2	18-1/2 turns no 22 (0.6-mm) wire on T50-6 form
L3	see text
RFC1, RFC2	Miller 70F475A1
RFC3	Miller 70F125A1

fig. 9. Iron-vane VFO for 15- and 20-meter transmitter.

on a small Fiberglas epoxy vane made from a piece of circuit board from which most of the copper was removed. The nut is supported on wires soldered to two small copper spots remaining on the vane. (One large spot might have changed the coupling pattern.) A better method would have been to mount the nut in its offset position by using a small plastic block.

For homebrew equipment, when multiband operation is desired, it's probably better to go back to early commercial practice and use a bandswitching VFO, since the penalty of separate frequency calibrations for each band may be less than that of using an extra mixer stage with frequency conversion and the accompanying difficulty of suppressing additional unwanted frequencies.

Fig. 7 is a circuit diagram of a VFO used as an LO in a 15- and 20-meter bandswitched portable CW receiver with a 9-MHz i-f. For 20-meters the oscillator operates in the 23-MHz region and for 15-meters in the 12-MHz region. This is the LO used in my "Minicruiser" receiver.2

Two switched coupling coils, L2 and L4, are placed on opposite sides of the rotatable couplet. Each is in series with a main inductance coil. These coils are L1 and L3.

Performance. The calibration curves of the receiver using this oscillator are shown in **fig. 8;** they are obviously far from linear. Even worse, in the 90-degree region such an arrangement can produce a small but annoying defect. As the powdered-iron core rotates past the minimum-effect position at 90 degrees and the brass slug rotates in, the iron slug may momentarily increase the coil inductance again to a greater extent than the brass slug decreases it, producing a tiny doubling back of the calibration curve. Similarly, the effect of the brass slug may predominate over a small interval on both sides of 90 degrees.

an iron-vane VFO

This VFO has some similarity to the previous one, except that the variable inductor in series with the main inductance consists of a semi-circular coil and a rotatable powdered-iron vane of the same general type as that originally designed for a QRP Transmatch by W1CER and K1KLO.[4]

Fig. 9 is the circuit for a two-frequency switched VFO used in the transmitter of the *Minicruiser.*[2] The generated frequencies are near 7 and 10.5 MHz, to be doubled to cover the CW frequencies near 14 and 21 MHz. **Fig. 10** shows some design details of an iron-vane tuning unit similar to the one used here.

Construction. I made the iron vane from a toroidal core 1 inch (25.4 mm) in diameter, of material said to be similar to Micrometals T2. I sawed the core in half and cemented it to a sickle-shaped flat plate from which the copper had been etched. Shaping the sickle was easy. I cut an approximate disk from the circuit-board material and turned it in a lathe to a diameter of 1 inch (25.4 mm), although without a lathe a tedious sawing and filing job would have been satisfactory. The material between the sickle and the "hub" was removed with bandsaw files and occasionally a small hacksaw.

The best adhesive for fastening the powdered-iron segment to the sickle was cyanoacrylic (like Eastman 910). To my surprise, epoxy worked badly, perhaps because it did not adhere to the lacquer on my particular core segment. Rubber cement was even better. The assembly withstood a voyage around South America and a truck trip from San Francisco to Denver.

I screwed the now-finished vane to the end of a 1/4-inch (6.35-mm) Formica shaft, which had been drilled and tapped for a 6-32 (M 3/5) screw. I made the shaft long enough to fit into the dial drive. A

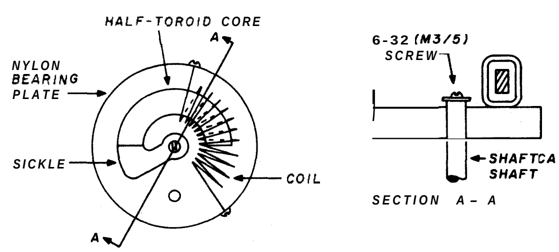


fig. 10. Design features of an iron-vane tuning unit. This unit was used in some tests of VFO H-2 in lieu of the copper-vane tuner of figs. 12 and 13.

good fit, essentially free of backlash, was made by drilling and tapping the dial drive collar for 4-40 (M3 screws with the shaft already inserted, using two screws at right angles and set at different lengths along the shaft.

To eliminate wobble of the iron vane as it moved in the pickup coil, I devised a bearing made of a circular piece of solid nylon. The material flows when hot, so the shaft hole had to be drilled with a 1/4-inch (6.3-mm) drill and then filed carefully for a tight fit to the Formica shaft. Very slow rotation, as in a back-geared lathe, can eliminate this problem. The nylon material provided a tight grip on the shaft, yet did not produce frequency jitter as the dial was turned.

In the design for the circuit of **fig. 9,** I wound the semicircular coil, L3, with five turns of no. 18 (1 mm) wire, using a rectangular mandrel 1/4 x 1/2 inch (6.35 x 12.70 mm) and cemented it, using epoxy, to the outside circumference of a small nylon bearing in the region where the bearing was mounted to an acrylic plate. In the design of **fig. 10,** I merely cemented the coil to the flat surface of the fairly large bearing shown, using Duro plastic-mending cement. I found it desirable to make the coil cover less than 180 degrees of arc, otherwise the end turns would act as stops (detents) as the core was rotated, with considerable potential for damage.

Performance. Fig. 11 is a set of calibration curves for the iron-vane VFO of **fig. 9** at the frequencies observed after doubling. They are quite linear over a fairly wide range. Irregularities are no doubt caused by imperfections in the hand-wound variable inductance.

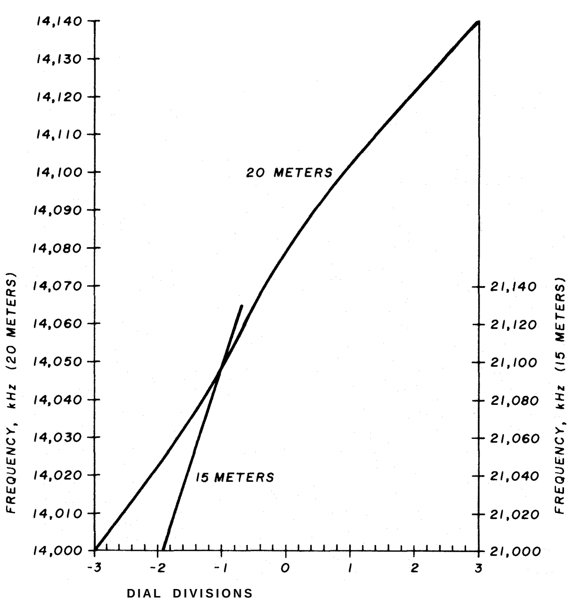


fig. 11. Calibration of "Minicruiser" iron-vane transmitter VFO.

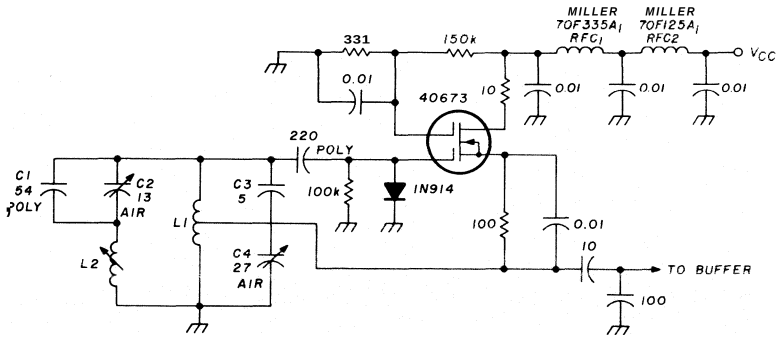


fig. 12. Circuit of inductance-tuned Hartley VFO **H-2. L1 and L2 are described in text. C4 is for calibration setting.**

a copper-vane Hartley VFO

In the design of the iron-vane VFO of **fig. 9,** the main inductance coil for each frequency is a powdered-iron toroid. Hoping to increase the temperature stability of the system, I decided to sacrifice some compactness and eliminate powdered-iron cores altogether. The main inductance coil now became an air-core toroid. The series tuning coil became an air-core half-toroid into which was rotated a sickle of copper tubing which, through eddy-current action, decreased the inductance (and also the Q) of the semi-toroid as more of the sickle was rotated into the coil. The effect was the opposite of that experienced with the iron-vane VFO. As a further innovation I decided to use the simpler Hartley circuit of **fig. 2. Fig. 12** is the circuit of my test VFO **H-2.** The earlier H-1 was similar.

Construction. For coil L1, I had evolved a simple means of making an air-core toroid as mentioned above. Coil of the order of 2 inches (51 mm) in diameter demonstrated Qs of the order of 180-200 at frequencies from 5 to 15 MHz, with shunt capacities of 60-100 pF.

Coils L1 and L2 of **fig. 12** are shown mounted on an acrylic coil board in **fig. 13;** their construction is described below.

First I constructed a spool consisting of two 1-inch (25.4 mm) acrylic disks separated by a cylindrical fiber spacer 5/16 inch (8 mm) in diameter and 0.400 inch (102 mm) long. The spacer was drilled for a no. 8 (M4) brass screw for holding the assembly together, and for fastening to a mounting board. The disks were drilled and tapped for short no. 4 (M3) brass screws to serve as terminals.

I made the main toroidal coil, L1, by inserting a 112-inch (12.7-mm) dowel with screw terminals into a lathe chuck and winding forty two turns of no. 14 (1.6 mm) soft-drawn enameled copper wire onto the dowel. Upon release, the coil I finally made was 40-1/2 turns, which was then carefully wrapped around the spool and held with a rubber band.

The next step was to cement the coil to the spool, using epoxy, although plastic mending cement was later found to be easier to use and worked just as well. I used this coil in experimental VFO H-2 for 5-MHz tests. It was reduced to 38-112 turns for 9-MHz tests, the purpose being to produce an appreciable gap to isolate the high-impedance end from lossy and temperature-unstable materials.

I made L2, the half-toroid for tuning, in a similar way except that eighteen turns were used for VFO H-2. I mounted a nylon bearing with a 114-inch (63.5 mm) shaft hole flat on the Plexiglas coil board. The outer diameter of the bearing was 314 (190.5 mm) and the height 5116 inch (79 mm). Then I wrapped the half toroid around the bearing to an angle slightly less than 180 degrees and also made it rest against the board. Next I cemented it into place. The copper-tube sickle was a half-circle of 114-inch (6.35-mm) tubing flattened at one end and offset from the shaft end by an insulated crank arrangement.

In some of the temperature tests described later, I replaced the air-core toroid of L1 by one of similar inductance and *Q*, consisting of a T50-6 core wound with twenty six turns of no. 24 (0.5 mm) enameled copper wire. Also, in some tests I replaced the copper-vane tuning unit at L2 with the iron-vane unit of **fig. 10.** Here coil L2 consisted of nine turns of no. 18 (1 mm) wire initially wound on a mandrel 1/4 x 318 inch (6.35 x 9.25 mm).

With different values of C1 I could make the VFO function at frequencies from roughly 4 to 12 MHz using essentially the same coil at L1, which was center-tapped. I did most of the experimenting near 9 MHz with C1 consisting of two 27 pF polystyrene capacitors in parallel.

Performance. Fig. 14 shows a frequency calibration of the copper-vane VFO H-1 around 9 MHz. The coils were similar to their H-2 counterparts, except that L1 had 41-1/2 turns and L2 had 17 turns. The linearity of most of the curve is evident. There was less success with temperature stability in this type of design, as described below.

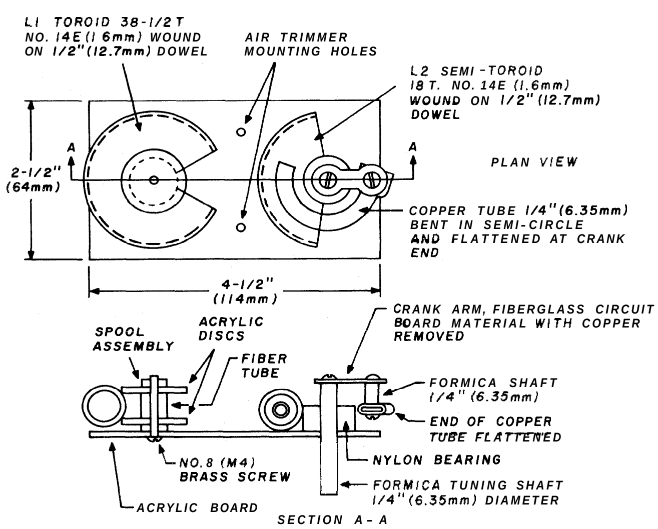


fig. 13. Coil board for copper-vane VFO. The etched circuit board is of the same size and is mounted above on l-inch (25.4mm) brass spacers.

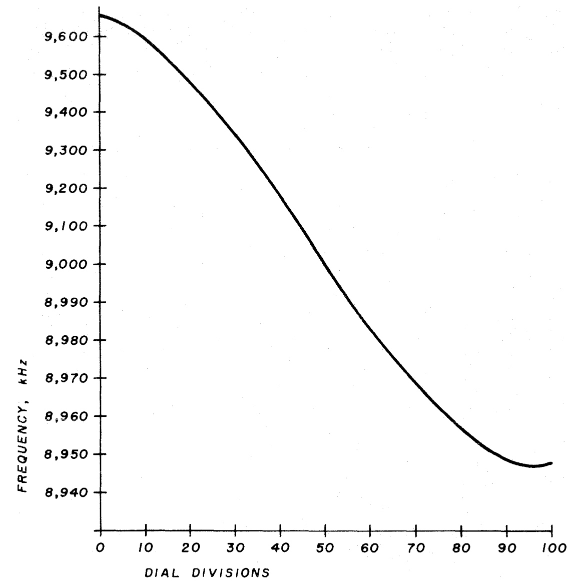


fig. 14. Copper-vane VFO H-1 frequency calibration.

temperature experiments

I had originally hoped to make readings of frequency drift with change of temperature of the VFO, but it soon became evident that this would not be an easy. matter. Any kind of structure, whether a VFO or a house, when subjected to temperature changes acts like a radio circuit to the extent that it has a time constant. However, whereas radio circuits may have time constants as short as several nanoseconds or as long as several seconds, a house being heated or cooled may have a time constant of the order of a day and a half. A piece of radio equipment may require many hours to come to something approaching temperature equilibrium.

Heat transfer. Heat introduced into a VFO from the outside gradually travels to the inside. If the heating takes place inside an insulated box, with the room temperature and the heat source constant, everything eventually will reach an equilibrium temperature. The setup I decided on only approximated the ideal described above.

I mounted the VFO box on an electric plate warmer and covered it with a large chassis taped to the heating surface. Leads for the dc power and the temperature sensor were brought out at one corner. The rf output, after passing through a buffer stage, was brought through a piece of RG-174 cable at another corner and fed to a frequency counter.

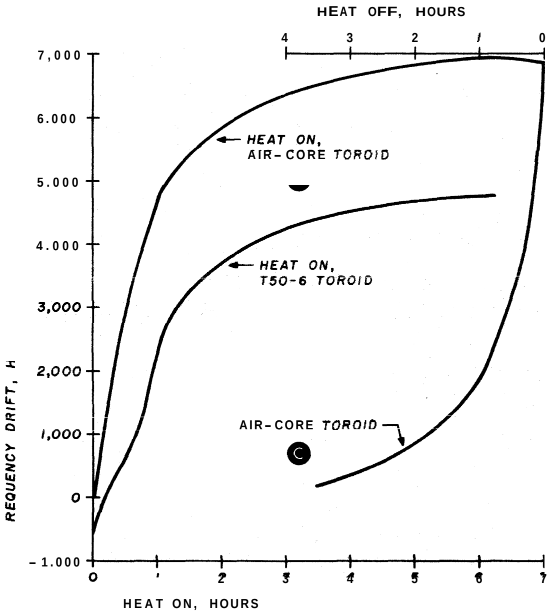


fig. 15. 5-MHz heat runs on VFO **H-2 with copper-vane tuning unit.**

Test procedure. For the temperature-sensing device, I used an IC made by Analog Devices, type **AD590.** This IC was fastened to the **VFO** box. The test procedure I used was to a) turn on the heat at the **LOW** setting when I thought the **VFO** to be in equilibrium with room temperature, and b) make periodic readings until the frequency stabilized. At this point, I assumed that the temperature of everything of importance inside the **VFO** was equal to the peak temperature reading of the run, about **34F (19C)** above room temperature.

Some of the circuit components were undoubtedly more temperatu're sensitive than others and some might have even had temperature coefficients of opposite senses. There was, of course, no simple way of knowing when each circuit component reached what temperature, nor which ones were doing what during the transient warmup process. Besides, to reach a single equilibrium temperature took the better part of a day. The best thing I could do was to plot frequency drift ***versus*** time for the procedure described above and repeat the test under the same thermal conditions after making changes in the components in which I was interested. These components were usually L1 and L2 of fig. 12. In tests of both iron-vane and copper-vane tuning units, I always placed each vane in the position of closest coupling with its coil. I assumed that polystyrene capacitors and air trimmers were consistent in their temperature behavior, and spent only a little time on mica capacitors, which apparently differ one from another.[1]

Test results. In fig. 15, curve **A** shows frequency drift ***versus*** time at 5 MHz, with the 40-112-turn air-core toroid as L1 and the 18-turn copper-vane tuning unit as L2 (fig. 12 circuit). C1 was a 220-pF polystyrene capacitor in parallel with one of 27 pF. The frequency drifted upward to 4700 kHz in the first hour, but it was not until the start of the sixth hour that the peak of 6900 Hz had been reached.

Curve B shows what happened after I turned the heat off and removed the **VFO** at the beginning of the seventh hour; the frequency decrease was very rapid at first. Since frequency drift is related to temperature, there was an obvious resemblance between the rise and fall of temperature and sf current as the dc voltage in an inductive circuit is switched on or shunted out.

Curve ***C*** (fig. **15)** is for the same oscillator but with the powdered-iron toroid in the L1 position. I was surprised to note that, after a small negative excursion, the frequency had drifted to only about 4700 Hz in six hours. There was later evidence, however, that the air-core toroid would have been more stable with a larger gap between the ends, although this did not make the air-core toroid competitive in the 9-MHz tests described below.

Fig. 16 shows the 9-MHz tests, with C1 at **54 pF.** For the test of curve **A,** I reduced L1 to 38-1/2 turns, but the copper vane remained as before. In this test the frequency drift was **6350** Hz in fourteen hours and still rising.

For curve B, I used the **T50-6** toroid in the L1 position, but continued to use the copper-vane unit at L2 as before. Here the frequency drift remained very small for the first twenty minutes, then went sharply negative to **– 1750** Hz, then went up again, crossing the zero axis at about 4-112 hours and reaching **1665** Hz at 14-1/2 hours, where it was still increasing.

I next tried the **T50-6** toroid in series with the iron-vane tuning unit of fig. 10. Curve ***C*** (fig. **16)** shows the results. Again, there was an initial large swing in the negative direction, to – 2000 Hz in the first hour. Then the frequency began to increase with time, crossing the zero axis at 7 hours and peaking at **400** Hz at 13 hours 45 minutes.

Curves B and ***C,*** being more irregular than curve **A,** show some undetected mechanical instability, probably in the **T50-6** toroid. However, the negative excursion and return of the two curves has some significance. Product literature issued by Micrometals for their toroids show curves of temperature coefficients ***versus*** temperature for all of their powdered iron, including no. 6 material. All have a transition from a positive to negative coefficient at 77F (25C). It's possible that, for curves B and ***C,*** the change of slope at about one hour occurs near where the T50-6 coil reached that temperature.

Fig. **15** and **16** are not directly comparable, because the air-core toroid was improved for the 9-MHz heat runs, as described above. Nevertheless, they do show that, contrary to expectations, the air-core toroids were not relatively immune to temperature changes. From the viewpoint of relative stability over hours, the combination of **C** might be termed best, with B next best. From the viewpoint of the first 20 minutes, **B** seems the most stable. However, the steep negative excursions of **B** and **C** are worse than the steep positive excursion of curve **A,** in the first few minutes after start.

In one brief test at 9 MHz, where I used all air-core coils and a 51-pF silvered-mica capacitor at C1, the temperature coefficient was negative. This test was discontinued for reasons mentioned previously.

Observations on coil construction. Thinking that the poor temperature behavior of the air-core toroid at L1 might be caused by the epoxy cement, I built an identical one using Duro plastic mending cement, which yielded essentially the same temperature curve. My guess was that perhaps the narrow diameter of the air toroid, relative to the wire diameter, was responsible for a greater proportionate reduction of the magnetic-flux-carrying cross-section area as the wire expanded with heat. In none of the tests did I try any temperature-compensation techniques because of difficulties in reproducing results when temperature-compensating components are introduced.[3]

conclusions

The Hartley oscillator uses fewer components than the series Colpitts, and may be adjusted more easily in an experiment. However it is possibly more prone to producing spurious oscillations and noise if not adjusted correctly. Apparently poor temperature characteristics of the Hartley oscillators with air-core toroids might be due to the fact that, because of spurious oscillations, it had not been possible to achieve isolation of the resonant circuit of **fig. 2** by bringing the n2 tap below the top of L1. In respect to isolation from temperature-sensitive transistor parameters, the Seiler circuit would probably have yielded the best temperature stability, with the series Colpitts as good but harder to adjust experimentally.

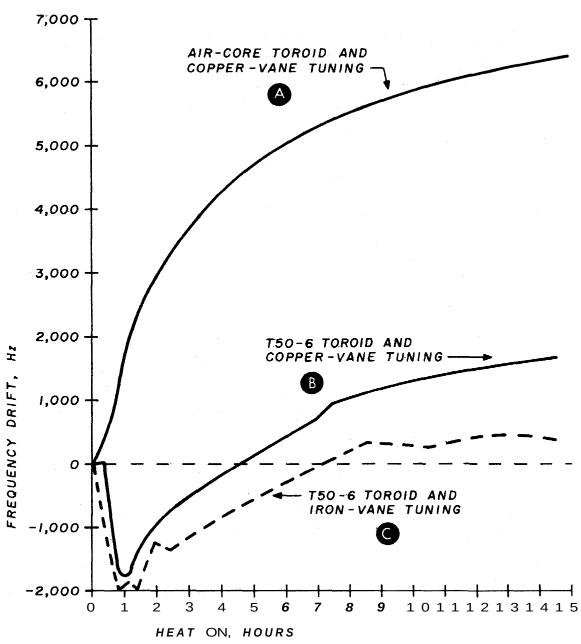


fig. 16. 9-MHz heat runs on VFO H-2.

Of the inductance-tuned VFOs the variometer VFO is an interesting antique; and the rotating-couplet VFO can be the most compact but yields a poor calibration curve. So far, air-core toroids are a disappointment in temperature behavior; one can make a practical VFO for portable use with an iron-core toroid and iron-vane tuning, especially if a second shield box is used. Frequent recalibration of a reference point against a simple crystal frequency standard is very helpful. Both iron-vane and copper-vane inductance tuning yield linear calibration curves.

My experiments over a number of years have demonstrated that the home constructor can build a useful VFO for either fixed-station or portable use without resorting to precision variable capacitors. In these experiments, frequency drift with temperature was greater than might be desired, but I believe that future experimenters should be able to determine whether improvements can result from the use of conventional air-core solenoid coils along with inductance tuning. Also, I regret that frequency-stability tests were not made on the Seiler or series-Colpitts circuit with inductance tuning. So there is plenty of opportunity for future learning.

The experimenter should certainly read the VFO references mentioned so far, besides many others, including Jim Fisk's early article[5] with its references going back many years. Bill Wildenhein's careful experiments[6] indicate that frequency drift can be conquered. In performing experiments, bear in mind one fact, which caused Reed Easton[3] to turn to a synthesis control technology: There are so many temperature-dependent variables that apparently identical VFOs may have different drift characteristics.

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ham radio

linear-amplifier cost efficiency

Consider gain per dollar rather than watts per dollar for a cost-efficient linear amplifier

Many Amateurs ask the question, "Should I invest several hundred dollars in a linear amplifier and what good will it do me?" The watts-per-dollar criterion is misleading because signal gain at the receiver does not increase linearly with power increase. A more valid measure would be gain per dollar. This article shows that the greatest signal gain-per-dollar cost is obtained at some easily calculated level of amplification. Amplification beyond this optimal level should be avoided by the cost-conscious Amateur.

background

It's well known that the increase in signal strength resulting from an increase in power is determined by the following formula:

$$dB = 10 \log \frac{P_2}{P_1} \qquad (1)$$

where P_1 is the initial power level output and P_2 is the increased power level output.

If we begin with $P_1 = 100$ watts output from the exciter and increase the power in steps of 50 watts, the increase in signai strength wiii be (for $P_1 = 100$ *watts*) as follows:

linear amplifier output (P_2)	increase (dB)	incremental increase (dB)
150	1.76	
200	3.01	1.25
250	3.98	0.97
300	4.77	0.79
350	5.44	0.67
400	6.02	0.58

An increase in output from 100 to 150 watts will increase the signal by 1.76 dB. Notice that as power increases the gain also increases but in ***progressively smaller steps.*** This is true for any level of power input and amplification. Each increase in power output results in a smaller increase in gain at an ***increased cost*** of energy and components. Clearly it's wise to stop short of the point where an increased cost would result in no significant incremental increase in signal strength.

efficiency

Most will agree that an increase in power will result in a greater cost of energy, components, insurance, space, and weight. We can define a cost-efficient amplifier as one that produces the strongest signal per unit power output, or the strongest signal per cost. In Pentagon terminology, a cost-efficient amplifier produces the "biggest bang per buck."

For example, if two signals of identical strength are generated from two amplifiers, the cost-efficient amplifier would be the one costing less. Another example is to compare the signal received from two amplifiers of equal cost. The one producing the greatest gain would be the most cost efficient.

Cost efficiency can be expressed mathematically:

$$\text{efficiency} = \frac{dB}{cost} \qquad (2)$$

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	DenTron 1200	DenTron Clipperton	ETO Alpha 76A	ETO Alpha 770X	Heath 201	Heath 221	Swan 12002	Swan 15002
power input (PEP)	1200	2000	2000	4000	1200	2000	1000	1500
gain (P_1 = 100W)	10.8	13	13	16	10.8	13	10	11.7
price (1979)	380	600	1495	3995	385	569	500	600
efficiency = $\frac{\text{gain}}{\text{price}}$ x 100	2.9	2.18	.87	.4	2.8	2.28	2	1.95

This article assumes costs increase proportionally but not linearly with the power level of the amplifier. I attempted to compare the selling prices of various amplifiers and to compute their relative efficiency. This proved to be a futile effort because prices are influenced by factors other than power, such as paying a premium for a name (Collins), paying for features not found on amplifiers of different manufacture, and not paying for costs of labor (Heath).

What could be done, however, is to compare amplifiers of the *same* manufacturer. This was done, and in every case the increase in power was accompanied by a commensurate increase in price and a decline in efficiency, where efficiency = gain divided by cost.

Above are calculations of comparisons of amplifiers manufactured by DenTron, ETO, Heath, and Swan which demonstrate this thesis.

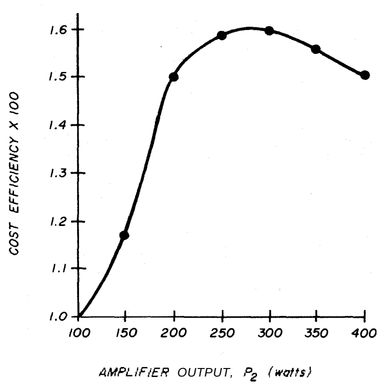

fig. 1. Cost efficiency obtained when increasing power-amplifier output. Best cost efficiency, in terms of costs of energy, components, insurance, space, and weight of the linear amplifier, occurs when amplifier output is about three times that of the exciter. In other words, $P_2 = 3P_1$ where P_2 and P_1 are the amplifier output and input power respectively.

where costs increase proportionally to power level. **Fig. 1** shows that the cost efficiency curve rises, levels off, and falls as a function of amplifier power. This is because of (a) the increasing costs that accompany an increase in power, and (b) the progressively smaller *incremental* increases in dB. Each additional dollar spent in power increases results in smaller and smaller signal increases. At some point the dB increase becomes insignificant but the costs continue to rise, resulting in reduced efficiency ***(for*** P_1 = ***100 watts):***

linear amplifier output (P_2)	increase (dB)	cost efficiency x 100
150	1.76	1.173
200	3.01	1.500
250	3.98	1.590
300	4.77	1.590
350	5.44	1.550
400	6.02	1.500

An examination of **fig. 1** shows that optimal cost efficiency occurs when the linear amplifier output is approximately *three times* that of the exciter. Because we're speaking of the ratio of two power levels, optimal amplification of three times exciter input is true for all power levels.

summary

We've defined cost-efficient amplification as that which gives us the biggest bang per buck. This occurs when $P_2 = 3P_1$. This won't give you the strongest signal on the air but it will give you the *strongest signal per dollar.*

This definition of efficiency might be modified to consider other factors. For instance, a weight-efficient transmitter could be defined as one having the highest ratio of dB to weight, thereby being useful for comparing mobile or portable equipment.

The cost-conscious Amateur should consider a linear amplifier that increases exciter output three times — all other things, such as the antenna, being equal. After this point, diminishing returns occur.

ham radio

vhf techniques

improved accuracy when measuring small inductances with a dip oscillator

The time-honored method of measuring inductance has been to parallel the coil with a capacitor of known value, determine the parallel-resonant frequency with a dip oscillator, and calculate the inductance from the formula

$$L = \frac{1}{4\pi^2 f^2 C}$$

For small inductance values this method is, at best, approximate because of **a**) uncertain calibration of the oscillator frequency, **b**) normal tolerance of the capacitor value, and **c**) lead inductance of the capacitor, which adds to the inductance being measured. The uncertainty of **a** can be eliminated by coupling a frequency counter to the dip oscillator when obtaining a dip. The effect of b can be minimized by using a 1 or 2 per cent capacitor, or by actually measuring the capacitance on a bridge or equivalent instrument. The problem of capacitor lead inductance, which can amount to 10 to 20 nanohenries even with minimum lead length, is solved as follows.

If suitable capacitance-measurment equipment is available:

1. Prepare one side of a small piece of double-sided copper-clad board as shown in **fig. 1**.

2. Clip off the leads of a dipped mica capacitor of any convenient value; 100 pF is suggested.

3. Carefully crack the conformal case of the capacitor by gradually applying pressure in a bench vise and then picking off the case fragments; be sure not to loosen the end crimps of the denuded capacitor.

4. Place the capacitor on the prepared board so that it straddles the bared strip, with the remainder of the leads facing upward, and solder the end crimps to the copper cladding.

5. Clip the remaining leads from the capacitor. You now have a leadless capacitor, with only its structure contributing any inductance.

6. Measure and record the capacitance between the two isoiatea sections of the board. (The measured capacitance will be greater than the capacitor value by approximately one-half the capacitance between the back of the board and either top section.)

If you do not have access to capacitance-measurement equipment, follow steps **1** through **5** above, using a small piece of ***single-sided*** copper-clad board. Assume the final capacitance to be the same as that of the capacitor.

The coil to be measured can be soldered onto the board so that it bridges the bared strip, placing it in parallel with the capacitor with virtually no added series inductance. If you must use a single-sided board, be sure that it is placed on a non-metallic surface before dipping the L-C combination; otherwise an error will be introduced.

Purists may want to use a chip capacitor to further reduce the parasitic inductance, but this is probably overkill at frequencies where a dip oscillator can be used.

Wilkinson power dividers

One of the accepted techniques used by Amateurs in up-conversion to the vhf through microwave regions is that of utilizing the same local oscillator (LO) for both transmitting and receiving. This requires either a coaxial relay to switch the LO from transmit to receive or a power divider. If the LO has sufficient power output to drive both the transmit and receive mixers, the power divider is the simpler approach. Such dividers are available commercially, but their price and that of a coax relay are comparable.

An elegant solution, both technically and economically, is to use a Wilkinson power divider. This handy circuit may be configured from discrete components or from transmission-line sections; the latter can consist of either actual coaxial line sections or microstrip. Regardless of the physical realization of the divider, it will provide two isolated outputs, at each of which one-half of the input power is available.

Fig. 2 shows a Wilkinson power divider circuit using lumped components that is useful to several hundred megahertz. The output R_o is the impedance of

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each load (R_L) and that of the load impedance which the divider presents to the input source. The values of the components which make up the divider are calculated by using the following relationships:

$$C1 = \frac{1}{2\pi f R_o}$$

$$C2 = \frac{C1}{2}$$

$$L1 = L2 = \frac{R_o}{2\pi f}$$

$$R1 = 2R_o$$

where f is the operating frequency in megahertz
C is in microfarads
L is in microhenries.

For example, a 50-ohm system at 116 MHz, C1 is 27.4 pF, C2 is 13.7 pF, L1 and L2 are 68.6 nH, and R1 is 100 ohms. Five per cent mica capacitors of 27 and 13 pF may be used (although bridging to closer tolerance is desirable), and the coils may be wound and measured using the technique described earlier.

Normal vhf wiring practices of using short, direct leads should be followed. Space the two output connectors so that C2 and R1 may be connected between them with minimum lead length. There should be no mutual coupling between L1 and L2; at higher frequencies a shield between the coils may be necessary.

Between about 300 and 1000 MHz, the values of discrete components become too small to be realizable, especially the inductances, which tend to approach transmission lines. Therefore a coaxial transmission line version of the Wilkinson divider is preferable, although it may be used at any lower frequency, limited only by the amount of space physically required by the line sections. This configuration is shown in **fig.** 3. As in the lumped-constant version, $R1 = 2R_o$. The two coaxial transmission-line sections, $T1$ and $T2$, are identical and are one-quarter

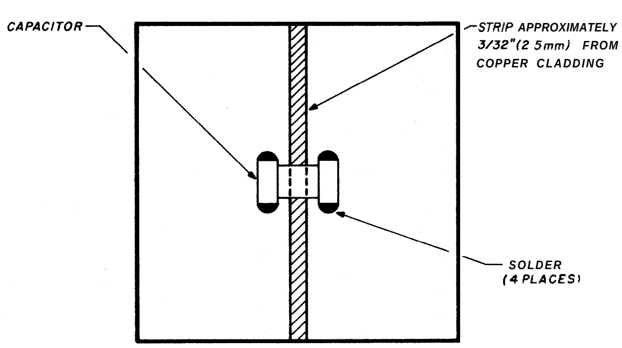

fig. 1. Leadless mica capacitor soldered onto copper-clad circuit board material minimizes series lead inductance.

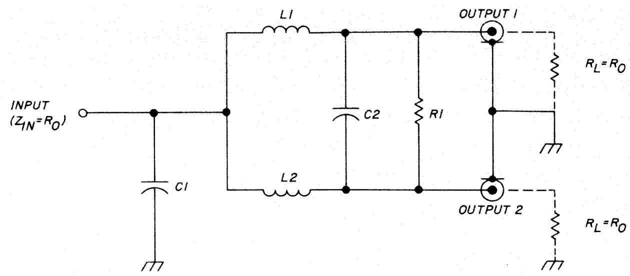

fig. 2. Wilkinson power divider using discrete components is useful below 300 MHz. Component values are discussed in the text.

wavelength long at the frequency of interest. The characteristic impedance, Z_t, of each line section is determined from the expression:

$$Z_t = 1.414R_o$$

and the electrical length, d_t, from

$$d_t = \frac{2951\, V_p}{f} \text{ (inches) or } d_t = \frac{7495\, V_p}{f} \text{ (cm)}$$

where f is the frequency in megahertz and V_p is the velocity factor of the cable.* For a 50-ohm system, the calculated value of Z_t is 70.7 ohms. Practically, any low-loss cable having a characteristic impedance between 70 and 75 ohms will prove satisfactory.

It is essential to ground the coax shields at both the input and output ends. If BNC connectors are used, they should be UG-260/U flange types rather than the UG-1094/U threaded variety. This will permit the transmission-line sections to lie closer to the connector mounting surface and reduce the length of the ground lead at each end. In addition, by closely spacing the flanges of the two output connectors, the spacing between the center pins will be such that a half-watt resistor can be placed directly on the mounting surface ground plane and soldered to the connector pins with minimum lead length.

At frequencies above 1 GHz, the circuit of **fig.** 3 is easier to construct using microstripline instead of coaxial cable. The length and width of the transmission-line section will depend on the dielectric material and thickness in the copper-clad board. G-10 glass-epoxy board is is adequate below about 1.3 GHz, but becomes excessively lossy at higher frequencies, where Teflon or Duroid should be used. Reference 1 contains the necessary information to determine the transmisssion-line dimensions when microstrip on G-10 board is to be used.

Both the discrete component and coaxial line versions of the Wilkinson power divider have been built for use at frequencies between 100 and 408 MHz. In

*V_p is 0.66 for solid polyethylene dielectric, 0.78 for foam polyethylene, and 0.695 for Teflon.

each case the two outputs were balanced within 0.2 dB, and the isolation between output ports exceeded 20 dB. The VSWR looking into any port with the others terminated was 1.1:1 or less.

terminating double-balanced mixers

The use of the commercially packaged double-balanced mixers, such as those manufactured by Anzac, Mini-Circuits, Vari-L and others, has become commonplace in Amateur design. Their convenience and relatively good performance, insofar as intermodulation products are concerned, make them useful in both transmitting and receiving upconverters.

Without going into the details of intermodulation distortion, which has been covered extensively in published literature,[2,3] it is sufficient to state that the intermod specifications of double-balanced mixers can be attained only if at least two of the three mixer ports are terminated in 50 ohms at *all* frequencies. If

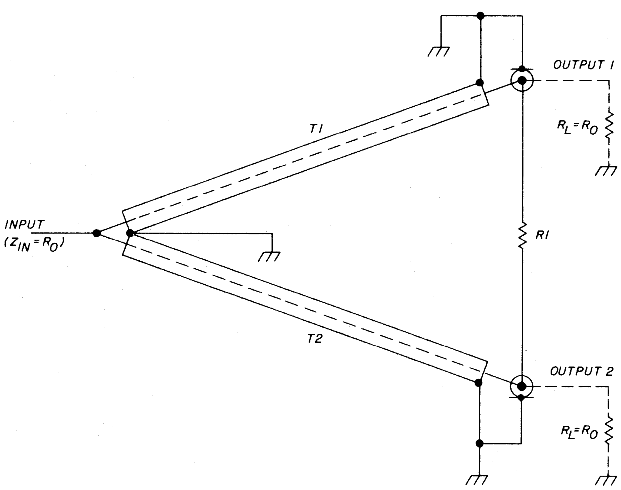

fig. 3. Transmission-line version of the Wilkinson power divider. Parameters of the coax sections are covered in the text.

only two ports are properly terminated, one of these must be the local oscillator input. Terminating the LO port is the easiest requirement to implement, since it entails only inserting a loss pad between the oscillator and mixer. A 2 or 3 dB loss pad is generally sufficient to provide a match, since the oscillator output impedance will normally have been optimized for maximum output into 50 ohms, thereby making its source impedance very close to that value.

Of the two remaining ports, the i-f output is the easier to match because it generally feeds an invariable, narrow-band load. Many techniques have been devised to ensure that the mixer i-f output sees 50 ohms at both the signal and the image frequencies, but most of them require two or more tuned circuits and 50-ohm terminating resistors. Another approach, suggested by Alan Podel (who designed many of the Anzac mixers), is to use a properly biased common-base amplifier following the mixer. This not only terminates the mixer output for all frequencies, but changes the gain of the mixer block from between – 6 and – 9 dB to one as high as + 13 dB.

Fig. 4 shows a double-balanced mixer directly coupled to a grounded-base NPN amplifier. (Note that a loss pad is connected between the local oscillator and the mixer LO input.) The input resistance of a grounded-base amplifier is approximately $28/I_c$, where I_c is the collector current in milliamperes. Therefore if the collector current is set to 0.56 milliamps with the 5 kilohm potentiometer, the mixer will see about 50 ohms at all frequencies. The desired intermediate frequency is selected by means of the collector tuned circuit, *L1* and *C1*. A regulated 12-volt supply is necessary to maintain a constant base current, and the 82.5k and 21.5k resistors should be metal film for temperature stability. The ubiquitous 1N914 ***idiot*** diode is also included to save the transistor in case of cockpit error (reversing the power supply polarity).

The type of transistor used at Q1 depends on the intermediate frequency and the desired gain. The gain-bandwidth factor (f_T) of the transistor should be eight to ten times the intermediate frequency for maximum stable gain. Transistors having the highest dc current gain (h_{FE}) at the lowest collector current will also yield maximum gain. In the circuit shown, the overall conversion gains from 432 to 28 MHz for several types of transistors, used with the same mixer, are tabulated below.

Q1	gain (dB)
2N708	10.5
2N3646	12.5
2N3692	2.5
2N3693	8.0

The collector current may be monitored by inserting a low-range milliammeter between the 12-volt supply and the mixer-amplifier unit. Set the wiper arm of the 5000-ohm potentiometer to the ground end, which will cut off the transistor. The meter will thus read only the bleeder current drawn by the 21.5k resistor in series with the pot. Then adjust the pot so that the meter indicates the bleeder current plus 0.56 mA. (The base current will also be measured, but it is negligible compared with the bleeder and collector current.)

It should be noted that **fig. 4** is not an error; the rf input signal is applied to the mixer i-f port, and the i-f output is taken from the rf port. This was done to provide a direct ground return for the emitter of Q1.

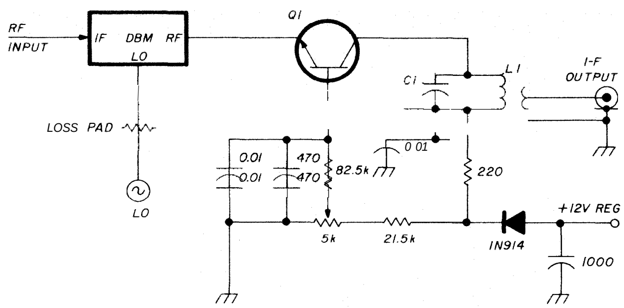

fig. 4. Schematic diagram of a double-balanced mixer with its LO and output ports terminated. L1 and C1 are resonant at the intermediate frequency; the text covers selection of the type of transistor. The mixer i-f and rf ports are reversed in this application.

As shown in **fig. 5,** the dc path from the rf port is directly through the transformer winding, but passes through the diodes from the i-f port. It can also be seen that an rf signal applied to either the i-f or the rf port will reach the same points in the diode quad with the same phase relationship. Therefore, except for the fact that the i-f port is usable down to dc, the two ports are interchangeable. As long as the desired intermediate frequency is above the minimum usable frequency of the rf port, typically 0.5 to 5 MHz for 500 MHz mixers, interchanging the ports has no effect.

VSWR measurements below 450 MHz

Measurement of VSWR using conventional slotted lines, such as those manufactured by Hewlett-Packard, GenRad, and General Microwave, is generally restricted to frequencies above 400 MHz because of the line length limitation. However, another technique is available which does not utilize anything more complicated than a signal generator, an SWR indicator (Hewlett-Packard model 415B or equivalent), and a resistive VSWR bridge. The use and applications of the SWR indicator have been published previously.[4] The resistive VSWR bridge, which has been used by many vhf experimenters for several years, was described by Joe Reisert in his article on antenna matching.[5] Although that article covered use of the bridge only to achieve minimum VSWR for antenna matching, the bridge can be used to measure the VSWR of any device from 3 to 450 MHz with a reasonable degree of accuracy, and does not require any sophisticated test equipment for calibration.

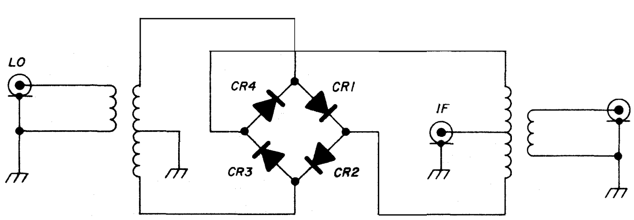

fig. 5. Schematic diagram of a typical commercial double-balanced mixer.

A few words are in order about the bridge, which is shown schematically in **fig. 6.** For best performance, resistors R1 and R2 must be matched to within one per cent, be virtually leadless, and should rest on a ground plane. The ground plane can be made of copper-clad board which has been notched to clear the connectors, and should be fastened in the enclosure just below the center pins of the four connectors.

Fig. 7 shows the test setup for both calibrating and using the bridge. The signal generator must be modulated at 1000 Hz, and must have sufficient output power to enable the SWR indicator to register 0 dB on its 30-dB range with J3 open-circuited or shorted. As large a pad as possible should be used between the signal generator and the bridge to keep the load presented to the generator as nearly constant as possible. If necessary, the sensitivity of the bridge may be increased by removing R3, which serves no useful purpose in this application nor in the one described in the original article.

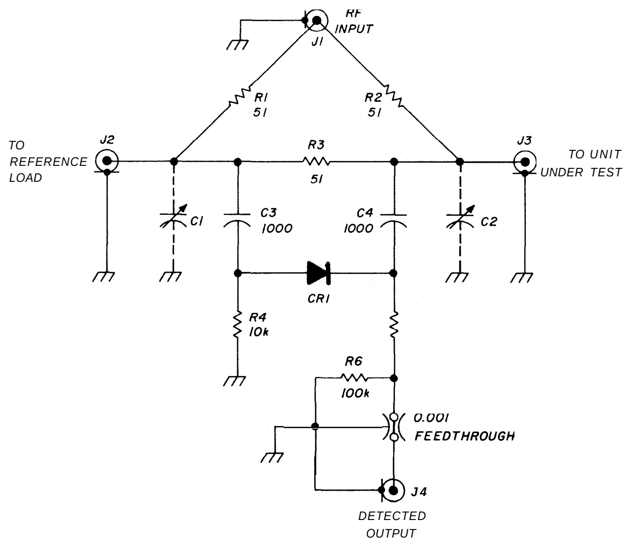

fig. 6. Schematic diagram of the VSWR bridge described in reference 5. Resistors R1 and R2 must be matched within one per cent. Either C1 or C2 (not both) is used to compensate for capacitive unbalance. CR1 is a 1N82 or equivalent germanium diode.

Before calibrating the bridge, the capacitive tab (C1 or C2) must be adjusted to yield a ***return loss*** of at least 40 dB at 450 MHz or at the highest frequency within your signal generator range, if less than 450 MHz. (I use the term "return loss" to indicate the relative balance of the bridge, as displayed on the dB scale of the SWR indicator meter, because it correstable 1. Calculated VSWR of open or shorted 50-ohm resistive attenuators.

attenuation (dB)	VSWR	attenuation (dB)	VSWR
1	8.7	11	1.17
2	4.4	12	1.13
3	3.0	13	1.11
4	2.3	14	1.08
5	1.92	15	1.07
6	1.67	16	1.05
7	1.50	17	1.04
8	1.38	18	1.03
9	1.29	19	1.025
10	1.22	20	1.02
		23	1.01
		30	1.002

ponds to return loss in a directional coupler, although the terminology is not absolutely correct in regards to an unbalanced bridge.) After the 0 dB reference level has been set on the 30 dB range, connect a 50-ohm load of known accuracy to J3. Then probe, with a small strip of metal touching the ground plane, the area adjacent to the pin of J2 or J3, to determine if C1 or C2 must be added. Solder a copper or brass tab at that point and bend the tab for minimum indication on the meter. If your load has a known VSWR of 1.1, the meter should read at least 25 dB below the reference level; if the VSWR is 1.05, the reading should be down at least 35 dB. A really good 50-ohm termination will result in a return loss of more than 40 dB.

To calibrate the SWR meter dB scale in terms of VSWR obtained with the bridge, a set of known mismatches is required. Since such mismatches are not readily available, unterminated or shorted coaxial resistive attenuators can be used instead. **Table 1** lists the calculated VSWR for open or shorted attenuators of standard values. A shorted attenuator is preferable, because the short eliminates fringing in an open connector, but either is sufficiently accurate in this application.

Because the bridge is an imperfect device, it is somewhat frequency sensitive. **Fig. 8** shows the theoretical curve of VSWR plotted against return loss in dB, as well as a set of curves taken on my bridge at various frequencies, using shorted attenuators as mismatches. Note that at low values of VSWR the errors are minimal, but increase as the VSWR increases. Fortunately, this is what we would have hoped for.

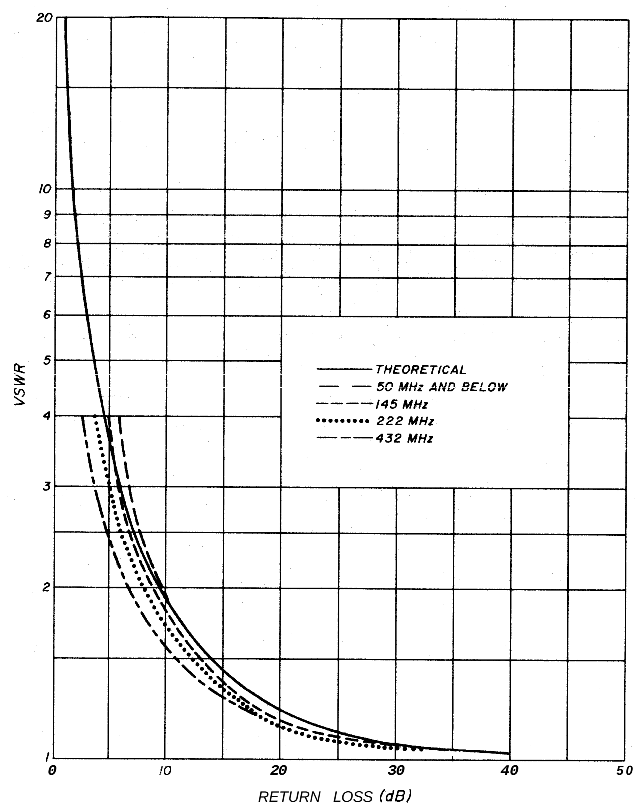

fig. 8. VSWR plotted against return loss for the VSWR bridge at several frequencies. Return loss is read on the SWR indicator shown in fig. 7.

To make your own calibration curves, establish the 0 dB reference level with J3 open, and determine the return loss for each known mismatch. It is also important to record the signal generator output and modulation percentage used for each calibration curve. When actually using the bridge, the same output and modulation level should be used. Otherwise an additional error may be introduced because the bridge diode is not in its square-law region at the high signal level required to set the 0 dB reference.

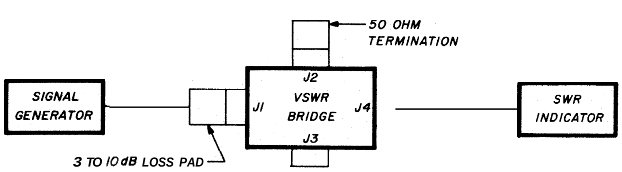

fig. 7. Test setup for calibrating and using the VSWR bridge. Known mismatches for calibration, or the device under test in actual VSWR measurements, are connected to J3.

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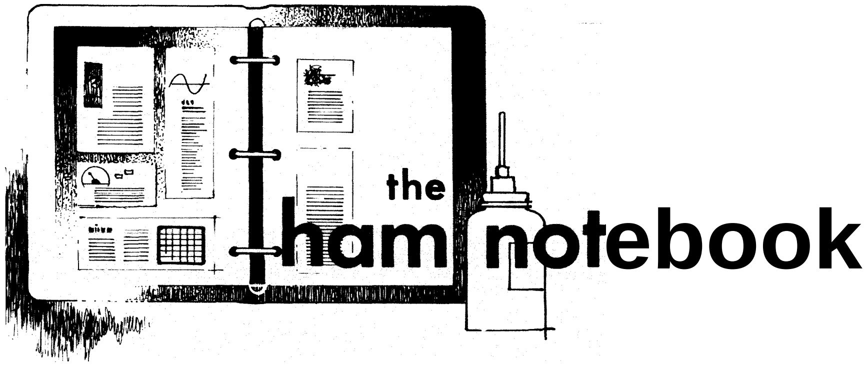

a method for measuring inductance or capacitance

Given a frequency counter and a calculator with buttons for square, reciprocal, and pi, there's an easy way to measure inductance in terms of a known capacitance, or a capacitance in terms of a known inductance. **Fig. 1** shows a complete setup for this

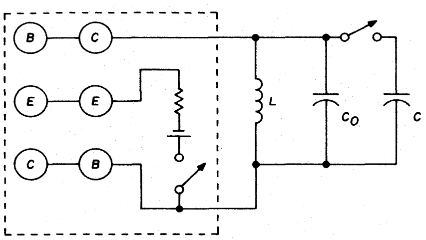

fig. 1. Arrangement for measuring inductance or capacitance in terms of known values. E, *B*, *C* refer to emitter. base and collector of a transistor pair. Plug equation into a calculator to determine the unknowns. Frequency is measured with a counter.

purpose. Inside the dotted enclosure is any sort of arrangement that will cause an inductance with parallel capacitance to oscillate. Other arrangements may well be better for the purpose, but I like the one shown because of its simplicity. Mounted in a small box, it is a handy gadget to have around as it will make almost anything oscillate. The letters **E,** B, C, refer to the emitter, base, and collector leads of a pair of transistors; those on the left are for one transistor, and those on the right for the other. The resistor value is far from critical. Anything over 1500 ohms will ensure that the battery drain will never exceed 1 mA.

Whatever the arrangement used inside the dotted enclosure, the procedure is simply to measure the frequency, f_0, with the switch open and f with the switch closed. Then use the calculator and the equation to determine L if C is known, or C if L is known. In the equation, frequency is in megahertz, inductance in millihenries, and capacitance in picofarads.

The essential feature of the method is that neither the value of C_0 nor the capacitance of the oscillation-generating device enters into the equation. So long as the latter does not change when the switch is operated, the results should be quite accurate.

Of course a little common sense should be used. Don't make C_0 too iarge compared with C, or the frequency change will be too small for accurate measurement. On the other hand, if C_0 is too small compared with C, the frequencies will be so different that the effect of the device may change as the switch is operated. With these extremes avoided, the measurement should be independent of the value chosen for C_0. Also look out for the effect of hand capacitance and oscillator drift. Open and close the switch several times to make sure the frequency readings repeat.

The method would hardly be practical without the counter and the calculator, which may be why I've never seen it described before.

Walter van B. Roberts, W2CHO/K4EA

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audio-driven DSB generators

One feature of a properly adjusted class-C amplifier is the linearity of its output power with respect to its input

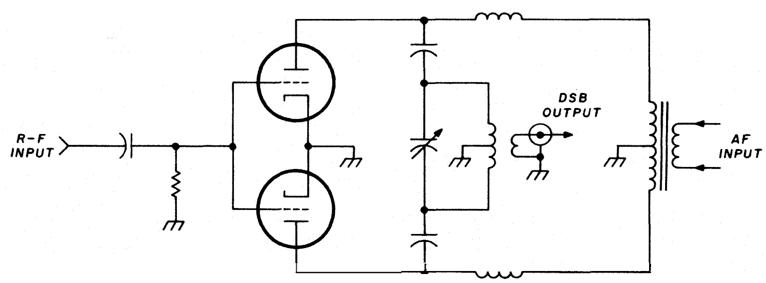


fig. 2. Suggested typical audio-driven DSB generator.

power. It should be applicable to many requirements if all the input power to the final stage of a DSB transmitter is provided by a high-power audio system, as suggested in **fig. 2.** Simple isn't it? No crystal filters, no phasing networks, and practically no standby power consumption. Great for portable operation!

So, why should one even consider DSB when every ham knows that SSB is better and uses less space in the rf spectrum? But wait a minute! Everything has its application — SSB for the crowded hf bands, of course; but on the higher-frequency bands there's a lot more room. And DSB can be received with phase-lock detection, which will provide AFC action — ideal for the probable channelized operation. When (and if) CB surfaces on frequencies in the hundreds of MHz, it will almost certainly be of a channelized nature, and AFC will probably be a requirement.

There are at least three different methods of receiving DSB. One is to use sharp filtering, which eliminates one sideband, and then treats the remaining signal as though it were SSB. Another involves phase-reversal techniques. When polarity reversals of the modulating frequency produce phase reversals of the rf signal, a corresponding phase reversal in the detection process will compensate, resulting in normal signals.

The third, and somewhat simpler method, is to phase lock an oscillator to the sideband signal on twice the i-f of the missing carrier. The oscillator signal can be divided by two with a simple diode circuit, and there's your carrier for re-insertion. This last method also results in the AFC effect previously mentioned.

Perhaps with the approach of more and more channelized operation as the higher-frequency bands are exploited, it would behoove us to investigate this little-used mode of transmission.

If you must use SSB, however, a power-efficient transmitter can be made by combining two of these DSB generators with appropriate phasing. Methods of doing this have been amply covered in the literature during the last twenty-five years and are not repeated here.

There is, however, an interesting angle I've not seen and which could have applications as a high-power emergency CW transmitter. The joker is that two-phase power would be needed, but an arrangement similar to **fig.** 3 should provide a respectable dc note when the input to the final amplifier is ac, direct from the mains! Saving the cost and weight of the rectifier and filter system just could be the deciding design consideration for some applications. It's worth thinking about.

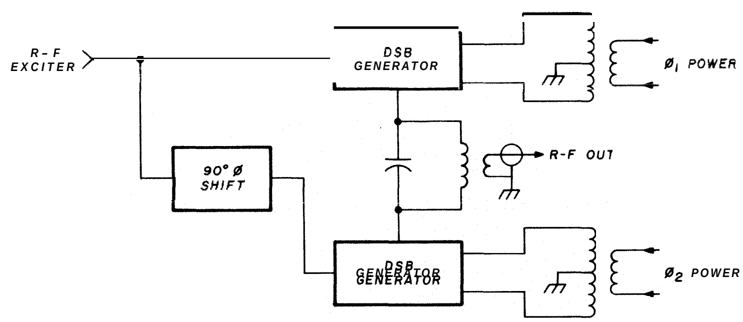


fig. 3. Possible transmitter without dc power.

Henry S. Keen, **W5TRS**

tuning aid for crystal-controlled vhf receivers

Aligning a crystal-controlled 2-meter or commercial high-band receiver using a VFO-based signal generator is tedious and exasperating because of generator drift and the difficulty of setting the exact frequency. For occasional use, the obvious alternative, a synthesized generator, is usually prohibitively expensive.

If one is available, a Regency The Touch synthesized scanner (Model ACT-T-16K) serves as a good substitute. The local oscillator in The Touch is 10.7 MHz below the programmed receive frequency on two meters and high band, and incidental radiation from the LO is audible at a distance of 30-40 feet (9-12 meters) on a tuned handheld with a rubber duck antenna. Thus, for a signal at 146.52 MHz, program in 146.52 + 10.7 or 157.22 MHz.

The Touch's priority feature provides an added bonus: If a nonpriority channel is used and the priority feature enabled in the manual mode, the periodic priority channel check gives a switched carrier with a distinct "chugging" sound that's easy to hear and tune to.

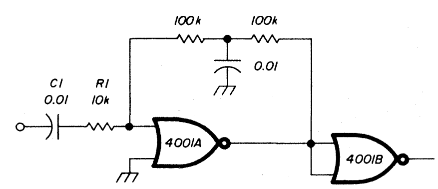
Without alteration of The Touch coupling is by radiated signal, and attenuation is provided by changing the distance. People in the area will affect received signal strength, so stand still while tuning.

The signal provided by The Touch this way will be delightfully stable and accurate enough for tuning. It will not usually provide a perfect frequency adjustment, but it will normally bring the receiver close enough to hear the repeater or base station for a final frequency tweak from the actual source.

David **McLanahan, WA1FHB**

no-adjust bias for VLF dip meter

When building the VLF dip meter converter described in the August, 1979, *ham radio,* it is possible to replace R2 with a fixed RC circuit. This modification, shown in **fig. 4,** provides no-adjust bias without sacrificing gain. In **fig. 4** the node between 4001A and 4001B automatically adjusts to the switch threshold of the first gate. Inherent matching in the chip means it also will be very close to the second gate's switch point.



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Observations & Comments

Living things change and *ham radio* is no exception. Since the passing of Editor-In-Chief Jim Fisk, I have been asked to take over his editorial page, at least during this difficult transition period. His is not an easy act to follow. Jim was close to the pulse of Amateur Radio, its problems, and the direction in which it's going. Jim did a superb job, and certainly all active Amateurs benefited by Jim's "second look."

My job is to try to carry on the precedent set by Jim in illuminating issues that affect Amateur operating, technology advances, and the future. I'm not as close to the immediate issues of Amateur Radio as Jim was, but that's going to change. I'll have to educate myself so that I can carry on *ham radio's* editorial page in the established tradition. I ask the support of readers in bearing with me.

I'd like to introduce in *observations and comments* some contributions by Amateurs who have something constructive to say. This material will reflect *your* ideas, problems, and what to do about them. If your contribution is positive, it could end up as a guest editorial. Here are some ideas:

Much has been published on FCC proposed rule making affecting Amateurs. If such PRM would have an impact on your sphere of interest, we'd like to hear about it. Give us the pros and cons from your point of view. If your contribution is in the best interest of Amateur Radio, we'll print it in *observations and comments.*

Do you have a club? We'd like to know about the problems you may have encountered in running a ham club so that others may benefit. What about a club paper? How do you run yours?

Say you're a contest operator. What can be done about the selfish attitudes of those who interrupt contest operation?

You've run across a new adaptation of current IC technology. Let's hear about it. Can it be adapted to Amateur Radio?

You don't like the restrictions on satellite communications. Why not? Do you have a better solution?

What about slow-scan TV and interference by SSB operation? The upper end of the 20-meter band is a good example. Is time-sharing the answer?

These are just a few ideas that come to mind. Others are welcome from our readers and will certainly be considered.

The object of this column is to present ideas and comments from our readers that will provide a positive thrust forward for other Amateurs. Reader contributions will be supplemented by editorial comment on current issues and their relationship to Amateur Radio. The idea behind this column is to present an image of what our readers think. Let's hear from you.

Alf Wilson, W6NIF
technical editor

comments

surplus tubes

Dear HR:

The comments by KB5EY and W6SAI in the December issue were very interesting. I, too, have a collection of 813 tubes and presently use one on 75 meters. I am building a near-kW transmitter which will use two in a class-C final.

But let me say something about those surplus, commercial and military, tantalum plate triodes, like the 100-TH and the 250-TH. They were advertised by Eimac many years ago as "gas-free." They can be bought (without warranty) for one or two dollars each at club auctions and flea markets. This is definitely not a "wad" of money. Interestingly, forty years after manufacture these tubes are still "gas-free." If not abused, and such abuse is readily discerned visually, they will perform as per specifications. I am now using a pair of Heintz and Kaufman HK-24 tubes that were made in 1939 in a class-C final on ten meter FM! No TVI, either.

My present collection includes such gems as the 35-T, 100-TH, 250-TH, and 450-TH — with spares. With these tubes 70 per cent (or more) efficient single-band class-C finals are planned. What old timer can deny the romance of ham radio that comes from a room softly illuminated with the glow from those tungsten filaments, and the nearly-white hot plates of a pair of (quite) fully loaded tantalum plate triodes?

Byron H. Kretzrnan, W2JTP
Huntington, New York

speech processors

Dear HR:

I read with interest the letter by Walter Schreuer of Maximilian Associates and the reply by Wes Stewart concerning his split-band speech processor. Since I have done some comparative on-the-air testing of the two designs, I think that my findings will be of interest.

Fcr my tests, I used three different speech processors: a Vomax, a N7WS split-band, and a quasi-logarithmic audio clipper. All three units were connected to a switching system that allowed instantaneous switching of the various units between the transmitter and microphone. The units were adjusted to provide the transmitter with an equal amount of drive measured by observing the metered ALC level.

Various tests were run with some of the local fellows on 10 meters as well as with DX stations on several bands. At the beginning of the contact, I explained the test I was about to run and asked the operator to note his preference, which unit he thought sounded the best, as well as which provided the most signal "punch." First, a transmission was made with no audio processing to be used by the receiving station as a reference. The three speech processing units were designated **A,** B, and **C,** and the stations were not told which was which until the test sequence was completed.

A test run was made by using unit **A,** then unit B, then unit **C;** in between testing the units I switched back to a "No Processing" mode for comparison. By keeping a record of these tests, I found that approximately 90 per cent of the stations preferred the N7WS design split-band processor. Operators who preferred the N7WS design said that although there was as much — or slightly more — than average power output with the Vomax, the N7WS design sounded "crisper." Monitoring my signal with a separate receiver, it sounded to me as though the Vomax suffered from a highiy restricted audio bandwidth, with the most notable point being the lack of low frequencies.

I've heard others on the air using the Vomax, and it seems that, while some operators sound excellent, others suffer from the same problem I observed during my tests. This leads me to conclude that the microphone audio response and/or timbre of the operator's voice will determine the audio quality when using the Vomax. (I should point out that only one Vomax unit was available for these tests; I would have liked to have been able to obtain another unit to see if it had the same characteristics.) The quasi-logarithmic processor, while contributing punch to the signal, did not fare well during these tests when compared with the other units because of its high percentage of distortion.

I make no claim that these tests were scientific or definitive; they were conducted only to satisfy my personal curiosity.

Gale A. Steward, K3ND
Quakertown, Pennsylvania

speed of light

Dear HR:

I would like to comment on the paper published in January, 1980 issue by Harold Tolles, W7ITB, regarding the speed of light. The measurement of the speed of light is a difficult task to accomplish, at best, to the precision quoted today. It should be noted that the speed of light has been given to at least six places since before 1930, i.e., 2.99796 x 10^8 meters per second (m/s). This value has, so to speak, converged to 2.997924580 x 10^8 m/s as recommended by the Committee on Data for Science and Technology, and the international Council of Scientific Unions (CODATA-ICSU) in 1973. The principal improvement in the knowledge of the speed of light has been the reduction in the measurement uncertainty from 4000 m/s in 1929, to 1000 m/s in 1951, to 100 m/s in 1963, to 1.2 m/s in 1973. Surely four parts per billion is adequate for ham radio work!

It should be noted that these speeds are in vacuum. When light propagates through any material (e.g., air), its speed is reduced by a factor of one over the material's index of refraction. The index of refraction of air, for example, is dependent on the temperature, pressure, and frequency of the EM radiation on a "point-by-point" basis (for most rf work, electron density distribution in the atmosphere plays a significant role), from which one can (reasonably) infer that the speed of an EM wave varies over the path of propagation. Finally, I suggest that we and Mr. Tolles not despair about *c* and simply use 3 x 10^8 m/s, which is good to 0.07 per cent; this value is adequate for all Amateur Radio requirements.

R. Barry Johnson, W4MLM
Rancho La Costa, California

receiver dynamic range

Dear HR:

WB6CTW is to be complimented for his fine article on measuring receiver dynamic range in the November, 1979, *ham radio.* His technique makes it possible for the average Amateur to make meaningful measurements with homebrew equipment. WD6FMG, N6ST, and I have been using Hewlett-Packard signal generators to perform similar measurements. Based on our experiences, I would like to offer several comments on this subject.

1. The measurement of dynamic range for either third-order intermodulation (undesired mixing of two interfering signals) or gain compression (overload by one interfering signal) is not particularly sensitive to the exact difference in the frequencies of the signals. Any difference between 20 and 100 kHz can be used. If the signals are too close together, i-f skirt rejection or local oscillator noise sidebands confuse the measurements; if they are too far apart, the rf preselector attenuates one or both of the signals.

2. Various receivers will show somewhat different results if tested on different bands. Just the same, if all of the receivers in **Table 1** of the article were tested on 40 instead of 20 meters, the ranking would probably be similar. Therefore, the cost of the crystals for the two oscillators could be saved by using some Novice band crystals from the junk box.

3. When performing the two-tone test, misleading measurements may be obtained with some receivers if the AGC is allowed to reduce the rf gain. The intermodulation signals should be kept weak enough that the S-meter barely moves. In some receivers, the stage that generates the intermodulation is a mixer or second amplifier stage that follows a stage with AGC. As the level of the two interfering signals is increased to the point that the intermodulation product appears out of the noise, passes through the i-f filter, and reaches the detector, the AGC will reduce the rf gain. Less signal reaches the intermod generating stage, and less intermod is produced than would be the case if the AGC were disabled. AGC, of course, also decreases receiver sensitivity.

The dynamic range of a receiver is the difference between the weakest signals it can detect, and the largest signals it can handle simultaneously. Unfortunately, different values for dynamic range will be measured depending on the definition of a "weak" signal. Some of the companies that publish dynamic range performance seem to use creative specsmanship. In one case, the measured value of dynamic range is much better if the sensitivity and intermodulation are compared for a weak signal of S-[illegible] than for a weak signal of S-1. Can you guess which value is the published value?

4. The article suggests that gain compression can be tested with only one signal. This procedure will often give false and inflated results. It is much better to tune the receiver to a weak signal; then adjust the amplitude of a second signal on a different frequency until reception of the first signal is impaired.

For example, a while ago we checked the performance of a popular synthesized 2-meter handheld transceiver. This rig was claimed to have 80 dB rejection 30 kHz away. The signal level for full quieting was measured. Next the receiver was tuned 30 kHz away. Then the amplitude of the signal generator was increased by approximately 80 dB, at which point the receiver was again fully quieted. From this test, one might suppose that the dynamic range was 80 dB.

The receiver was next tested using two signal generators. The first one was tuned to the receiver frequency, frequency modulated with a 1 kHz tone, and adjusted in amplitude until full quieting was achieved. The second signal generator was tuned 30 kHz away and increased in amplitude until the 1 kHz modulation of the first signal became noisy. The difference in the two signal strengths was only 56 dB . . . a lot less than the 80 dB measured the first way!

5. Based on testing one of each model, the TS-520S is much improved over the TS-520 and even slightly better than the TS-820. Owners or prospective owners of the TS-520S should perform their own measurements before they panic.

Paul A. Zander, AA6PZ
Los Altos, California

presstop

A LAW PROHIBITING "INTERFERENCE by Radio Transmitter" has been enacted by the Township of Winslow, New Jersey, in the aftermath of TVI/RFI problems experienced by a local Amateur, WB2SZK. Just over a year ago a neighbor complained of TVI plus interference to his stereo, electronic organ, and intercom. Stubs cured the TVI problem, but filters were only partially effective on the other equipment. WB2SZK, after urging the neighbor to seek manufacturer's help with the remaining problems, was summoned to court in November for violation of a township nuisance ordinance and ordered to stay off the air for 30 days pending FCC inspection. When he refused, he was fined $250 and costs.

The FCC's Inspection in early December gave WB2SZK a clean bill of health, and the complainants were so advised. On December 19 the township then adopted a new ordinance, Chapter 50-10.2, that makes it unlawful to transmit any radio signal that "...causes or creates electrical, visual or audible interference..." or "...annoys, disturbs or endangers the comfort, repose, health, peace, safety or general well being of others within the township." Any interference with "...receiving sets, musical instruments, phonographs or other machine,.." is included under the ordinance's omnibus coverage.

WB2SZK Was Again Summoned for hearing, this time under the new ordinance, in February. On his request the FCC submitted a 1977 Public Notice citing federal pre-emption of the control of radio transmissions to township officials, but to no avail. His attorney was able to obtain an interim injunction from the New Jersey Superior Court halting prosecution under the new ordinance, but at a May 2 hearing the Superior Court judge upheld the township ordinance on the grounds that there is no specific federal pre-emption of control of radio communications — it's only implied.

Since This Decision Contradicts previous decisions on the pre-emption question, it sets a dangerous precedent and must be challenged. Over $1000 (including $500 from the Mt. Airy Pack Rats) has already been spent, and the necessary appeal in Federal Court will cost much more. Contribution checks made out to Harry B Stein, W3CL, with the notation "Randy Bynum Defense Fund" can go to 2087 Parkdale, Glenside, Pennsylvania 19038.

A NEW COMMUNICATIONS ACT REWRITE bill, S-2827, has been introduced in the Senate by the Senate Communications Subcommittee. Combining the better and less controversial ideas of the previous rewrite proposals (S-611 and S-622), the new bill includes the 10-year license term, authorizes the FCC to require TVI rejection standards for TV receivers, and — most important — gives the FCC authority to delegate license examination authority to nonemployees.

The Full Text Of S-2827 appears in the June 13 issue of the *Congressional Record*. Because of the very pro-Amateur-Radio aspects of this revised bill, Amateurs are again urged to write their Senators and Representatives as well as the Subcommittee Chairman, Sen. Ernest F. Hollings (D., South Carolina), and minority leader, Sen. Barry Goldwater. A complete list of the Senate Committee members who are directly concerned with S-2827 appears in the May, 1980, *QST* editorial. Some Washington observers believe that, with public support, S-2827 has a good chance of being passed by Congress this year.

SSTV AND FACSIMILE WOULD BE permitted on all Amateur voice frequencies above 3775 kHz, under a Notice of Proposed Rule Making adopted by the Commission June 3. Personal Radio Docket 80-252 proposes dropping the present subband restrictions on SSTV, meaning that Generals as well as Advanced and Extra class licensees would be able to use that mode on 80 through 15 meters. Facsimile would be permitted on the same frequencies thus opened to SSTV. 160 was not included in the NPRM at this time, as that band is still shared with Loran, Amateur Radio being a secondary user through 1981. Under the proposed rules change, bandwidth for either mode would be limited to SSB bandwidths below 50 MHz, and to AM bandwidths above.

Comment Due Date is September 22, with Reply Comments due October 22.

FCC'S NEW EXAMS SCORED AN "A" with the first group of applicants who took them recently. According to instructors from several parts of the country, their students rated the new FCC efforts well written, unambiguous, and closely related to the new FCC study guides. They also liked the practical emphasis on operating procedures in the Technician/General-class exams with more technical subjects covered in the higher class tests. Congratulations to Jay Jackson, AF4O, and the other FCC staff members who are responsible for the new exams.

HAM RADIO'S NEW TECHNICAL Editor is Alf Wilson, W6NIF. Alf is no stranger to *ham radio*, having been Jim Fisk's technical right-hand man since the magazine's early days. He will be working on *ham radio* through the summer to help ensure its continued technical accuracy while the search for a permanent editor continues. Prior to his retirement several years ago, Alf was a Technical Publications Specialist for General Dynamics in San Diego.

light-emitting diodes: theory and application

A digest of LEDs — how they work and how they're used in many of today's electronic circuits. Also included is a simple logic probe for testing digital circuits

Learning the theory of the electronic devices we use can be greatly enhanced by putting the theory to work with a weekend construction project. This is just what we have in this article: some basics, a few applications, and a very easy construction project. A few hours invested should yield a good understanding of light-emitting diodes (LEDs) and a unique logic probe that fits well with today's technology.

Initially these LED devices were too expensive for the Amateur or experimenter, but their wide acceptance and availability now make them a workable addition to our hobby. As Amateurs continue their transition from vacuum tubes, through transistors and into integrated circuits, the LED will become more valuable as an indicator, display, or active component.

For practical applications we can think of the LED indicator, or the individual segments of a LED display, as being the same as any other diode; much like the power diodes used as rectifiers, or the small-signal diodes used as detectors. As with any diode, we must operate within the parameters of the device as specified by the manufacturer. Under these conditions the LED will function well.

To illustrate the relationship between a silicon diode and a LED, I've shown both devices under similar conditions in **fig. 1**. Both are forward biased, and the current through each is limited by the series resistor, R1. The voltage drop across the silicon diode, or its threshold voltage, is approximately 0.6 volt. If the value of R1 is changed, the current through the diode will vary, but the voltage across the diode will remain fairly constant. In effect, the diode will act as a voltage regulator. The diode could be used as a shunt regulator in the same manner as a voltage-regulator tube found in many receivers and VFOs of the past. In semiconductor circuits, one or more of these diodes are often used as regulators and clamps, making good use of this threshold action.

The same conditions exist for the LED except that the threshold voltage is higher, usually about 1.6 volts for the red LEDs. As with the silicon diode, varying the series resistor will vary the current through the LED and thus its intensity, but the voltage drop across the LED will remain fairly constant at 1.6 volts.

From **fig. 1** it can be seen that all we must do to activate a LED is apply forward bias and limit the current to a safe value, as specified by the device manufacturer. As shown in **fig. 1**, a simple application of Ohm's law is all that's needed to use the LED as an indicator. An understanding of the basic principles of the diode will enable you to get started in the applications of LEDs.

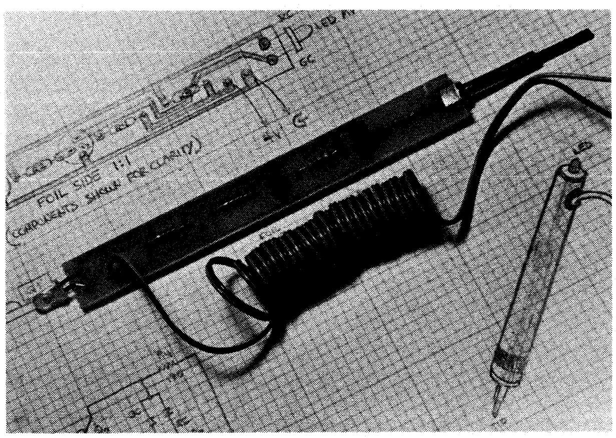
Fig. 2 illustrates the parameters of a typical red LED. As shown, 100 per cent of the rated intensity is achieved with 1.6 volts at 0.020 ampere. These parameters were used to calculate the series-resistor value in **fig. 1**. The threshold voltage for colors other than red vary from this norm. Typical values are, for amber devices, approximately 2.0 volts. For green units threshold approaches 4.0 volt level.

These numbers are adequate approximations for general use and will provide a starting point for unknown surplus units. The amber and green LEDs generally require more current than the red LEDs to achieve equal intensity. However, our eyes are very understanding in this area, and the current through the diode can vary over a formidable range without a profound effect on the indicator's appearance.

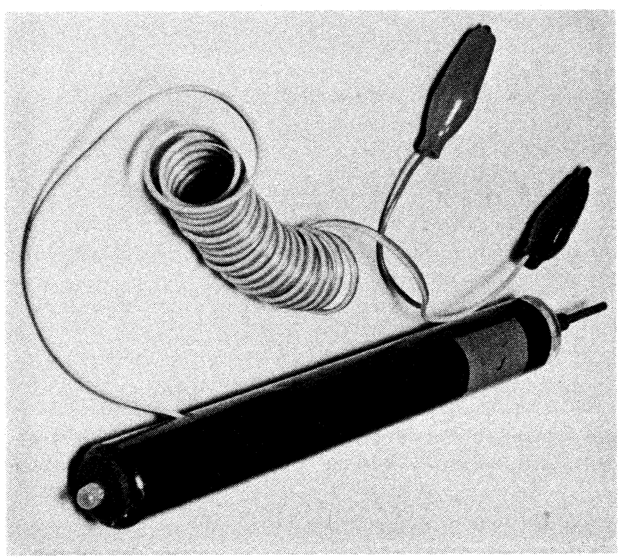
By Ken Powell, WB6AFT, 6949 Lenwood Way, San Jose, California 95120

applications

We tend to think of LEDs as low-voltage devices primarily associated with transistors or integrated circuits. However, with suitable current-limiting resistors, these devices will function equally well in higher-voltage circuits. The type 4403, for example, has an isolation voltage rating between the leads and case of 300 volts. This rating allows the device to serve as an indicator in moderately high voltage circuits.



Complete logic-probe board ready for assembly.



A logic probe for testing digital circuits. A few parts and as many pleasant hours at the workbench produced this useful and handsome piece of test gear.

Voltage monitor. Fig. 3 illustrates this principle by the use of a LED to monitor the power-supply voltage in a tube-type receiver. The calculations required to determine the value of the series resistor is shown;

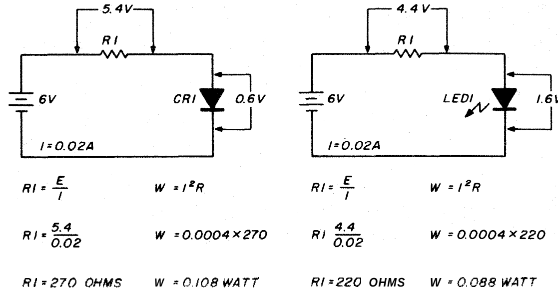


fig. 1. Silicon diode and LED under similar operating conditions: each is forward biased, and current through each device is controlled by series resistor R1. Threshold voltage across the silicon diode is about 0.6 volt, whereas that across the LED **is** somewhat higher, usually about 1.6 volts **(for** red **LEDs).** Simple application of Ohm's law determines device operating parameters.

note that this application is no different than the basic circuit discussed in **fig. 1.** From this application it can be seen that we're not limited to low-voltage or solid-state circuitry in the use of LED devices. With higher voltage applications, the series-resistor value becomes greater along with increased power requirements; but the physical size is still within reason, and the looks are more in keeping with today's technology.

The majority of LED applications encountered are in dc circuits, but these devices aren't limited to dc and make excellent indicators for ac if some precautions are taken in their application. One of the parameters not yet discussed is the reverse-voltage specification. The reverse breakdown voltage of LEDs is

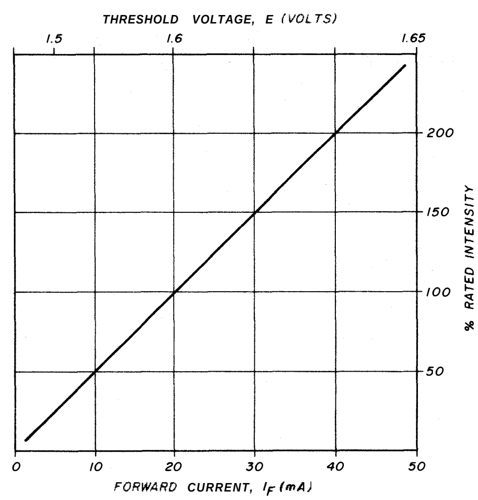


fig. 2. Intensity of a typical red LED as a function of threshold voltage and forward current. One-hundred per cent of rated intensity occurs with **1.6 volts** at 20 mA. These data were used to calculate the series resistor, R1, in fig. *1*.

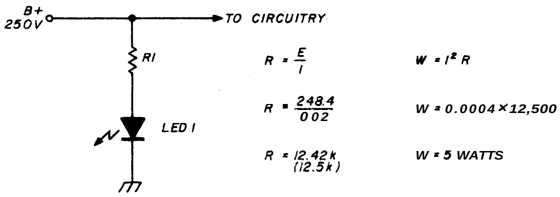


fig. 3. A voltage monitor that can be used, for example, in a power supply of moderate voltage. Calculations for series-resistor values are shown.

generally low, so particular attention must be paid to this specification. The reverse breakdown voltage for the type 4403 is typically 3 volts, meaning that the reverse bias voltage must be kept below this level.

Filament-circuit monitor. In **fig. 4** an LED is used to monitor a filament circuit, and a diode is placed inversely in parallel with the LED to limit the reverse voltage applied to the LED to approximately 0.6 volt, or the threshold of the silicon diode. This diode will protect the LED during negative excursions of the filament voltage. As in the previous circuits, the limiting-resistor value is calculated with Ohm's law as shown in **fig. 4A,** but the power must be calculated as in **fig. 4B,** since the current flow will be greater on the negative half cycles when shunt diode CR1 is conducting. That is because of the difference in LED threshold voltage and that of the silicon diode. In this case, the difference in power is minimal but should be taken into account, particularly with higher source voltages and higher threshold LED devices, such as amber and green units. The reverse-voltage parameter is an important factor and must be kept in mind when thinking about LED applications.

Color transitions. Another type LED, the MV-5491, is illustrated in **fig. 5.** This unit is actually

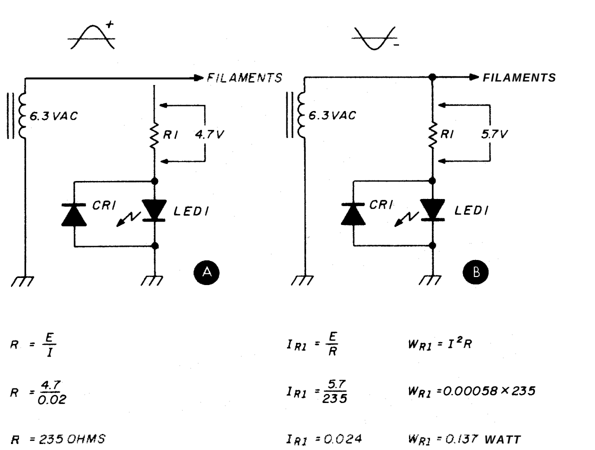


fig. 4. Using an LED to monitor a filament circuit. The limiting-resistor value is calculated as in (A), but the power dissipation must be calculated as shown in (B).

two LED junctions in a single package placed inversely in parallel. One of the parallel diodes is red; the other is green. As with the LEDs discussed earlier, the threshold voltages of the two diodes are different, so a bit more thought is required in their application.

With the MV-5491, the color can be changed by reversing the voltage applied to the diode pair; and with the application of an ac voltage, an alternating color that approximates yellow can be obtained. This device can achieve four states: red, green, amber, and off.

In **fig. 5A** the red diode is conducting and the limiting resistor, calculated for a specification of 1.65 volts at 0.20 ampere, controls the current through

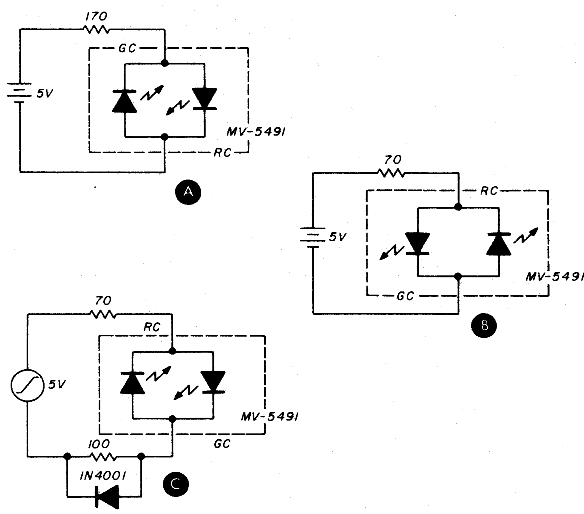


fig. 5. Example of a dual-diode LED, type MV-5491. This device is actually two LED junctions in single package placed inversely in parallel. Sketches (A), (B), and (C) show, respectively, how the LED colors are generated. (A) shows the diode pair in a red configuration; (B) shows green. With the application of an ac voltage and the use of a compensating diode, as seen in sketch (C), an amber color can be generated.

the diode. In **fig. 5B,** the green diode is in conduction, and the limiting resistor has been chosen to provide 3.0 volts at 0.020 ampere. Because of the difference in the specifications of the red and green diodes, external components must be used to provide compensation. **Fig. 5C** illustrates this compensation in the form of a silicon diode that will shunt the 100-ohm resistor when the green LED is conducting. Reversing the polarity of the input voltage will reverse bias the silicon diode, placing an effective 170-ohm resistance in series with the red LED. In this manner the correct voltage and current can be furnished to the dissimilar LED junction.

LED drivers. So far in the discussion all the LED applications have been of the static type. To make these units dynamic, or to turn them on and off with signals, a switching device must be added in series with the LED to control the current flow. In **fig. 6** a transistor switches the LED on and off with a signal or logic level to be monitored. In **fig. 6A** the voltage across each element of the circuit is shown with the transistor in conduction and the LED indicator lighted. The calculation for the current-limiting resistor is the same as in previous illustrations except for the added voltage drop of the transistor, usually on the order of 0.2 volt. As in previous illustrations, 1.6 volts appears across the LED and 3.2 volts across the limiting resistor. In **fig. 6B** the circuit is shown with the transistor cut off and the LED indicator extinguished. In **fig. 6C** a similar circuit is depicted for use with a negative power supply.

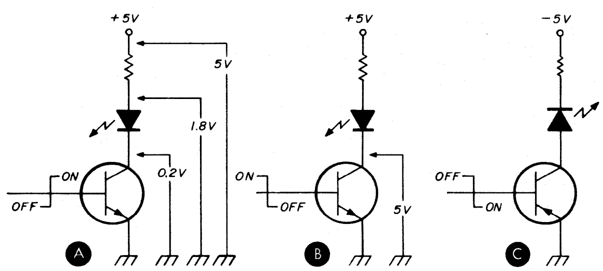


fig. 6. Dynamic application of LEDs. A transistor controls current flow; that is, a switching device is used. (A) shows the transistor conducting (LED illuminated); (B) shows the transistor cut off (LED extinguished). In (C) a circuit is shown for use with a negative power supply. In these applications the device is called an "LED driver."

Integrated circuits lend themselves well to driving LED devices, and six LEDs can be controlled from a single IC package. The SN7406 and SN7407 are well suited for this application. The 7406 is a hex-inverter with each of its output circuits rated at 0.040 ampere and 30 volts. As shown in **fig. 7,** the 7406 will cause the LED to conduct when conditioned with a high, or positive, input. The 7407 is the same basic package but is a noninverting circuit, so a high input will yield a non-conducting, or high output, extinguishing the LED.

In applications requiring more than one LED indicator, the 7406 and 7407 form a very compact and cost-effective circuit. They are useful for adding monitors to keyboards and for data bus applications. The calculations for the current-limiting resistors in this application are identical to those discussed earlier. Just about any open-collector TTL IC will work well as an indicator driver; and for practical applications, the LED current can be limited to 0.010 ampere to

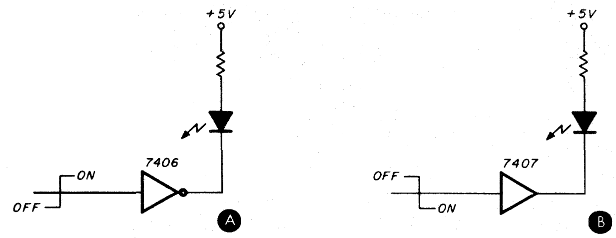


fig. 7. In this example of LED drivers using ICs, as many as six LEDs can be controlled from a single IC package. (A) and (B) show, respectively, drivers with inverting and noninverting inputs.

reduce current use. This will yield adequate light output in virtually all situations.

Driving a dual LED, such as the MV-5491, is a bit more complex. But the result of the LED changing color with changes in the signal is more dramatic and can be accomplished with ICs. **Fig. 8** shows a driver circuit for dual LEDs using one mini-dip, type SN75452, and one section of a SN7404 inverter.

The calculations for resistor values are the same as those in previous illustrations, but the power requirements are a bit more. With the driver input low, the input to IC1A is high and its output is low. In this state 4.8 volts will be dissipated across the 220-ohm resistor, placing the green cathode (GC) at 0.2 volt. IC1B, with its low input, will have a high or nonconducting output. This action will allow the green diode to be forward biased, and 1.8 volts will be dissipated across the 100-ohm resistor. When the circuit input goes high, IC1A and IC1B outputs will change state, and 4.8 volts will be dissipated across the 100-ohm resistor. The red diode will go into conduction, and 3.15 volts will be dissipated across the 220-ohm resistor. Rapid transitions of the input signal will alternate the red and green LEDs and form a somewhat amber indication.

The dual-LED indicator makes a very nice display. The ability to display a number of states with a single

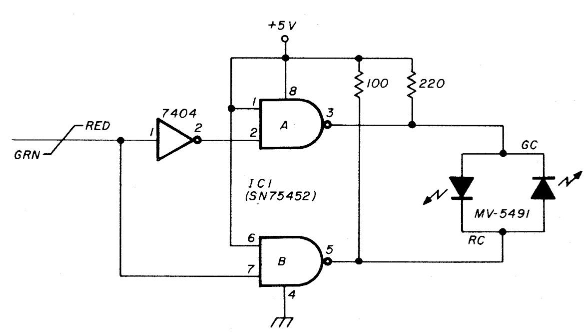


fig. 8. A dual LED driver using the type SN75452 IC. Calculations for the resistor values are the same, but the power requirements are a bit higher.

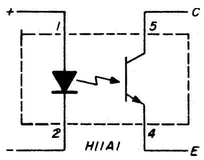


fig. 9. Application of the LED as an optical coupler. In this case the LED has been packaged with its light output focused onto a photo transistor. In this configuration the device can provide a signal path while providing electrical isolation of more than 2500 volts.

indicator is very useful, particularly in digital applications such as circuit monitoring. The IC driver circuitry could be replaced with discrete transistors if desired, and no doubt many applications for this unique device will be implemented by Amateurs and experimenters. As with many other devices, these units were expensive when first developed but are readily available at low cost today. The MV-5491 is packaged in the standard T-1% package and uses the same mounting hardware as most LEDs.

the optical coupler

All of the LEDs previously discussed have been indicator types, so now is a good time to take a break and look at another LED device known as the optical coupler. The coupler uses the LED for a much different purpose – that of isolation. This device can provide a signal path while furnishing electrical isolation in excess of 2500 volts. Again it is the basic LED with virtually the same parameters as those of the LED indicator. However, in the coupler configuration the LED has been packaged with its light output focused on the sensitive surface of a photo transistor **(fig. 9)**. In this manner, the current flow through the LED will provide base bias for the transistor through an optical path within the package, providing signal coupling while maintaining physical isolation.

While I've not seen these devices used extensively in Amateur gear, I have used them in keyer circuits to isolate the keyer from the transmitter. They are a very effective device in this application and remove the grid-block keying voltage, which is sometimes at a fairly high potential, from the paddles. This application is depicted in **fig. 10.** The optical coupler, or opto-isolator as it is also called, is used extensively in medical and data-processing equipment. I think that, as Amateurs become more aware of the unique properties of this device, many new and worthwhile applications will ensue.

The coupler in **fig. 10,** type H11D1, is packaged in a six-pin mini-dip, so it lends itself well to today's construction techniques. Keep this device in mind for both safety and noise reduction applications.

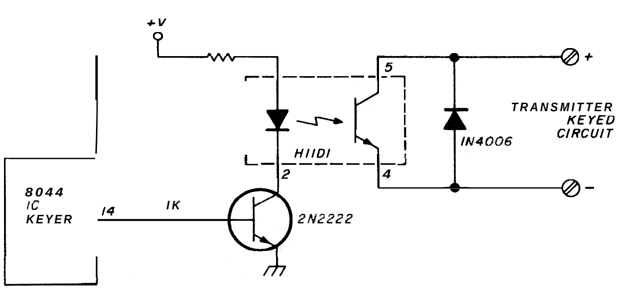


fig. 10. Use of the LED optical coupler in an Amateur keyer. Device isolates the IC keyer from the transmitter, removing grid-block keying voltage, which is sometimes rather high, from the keyer paddles.

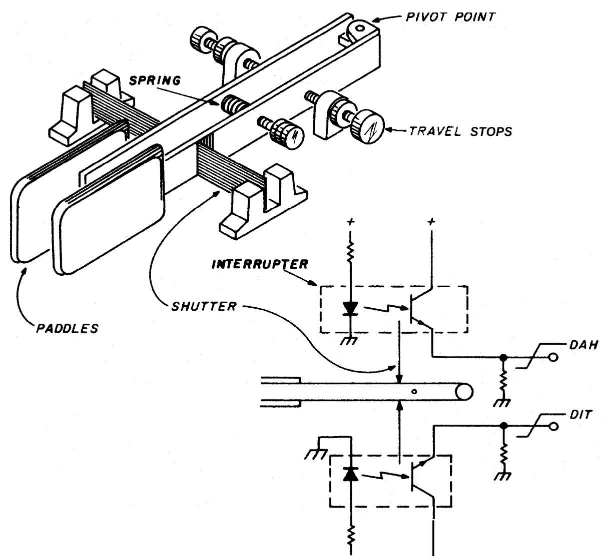


fig. 11. A contactless keyer, or "optical paddle," using an LED optical interrupter. The optical interrupter is similar to the optical coupler, except that the light path from the diode to the photo transistor can be interrupted with a shutter or other mechanical device.

optical interrupter

Another interesting device in the LED family is the optical interrupter. Its construction is similar to that of the optical coupler, except that the light from the diode to the photo transistor is accessible, meaning that the light path can be broken or interrupted with a shutter blade or other mechanical device. When this action occurs, the output transistor will be cut off, which allows an easy mechanical interface to electronic circuitry. I've not seen this device used to any great extent in Amateur applications, but I think a contactless keyer would be a good item to begin with. I've shown one in **fig. 11** and hope to build a device such as this in the near future. The ease with which it will interface digital logic makes it a natural.

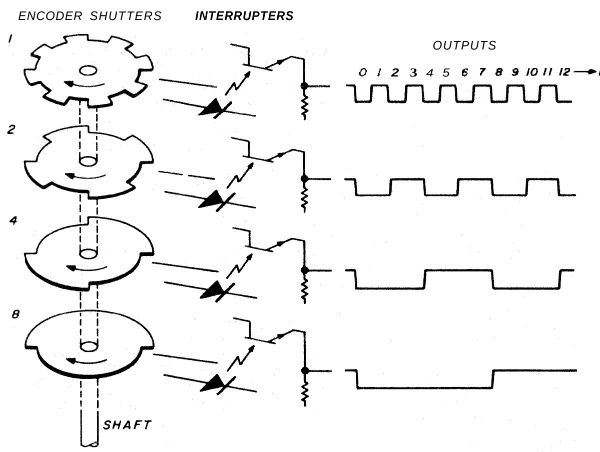


fig. 12. Example of a shaft encoder using the optical interrupter. No electrical contacts in the encoder mean increased reliability and low maintenance. BCD outputs are shown for each encoder shutter. A neat way to encode your antenna rotator for digital readout!

shaft encoder

Another application comes to mind and seems quite reasonable, since the once-plentiful selsyns are rather difficult to find these days. This application is a position indicator or shaft encoder. Using digital techniques, you could easily get sixteen discrete positions with only four optical interrupters. Adding another interrupter would double the resolution; this could be carried out to any degree desired.

By arranging the shutters and interrupters in a format such as shown in **fig. 12,** you would obtain a four-wire BCD output. This could be carried out to an eight-digit arrangement and applied to the input of a micro processor if you really wanted sophistication and had an extra eight-bit port on your micro. This would yield 256 discrete outputs or positions.

Put a device like this under your antenna rotator or weather vane and you would have a real winner! The same scheme could be used for 180-degree capacitors and multi-turn inductors for remote tuning. This low current, contactless device could form the basis for some interesting, reliable equipment.

LED displays

The seven-segment display is probably the most widely used LED display available today. This popularity has made the price right for the Amateur. The display can be thought of as seven individual LEDs placed in a single easy-to-use package and arranged to provide a numeric output. Each individual element or segment has parameters that are similar to those of the diodes discussed earlier. These devices are available in a common-anode configuration, which is generally used for applications involving a positive power supply and the common-cathode configuration usually associated with negative supply designs. **Fig. 13** illustrates both types and the physical relationship of the individual diodes as viewed from the front or display side.

This illustration shows how the characters are formed by forward biasing the individual diodes, or segments, in a prescribed manner. The character

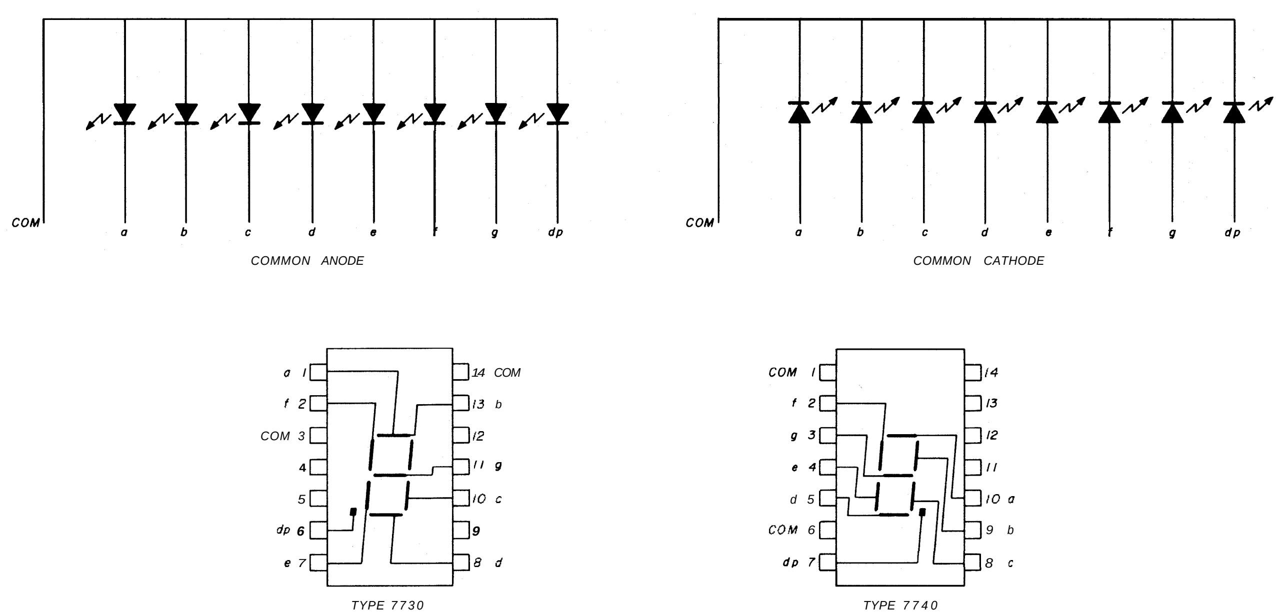


fig. 13. Examples of seven-segment LEDs designed for positive power supplies (common anode) and negative supplies (common cathode). The characters are formed by forward biasing the individual diodes, or segments, in a prescribed manner. These displays have decimal points (DP). Units are available with left-hand, right-hand, both, or no decimal point. Colors are red, green, and yellow (amber).

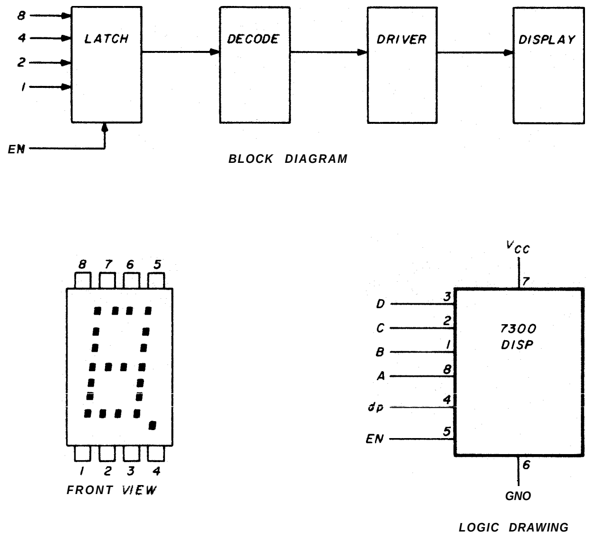


fig. 14. Integrated displays have an IC built into the package. The IC provides decoding, storage, and a display driver. Cost, however, is five to ten times that of the seven-segment type on the surplus market. Device is a dot matrix type and accepts a four-wire BCD input. Others are available with bar-type LEDs such as used in the seven-segment displays.

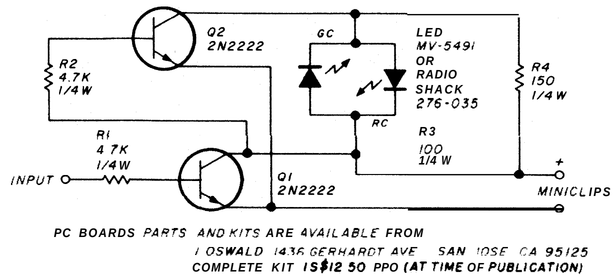


fig. 15. Schematic diagram of a logic probe that can be built in a few hours with readily available components. Logic probe output indicator yields three discrete states according to input signal or level. Power is borrowed from the circuit under test.

"O" would be formed by forward biasing the elements labeled "a" through "f." The displays used for this illustration have decimal points (dp), and units are available with left-hand, right-hand, both, and no decimal point. As with the other LEDs discussed, seven-segment displays are available in red, green, and yellow; as with the indicator units, voltage and current parameters vary accordingly.

A number of physical sizes and package configurations are available, and the price is often less than a dollar per digit on the surplus market. The units used in the illustrations are configured in a 14-pin DIP package and furnish a character height of 0.3 inch (7.6mm). This seems to be an adequate size for most projects, and bezels are readily available to give your project a finished look.

the integrated display

The integrated display unit is a display much like the seven-segment LED, with the addition of an IC built into the display package. The IC provides decoding, storage, and display driver — all within the display package, and requires no more space than the seven-segment display. All these functions being performed by the display unit make it much simpler to use and reduces component count considerably. As can be expected, there's a hitch: the cost of the display is five to ten times that of the seven-segment type on the surplus market. This sounds very high indeed, but in figuring the total expense of the components required to do the functions of the integrated display, it's often cost effective, particularly when space is at a premium.

The integrated display illustated in **fig. 14** is a dot matrix type and accepts a four-wire BCD input. Other integrated displays are available with bar-type LEDs such as used in the seven-segment display. Various types of logic configurations are available such as counters, latches, and hexidecimal decoders. These devices interface TTL logic very well and can make the design and construction of counters and similar devices relatively easy.

logic probe

The logic probe is a good application of the theory discussed earlier and makes an ideal instrument for digital testing. It can be constructed in a few hours using readily available components. The output indicator of the logic probe yields three discrete states according to the input signal or level. The probe borrows power from the circuit under test and is designed to function with the popular TTL logic families.

The circuit of the logic probe is shown in **fig. 15** and is a variation of the dual LED circuit in **fig. 8.** When power is applied to the probe through the power leads, and the tip or input is touched to a low level or ground, Q1 is cut off. This condition will cause Q2 to conduct since the base is positive with respect to the emitter. With Q1 cut off and Q2 conducting, the green diode of the dual LED will be forward biased, yielding a green output from the LED. Touching the probe tip to a high level will cause Q1 and Q2 to complement, and the red diode will be forward biased, yielding a red output from the LED. An

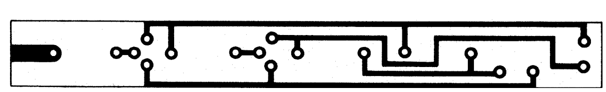


fig. 16. Full-size foil pattern for the logic probe.

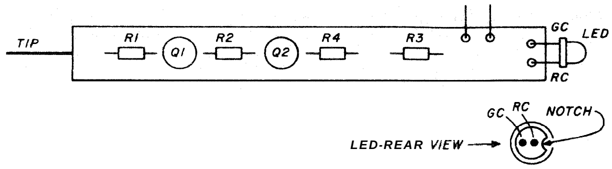


fig. 17. Logic-probe component layout.

alternating signal will cause alternating conduction of the red and green diodes and will yield an indication approximating amber. In this manner both static and dynamic signals can be traced with the logic probe.

Printed circuit construction is used for the logic probe, and a full-size foil pattern is shown in **fig.** 16. After etching and drilling the PC board, the components are mounted as shown in the component-side view, fig. 17. **Fig. 18** is a sketch of the probe assembly. I used the components from an Eico demodulator probe for construction, but any plastic or plexiglass tubing with approximately ½-inch (13-mm) ID will suffice. The end caps can be cemented in place after the PC board is slipped into the tubing, and your logic probe should be ready to go to work.

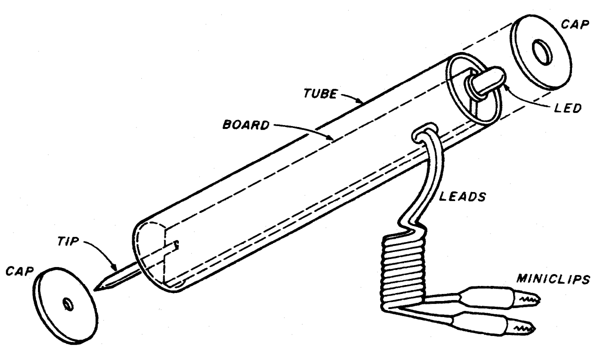


fig. 18. The logic probe. Any plastic tubing about ½ inch (13 mm) ID will suffice for the probe container.

summary

I've enjoyed building and using the logic probe and certainly found LED devices to be interesting and a very useful addition to equipment designs. As stated earlier, they are no more complex than any other diode and can be a lot more fun to use. The digital revolution is rapidly gaining acceptance in the world of Amateur Radio, and the logic probe and associated LED theory will help you to accept and enjoy the benefits we will all gain from this new technology.

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ham radio

measuring signal-strength

Receiver S-meter readings are no guarantee of true signal strength — here's how to quantify this controversial subject

Amateur radio transmitters require many measurements, most of which are obtainable from one or more internal or external meters. Some meters have two or more scales, which are selected by switches. Readings may include grid current, plate current, high voltage, relative power, SWR, forward power, reflected power, and ALC level. Other readings may be taken with rf ammeters, rf voltmeters, frequency meters, and impedance bridges. Also, field-strength indicators may be used to adjust the dimensions of fixed or mobile antennas. These indicators are basically short-range portable receivers with loop or whip antennas and meters with arbitrary uniform scales. They are often called field-strength meters despite their inability to receive and measure distant radio signals.

Amateur radio receivers, unlike transmitters, have only one meter and no meter switch. Transceivers also have only one meter, but a switch provides two or more transmitter readings and a receiver S-meter reading. An S-meter gives relative values of signal strength on a uniform 0 to 9 scale, followed by a dB scale. It reflects the last i-f stage level of all signals by responding to the average agc voltage, and is not related to af gain. Readings are based on maximum rf gain and peaked tuning of an antenna or preselector circuit.

S-meter problems

In general, S-meter readings depend on three variables: signal field intensity, antenna characteristics, and receiver gain. Unfortunately, readings bear no fixed relationship to antenna voltage, as receiver gain varies with various makes, models, units, frequency bands, and parts of bands. Also, there is no accepted standard of input voltage for S9 readings, so various manufacturers have chosen signal-generator outputs of 25, 50, or 100 microvolts for calibration at a specified frequency. Hammarlund *Superpro* receivers for military use were made to read S9 with 50 microvolts input at 3.5 MHz, and modern receivers are not very different.

A typical receiver has a one-milliampere meter in a bridge network containing a potentiometer, which balances no-signal plate or cathode current during zero adjustment. *Superpro* receivers had a 200-microampere meter with an adjustable 1000-ohm shunt for sensitivity adjustment, and also had an AVC amplifier preceding the AVC rectifier. Modern receivers usually avoid these features. Zero and sensitivity adjustments often vary with time.

Although S-meter scales are uniform, S-units are

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not. One manufacturer states that a change of one S-number on the meter indicates a change in signal strength of approximately two to one (6 dB). Another manufacturer states that each S-unit indicates a 3-dB increase in signal strength. Both statements are thus questionable. An instruction book for a Collins receiver defined neither S9 input nor S-meter steps, but a calibration showed 50 microvolts and steps of about 1.4, or 2 to 1 in voltage ratio, or 3-6 dB, on the 80-meter band.

Since S-meter readings can be either optimistic or pessimistic, and give only qualitative information at best, it follows that quantitative readings would be more desirable. Field intensity, often called field strength, is the only true measure of signal strength Broadcast-station operators often use field-intensity measuring equipment because they must submit field intensity data to the Federal Communications Commission. This data shows the extent of service areas and interference areas for other stations using the same frequencies.

field-intensity meters

Despite the vagaries of S-meters, Amateurs have avoided using field-intensity meters because of their cost and complexity, and also because Amateurs have no service area and interference area limits. Unlike broadcasters, Amateurs can enhance their service areas by changing frequency, power, and time of operation. Likewise, Amateurs can avoid interference by changing the same variables.

A field-intensity meter is basically a combination of a local oscillator or signal generator of known output, an attenuator, a portable receiver containing an output meter, and a loop or whip antenna of known effective height or length. Another antenna can be used if it is calibrated by comparison with the loop antenna. It follows that, with certain additional equipment, an ordinary receiver can serve as the major component of a field-intensity measuring set.[1]

High accuracy is unnecessary in field-intensity measurements of distant radio stations. Propagation conditions, such as fading, vary considerably with time. A receiving station may intercept ground waves, sky waves, or a variable combination of both, which results in multipath interference. Sky-wave fields can be measured with a horizontal antenna, regardless of whether horizontal or vertical polarization is used at the transmitter, since practically equal amounts of both types of polarization are present in the incident ionospheric field.

Radio waves are travelling electric and magnetic fields that are perpendicular to each other and to the direction of propagation. Both fields convey equal power at a distance from their source, so either field may be considered as a measure of total radiation. It is customary to measure the electric field component, which may be horizontally or vertically polarized, according to the direction of that field. Measurements of field intensity are usually expressed in units of rms microvolts or millivolts per meter at some point, which are really potential gradients.

When a radio wave strikes an antenna, it induces a voltage equal to the product of the field intensity and the effective height or length of the antenna. Thus, if the antenna voltage and effective height or length are known, the field intensity is given by the quotient of their values. Algebraically, since

$$V = EH \text{ or } EL, \quad E = V/H \text{ or } V/L \qquad (1)$$

Therefore, the measurement of field intensity depends primarily on the measurement of a small, variable antenna voltage. This is usually done by substituting a measurable and attenuated local signal in place of the real signal and noting the attenuation for equal output from a receiver. Another, less used, method compares the voltage induced into the antenna by the desired signal field with that from a standard local field. In general, an incoming signal is measured by comparing it with a known calibration signal.[2]

Any kind of signal can be measured with the aid of a receiver. An S-meter will serve as a comparison output meter, as it responds to a-m or fm carrier levels of single-tone SSB modulation and shows about one-half maximum output for CW signals. Audio gain and output are of interest during tune up and zero beating the real and artificial signals for frequency match, but are otherwise irrelevant.

If the real and artificial signals are equal and the antenna characteristics are known, it should be easy to determine field intensity. In principle, a calibrated signal generator is a substitute for the antenna and its output is adjusted to give the same output-meter reading as the real signal. But it is hard to design an attenuator that is accurate at high radio frequencies. This problem was avoided over fifty years ago by putting an attenuator before the i-f amplifier in a special set, which included a local oscillator and calibrating means. The theory and operation of such a set was covered in an IRE paper and will not be discussed.[3]

effective height

The effective height of the receiving antenna must be known to complete a field-intensity measurement. A loop antenna is a standard of comparison because its effective height can be calculated. The relation is as follows.[4]

$$H = 2\pi AN/\lambda = 2.0944 \times 10^{-5} fAN \text{ meters} \qquad (2)$$

where A is the loop antenna area in square meters
N is the number of turns of that area
λ is the wavelength in meters
f is the frequency in kHz

The effective height of a grounded quarter-wave vertical antenna is not the true height, H, but $2H/\pi$ or $0.6366H$. It is the height of an imaginary conductor, which carries a uniform current equal to that at the base of a real conductor and radiates the same field along the horizontal. The effective height of a much shorter vertical antenna is about one-half of the true height.[5]

The true length of a horizontal half-wave dipole antenna is $143/f$ meters, where f is in MHz. This is 95 per cent of a half wave in space, but the effective length is only 62 per cent of that value, or 65.26 per cent of the true length, or $94.3/f$ *meters*. If a horizontal dipole is used to measure field intensity, the horizontal component of the field is quite independent of ground characteristics, which differ with location.

The effective lengths of horizontal dipole antennas designed for the 20, 40, and 80 meter phone bands are about 6, 12 and 24 meters respectively. Thus, 50-microvolt S9 signals would produce field intensity readings of approximately 8, 4, and 2 microvolts per meter. Other calibrations and S-readings could be used, with the assumption that S-units are 6 dB apart or 2 to 1 in voltage ratio. This shows that, with the aid of a calibrated signal generator, an ordinary receiver can serve as a substitute for a field intensity meter if accuracy is unimportant.

Broadcast stations use commercial field-intensity measuring equipment and take readings at various distances in various directions, called radials. Field intensity is plotted against distance along each radial on log-log coordinate paper and a smooth curve shows the distance to any value of interest. Distances so obtained from all radials are plotted on polar coordinate paper or map to show the contours of various field intensities, hence service areas. Field intensities of two interfering stations may be plotted against distance in opposite directions on ordinary graph paper to show regions between them where interference may occur.

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ham radio

More Details? CHECK–OFF Page 94

tone alert monitor

Design and construction of a tone call system for Radio Amateurs interested in emergency communications

Recently it has been emphasized many times that the need for Amateur Radio communications in disaster situations is very real and necessary. Many groups of Amateurs and clubs have spent money for equipment and much time in drills and practices to make themselves ready for emergency communications.

One problem that continually bothers Amateur operators who are striving to maintain their proficiency is the need to monitor one or more radio communications links. Such monitoring is necessary so that a central agent can quickly "call up" operators who can provide communications for disaster services. Many Amateur operators would be more willing to monitor for the call up if it were not for the task of listening to a party line such as a 2-meter repeater, national calling frequency on 2 meters, or other busy radio frequencies.

alert monitor

This problem of requiring an individual to monitor for an alert, either real or during practice, has inspired the design of an Alert Monitor that will alert each station without the operator having to listen to the party line chatter associated with most frequencies. The Alert Monitor is also handy for those using a receiver in a busy household or in an automobile, where a busy repeater sometimes can be annoying. The monitor can be used with almost any receiver capable of passing normal audio-frequency information. It is not recommended for SSB transceivers that can't hold tight frequency stability.

This Alert Monitor also can alert groups of stations simultaneously (Group Call), or it can alert one station at a time. Tne Alert Monitor operates with standard *Touch-Tone** audio. It uses two digits for normal operation and a long, single digit for the Group-Call feature.

circuit description

Fig. 1 shows the schematic. The audio is applied to the Alert Monitor through the audio-input jack and a 2.2k resistor. Audio amplitude is limited by two 1N914 back-to-back connected diodes. A 5k pot adjusts the audio input level, which is coupled to the tone decoder ICs by 4.7-μF capacitors. U-1/U-2 decode the high and low tones for the first *Touch-Tone* digit, and U-3/U-4 decode the tones for the second digit. The resistor and capacitor at the output of each decoder IC provide a delay between the time that the decoder IC senses a tone and the time that the NOR gate will respond. This prevents falsing on other audio that may be present in the decoder.

When the first digit has been received, a logic high is applied to U7A pin 1 for 2.5 seconds through timer IC U6. A logic high must be applied to U7A pin 2 during this 2.5 second period for the Alert Monitor to operate. Therefore, the second digit must be received within 2.5 seconds of the first for a valid call. This sequence ultimately sets U10B and causes Q1 to conduct. This, in turn, causes relay K1 to transfer the audio to the audio-output jack. In this condition, the audio signal is connected to the speaker, and the receiver can be monitored for the "Alert Call."

For continuous receiver audio monitoring, the monitor switch can be connected to the monitor positions and the communications receiver can then be used normally without the Alert Monitor defeating the receiver audio.

The Group Call feature works by the sending station (encoder) transmitting the second digit of the

***Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.**

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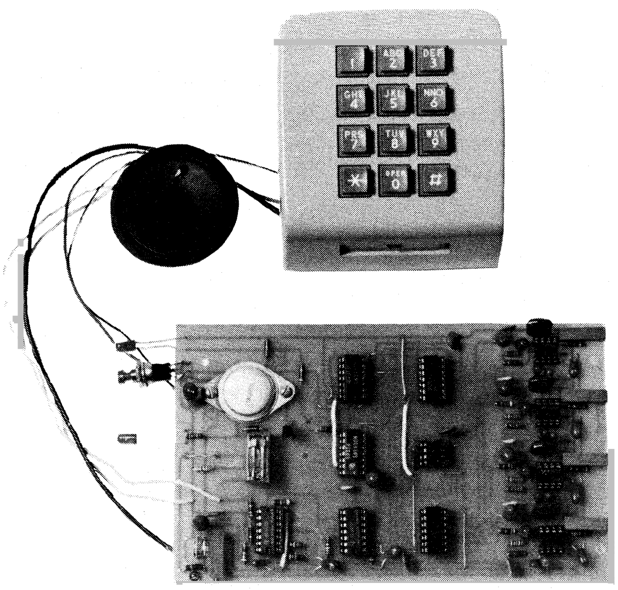
Control box of the prototype Alert Monitor.

two digits for 6 seconds or longer. In this mode, U-1/U-2 don't respond. The signal path now appears in the form of a pulse at U8A pin 6 and will, after 6 seconds, produce a logic high at UIOA pin 3. If, in fact, the second tone is still being received, it will cause a logic high to be applied to UIOA pin 2. This action will enable the Group Call Detector to produce a logic high, which will eminate from UIOA pin 5. This high is applied to U5D pin 12. U5D is a NOR gate; and from this point on, the operation is the same as when two digits were received.

Whenever U10B is set, either by receiving the correct two tones or the Group Call tone for 6 seconds minimum, three reactions take place:

1. The audio is applied to an external speaker as described earlier.

2. A call indicator is illuminated. This is an LED mounted on the front panel.

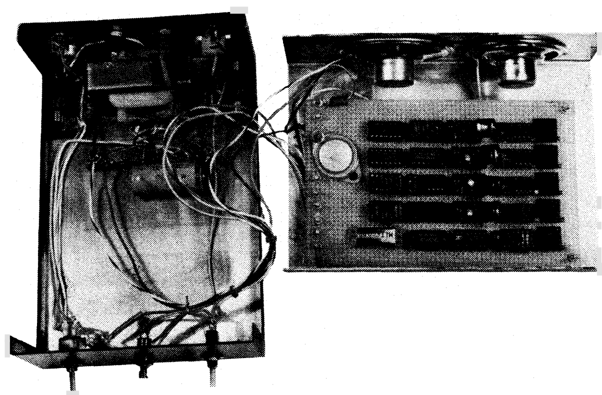


Alert Monitor built on PC board, shown here ready for alignment and testing.

3. The Alert Monitor oscillator is triggered for 4 seconds, producing a two-tone "twee-dell" audible signal from the Alert Monitor.

The reason for the visual call indicator is in case a call has been received while the operator is out of hearing range. Upon returning, if the call indicator is illuminated, the operator knows that an alert tone has been received, even though the receiver may be squelched or silent at that particular time.

To put the Alert monitor back into the alert mode, press the RESET push-button to reset UIOA and B and to extinguish the call indicator LED. It will also disable the audio path through K1 contacts. This action will again silence the receiver speaker until the next proper alert tone is detected.



Complete prototype Alert Monitor showing perf-board construction, power supply, speakers, and minibox enclosure.

power requirements

Power requirement for this unit is from 8-20 Vdc at approximately 150 mA. This makes the unit adaptable for mobile operation, and the low current drain should not present any problems so far as leaving the receiver and the Alert Monitor on while away from the automobile. The prototype was built to include a dc power supply which also houses the Alert Monitor speaker as well as a speaker for receiver audio. A power supply is shown in **fig. 2.**

construction

The original Alert Monitor was constructed on a fiberglass perf board using wire-wrap sockets. All discrete components except variable potentiometers, the relay, and the capacitors associated with U12, were mounted on 14-pin header plugs. **Fig.** 3 shows the wiring of the discrete components to the ICs.

No special construction practices were observed, except that all V_{CC} connections for the ICs were carried back to the power source at U12 pin 2. Each IC

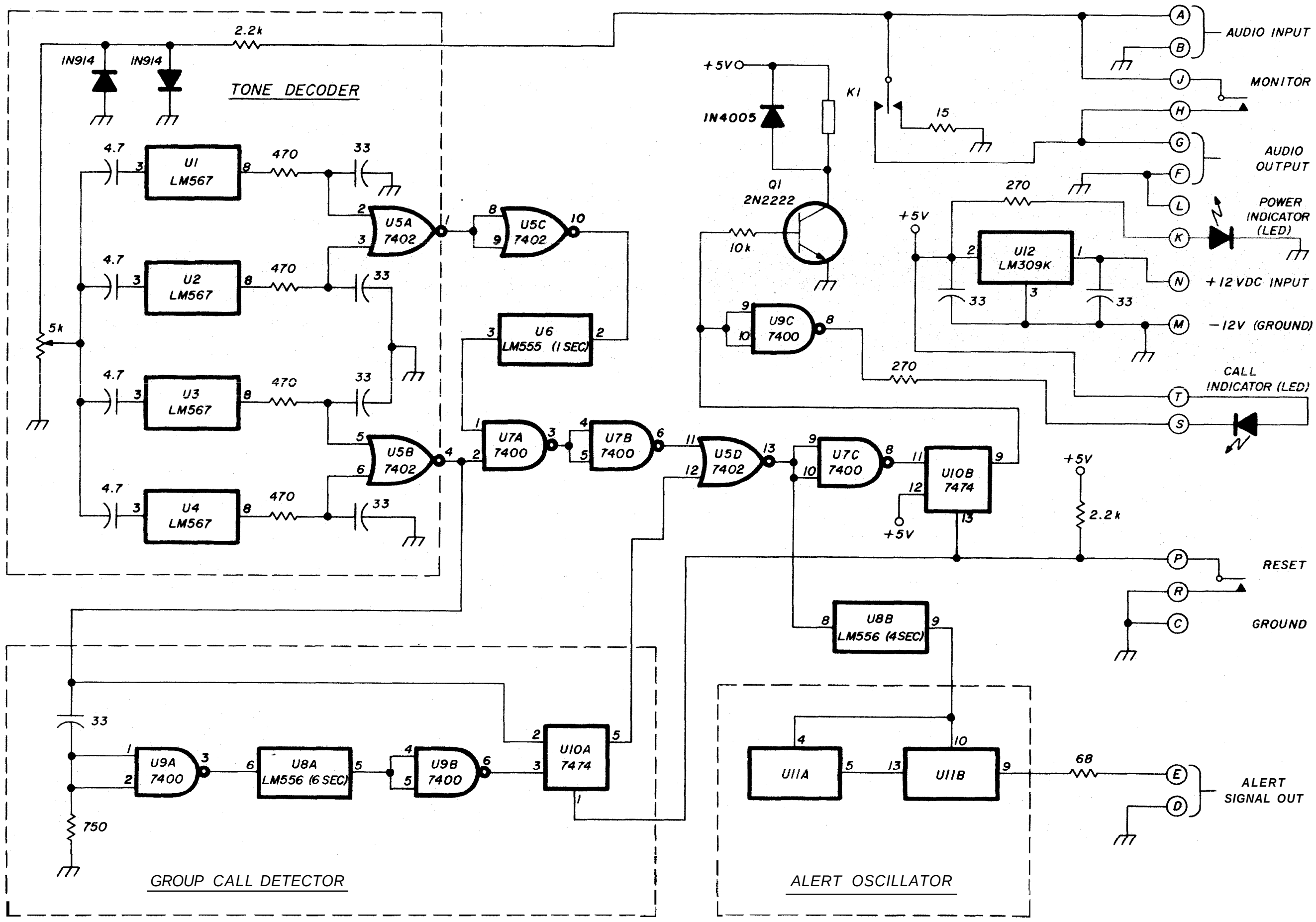


fig. 1. Alert Monitor schematic. Circuit operates with standard *Touch-Tone* audio. Two digits are used for normal operation and a long, single digit for Group Call. Original circuit was constructed on a fiberglass perf board using wire-wrap sockets, but a PC board is now available (see text).

also had its power ground returned to the central ground input point at the case of U12 to reduce transients on "daisychained" power leads, which could cause unstable operation.

Bypass capacitors across the plus and minus power connections of the ICs may be added if necessary. No heatsink was provided for the LM309K (U12). At the low-current level required for the unit, it was found to be unnecessary. The 5.0k trim-pots used for frequency adjustment and input-level adjustment are all multi-turn potentiometers. These are necessary, especially for the frequency-adjusting pots, for the vernier action needed to "tune" the LM567s to their correct frequency.

Should you not wish to use the wire-wrap method of construction, a PC board is available.* The board is single-sided and requires very few jumpers. I prefer this method of single-sided PC board because it's much easier to use or copy. All component parts used for both prototypes (perf board or PC board) were purchased from Jarneco Electronics.† Of course, parts from other suppliers are acceptable, but this attempt at standardization should help the builder. The only part I did not get from Jameco was the reiay. It was purchased from Radio Shack, and the part number for this item is 275-215.

alignment

Several methods for alignment of the tone decoder section are available. The fixed-value resistor connected to pin 5 of each LM567 in series with the 5.0k variable resistor and the 0.1-μF capacitor connected to pin 6 are the frequency-control elements for each single-tone detector (LM567). A good-grade Mylar capacitor should be used in this circuit to prevent frequency drift with temperature or internal leakage. If an oscilloscope is available, it can be connected to

*Order boards from MJW Boards, Route 2, Box 167-C, Berryville, Virginia 22611.

†Jameco Electronics, 1021 Howard Avenue, San Carlos, California 94070.

pin 8 of the tone decoder IC to be aligned. An audio signal of the correct frequency is applied to the audio input by connecting the Alert Monitor to the receiver with which it's to be used. Then a ***Touch-Tone*** signal is transmitted in the normal fashion from a transmitter.

alignment procedures

The procedures described should, of course, be done with a dummy load connected to the transmitter to avoid interference:

method 1

1. Set the audio level at the receiver for the normal listening volume.

2. Set the audio-input-level control at midrange.

3. Adjust the 5k frequency-control pot until a logic low appears at pin 8 of the first decoder IC under test.

4. Reduce the audio level by the 5k level pot in increasing amounts while adjusting the frequency-control pot to maintain the logic low output until no further sensitivity can be obtained for the IC under test.

5. Increase the audio level by the 5k level pot, with the oscilloscope moved to the next IC.

6. Adjust this IC in a like manner to the next ***Touch-Tone*** frequency being received.

As an alternative method, the LED connected to the tone decoder IC, pin 8, may be used to indicate that the tone decoder has received the correct tone at a sufficient level. As the LM567 detects (or locks onto) the input tone, the LED will light. (Make sure of

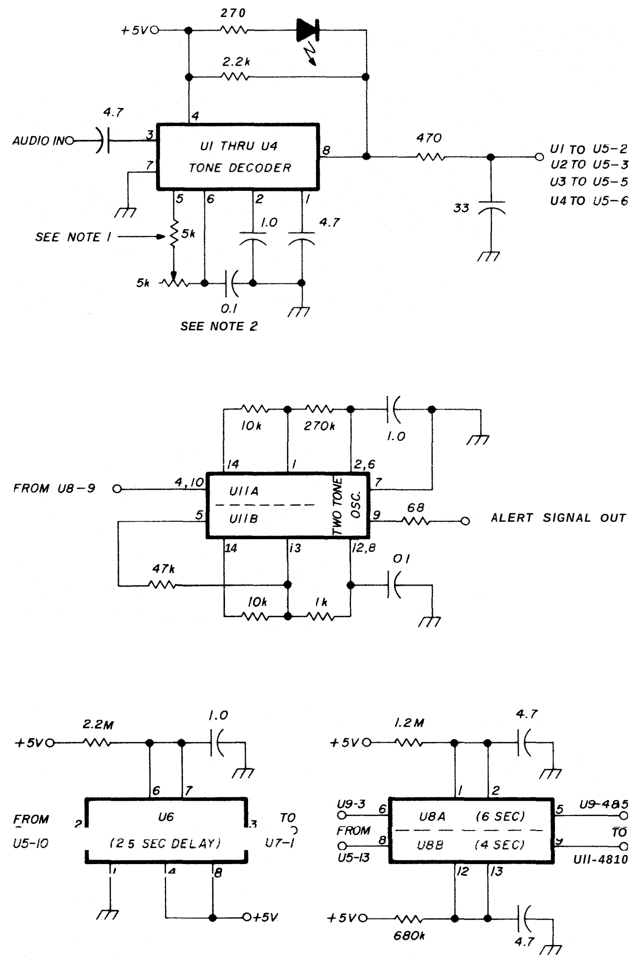


fig. 3. Discrete-component wiring to the ICs. The resistor connected to pin 5 of the tone decoder IC in series with the 5-k pot and 0.1 cap connected to pin 6 are the frequency-determining elements for each single-tone detector. A good-grade Mylar cap should be used in this circuit to prevent frequency drift with temperature.

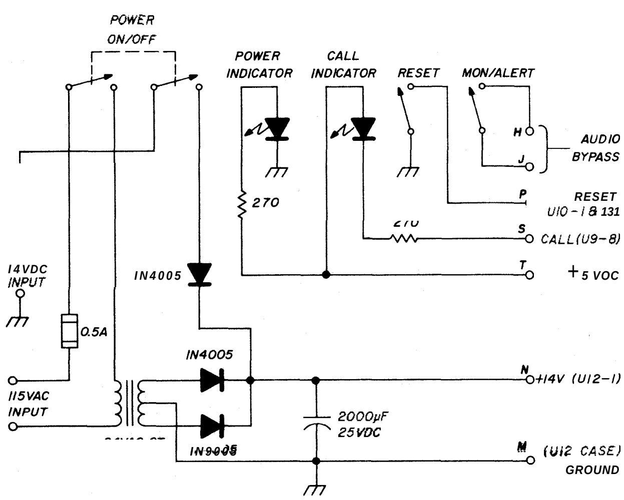
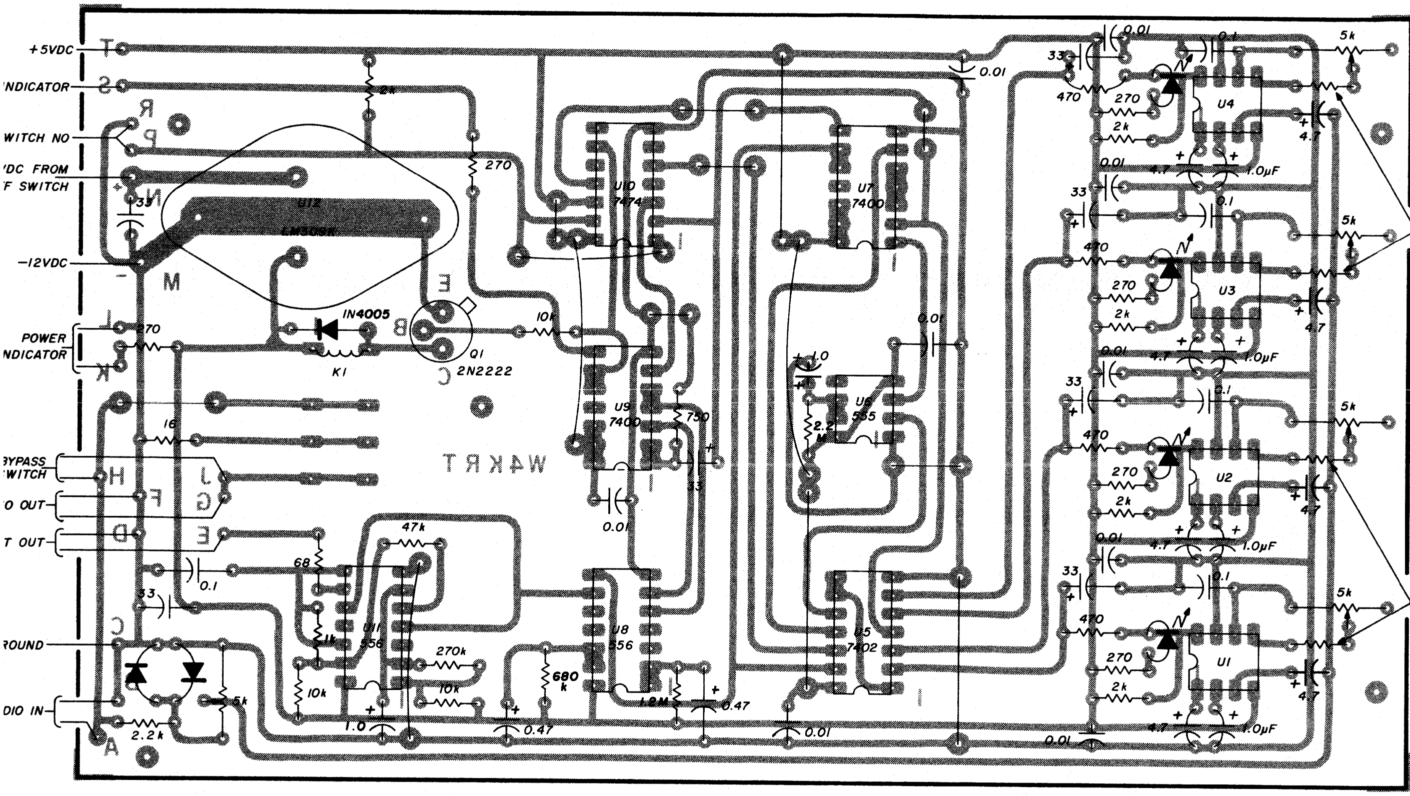


fig. 2. A power supply for the Alert Monitor.

correct polarity for the LED). When no further refinement of frequency can be made by the frequency-control pot, you'll notice that the IC will oscillate between a logic high and low. If you're using the LED method for alignment, the LED will start to dim at this point. The LEDs and the associated 270-ohm resistor may be eliminated for each tone decoder on the PC board if not used as an alignment indicator. This will not affect the performance of the circuit.

Each ***Touch-Tone*** digit consists of two audio frequencies. Therefore, two decoder ICs are necessary for each digit being decoded. Through experimentation I've found that 11k will allow centering the variable pot for the lowest frequency tone, and 2.4k will allow adjustment at the highest frequency. Exact calculation of the time constants for the RC combinations can be found in the data sheets provided by the



.NOTE: FREQUENCY DETERMINING RESISTOR—SEE TEXT

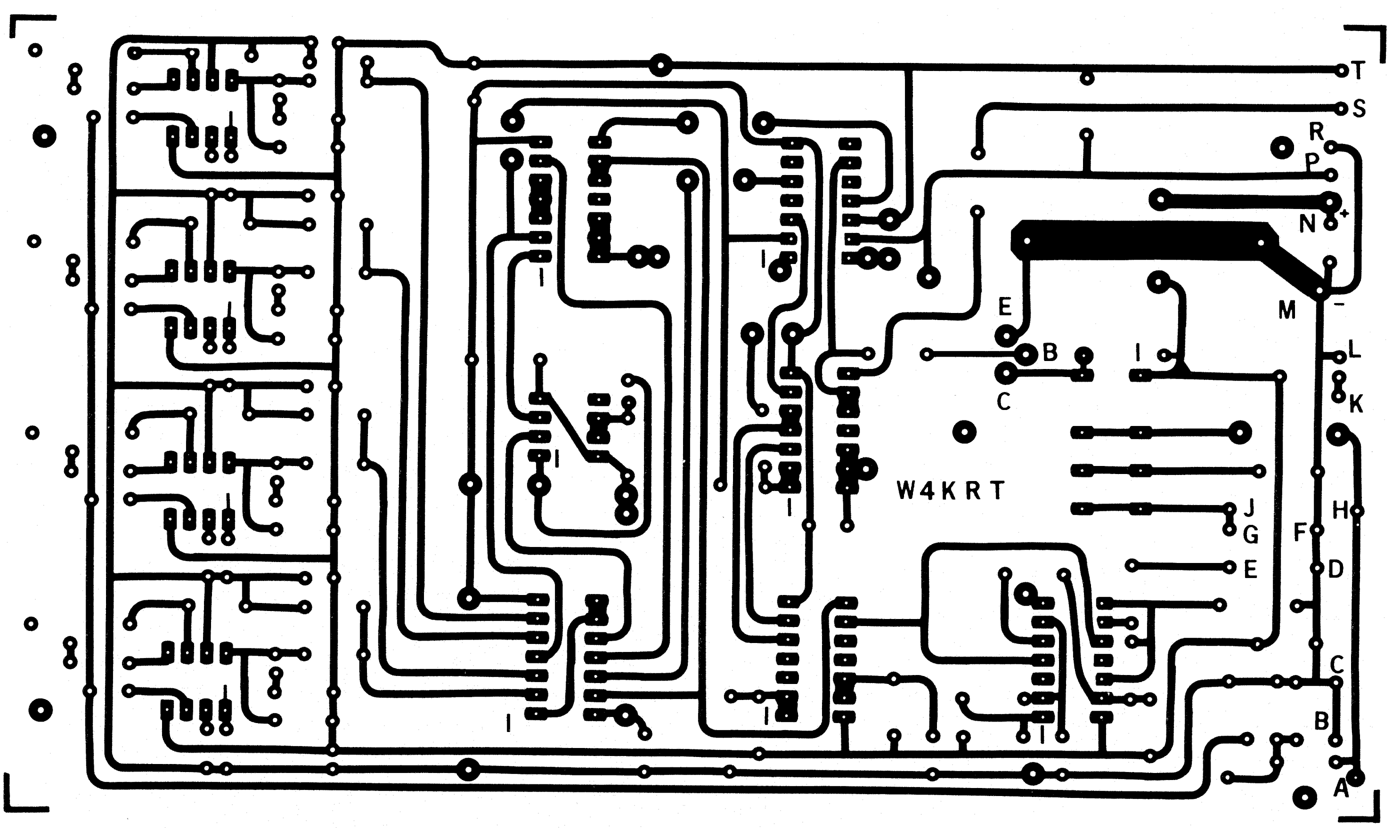


fig. 4. Above, component placement on PC board. Below, Alert Monitor PC board, foil side.

table 1. Relationship between frequencies produced by the various *Touch-Tones.*

digit	high tone (kHz)	low tone (kHz)
1	1209	697
2	1336	697
3	1477	697
4	1209	770
5	1336	770
6	1477	[illegible]
7	1209	852
8	1336	852
9	1477	852
*	1209	941
0	1336	941
#	1477	941

manufacturer. See **table 1** for the frequencies produced by various touch tones.

acknowledgments

I'd like to thank all of my Amateur friends who gave suggestions and encouragement for this project. In particular, I'd like to thank KA4GCF, Bob, for his help with the manuscript; WB4HID, Harry, for his help with construction tips; Don Bungard for his help with the PC board; and KA4HOP, Janie, my XYL, for her many hours of assistance in testing the Alert Monitor and for her typing work.

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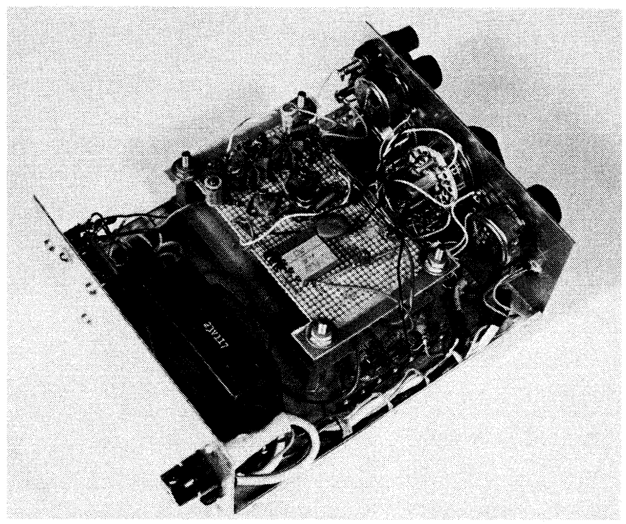
ham radio

More Details? CHECK — OFF **Page** 94

integrated circuit function generator

A function generator based on the Exar XR-2206 which features multiple waveforms over the range 10 to 100 kHz

If you are a builder, experimenter, or just like to maintain your own equipment, I am sure you appreciate the value of a good signal source. Ideally, the perfect signal source should be able to provide whatever frequency, amplitude, waveform, and impedance level required for the job at hand. But, unfortunately, like most things in life, we must compromise a little. The classical signal source was either an audio generator or an rf generator. At best, it produced a sine wave and, in the case of the audio generator, perhaps a squarewave. In recent years, a third entry has appeared on the scene. It is called a function generator and is distinguished primarily by its ability to produce a variety of waveforms (functions) through the audio and sometimes low rf frequency ranges. Its repertoire usually includes the sine, square, triangle, and perhaps saw-tooth or pulse output waveforms. The spectral purity of a particular waveform is usually not quite as good as that from a comparably priced generator that is designed to produce only one waveform, but is usually more than adequate for general purpose use.



Internal view of the function generator showing perf-board.

Until recently, high cost has relegated the function generator to the laboratory, but now several new integrated circuits have made it possible for the Amateur to construct his own for a very modest investment. The function generator described in this article uses the XR-2206 integrated circuit manufactured by Exar. This IC can supply a sine wave, triangular wave, square wave, and a 50 per cent duty cycle positive pulse and has the options of a-m and FSK modulation. The frequency can be varied from about 1 Hz to above 100 kHz and the output waveform, although not lab quality, is suitable for most Amateur applications. The manufacturer states that the sinusoidal distortion is less than three per cent over the

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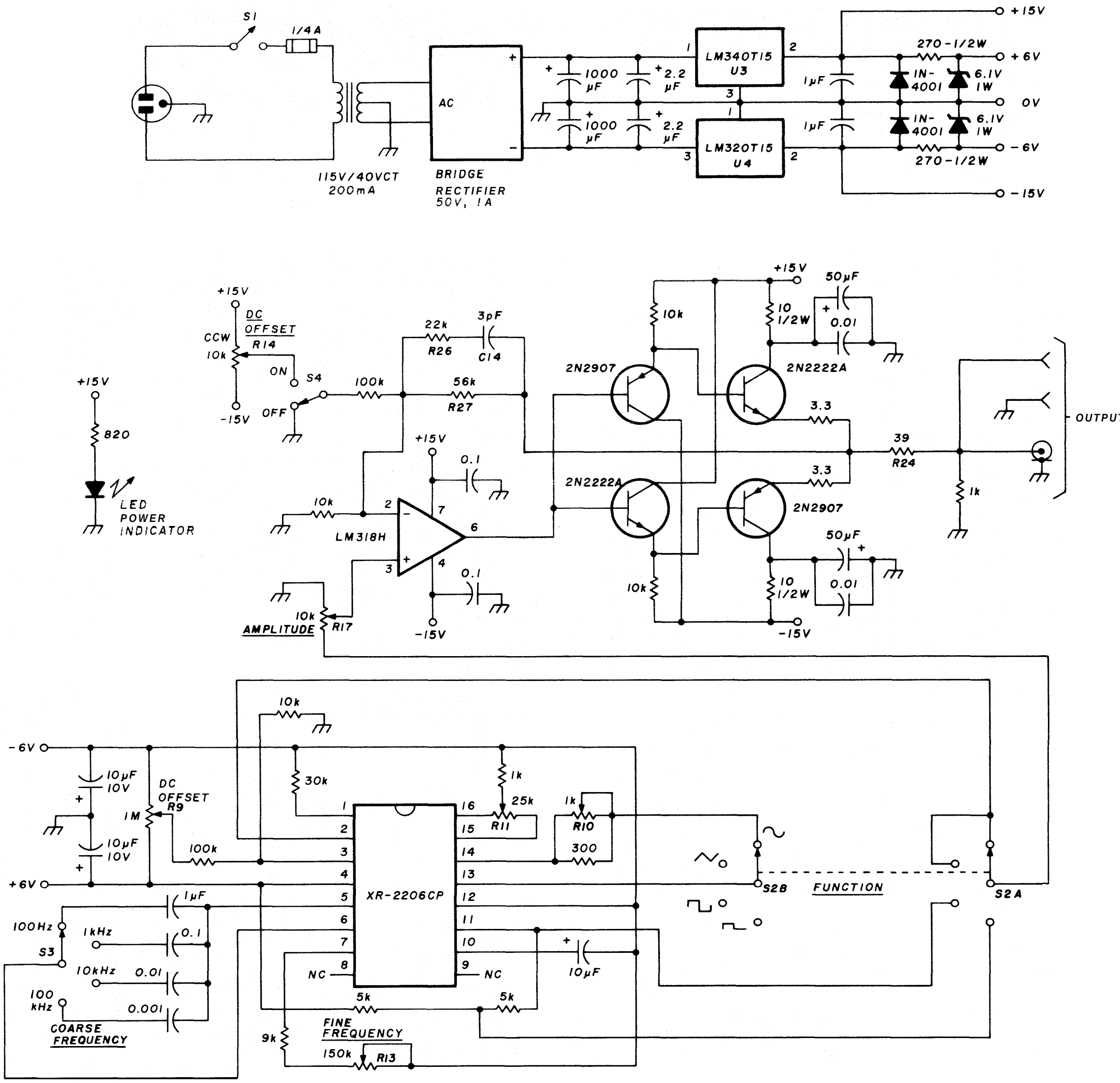

fig. 1. Complete schematic diagram of the function generator based on the Exar XR-2206 integrated circuit.

entire frequency range; this isn't bad considering the versatility and circuit simplicity.

I added a suitable power supply and power amplifier to the basic XR-2206 and came up with a very handy addition to my home workshop. The power amplifier increases the amplitude range, lowers the output impedance, and provides a means of introducing an adjustable dc offset level to the basic waveforms.

circuit description

The oscillator circuit is lifted almost entirely from Exar's application data on the XR-2206. I elected not to incorporate the a-m and FSK inputs, although they would be simple to include. R13 controls the frequency and is panel mounted. R9, R10, and R11 are all trimpots mounted on the circuit board. R10 adjusts sine wave distortion. R11 controls sine wave symmetry, and R9 is adjusted to eliminate any residual dc component in the sine and triangular waveforms. With the components shown in the schematic (see **fig. 1**), the frequency can be varied from about 10 Hz at the low end to slightly over 100 kHz at the high end in four overlapping ranges, each with a 100-to-l ratio. S3 selects the frequency range and S2 selects the waveform. R17 is the amplitude control; it is panel mounted and controls the signal level fed to the LM318 operational amplifier which in turn drives the power amplifier stage. R26, R27, and C14 provide negative feedback around the entire amplifier section, as well as frequency compensation. R14 and S4 provide an adjustable dc component for the output waveform. R24 brings the output impedance up to the vicinity of 50 ohms. The power supply uses two three-terminal regulators and although the positive and negative voltages do not track, it has proven quite satisfactory.

construction

I built my generator in a small aluminum box with most of the circuit, with the exception of some of the larger power supply components, mounted on perfboard. U3 and U4 use the cabinet as a heatsink and the two output transistors are equipped with small push-on heatsinks.

adjustment

Initial adjustment consists of setting R9, R10, and R11 to their optimum settings. Switch S4 to the off position and select the sine wave. Be sure that the amplitude control (R17) is set so that the output waveform is not clipped when viewed on an oscilloscope. Alternately, adjust R10 and R11 for the least sine wave distortion and best symmetry. Adjust R9 to eliminate any dc component and repeat the process until no further improvement can be seen. A distortion analyzer would be helpful for these steps if one is available, but the "eyeball" technique gives fairly good results. Make these adjustments at a frequency of about 10 kHz.

operation

Operation is fairly straightforward with the exception that the square wave will have a slight dc component with S4 in the off position. This is easily eliminated by switching on S4 and adjusting R14 to eliminate the offset. The waveforms deteriorate slightly around 100 kHz, but are still useful.

For the modest investment in time and material, this little generator has proven to be a very handy addition to my shop. I'm sure that some readers may have suggestions for improvements and modifications, and I would be most interested in hearing from them.

I would like to thank Mr. Ralph Flagel for his help with the photographs.

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ham radio

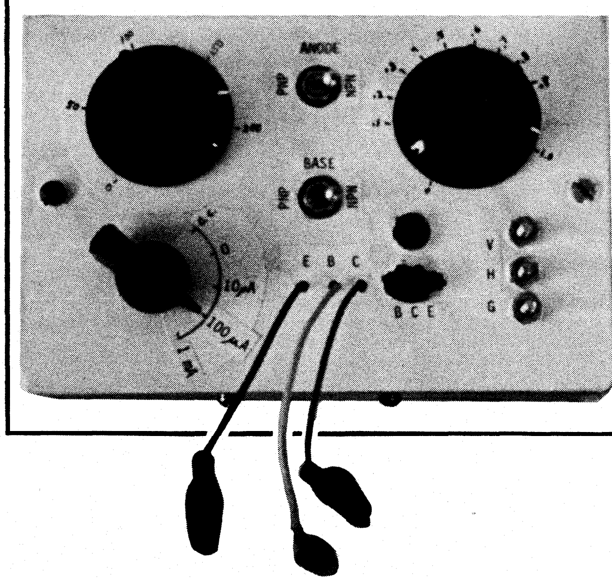
semiconductor curve tracing simplified

When used with a good scope, this versatile circuit provides a wealth of information about any unknown semiconductor

Most experimenters have accumulated a vast assortment of semiconductors, many unmarked, undecipherable, or otherwise of unknown characteristics. An ohmmeter can provide some information about them, such as which transistors are PNP and which are NPN; it can also distinguish silicon from germanium, and establish diode polarity. But to really gain detailed information requires a curve tracer and scope. A curve tracer will reveal nearly all low-frequency transistor parameters such as current gain, breakdown voltages, and input-output impedances. It will also identify, at a glance, zeners, tunnel diodes, and other specialized semiconductors, and give much valuable information about their operating characteristics.

A laboratory-type curve tracer usually injects a staircase waveform of current into the base of the unknown transistor, then displays a complete family of collector current waveforms on the screen.[1] This is an ideal arrangement, but not exactly simple. The curve tracer described here displays only one curve at a time; to look at an entire family requires the turning of a knob. I consider this a small price to pay for simplicity. It also has the advantage of providing a

The complete curve tracer unit. Upper lefthand knob controls peak collector (or anode) voltage swing. The knob on the right is the base current adjustment potentiometer, R2.



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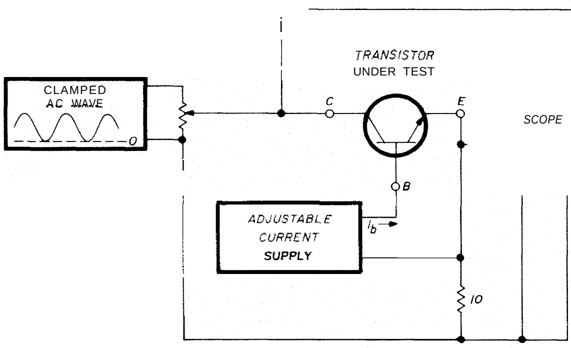


fig. 1. Simplified diagram of the curve tracer.

continuously adjustable base current, which permits the display of any collector curve, rather than only discrete values. It's sometimes nice to be able to see "between" those discrete curves.

theory

Fig. 1 is a simplified diagram of the curve tracer. The clamped ac supply swings the collector voltage from zero to a peak value determined by the potentiometer setting. This voltage is also applied to the scope horizontal input. Vertical deflection is provided by sampling current through the 10-ohm resistor in series with the emitter. The result is a trace of collector current vs collector voltage. The small voltage drop across the 10-ohm resistor does not significantly affect collector-to-emitter voltage; the error is only 1/10 volt for 10 mA of collector current. The 10-ohm value permits current readings to 100 microamperes per cm when the vertical gain is turned up to 1 mV per cm.

If greater sensitivity is desired, a 100-ohm resistor could be used, at the expense of a tenfold increase in error. The error in V_{CE} can be avoided by using a scope with completely independent horizontal and vertical inputs or by tolerating a downward deflection for increasing current. But since most scopes have one terminal common to both inputs (ground), and give a positive deflection upwards, the circuit was designed for this type of scope.

The clamped ac voltage for the collector is created by the simple circuit of **fig. 2A.** This voltage is positive throughout the cycle **(fig. 2B)** except for the brief interval, T_c, when the diode is conducting. During this interval, the waveform swings slightly negative by an amount equal to the diode barrier potential, about 0.7 volt.

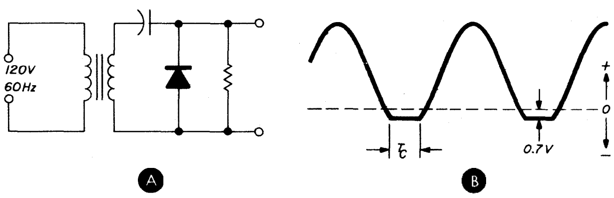


fig. 2. The simple circuit shown at A will produce the clamped ac waveform shown at B.

If the transformer and diode were perfect, the charge time, T_c, for the capacitor would be zero, and the output would be a pure sine wave. But since they are not, the waveform dwells at approximately – 0.7 volt for the time it takes to charge the capacitor. In my unit a very small transformer was used, which resulted in a T_c of about ***3*** milliseconds; with a larger transformer it would be less. The dwell time results in a bright spot at the beginning of the trace — not really a disadvantage since it makes the origin easy to identify. The spot can be made brighter by omitting C2 in **fig.** 3.

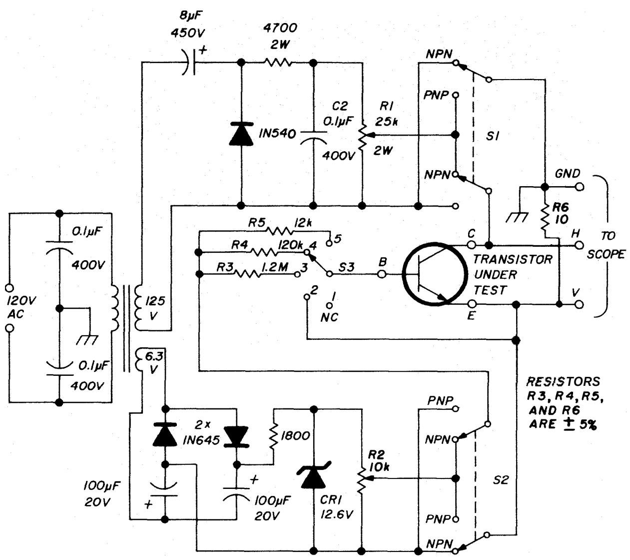
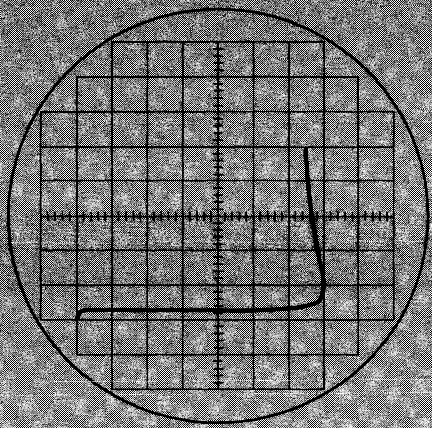


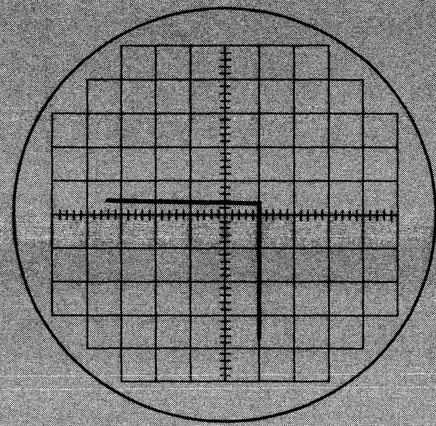
fig. 3. Schematic of the tracer. The power transformer can be a Stancor P8181 or an Allied PS8415.

the circuit

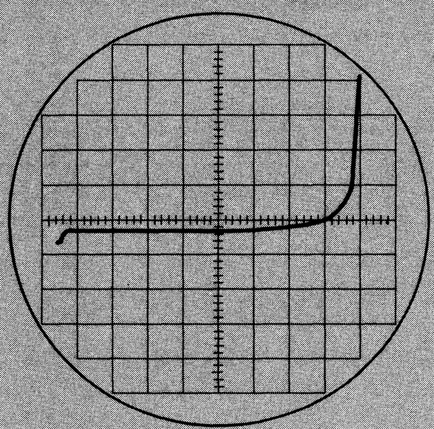
The complete circuit is seen in **fig.** 3. A small "one-tube" power transformer of the type used on early uhf TV converters, boosters, and other one-tube devices is used for the two power supplies. The "plate" winding is used for the clamped ac supply, and the 6.3 Vac "filament" winding is used for the constant-current base supply. The latter uses a full-wave voltage doubler and is regulated to a constant 12.6 volts by zener diode *CR1*. A voltage of 12.6 rather than 12.0 volts is used because of the 0.6-volt base-to-emitter voltage drop in a silicon transistor. Since this drop is about 0.2 volt for germanium transistors, base currents will be about 4 per cent higher than indicated when measuring germaniums.



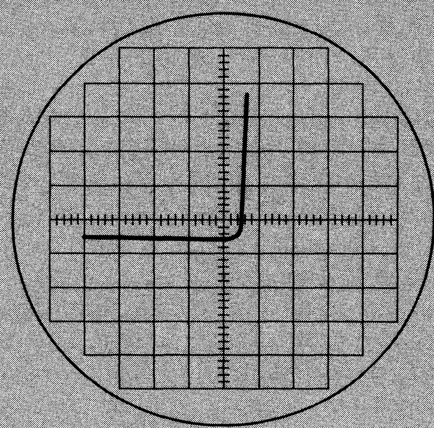
Collector characteristic curve of a 2N498 with base current of 9 μA. Scales are 20 V/cm horizontally and 1 mA/cm vertically. Notice the negative resistance effect in the avalanche breakdown part of the curve at about 120 collector volts.



Sharp breakdown characteristic of a 1N1527 zener diode at approximately 20 volts. Scales are 5 v/cm horizontally and 1 mA/cm vertically.



Collector characteristic curve of a 2N2369A with 8 μA of base current. Scales are 2 V/cm horizontally and 1 mA/cm vertically.



Reverse characteristic of a 1N914 showing avalanche breakdown at −90 volts. Scales are 20 V/cm horizontally and 1 mA/cm vertically.

Switch S3 provides three ranges for the base current adjustment pot, R2. The maximum currents, 10 μA, 100 μA, and 1 mA, are determined by resistors R3, R4, and R5. These resistance values ideally should be close tolerance, 5 per cent or better, to ensure the precise values of maximum base currents indicated on S3. The scale of R2 is calibrated 0-1.0 in divisions of 1/10 to indicate the fraction of base current shown on S3. For instance, if S3 is set to 10 μA, and R2 is at 0.4, base current would be 4 μA.

Switches S1 and S2 are polarity-reversing switches to accommodate either PNP or NPN transistors. You might wonder why these two switches are not combined into one 4-pole switch. The use of two separate switches makes possible the testing of depletion-mode fets with no further increase in complexity. For instance, an N-channel fet is tested by placing S1 in the NPN position and S2 in the PNP position. The fet source, gate, and drain are connected to the curve tracer emitter, base, and collector terminals, respectively. The result is a positive drain supply and a negative gate voltage; the latter adjustable by R2. Negative gate voltage can be measured with a VTVM connected between gate and source, or base to ground terminals on the tracer.

For the $I_b = 0$ position of S3, the base is grounded. Switch S3 also provides a fifth position in which the base is left floating, which is handy for checking BV_{CEO} and I_{CEO}.

construction

This unit was built into a 2 x 4 x 6 inch (51 x

102 x 152 mm) chassis without much crowding of components; parts placement isn't critical.

A variety of transistor sockets can be wired in parallel to accommodate the various transistor basing arrangements in common use. It's also recommended that alligator clips on short leads be included for those diodes and transistors that will not fit into conventional sockets. These three leads should preferably be of different colors and clearly labeled E, B, and *C.*

Wirewound pots are recommended for R1 and R2. Both these controls have hand-drawn scales. The scale of R1 indicates collector (or anode) voltage in peak volts; the range is zero to roughly 200 volts peak.

operation

For best results the tracer should be used with a laboratory-type scope that has accurately calibrated vertical and horizontal deflection. I use a Hewlett-Packard 130B, which works beautifully in this application.

Nearly all work is done with vertical gain set at 10 mV per cm; the scope then reads 1 mA per cm vertically. Horizontal sensitivity, however, ranges all the way from 0.1 volt per cm to at times as much as 50 volts per cm.

Since the curve tracer operates at 60 Hz, it will tell you nothing about the frequency limitations of your transistors. To determine high frequency performance, an rf transistor tester is recommended.2.

The current-limiting resistors in series with the base make transistor damage unlikely. However, damage is possible if R1 is run too high. When testing an unknown device, it's wise to start with R1 set near minimum and keep an eye on the scope screen as the control is gradually advanced.

At times you'll notice that part of the characteristic curve drifts upwards. This is usually due to heating of the device under test. It doesn't normally occur unless the amount of heat generated is a significant fraction of the maximum device dissipation. It's particularly apparent with germanium transistors because of their considerable temperature sensitivity.

You'll also notice the characteristic curve occasionally takes the form of a very elongated loop rather than a single trace. This is because of what is called "temperature hysteresis" and is caused by cyclical heating of the transistor junction during collector voltage peaks.[3].

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ham radio

More Details? CHECK — OFF Page 94

digital logic probe

A discussion and severai designs for TTL and CMOS logic probes, featuring short pulse type memories

When electronic equipment consisted primarily of analog circuitry, most maintenance and troubleshooting could be handled with a simple volt-ohm-meter and some common sense. The VOM was the one instrument that could always be found on the bench of any ham or electronic experimenter. In addition to being generally useful, the VOM had much going for it. It was relatively small and easily handled. It was generally affordable even by Radio Amateurs of very modest means. Although it wasn't a precision instrument, if one knew how to use it, very good results could be obtained. By and large, it is still a most useful tool, but it's one whose relative importance has considerably diminished.

Complementing the VOM today, as the general purpose test instrument of digital circuitry, is the logic probe. Like the VOM, the logic probe is easy to handle, convenient, inexpensive, and, when used intelligently, capable of furnishing the needed troubleshooting information.

There isn't much to a logic probe. It is simply a device that will indicate the "state" of an accessible point in a digital circuit. Using some sort of quick response display, like an LED for example, the probe will indicate whether the voltage at the test point is high, low, or, perhaps, alternating. Does every ham and electronic experimenter need one now? Well, that's pretty much for the individual to decide, a decision to be based on what other equipment is around the shack and whether he does his own maintenance or pays someone else to do it.

There is no getting around the fact that each new piece of electronic gear hitting the market contains more digital circuitry than did its predecessor. Not too long ago the digital circuitry in the typical ham shack might have been limited to that in the electronic keyer or, if there was a new and expensive oscilloscope, in the trigger circuit of the horizontal sweep. Today, digital circuitry is the heart of the fully synthesized, VFO, of frequency counters, modern capacitance meters, fashionable readouts and displays, to name a few places, and it's becoming ever more commonplace.

For troubleshooting these digital circuits, it is pretty tough to find better instrumentation than the simple logic probe. Coupling that observation with the fact that a generally adequate logic probe can be quickly and easily assembled at a cost of anywhere between a few quarters and a few dollars, depending upon the status of your "junk" box, it's hard to justify not having one around the shop bench or shack. Just how elegant or sophisticated a probe is needed for any given shack is easily determined and, using one or more of the following circuit ideas, built.

If all of the digital circuitry currently in your shack operates on +5 Vdc and you have no particular interest in detecting very short pulses (10 μs or less) occurring at very low frequencies, then the very simplest of TTL (transistor transistor logic) probes should suffice. If your equipment is all in the 5-volt category, but there is a reasonable chance that you'll be looking for fleeting pulses, troubleshooting the triggered sweep of a modern oscilloscope is a prime example, then the logic probe should be slightly more elegant. The probe will still be based on a TTL integrated circuit, but some technique for capturing those elusive pulses need be added.

Some of the newest frequency synthesizers and frequency counters are built around LSI, or large-scale integrated circuits. Many of these are made up of mosfet rather than bipolar transistors. They may be operating on any voltage between about +4 and

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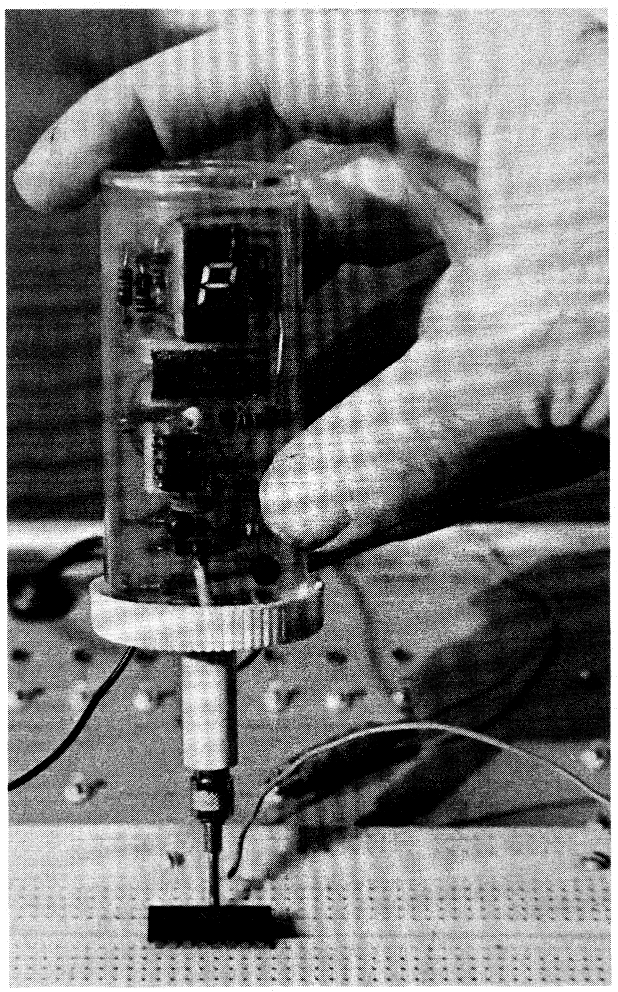

View of the digital logic probe showing the mounting and connections within the pill bottle.

+15 Vdc. A logic probe suitable for working on these circuits must be able to cope with the entire span of voltages. Such a probe is going to entail more than the basic probe discussed above. If the capability to detect very short pulses or noise spikes is desired, it will be still more elegant. However, it will be only slightly more complex. In fact, a probe capable of handling the full spectrum is so simple and little more complex than the simplest one that the only reasons for building the lesser one is that it's all you need for the foreseeable future, every part needed for it is in hand, and you have time to build it right now! Well, this very simple probe, the circuit of which is shown in **fig. 1,** has done yeoman service for me over five years and has on only the fewest of occasions been inadequate to a task at hand. It's not to be sold short.

simple logic probe

I previously pointed out that a logic probe is simply a device that will indicate the "state" of an accessible point in a digital circuit. Thus, the probe must have some sort of a readout, and its readout should be unambiguous. The probe ought to be convenient to use, and it must not upset or influence the circuit being observed. **Fig. 2** is a circuit diagram for a very simple logic probe. It should be adequate for most purposes, just as long as its use is limited to 5-Vdc circuits. With a minor exception, it's the circuit of a kit that was widely available a few years ago.

Isolation between the logic probe and the circuit under test is provided by the 10k resistor in the base of Q1, which can be any NPN switching transistor with a low-current beta equal to or greater than about 50. A 2N2222 would be just fine. All of the logic and the display drive is provided by an inexpensive 7404 hex inverter. The two LEDs make up the readout. LED A is turned on when the probe sees a high, LED B a low. If the test point has an alternating voltage, the two LEDs will blink alternately or appear to be both on depending upon the pulse frequency. Power to operate the probe is taken from the circuit under test. As each inverter of the 7404 can source about 15 to 20 mA, there's adequate current to drive the LEDs to a reasonable brightness. The current limiting resistors protect the 7404 and the LEDs. The whole thing is assembled in a salvaged plastic pill bottle or the like.

If the two LEDs are of different colors this logic probe, exactly as described, could be satisfactory. However, if only LEDs of the same color are available, the resultant readout might be something less than desirable. The logic probe is small, hand-held, and is frequently used to probe densely packed circuitry. It isn't advisable to let one's eyes stray too far from the tip of the probe if the risk of accidentally shorting a couple of pins together is to be avoided. This means that it is most advantageous if the readout is such as to permit reading "out of the corner of the eye." A distinctive difference in the readout for high or low is most desirable. An elegant and inexpensive way of accomplishing this is shown in the

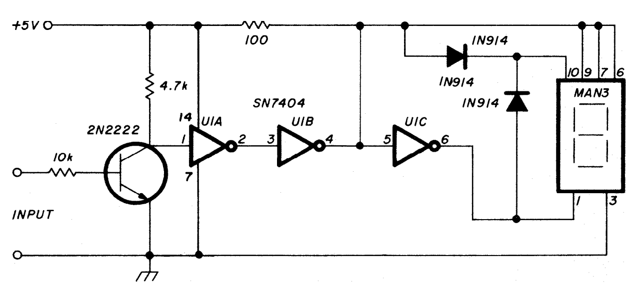

fig. 1. Schematic diagram of a simple logic probe for use with strictly 5-volt circuitry.

circuit of **fig. 1.** MAN 3A or MAN 3M LED readouts, or their equivalents, are available for as little as $1.00 (or less) on the surplus market. One of these devices coupled to the output of the SN7404 hex inverter, as shown in **fig. 1,** makes for a most ingenious little display. A high at the probe tip causes the hex inverter to drive the numeric display so that a *1* appears. A low at the probe tip results in a *0* being displayed. If there is a pulsating signal at the test point you will see a *P*.

I did not originate this circuit; it is merely one of several that were available as inexpensive kits over the past several years. The output transistors of the hex inverter source the current to drive the LED segments. As this circuit uses no current-limiting transistors between the 7404 and the display, a short in the latter will likely destroy the inverter. That's the primary weakness of an otherwise very clever circuit.

The MAN 3A was used because it was available. A common-anode display could be used with, of course, appropriate interchange of connections. The significant point is that the output transistor is capable of sinking up to 30 mA, so that you can use either a common-cathode or common-anode display. This is not true, as will be discusses later, if the TTL device is replaced with an MOS device. Referring again to the probe shown in **fig. 1,** the two leads are connected to the V_{CC} and ground of the circuit under test, and the probe is held against the test point. **Fig.** 3 shows the printed circuit board layout as seen from the foil side. The overall size of the board is tailored to fit snuggly into the pill bottle so it may be necessary to make some slight changes to the printed-circuit board layout to fit any given plastic bottle.

pulse memory

When the instrumentation target is a very fleeting positive going pulse, such as in the previously suggested example of the trigger circuit of a modern oscilloscope, or if you are trying to ferret out some suspect random noise pulses in that new desk top computer, the logic probe must see and retain the high long enough to produce a visible signal on the

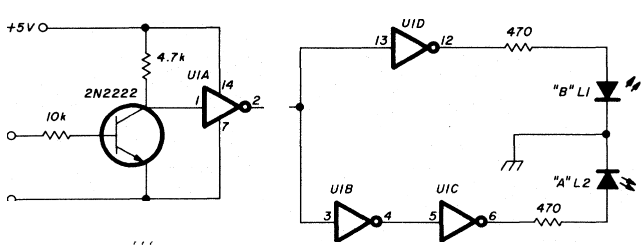

fig. 2. Schematic of the basic logic probe which used single LEDs to indicate either a high or low logic level. A cyclic signal will cause the LEDs to flash or both appear to be on due to the repetition rate.

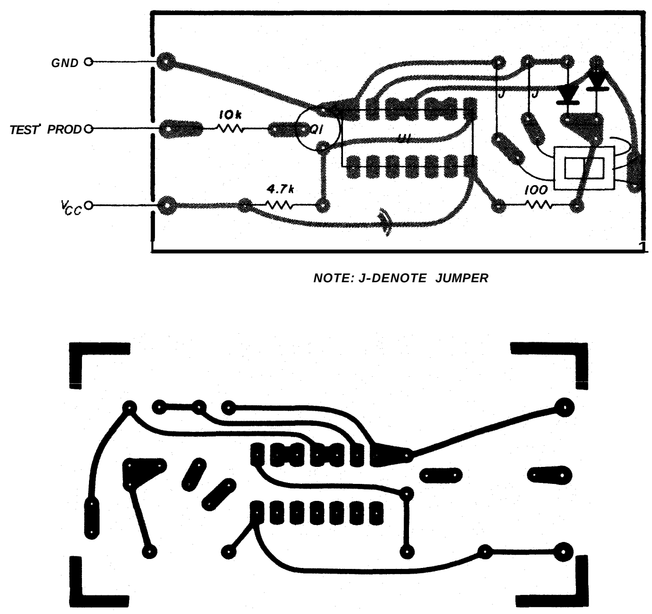

fig. 3. Etching pattern and parts placement diagram for the simple TTL logic probe.

LED display. An acceptable way of accomplishing this pulse "stretching" is through recourse to a one-shot multivibrator as a memory circuit. A very short incoming pulse, too short to be seen directly on the LED, is fed by the input amplifier of the probe to the input of the one shot, triggering it on. The output of the multivibrator is placed in parallel with one of the normal outputs of the hex inverter. The pulse duration at the output of the multivibrator is tailored by the time constants of the circuit so as to ensure a visible signal. Where the logic probe is to be used only on a +5 Vdc supply, and if there is an SN74121 in the junk box, a possible probe circuit is shown in **fig. 4.** The length of the output pulse duration needed to yield an acceptable display can be determined experimentally by varying R1, C1, or both.

versatile logic probe

An alternative short term memory uses the ubiquitous 555 timer. The connections for the 555 as a pulse stretcher is shown with a CMOS rather than a TTL hex inverter. Nevertheless, it can be used with the latter in exactly the same way. This, the most flexible and elegant logic probe, is presented in **fig. 5.** It can be used with any logic circuitry operating between approximately 4 and 15 Vdc. This means the logic probe can be used for RTL, DTL, TTL, or CMOS; in fact, for about any existing digital circuitry except for I^2L. On the lower side, the voltage limitation is the efficiency of the LED display. With prime LEDs, it might be possible to work down to about 3 volts. On the high side, the limitation is the upper limit on V_{DD} for the 4049 CMOS hex inverting buffer used as the logic chip and display driver. If a Fairchild F4049 is used, V_{DD} could safely go as high as 18 Vdc without damaging the probe. At any rate, it is most unlikely that the user will be confronted by voltages exceeding the range of 4.5 to about 13.8 volts so any 4049 will be acceptable and almost any surplus common-anode, seven-segment readout will work satisfactorily.

Being aware that the 74C04 and CD4069 CMOS hex inverters are pin compatible with the TTL 7404, one might well ask why go to the trouble of redesigning the circuit. Why not just replace the 7404 of the previously described logic probe with its CMOS counterpart and let the circuit go at that? There are two reasons why this cannot be done. The first has to do with the nature of the output mosfet of the CMOS chip, the second with the current limitation of the LED display.

In discussing the logic probe built around an SN7404 TTL hex inverter, note was made that the output transistors could each source about 15 mA or sink 30 mA. This permits the designer to select either a common-anode or common-cathode seven-segment LED display with the full confidence that there will be adequate current for safe direct drive of the LEDs. The 74C04 or 4069, on the other hand, are specified as being able to sink or source considerably less than 1 mA for +5 Vdc V_{DD} operation and 1.5 to 2 mA for 15-volt operation. The chip might be used with a common-cathode display, but the light intensity would be low. Used with a common-anode display, the CMOS output stage would quickly fail if it were forced to sink enough current for the LED to be acceptably visible, a function of the voltage applied to the common anode of the display and the size of the current-limiting resistors.

The problem is circumvented by turning to the CMOS 4049, a hex inverting buffer. These CMOS buffers provide both the necessary logic for the probe and a high current output capable of safely driving the LED load. It is not, however, as flexible as the TTL 7404. The CMOS buffer will typically sink about 5 mA with a V_{DD} of 5 volts and about 20 mA for a 15V V_{DD}. Under the same operating voltages, it will source only 1 to 3 mA. Thus, the TTL design option of using either a common-anode or a common-cathode configured display is closed; only a common-anode device can be used. How this is done is shown in the circuit in **fig. 5.**

The other major concern when designing the logic probe for this very wide range of operating voltages is the current limitations of the LEDs themselves. The generally useful current range of most LEDs is about 2 or 3 to 1. That is, starting with no current through the LED, current is gradually increased until first, the light output is barely adequate to be seen in a lighted room and then second until the LED fails. The current at failure will be about 2 to 3 times that at "visible." By the way, this isn't offered as a "scientific truth," but rather as an observation based on experience and generally supported by pertinent specification sheets. Q2 and ZD1 in **fig. 5** provide a voltage regulator whose output is applied to the common anode of the display. As the applied voltage at the V_{DD} lead of the probe is varied between +5 and +15 volts, the voltage at the output of the regulator varies between 4.4 and 6.2 Vdc. In the path between the output of the regulator and ground there is the 1N914, across which there will be about a 0.6-volt drop, the LED itself, which will account for a drop of 1.7 volts, and the current-limiting resistor which must make up the rest of the drop. The variation in voltage across the resistor, for the 4.4 to 6.2-volt swing, will thus be 2.1 to 3.9 volts, considerably less than 2:1 range. The LED current will be limited to the same range, one that is quite safe.

Three 1N914 diodes are shown between the 4049 and the LED readout in **fig. 5.** These diodes perform several functions so, unlike the diodes in **fig. 1,** cannot be replaced by slightly larger current-limiting resistors. This probe is designed to be used with operating voltages as high as 15 volts. Under this condition, and when the output of the buffer is in the high state, the output will approach 15 volts. Meanwhile, because of the voltage regulator, the anodes of the LEDs are close to 5 volts. The 1N914s protect the LEDs from what otherwise would be about a 10-volt reverse voltage, some 4 to 7 volts more than the maximum permitted according to the manufacturer's specifications. A second function of the diodes, or at least two of them, is to isolate the output mosfets of

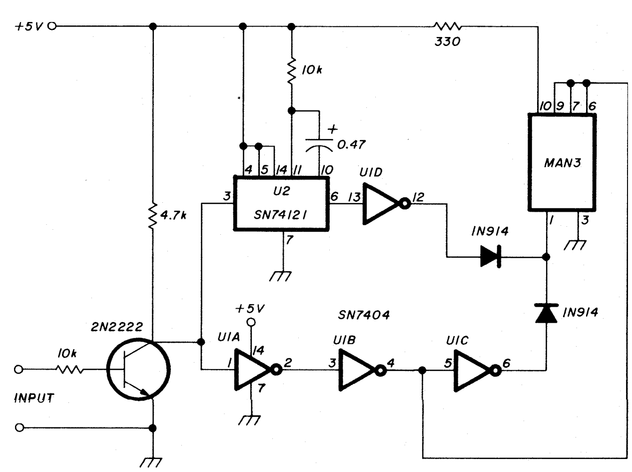

fig. 4. Diagram of the TTL probe with a short pulse memory. The monostable is used to capture any short-duration pulses for display on the LED display.

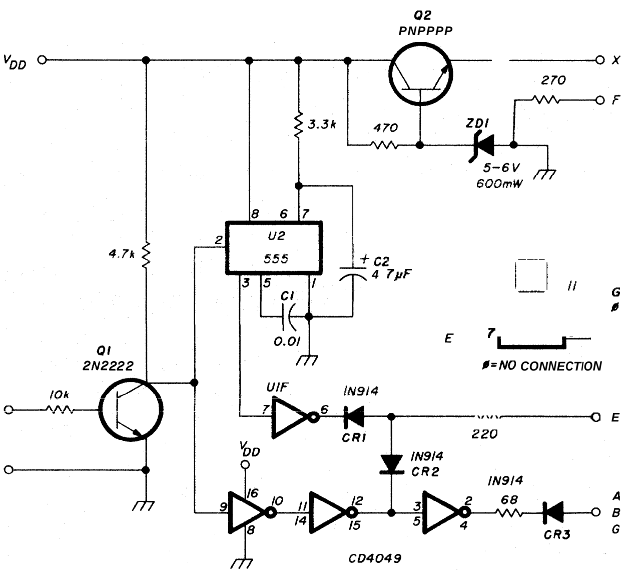

fig. 5. Schematic of the CMOS **type logic probe with a short pulse memory. A 555 timer is used as the memory element; a common-anode display must be used in this version.**

the inverter buffers from each other. Either of two inverters may go low to turn on segment E of the LED while the other is high. The diode isolation permits this to occur without risk to the 4049.

Unused inputs of CMOS ICs are never allowed to float. They are tied high, low, or to a used input. In the design of the circuit of **fig. 5** the inverters were simply paralleled as necessary so that no inputs were allowed to float.

Just as the pulse stretcher for the TTL-based logic probe design could have been a 74121, this CMOS-based design could as well be a CD4047A monostable/astable multivibrator. The 555 timer was used because it's smaller and was available; it is also less expensive. There's nothing unusual about the employment of the 555; the one-shot configuration is right out of the book for a negative going trigger input and a one-shot stretched output. The output pulse length is given as 1.1 RC where the RC applies to the resistor between V_{DD} and pins 6 and 7 and the capacitor between this point and ground. The component values shown on the schematic were found, experimentally, to give a pulse that was just long enough to barely flash the Litronix readout. For test purposes, 0.25-microsecond pulses were generated at a pulse frequency of one pulse per second. Readout visibility was very acceptable. The test circuit is described briefly at the end of this article.

Obviously, the readability of the output for a stretched pulse can be enhanced simply by increasing the RC time constant in the 555 timer circuit. In designing the probe, however, the duration of the stretched pulse was deliberately kept to the useful minimum; the probe readout differentiates between short pulses or noise pulses at low frequency reoccurrence rates and low frequency "clocking" phenomenon.

With a low-frequency, alternating state signal at the probe point (10 Hz or less), the readout will alternate between *1* and o. At a higher frequency and in particular where the duty cycle is between 20 and 80 per cent, the eye of the observer is fooled into seeing a steady P. For very short positive-going pulses at low-frequency rates, the display is a brief P followed by an extended ***o***. At higher frequencies, the display takes the form of a fairly bright o with a dim staff to form the ***P.***

None of the described logic probes will indicate the presence of a brief negative-going pulse. This is a design limitation accepted because I have never found need for that capability and because providing for it does cause some additional circuit complication. If the added capability is required, it can be achieved by modifying the CMOS logic probe as follows. Replace the 555 timer with a 556 dual timer. Isolate the two paralleled hex inverters. Connect the input of one of these inverters to the collector of Q1 and its output to the input of the added timer. Connect the output of the added timer to the input of the other freed inverter and its subsequent output to the cathode of an additional 1N914. The anode of the 1N914 connects to the anode of the existing 1N914 in the circuit coupled to the three LED segments that

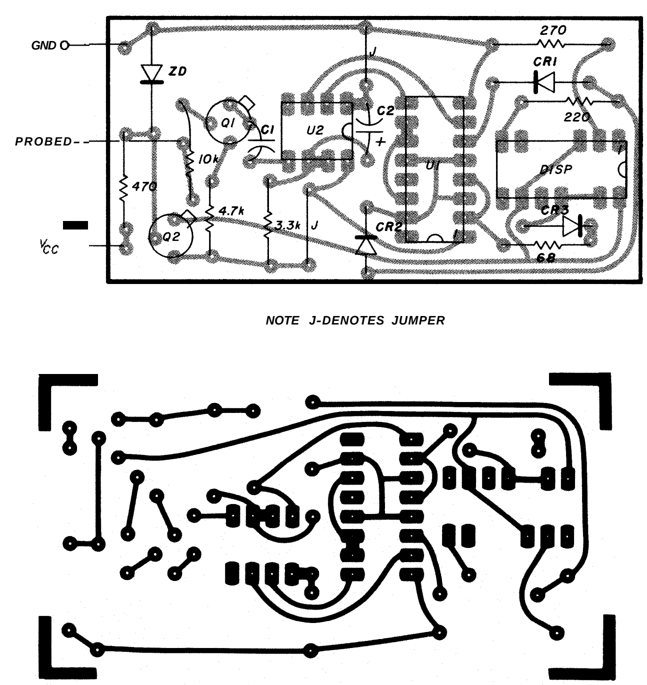

fig. 6. Printed circuit board layout and the parts placement diagram for the CMOS **logic probe.**

make up the switched element of the *o*. In the presence of an occasional negative-going pulse, the display should be a **I** changing to a P when each pulse appears. The passive components associated with each half of the 556 should be the same as those shown for the probe in **fig. 5.**

Having decided upon the circuit to be implemented, the next step is to collect the pill bottle to be used as the case. This is an important step because the size and shape of the pill bottle will determine the size and layout of the circuit. **Fig. 6** shows the circuit board layout of the CMOS logic probe described in detail above. It will be useful if your pill bottle will take a 1¼ x 2%-inch (3.0 x 6.0-cm) board.

Printed circuit board techniques were used for both probes shown in this article only because it was convenient to do so. Wire wrap techniques or even point-to-point wiring on sockets mounted in perf board would be just as good. The logic probe is fundamentally a low-frequency device. it would be difficult to find a poor construction technique as long as the workmanship is good!

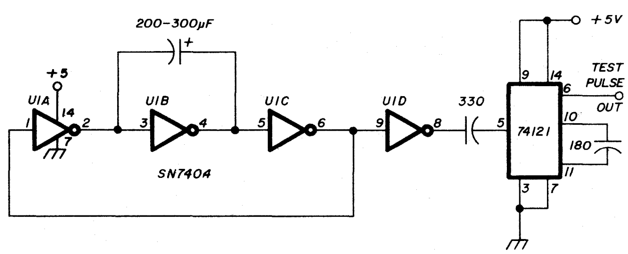

fig. 7. *Pulse* generator to test the short pulse memory capability of the logic probes.

Test of the completed circuit for all but the pulse stretching feature is easily accomplished. Connect the V_{DD} and V_{SS} or V_{CC} and ground wires, depending upon your choice of CMOS or TTL, to an appropriate power source. If everything is working properly the readout will display a ***o.*** Touch the probe to the positive voltage terminal of the power supply and the o should change to **I**. The circuit shown in **fig. 7** will test the pulse capture feature if one has been included. Pulse length of the output pulse of the 74121 is approximately $1400 \cdot C1$ seconds. If C1 is 180 pF, the pulse at the test point would be $1400 \cdot 180 \cdot 10^{-12}$ second or approximately 0.25 microsecond.

The logic probe is a very practical instrument to have around the shop or shack. If you don't have one and can squeeze out an evening, try one of the circuits presented here. It won't be long before you begin to wonder how you ever managed to get along without it!

ham radio

challenge for microwave-antenna designers

New ideas are needed for low-cost, efficient microwave antennas for satellite TV reception

Attention antenna-design enthusiasts: Will you be the person who develops a novel idea for low-cost antennas for satellite TV reception? There's a real challenge waiting to be met, and this challenge is a prime opportunity for Amateur Radio operators to make a significant contribution to an important new and growing segment of space-communications technology. Conventional parabolic-reflector antennas are too costly – new ideas are needed to reduce the cost of antennas with **40-50** dB gain at **4 GHz**.

TV receive-only terminals

There is a TV technological revolution underway, and the advanced hobbyists who are a part of that revolution have been searching for a gallium arsenide fet low-noise amplifier or a low-cost 12-foot (3.7-meter) parabolic reflector antenna for a home TV receiver terminal. Many hobbyists have succeeded in obtaining enough gain and a decent signal from a backyard antenna aimed at one of a dozen – soon to be more – satellites transmitting from above the equator.'.

These installations are called TV receive-only terminals, or home TVROs. Home TVROs are already the province of skilled and dedicated experimenters. The number of private terminals in operation is difficult to determine since very few are licensed, but some estimates are as high as 1000.2

Video programming on domestic satellites currently offers a great deal in the way of quality entertainment and information, and it will get even better in the future. Satellite distribution of these TV programs to private and shared receiver terminals allows anyone in this country to participate in a new and entertaining form of communications. Technical sessions at WESCON by J.C. Bacon,' J. Kinik,[2] H.P. Shuch,[3] and H.T. Howard[4] form the basis for much of the information in this article.

FCC deregulation

Much is happening in Washington to help give people these new opportunities.[1] Bills in both the Senate and House to rewrite the 1934 Communications Act have been introduced, deregulation trends are underway at the FCC, and the Executive Branch has initiated several actions to expedite these efforts.

The Justice Department has responded to the FCC in favor of deregulation of receive-only earth stations stating that the language of the 1934 Communications Act, which created the FCC, authorized regulation of transmitting devices – not receive only.[1]

The Federal Communications Commission decided on October 18, 1979, to drop its licensing requirements for satellite receiving stations.5 The action eliminates the requirement that persons constructing a receiving antenna have it coordinated to eliminate interference. It also ends a requirement that they

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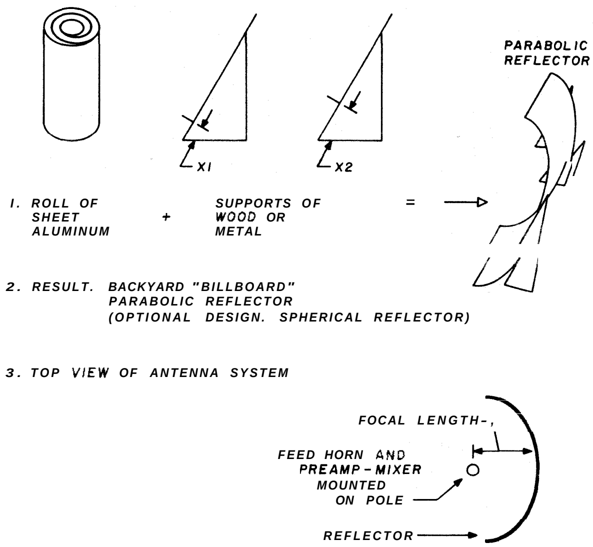

fig. 1. Billboard parabolic reflector construction for low-cost TVRO antenna.

obtain a construction permit and ultimately a license to operate the receiving station.

Those who want licenses, to obtain government protection from interference with the signals they receive, will still be able to apply, the commission said. But it also said the licensing will be entirely optional. The FCC said it took the action to eliminate the costs of the licensing process for builders of receiving stations and to end delays involved in obtaining a license.

FCC Chairman Charles D. Ferris noted that, while operators of unlicensed stations will not be protected from interference, this can normally be eliminated by relocating the station slightly, which is usually less costly than obtaining a license.[5] Operators of receiving stations will still have to obtain permission from the operators of the sending satellites to receive their transmissions.5

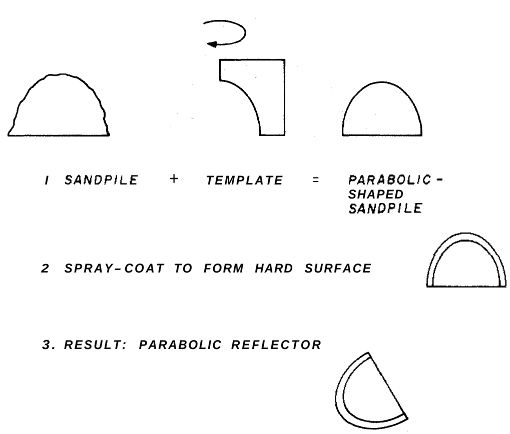

fig. 2. Sandpile parabolic reflector construction for low-cost TVRO antenna.

low-cost novel designs are needed

In January, 1977, the FCC ruling allowing the use of 4.5-meter diameter antennas for TV receive-only earth terminals created a strong need for low-cost antennas in the 4.5 to 6 meter size range for the first time.2

Two suppliers, Scientific Atlanta and Anixter-Mark, have invested in tooling to stamp out panels to the correct curvature on a mass production basis.2 Another supplier, United States Tower Company, has combined a fiberglass reflector with an aluminum

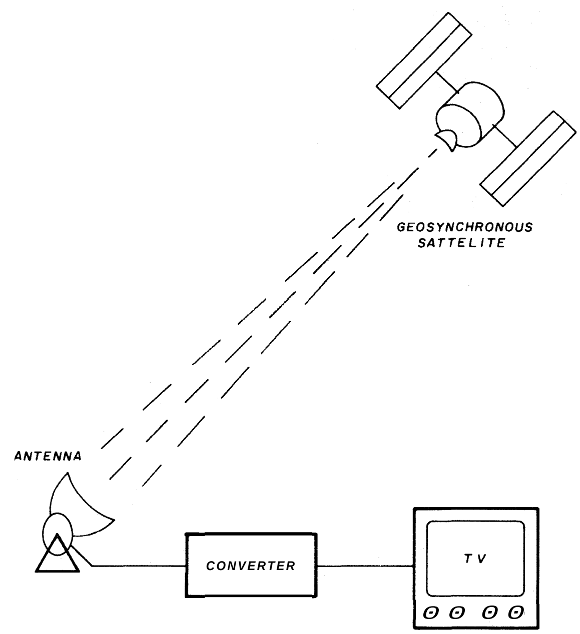

fig. 3. Typical TV receive-only (TVRO) terminal for receiving signals from geosynchronous communications satellites. Problem: Design a low-cost antenna of good structural integrity and adequate rf performance.

backup structure to realize a more cost-effective design. The nominal current price levels for the lowest-cost designs offered by the commercial suppliers are $1500 for 3-meter diameter, $4000 for 4-meter diameter, and $6000 for 5-meter diameter antennas.2 These antenna costs must be reduced, through novel antenna designs, so that the total cost of a complete TVRO terminal can be kept low enough to be afforded by nearly everyone.

satellite TV signals

The technological revolution which makes possible the distribution of television programming via satellite is based on receiving DOMSAT (Domestic Communications Satellite) signals.[3] The downlink band used by most North American DOMSATs is 500 MHz wide, and for a given antenna polarization there will be present up to twelve video carriers spaced 40 MHz apart. These signals are of extremely low amplitude, and this complicates the design of TVRO antennas.

It has been shown that, for the illumination contours typical of most North American DOMSATs,[3] an optimum private-terminal antenna will exhibit on the order of +41 dBi gain.' Given the signal power (EIRP) and path loss numbers listed in **table 1,** the signal level available to the low-noise amplifier will be on the order of – 90 dBm.

technical requirements

The dominant requirement for private TVRO antennas is low cost, while of course retaining reasonable structural integrity and adequate rf performance. These three considerations establish the baseline for a set of requirements, but the process of arriving at a set of such requirements is one which is more practical than scientific — thereby creating a challenging opportunity for ham radio operators and experimenters who can make a significant contribution to an important new and growing segment of space-communications technology.

A low-cost design must not require high-cost fabrication methods, tooling, or labor. It should also have a minimum weight and volume to keep costs down, and it should be designed with ease of installation in mind to avoid expensive hoisting equipment. Total installation time should be kept to a minimum. An additional requirement is that the design should be amenable to kit construction techniques. These goals, if met, can be combined with gallium arsenide integrated circuit technology6 to develop a superior TVRO terminal.

table 1. Typical DOMSAT **signal characteristics (from reference 3).**

video carrier	
channels	24
adjacent channel spacing	40 MHz
orthogonal channel spacing	20 MHz
frequency band	3.7-4.2 MHz
peak deviation	10.25 MHz
maximum video frequency	4.2 MHz
pre-emphasis curve	CCIR 405-1
audio subcarrier	
frequency	6.8 MHz
peak deviation	75 kHz
maximum audio frequency	15 kHz
pre-emphasis time constant	75 μSec
composite	
EIRP	+65 dBm
path loss	– 196 dB
99% power bandwidth	36 MHz
received spectral density	– 206 dBm/Hz

"Gain referred to an isotropic source. **Editor.**

table 2. Technical performance goals for 3.7-4.2 MHz TVRO antenna.

	required minimum	desired
antenna gain	40 dBi	45 dBi
rf efficiency	45%	55%
wind survival	75 mph	75-100 rnph

new design ideas?

The optimum low-cost TVRO antenna has yet to be designed and developed. This is an active area of research, and some new ideas are beginning to emerge. One idea, shown in **fig. 1,** requires only low-cost materials and requires no expensive metal shaping. A second idea is illustrated in **fig.** 2. This clever design is based on an idea by John, K6EJF, who suggested using spray-on material of the type often used for coating swimming pools. The template can be made of plywood and a guide-pin or rod can be driven into the sandpile for attachment to the template.

New microwave antenna designs and discoveries are bursting forth at a rapid rate. An example of this is the discovery that the snow sled saucers sold as kids' toys exhibit 22 dB of gain at S band![4] A similar antenna was constructed from a child's 25-inch (64-cm) snow sled saucer and a feed horn was made from a one-pound coffee can. The saucer is not a true parabola but is close enough to give 15+ dB of gain at 2 GHz.[7]

Will you be the person who develops a new idea for low-cost antennas for satellite TV reception? I'd like to hear from you if you have a clever idea for the construction of a new low-cost antenna. Your information may be useful during the preparation of a subsequent article on the subject of TVRO terminals. Write to Dr. D.H. Phillips, 1345 Arizona Avenue, Milpitas, California 95035.

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ham radio

calculating the cascade intercept point of communications receivers

New equations for calculating a receiver's cascade intercept point are a powerful design tool

On today's heavily used Amateur bands which have many extremely strong signals, receivers with high dynamic range are required. Many articles have treated the considerations and problems of designing a high dynamic range receiver from the circuitry point of view, but a systematic approach to receiver design seems to be lacking.

receiver system design

The best way to approach a receiver design problem is with a block diagram. By identifying the various functional blocks in a receiver, the critical parameters for dynamic range (input intercept point, noise figure, and bandwidth) can be predicted for the overall receiver.

Dynamic range may be defined as

$$DR = \frac{2}{3}(IP_{i3} - MDS) \tag{1}$$

where DR = spurious-free dynamic range, dB
IP_{i3} = third-order input intercept point, dBm
MDS = minimum detectable signal (noise floor), dBm

For a system at room temperature the minimum detectable signal is

$$MDS = -174\,dBm + NF_t + 10\log BW_n \tag{2}$$

where NF_t = overall system noise figure, dB
BW_n = system noise bandwidth, Hz

Note that the system noise bandwidth is usually well approximated by the 3-dB bandwidth of the narrowest filter in the system. The total (or cascade) noise figure of a system is

$$F_t = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_2 G_1} + \ldots \tag{3}$$

$$NF = 10\log F_t \tag{4}$$

The minimum discernible signal can be calculated from the last three equations, but some method is needed to predict the system's input intercept.

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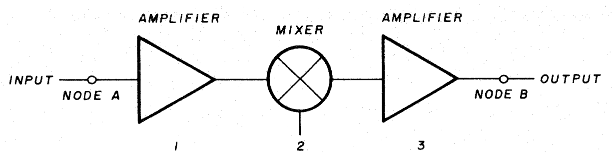


	1	2	3
Gain	+ 10 dB	− 7 dB	+ 20 dB
IP_{o3}	+ 20 dBm	+20 dBm	+ 50 dBm

fig. 1. Simple converter used to illustrate cascade IMD equations. Note that although only two nodes are specified, there are two other nodes (on either side of the mixer) that could be used for intercept point calculations.

cascade intercept point

To obtain the intercept point for a system, the intercept points of the various functional blocks will be combined in such a way as to predict the input or output intercept of a system.

There are two ways of approaching the cascade intercept point equations. The first is to assume the intermodulation products are coherent; when the products are coherent their voltages wil add in phase. The second approach is to assume the intermodulation products are non-coherent; in this case their voltages will combine as a sum of squares.

The assumption of coherence will always result in the lower predicted intercept point and for most Amateur applications is the preferred approach. There are situations, however, where the assumption of non-coherence is reasonable; the most obvious situation is a microwave system where phase shifts between system elements may place the products out of phase.

coherent summation

The equation for coherent summation* is

$$IP_t = \frac{20}{n-1} \log \left(10^{\frac{-(n-1)IP_1}{20}} + 10^{\frac{-(n-1)IP_2}{20}} + \ldots\right)^{-1} \qquad \textbf{(5)}$$

where n is the order of the intermod (2 for second order, 3 for third order, etc.), and

IP_m is the reflected intercept (of the appropriate order) of the mth element

All the intercepts of the various system elements must be reflected to a single node. The example in **fig.** 1 will help clarify this.

First a table must be drawn up **(table 1)** that contains the reflected intercept points. Note that input intercept plus gain equals output intercept $(IP_i + G = IP_o)$.

Substituting the information in **table** 1 into **eq. 5,** the input intercept (node A) turns out to be $IP_{it} = +9.14\ dBm$. The output intercept (node B) is $IP_{ot} = +32.14\ dBm$.

table 1. Table of reflected intercept points for the system of *fig. 1*. This listing not only gives the intercept information, but also pinpoints the weakest elements in a system, in this case element 1.

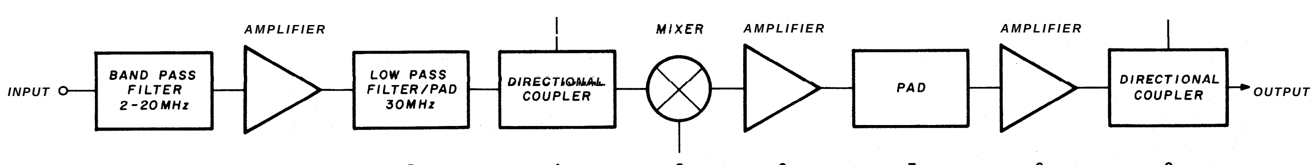
	reflected intercept point	
element	**node A**	**node B**
1	+ 10 dBm	+ 33 dBm
2	+ 17 dBm	+ 40 dBm
3	+ 27 dBm	+ 50 dBm

non-coherent summation

For the case of non-coherent summation the cascade intercept point equation is

$$IP_t = \frac{10}{n-1} \log \left(10^{\frac{-(n-1)IP_1}{10}} + 10^{\frac{-(n-1)IP_2}{10}} + \ldots\right)^{-1} \qquad \textbf{(6)}$$

If non-coherent summation was assumed for the system of **fig.** 1, the input intercept becomes



	1	2	3	4	5	6	7	8	9
Gain	−0.43 dB	+ 11.08 dB	− 1.3 dB	− .25 dB	− 7.0 dB	+ 10 dB	− 4.0 dB	+ 9.5 dB	− 0.25 dB
IP_{i3}	—	+ 33.92 dBm	—	—	+ 29.75 dBm	+ 22.0 dBm	—	+ 22.5 dBm	—
IP_{i2}	—	+ 63.92 dBm	—	—	+ 62.0 dBm	—	—	—	—
NF	0.43 dB	4.5 dB	1.3 dB	.25 dB	7.0 dB	4.5 dB	4.0 dB	5.0 dB	.25 dB

fig. 2. Block diagram of a receiving converter built by the author with performance data for each of the elements in the system. When a dash is used in place of data, it indicates no contribution from that particular system element.

$IP_{it} = +9.91\ dBm$ and the output intercept becomes $IP_{ot} = +32.91\ dBm$. In this simple example the results from eq. 5 and **eq.** 6 are nearly identical. In general, however, as the system becomes more complicated, the difference between the two will be much larger.

receiving converter

The block diagram of fig. 2 represents a receiving converter I built. It provides a convenient example system for comparing the predicted and measured parameters. Using the data in fig. 2, a table of reflected intercept points can be drawn up (table **2**); by using the formula for coherent summation (**eq. 5**) the resultant input intercept can be calculated, and from **eq.** 4, the system noise figure can be calculated.

table 2. Calculated intercept point for the receiving circuit shown in *fig.* 2. Measured intercept point is shown in parentheses for comparison and shows good agreement.

element	reflected intercept point IP_{i2}	IP_{i3}
2	+64.35 dBm	+34.35 Bm
5	+52.90 dBm	+20.65 dBm
6	—	+19.90 dBm
8	—	+13.90 dBm
IP_{it}	+50.84 dBm	+12.22 dBm
	(+54.0 dBm)	(+13.0 dBm)

Using eq. 4 the predicted system noise figure is 7.1 dB; dynamic range can now be predicted. If a 500-Hz filter were placed after the system shown in fig. 2, then the $MDS = -139.91\ dBm$ and the dynamic range would be 101.42 dB.

Two-tone and noise figure tests were run on the system of fig. 2 which resulted in measured $IP_{i2} = +54.0\ dBm$, $IP_{i3} = +13.0\ dBm$, and $NF_t = 6.9\ dB$. These tests show excellent agreement with the predicted data.

conclusion

The cascade intercept point equations, when used in coombination with the cascade noise figure equation, provide a powerful tool for rf system design. Derivations of **eq.** 5 and **eq.** 6 will be sent to interested readers upon receipt of a self-addressed, stamped envelope. With these equations a receiver designer can predict a system's characteristics without investing in any hardware.

acknowledgments

My thanks to Rich Phillips of ARGOSystems (formerly of ESL, Inc.) for initially pointing out the utility of these equations. My thanks also to Wes Hayward, W7ZOI, for kindling my interest in receiver design.

ham radio

More Details? **CHECK—OFF** Page 94

diode frequency divider

Using diodes as voltage-variable capacitors to produce a sine wave at one-half the input frequency

The use of diodes as frequency multipliers, and particularly as doublers, has been well known for a good many years. Such applications are covered quite well in a recent ARRL publication1 and provide many ideas to the Amateur builder and experimenter. However, one application for diodes that I don't recall finding in an Amateur publication is their use in a frequency divider, with what is practically the same circuit!

Consider **fig. 1A**, which is a standard full-wave rectifier. It is familiar to just about every ham as is the output waveform, **fig. 1B**. The fundamental frequency has been cancelled by the full-wave circuit, and a quite respectable frequency doubler is the result.

frequency-divider circuit

Now let's change the circuit slightly to that of **fig. 2.** We will feed a signal of frequency **f** through a blocking capacitor to the diodes, reversing the direction taken when the circuit was a rectifier. There is another difference, as well, as the clamping action of the diodes builds up a bias voltage across the blocking capacitor, and the diodes are operating in their nonconducting range. Furthermore, the center-tapped transformer has become a center-tapped tank circuit by the addition of a capacitor that tunes the circuit to $f/2$, one-half the input frequency.

Operating in their nonconducting range, the diodes present only their junction capacitance to the circuit and are now regarded as capacitors. Even though both diodes may be of the same type and rating, their junction capacitances will not be identical, so a voltage will build up on one end of the tank circuit. The voltage on the other end will be of opposite polarity, increasing the reverse voltage on that diode, further increasing any differences in the capacitances. When the next input pulse appears, the polarity of the voltages on the tank circuit will have reversed due to the flywheel effect of the tuned tank. Thus, successive input pulses will affect alternate ends of the tank, resulting in a signal of half the input frequency. Input power transfer will probably

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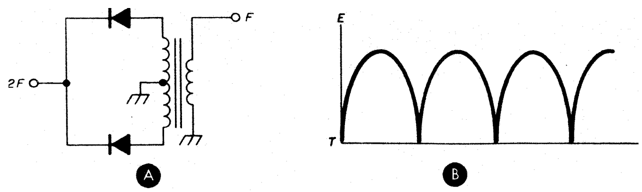

fig. 1. Standard full-wave rectifier (A) and the typical output waveform it produces (B).

be improved if some form of input matching network is used.

The power limitations of this circuit are a function of the diode characteristics, such as junction capacitance, leakage, and peak inverse voltage. Of course, power varactors are available at several bucks a throw, which would probably be more predictable in their operation; but silicon diodes out of your junk box may be pressed into service for a tryout.

suggested applications

A frequency divider such as this should offer a quick means of giving 160 meters a whirl, using the 80-meter rig, and without the necessity of building a separate new transmitter! My first use of the circuit involved a pair of top-hat diodes to reduce a 910-kHz

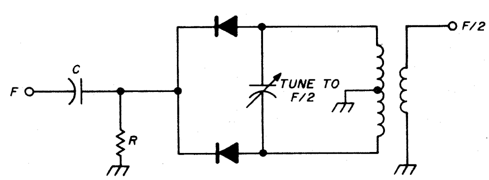

fig. 2. The diode frequency divider. A signal, *f*, is applied through a blocking capacitor, which reverses the signal direction. The center-tapped transformer is now a center-tapped tank circuit with the addition of a variable capacitor. Circuit delivers a good sine wave with good efficiency.

oscillator signal to 455 kHz in experiments with DSB reception. Unless the diodes are badly mismatched, it works right off. The value of resistor ***R*** of **fig. 2** should be quite high, as we are interested only in biasing the diodes into the non-conductive region. With a good pair of diodes, resistances up to 100 k would seem a reasonable figure.

This circuit delivers a good sine-wave signal with good efficiency, as there is little in the circuit to dissipate input power when the diodes function as voltage-variable capacitors.

reference

1. Hayward and DeMaw, ***Solid State Design for the Radio Amateur,*** Chapter 3, ARRL, Newington, Connecticut.

ham radio

More Details? CHECK—OFF Page 94

an accurate and practical AFSK generator

Putting the Exar XR-2206C IC to work in a circuit for RTTY enthusiasts

It's **no** secret that Amateur RTTY is enjoying a huge rate of growth and energy. Much of this interest can be directly traced to the newer video display type of TTY terminal and its quiet fascination. This same upsurge has caused many old-time RTTYers to dust off the mechanical machine and join in. Regardless of the type of terminal used, electronic or mechanical, the operator must provide the tone demodulation and FSK generator between his terminal and the radio gear. Most high-frequency stations use an AFSK audio input to an SSB transmitter, thus creating a need for a good AFSK generator. A great many circuits have been developed to fulfill this need, both simple and complex.

To fulfill my need for an AFSK generator, I looked over what had been designed and found either the 555 IC type oscillator or the crystal-controlled system. The former is not known for best stability with time and/or temperature, and the latter sometimes deserves a Nobel prize for complexity and would not fit the space requirements in my new converter.

Some time ago a data sheet came across my desk on a function generator in one IC package, made by Exar.* If that data sheet was to be believed, my answer was in the XR-2206C. After thoroughly testing the final circuit (fig. 1), I believe this AFSK generator is the most accurate and simply practical circuit possible considering stability, space requirements, and cost.

the XR-2206C IC

The device is a function generator designed for instrumentation and communications use. It will operate from a single supply range of 10-26 volts, or a split supply of ±5 to ±13 volts. Its stability is excellent; drift rate is 20 ppm/°C. It produces very-low-distortion sine, square, triangular, ramp, or pulse waveforms. And it's ready made for FSK operation with a built-in switch to select between two timing resistors for two-frequency output. In this FSK operation, the output is phase-continuous during frequency transitions, so distortion never results during switching (a common source of trouble for many of the simpler circuits).

AFSK generator

Fig. 1 shows the simplicity of the AFSK generator. It has six trimpots (four for setting frequency and two for IC controls). Supply voltage indicated is ±12 volts. This was the supply in use for my converter, and it was borrowed to operate the AFSK generator as well.

Sinewave output from the XR-2206C is selected by connecting the 200-ohm resistor between pins 13 and 14. (If this resistor is removed, the output becomes triangular.) Pin 1 is used to set overall gain in this circuit by trimpot R5. The dc offset of an internal amplifier is set by trimpot R6 at pin 3. The 1-μF tantalum capacitor bypasses an internal reference voltage at pin 10. The value isn't critical, but a tantalum type is definitely needed here.

stability considerations

The IC data sheet indicated that, for optimum temperature stability, the timing resistors should be as close as possible to 10k. The timing-capacitor value is then adjusted to yield the desired output frequency.

Working through Exar's formulas, a capacitor value of just less than 0.05 μF is required, so I connected two 0.022-μF caps in parallel; the result turned out to be just right.

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*R-Ohm Corporation, Exar Integrated Systems, P.O. Box 4455, Irvine, California 92664.

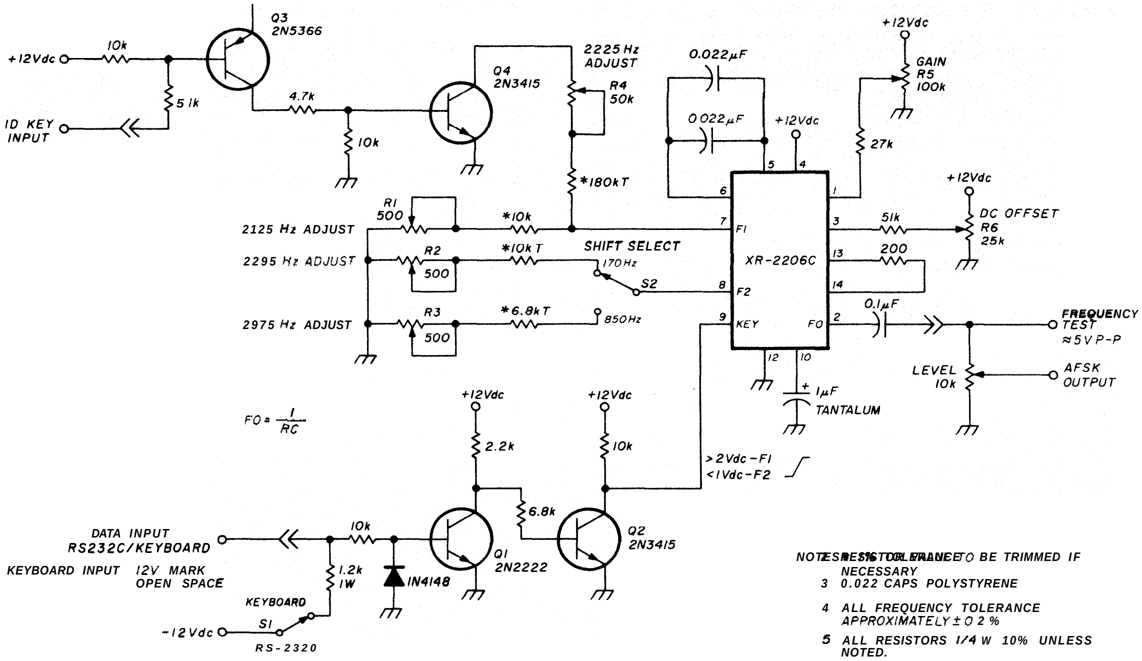

fig. 1. Schematic diagram of the AFSK generator. Circuit is built around the XR-2206C, which is a function-generator IC designed for instrumentation and communications use. This device has excellent frequency stability and low-distortion output. The circuit features optional CW IDENT circuit. Values of timing capacitors are critical; they must be polystyrene for maximum stability.

The timing-capacitor combination is connected to pins 5 and 6. These caps ***must*** be polystyrene for best stability. Do not use disc or Mylar caps in this application.

The timing-resistor networks connect to pins 7 and 8, with pin 7 being the F_1 frequency and pin 8 the F_2 frequency. We'll designate F_1 as mark (2125 Hz) and F_2 as space (2295 or 2975 Hz). Two resistors are selected by S2 for either 170-Hz or 850-Hz shifts. Obviously, if only 170 Hz shift is needed, delete the unnecessary components.

CW IDENT

I desired a CW IDENT feature, so I added Q3 and Q4. When the ID key input is pulled to ground or nearly so, R4 and 180-k resistor are in parallel with the 2125-Hz timing resistors, which shifts the output frequency upward 100 Hz for identification. Again, if this feature isn't needed, simply delete this little circuit.

Frequencies F_1 and F_2 (mark and space) are switched by the input at pin 9. If the level at pin 9 is greater than about 2 volts, F_1 is selected; if the level is less than about 1 volt, F_2 is selected.

input level translator

Q1 and Q2 act as an input level translator and switch for either RS-232C or mechanical keyboard inputs. This feature allows the generator to be used by the computer world as well as by traditional equipment. S1 is opened for RS-232C input signals, which will switch between +10 and −10 volts. A keyboard should be wired as shown to +12 volts (or the A+ level being used), and S1 is closed. This action applies −12 volts through the 1.2 k resistor to the keyboard, which applies 24 volts at 20 mA across the keyboard contacts. A high level at the DATA INPUT will cause F_1 to be selected; a low level (or ground) will cause F_2 to be output.

construction

Construction of the AFSK generator can be by any method convenient to the builder. Layout is anything but critical. Leave room to trim the fixed timing resistors from pins 7 and 8 if necessary.

tune up

Tuning the generator will require a frequency counter and oscilloscope. Frequency setting could be done by applying the output through a known accurate tone demodulator and tuning for maximum output levels, but a counter sets frequency precisely. A scope will be needed to adjust for minimum distortion and best waveform.

Begin tuning by opening S1 and setting all trimpots to mid range.

1. Open the ID KEY input, if used. Apply a positive-level (+10 or +12 volts) to the DATA INPUT point, then apply voltage power to the circuit. There should be some kind of waveform at pin 2 or at the output side of the 0.1-μF coupling capacitor.

2. Initially, adjust R5 and R6 for best waveform. With a 12-volt supply, the output level will be around 5 volts p-p.

3. Next, adjust R1 throughout its range to determine if 2125 Hz can be set with the values as shown. If not, center R1 and trim the 10k fixed series resistor for about 2125 Hz; then readjust R1. If the 10k resistor value must be shifted by more than 5 per cent, trim the timing capacitors with other values. Try to keep the 2125-Hz resistors as close to 10k as possible. Once the mark frequency has been set, do not readjust R1 for any other frequency.

4. If the CW IDENT circuit has been added, ground the ID KEY input. Adjust R4 for 2225 Hz, trimming the 180k fixed resistor as required.

5. Next, either ground the DATA INPUT point or apply a negative voltage level.

6. Set S2 to the 170-Hz shift side and adjust R2 for 2295 Hz.

7. Trim the 10k fixed resistor with a 150k resistor to begin with, and trim from there.

8. Repeat the procedure with S2 in the 850-Hz position, adjusting R3 and trimming the 6.8k resistor as needed.

9. After all four frequencies have been initially adjusted, let the generator run for an hour or so, then carefully reset each frequency. A drift of only ±2 Hz can be expected over a long-term period and over a wide temperature swing.

10. Carefully look at the output waveform and adjust R5 and R6 for the most perfect and smooth sine waveform possible. Set gain trimpot R5 for ***just less*** than maximum perfect waveform level.

closing remarks

Considering the space this generator consumes inside a typical RTTY converter cabinet, and the fact that its stability is better than 0.2 per cent over a wide temperature and time range (if 1 per cent resistor and 2.5 per cent polystyrene capacitors are used), this circuit offers much in terms of simplicity and accuracy.

ham radio

More Details? CHECK-OFF Page 94

notes on the EIMAC 5CX1500A power pentode

Some operating tips on the use of this tube in ham gear

Ask any experienced ham to list the tubes most likely to be used in a linear amplifier in the Amateur service, and chances are that the 5CX1500A won't be mentioned.

And, small wonder. In this day of zero bias, high-mu triodes, the 5CX1500A doesn't really seem to be the ideal tube for ham use. It's expensive (about $500); its socket is expensive and, in this day of single-power-supply-tubes the 5CX1500 requires *three* power supplies plus filament voltage.

the case for the 5CX1500A

So, why even consider it here? For two reasons. First, at least one manufacturer of high-power linear amplifiers, Tempo, uses this tube in their Model 4K. The typical 4K owner may not know enough about his final for his own peace of mind. Second, this tube is rather common in the broadcast service, and some of these tubes, with reduced emission, have become available at reasonable prices at swapfests and such." It is for these reasons that this article is presented.

background

The 5CX1500A was designed by EIMAC about a dozen years ago at the request of an American manufacturer of fm transmitters for the 88-108 MHz broadcast band. The need was for a final amplifier or driver tube that would operate in the 2-3 kW power range, Class B or C service, with good stability and ruggedness, a reasonable life expectancy (about 8000 hours in continuous commercial service [CCS] and high efficiency. The original 5CX1500 was the result.

There are some problems, of course. The '1500A is a beam-power pentode, and because of its parameters, which were dictated by the specifications listed above, the tube is very difficult to construct. As a result, EIMAC is still the only manufacturer of the tube in the world. Couple this with the fact that every major manufacturer of broadcast fm transmitters in this country uses the 5CX1500A in all their transmitters designed for this power level, and you can guess the rest. The tube is not always readily available. As of this writing, however, that situation has not existed for some months.

early tube problems

Several years ago the assembly line for the 5CX1500A was moved from EIMAC's main plant in California to a new facility in Salt Lake City, Utah. Shortly thereafter, problems arose. Tube life in the field began to drop, particularly in rf driver service, but later in all fm broadcast service. Tubes began to lose emission to the point where they had to be replaced after about 3000-4000 hours. EIMAC and the broadcast equipment manufacturers began to research the problems, and two differing causes began to emerge. There was one comon denominator: the filament was being "poisoned" by gas."

First it was discovered that the tube, especially when used in rf driver service (CCS), was being loaded much too lightly. This action resulted in high rf circulating currents in the tube, particularly across the aluminum oxide insulating ring between the suppressor ring and the anode. As a result undue heating occurred, which caused minute cracking of the ring. These cracks are not visible to the naked eye, except through the use of the special dye applied to the ring.

Second, investigators found that, on tubes made between mid-1975 and early 1979, the metal alloy used in the construction of the screen grid emitted

*Another source of these and other high-power tubes is your local fm broadcast or TV station. The tubes are replaced after a specified number of operating hours and in many cases have a lot of life in them, especially when used in the Amateur service. Contact your local station and talk to the engineer in charge. **Editor.**

*An interesting sidelight on the problem of contamination of thoriated tungsten filaments by gas is discussed in reference 1. **Editor.**

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excessive levels of carbon-monoxide gas, which is lethal to your typical 5CX1500A cathode. EIMAC corrected that problem early in 1979, and tubes manufactured after that date show no ill effects from that quarter.

operating tips

Now, in practical terms, what does this all mean to present and potential users of the 5CX1500A? Here are some tips on its operation that might help.

1. Load the tube as heavily as possible, consistent with the ability of your power supply to deliver the extra current. If your plate impedance is over 6000 ohms, it's too high! Reduce this impedance as much as possible by decreasing plate voltage and increasing current to reduce circulating rf currents in the tube.

2. If your tube is beginning to go "soft," determine if the problem is loss of cathode emission. To do this, record your present current drain on all tube elements that show current, and compare them to your observations when the tube was "fresh." If *all* currents are down, then the problem is low cathode emission caused by poisoning (contamination). These currents must be determined with drive power applied. Dc values alone will tell you nothing, since at radio frequencies the peak current drawn by the cathode may be 2-3 amperes. It's the inability of the cathode to deliver *that* amount of current that causes the tube to be considered "soft," no matter what the dc values may be.

3. Do *not* try to increase the cathode voltage above 5.1 volts to increase emission. For every 114 volt the filament is increased over its specified value, expect your tube life to drop in half (i.e., 5-114 volts, 4000 hours on an original 8000-hour tube; 5-1/2 volts, 2000 hours, *etc.*). Remember, this is a "carburized" thoriated tungsten filament. At the present state of the art, if the filament opens up, it can't be rebuilt.

4. If the tube is too "soft" to live with but seems to be OK otherwise (no short circuits), it can be rebuilt for about half the cost of a new tube. If indeed the cathode has been contaminated as determined in 2. above, you can ship it to Econco Broadcast Service, 1302 Commerce Avenue, Woodland, California, 35695. Unlike the process used in rebuilding other tube types, rebuilding the 5CX1500A doesn't usually require replacement of its grids. Instead, as EIMAC informs me, a process called "recarburization" is used. The tube seal is broken and a gas with a high carbon content, such as methane, is admitted. The gas-loaded tube is then fired in an oven, or its filament run at 120 percent voltage for a few seconds, during which time a new carbon coating is deposited onto the cathode. The tube time is then re-evacuated and resealed. Generally, if the tube has no other problems, the renewability rate is in the 80-90 percent range. If your tube loses here, you pay nothing more than shipping charges one way. By the way, this rebuilding process can be done more than once, thereby increasing the life of the tube in your rig.

5. If your 5CX1500A develops a short circuit, it will yeneraliy be from cathode to control grid. That's a pretty safe statement, considering the fact that the control grid is located a mere four ***mils*** from the cathode (the grid wire mesh is one mil thicker than that!). As the tube ages, the cathode can get brittle, and a strand from the filament may break away and fall across the grid.

Don't dismiss the idea of burning out the short in this case. I've heard of this happening several times, and a car battery is ideal for the purpose. If the short does not disappear, you'll have to admit that, with a little polishing and a walnut base, the tube makes a nice looking award for "Ham of the Year" at your local radio club. (This is particularly true of tubes manufactured before 1980, which are silver plated. EIMAC has changed the outer plating of the tube to nickel for cost reasons. There is no noticeable electrical effect on the tube.)

summary

The 5CX1500A tube may not be the best of all possible worlds for linear amplification in the Amateur service. However, for those who want a stable, very conservative amplifier for up to, say, 225 MHz, or for those who already are using this tube in such an amplifier, I hope this article has shed some new light on a tube that few Amateurs seem to know muchabout. The interested reader is referred, for further information, to "The Care and Feeding of Power Grid Tubes," by EIMAC. Data sheets for the 5CX1500A are also available from EIMAC.*

acknowledgments

My thanks to Ken Atkinson at EIMAC/Salt Lake City, Ed Numerych at EIMAC/Chicago, Dave Gilden, formerly of CECO, Ray Shurtz at ECONCO, and Bill Hoyt, Lou Pifer, and Bob Gorjance at Harris Corporation, for their help in providing information for this article.

reference

1. Alf Wilson, W6NIF, "Rejuvenating Transmitting Tubes with Thoriated-Tungsten Filaments," *ham radio,* August, 1978, page 80.

*EIMAC division of Varian, 301 Industrial Way, San Carlos, California 94070.

ham radio

base-loaded vertical antenna for 160 meters

No room for a 160-meter beam? Try this vertical antenna which can be easily made from readily available materials

Much has been written and discussed on the best antenna for 160 meters. The most popular solution seems to be to tie the ends of an 80-meter antenna together and feed the system on 160 meters with an antenna matching network. Some Amateurs string up an inverted L antenna. Both work fine for local contacts, but if you really want to work across the country, the vertical antenna is best.

the case for a vertical antenna

Several 160-meter enthusiasts use phased vertical antennas. Their signals are outstanding all year around compared with signals from other antennas. For those who don't have room for a beam, the top-loaded vertical is the next best. The loading coil should be wound with no. 10 AWG (2.6 mm) wire. It requires a long coil and an extended tube for adjustments. I tried such an antenna, but the assembly swayed back and forth like a pendulum, and the nylon guys would not remain tight enough to hold it. After the coil broke off, I experimented with a base-loaded vertical.

base-loaded vertical

I was surprised that my signals seemed to be equally good, but not before some testing of the wire size used on the base coil. My vertical uses a 32-foot (9.8-meter) length of aluminum irrigation tubing, which is 2 inches (50 mm) in diameter. (It cost $20.00.) The tube was set on a beer bottle for an insulator and guyed with nylon rope. This assembly was backed up by burying a 6-foot (1.8-meter) length of 4 x 4 lumber into the ground and using insulators and wood blocks to secure the tube to it (**fig. 1**).

To resonate the tube to 160 meters, a series capacitor and coil were first tried. However, I was told it would be better to just use the coil. First tried was a wire coil, but later a coil made from 3/16-inch (5-mm) diameter copper tubing was substituted, and the signal increased by 1 dB.

coil construction

The inductance was wound with 3/16-inch (5-mm) diameter copper tubing which cost $9.75 for a 50-foot (15.25-meter) coil. I used a 4-inch (102-mm) diameter pipe as a mandrel and wound a coil of forty turns.

Next three pieces of plastic were cut 1 inch (25.4 mm) wide and 1/4 inch (6.5 mm) thick for the length of the coil. Holes were drilled in these strips with a drill just over 3/16 inch (5 mm), so that the copper would slide through it easily. The first hole was 1/4 inch (6.5 mm) from the end.

The holes were cut with a drill sharpened like a sheet metal drill so that it did not shatter as it came through the plastic. A small hole could have been drilled, then a large drill put through half way on each side. That takes patience. Once the pieces are snaked onto the coil and spaced, they are treated with coil dope. The coil was rugged enough to be mounted on insulators and put into a wooden dog house at the base of the antenna.

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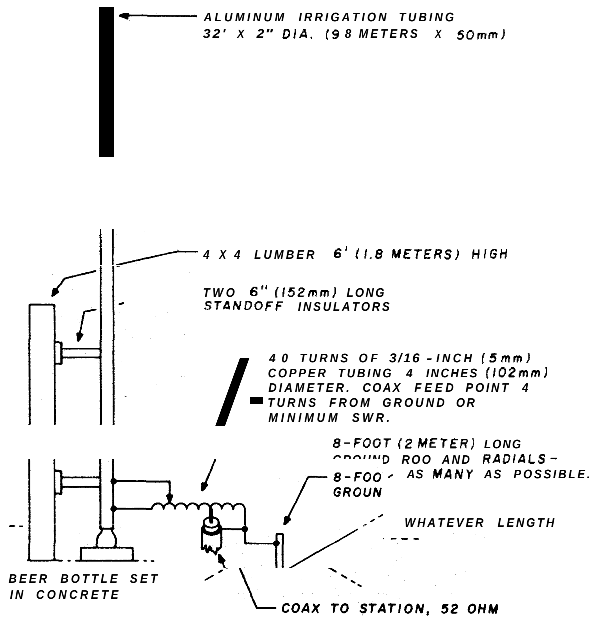

fig. 1. Construction details of a base-loaded vertical antenna for 160 meters. Antenna performs as well as a top-loaded affair and is much more stable and easier to construct. Use as many radials as possible in the ground system.

tune up

Antenna tuning was accomplished by leaving the feeder off and grid dipping the coil with it all in place. The coil was then tapped for resonance at 1820 kHz. Wide copper straps can be formed around the 3/16-inch (5-mm) copper and soldered once the proper place is found. The next step is to vary the tap for the 50-ohm feeder from the bottom of the coil for minimum SWR. Mine came out at the fourth turn up from the bottom. (This will depend on your ground system.) I used an 8-foot (2-meter) rod driven into the ground at the antenna base and four or five radials of various lengths pushed into the grass. None are over 30 feet (9 meters) long, but make them as long as you can and use as many as possible. The more the better on 160 meters."

performance

We have a daytime 160-meter net here in California, and records are kept of signal strengths up and down the coast. At 11 AM Sunday mornings Santa Barbara checks in with signal reports. I can say this antenna receives and sends equally as well as my old top-loaded affair. I've worked the East Coast with it and am pleased to report that it is more stable and easier to construct. All in all, it seems like the best answer to many 160-meter antenna problems — if you can't have a phased array.

'Or on any frequency. **Editor.**

ham radio

More Details? CHECK — OFF **Page 94**

digital capacitance meter

Easy to build digital capacitance meter for the home shop features ranges from 1000 pF to 100 μF

Amateurs who build or service electronic equipment sooner or later encounter the situation where replacing a capacitor with a "larger" one produces the wrong results: power supply ripple worsens or the time constant of a timing circuit decreases when it should increase. Highpass or lowpass audio might have their actual 3-dB rolloff points at 200 Hz instead of the intended 300-Hz point. Such differences often occur because the actual value of the capacitor used is different from its marked value. The best performance of narrow bandpass filters and notch filters is obtained when matched capacitors of exactly the same value are used. There are many good "100-for-a-dollar" capacitor buys available, but they often included unmarked or house-numbered units. Those 25-cent, 68-μF capacitors I bought at a hamfest were actually 6.8 μF – the reason, no doubt, they were only 25 cents!

Capacitors are among the most common components used in electronics. Most users assume that the value marked on the capacitor is its actual value; specifications simply guarantee a minimum value. Most electrolytics, for example, are specified to be within +80 to –20 per cent of their indicated value. There are a few that are within ±10 per cent of their marked value; some small capacitors are available with 1 per cent and 5 per cent tolerances. The true value of a capacitor is not important in some cases, such as audio bypass applications, while in other applications the capacitance must be accurately known to produce the desired results.

The digital-capacitor meter presented in this article was built to preclude the type of problems described above. It measures capacitors from 0.001 μF to 999 μF in six ranges, with accuracy of about 1 per cent. The three-digit display has the decimal point correctly positioned as the ranges are switched. The circuit uses low-cost components which are readily available. It requires no difficult adjustments for reliable operation and is easy to duplicate with the printed circuit board layout shown. The meter requires about 100 mA from a 5-volt regulated source, so it lends itself to battery operation if desired. The circuit includes a flashing overflow indicator.

circuit description

The circuit is based upon a digital counter that counts a reference oscillator. The input to the counter is gated by the C_x monostable which has its period determined by the capacitor to be measured.

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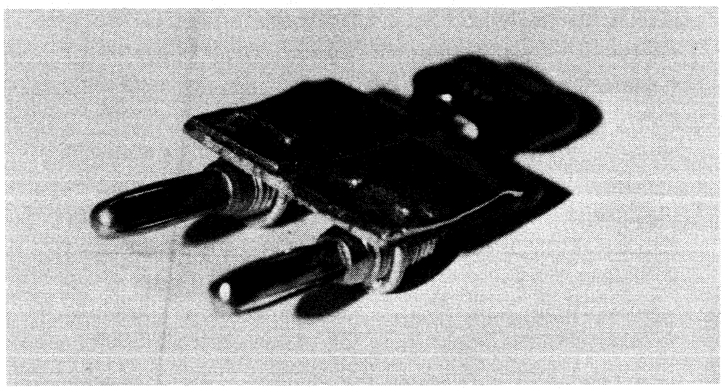
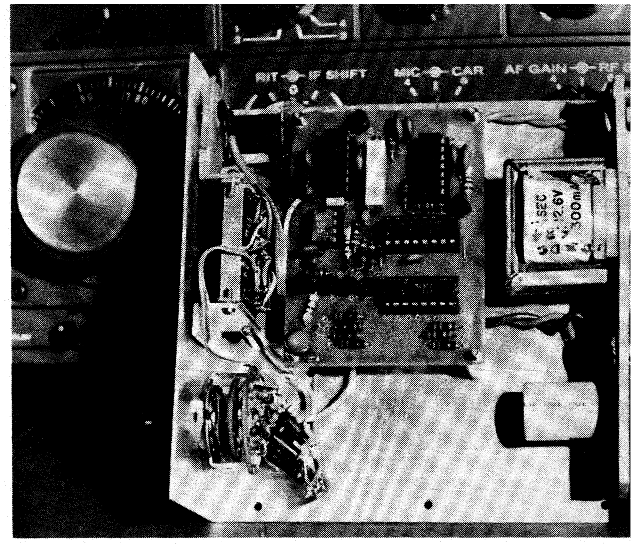
Construction of the short lead adapter.



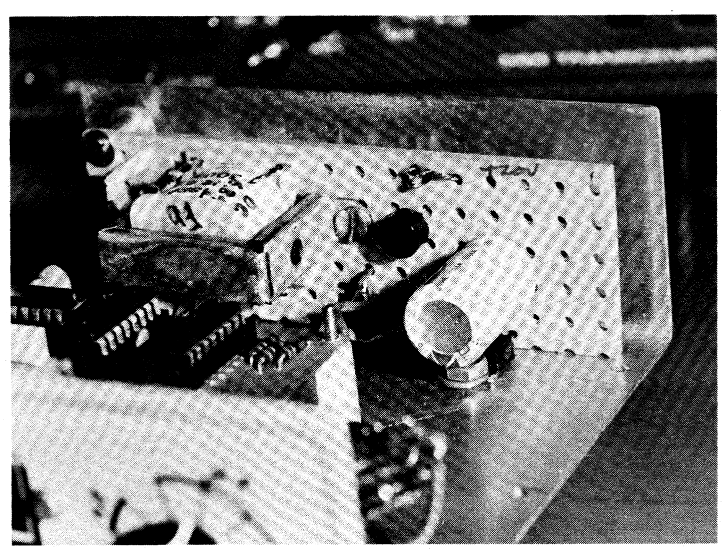


An interior view showing arrangement of the display cuit board, power supply, and range switch. This was a **totype** circuit board which has its overflow circuit rnou below the main board.

Digi Cap

Two different meters showing suggested range switch labeling for right-hand decimal displays per the text (on top) and left-hand decimal displays. The unit with the small display (top) was used to develop the circuit. The bottom unit was built by WA4RVN to verify the circuit reproducibility and performance consistency.

Closeup view of the point-to-point wired power supply. The 7805 voltage regulator is snugged beneath the **1000-μF** filter. Yes, the capacitor was measured before use. Would you believe 998 μ**F?**



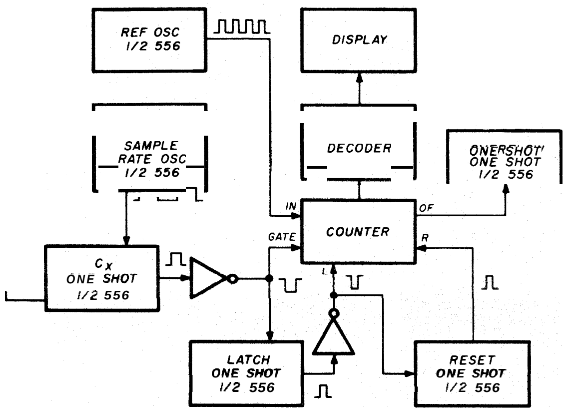


fig. 1. Functional block diagram of the digital capacitance meter. The meter is based upon the 14553 counter. The other ICs provide the necessary gating for the oscillators and display functions.

The functional block diagram is shown in **fig. 1.** About twice a second, the sample rate oscillator triggers the C_x monostable circuit. This monostable output is inverted and applied to the counter control gate. The duration of this control gate input is directly dependent upon the value of the capacitor being measured. If the reference oscillator input to the 14533 IC counter is at the proper frequency, the resulting display will indicate the value of the capacitor. One half of a 556 dual timer serves as the sample rate oscillator, while another 556 dual timer is used as the C_x monostable and reference oscillator.

The 14553 counter chip contains all the circuitry to count and multiplex three digits. It has built-in latch and reset functions and an input control gate. The counter chip's BCD output is applied to a single seven-segment decoder which drives the multiplexed LED displays. The required latch and reset functions are provided by another 556 dual timer with each of its sections operating in the monostable mode. The latch signal is applied to the 14553 at the end of the input gate enable period to store and display the accumulated count. Immediately thereafter the reset signal is applied. The 14553 holds the outputs for the displays, even though the internal counters have been reset, until the latch signal is again low. The latch signal goes low only after the capacitor value has been measured again. This produces a constant or steady display that does not flicker or count up to the final value.

The circuit timing diagram is shown in **fig. 2.** The overflow signal from the 14553 is applied to one half of a 556 dual timer to provide an overflow indication. The timer is run as a monostable to produce a flashing LED overflow indicator. **Fig. 1** shows wave forms at significant locations and indicates the direction of information flow in the circuit. The complete schematic diagram is shown in **fig.** 3.

Construction is uncomplicated when using the printed circuit board. **Fig. 4** shows the location of components on the board, while **fig. 5** shows the circuit board foil pattern. Careful examination of **fig. 4** will reveal the location of the numbered and lettered points to be wired to the display and the range switch. These points are shown on the schematic for easy reference. Switch wiring is shown in **fig. 6.** Points **X**, Y, and **Z** are not used.

The circuit uses a common-anode multiplexed display. The seven 82-ohm resistors near the 7446 decoder are the recommended value for displays that require around 10 mA per segment. The suggested value for displays rated a?5 mA per segment is 150 ohms. These values can be varied to achieve the desired display brightness. One unit was built without the seven current limiting resistors (to achieve the maximum brightness) and has worked without any LED burnout problems.

None of the circuit component values are critical, but best performance can be obtained with a good quality capacitor, preferably plastic, for the reference oscillator. This particular capacitor is the 0.001-μF capacitor located near the 100k pot and connected to pins 2 and **6** of U2. Q1 is used to boost the current-handling capability of the C_x monostable (U2) and should have low capacitance and a power rating of ½ to 1 watt. A 2N3906 will work with good results. Transistors 01, Q4, Q5, and Q6 are PNP transistors, while Q2, Q3, and Q7 are NPN transistors; 2N3906s and 2N3904s can be used, respectively. Q4, Q5, and Q6 should be installed so that their emitters go to the 5-volt land, bases go to the 1 kilohm resistors, and their collectors to the anodes of the display. The overflow LED is connected with its anode to point **F** on the circuit board and the cathode to ground.

A well-regulated, 5-volt power supply capable of 100 to 150 mA is required. **Fig. 7** shows a schematic for a suitable supply. Point-to-point wiring on a insulated board is an easy way to build the supply.

Care should be taken to keep the wiring between 01, the range switch, and the C_x input jacks as short as possible and away from the 60-Hz ac line.

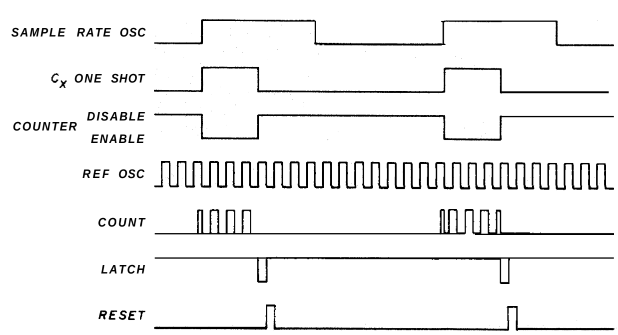


fig. 2. Timing diagram of signals in the capacitance meter.

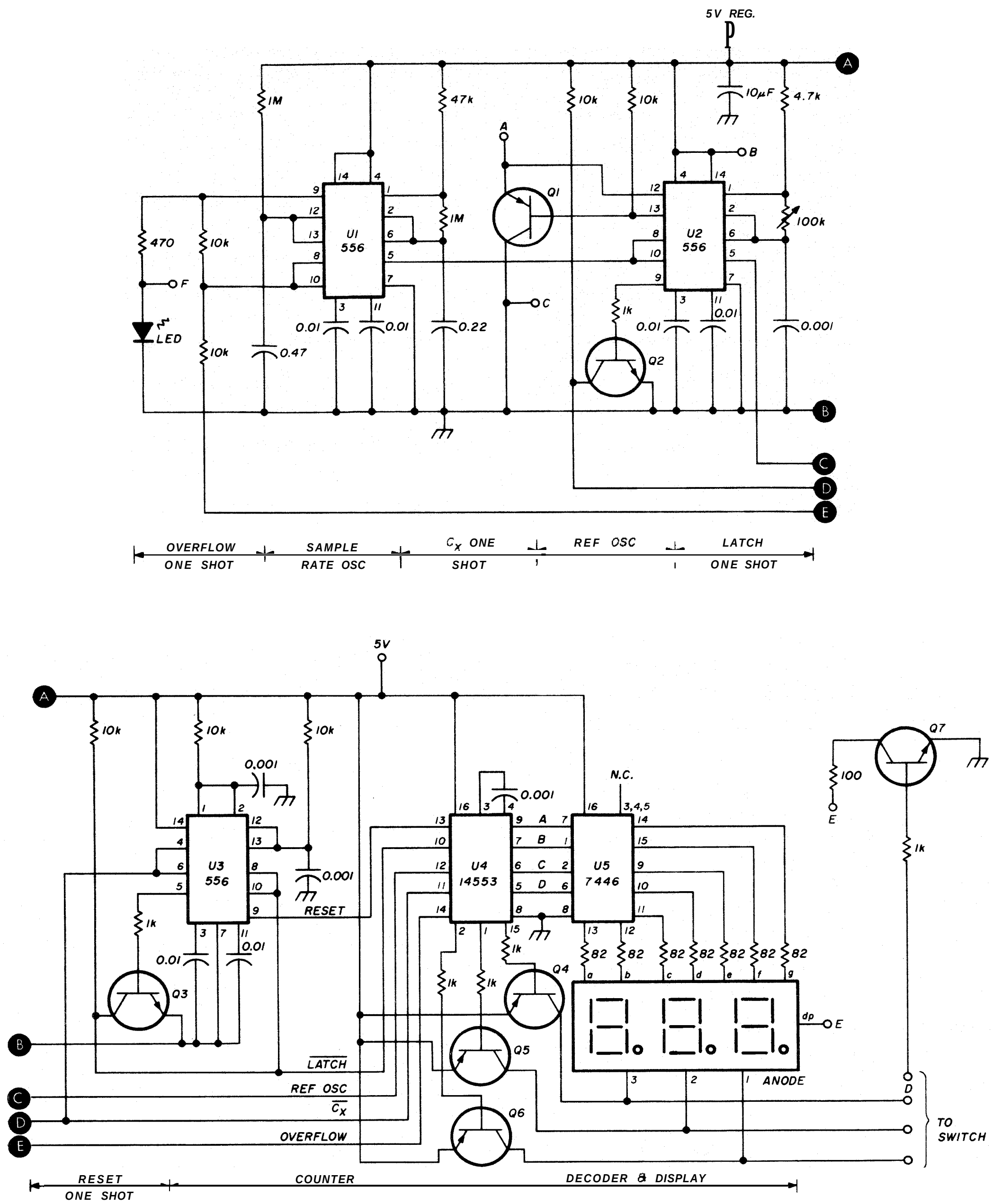


fig. 3. Complete schematic diagram of the digital capacitance meter. Suggested types for the transistors are given in the text. The current requirement of the meter is approximately 100 mA, small enough that a battery supply can be used for field use.

checkout and calibration

The circuit board should be completed and all wiring connected to the display, overflow indicator, and range switch before starting checkout. Make sure that the power supply is delivering 5 volts and is properly connected to the circuit board. At power turn on, the display should light and the overflow indicator should flash once. The display should show 000 or 001 with no connection at the C_x input. With a short across the C_x input, the display should show a number, say 433, and the overflow indicator will flash continuously. This number should not change when the range switch is moved to other positions. The display should show a number of 000 to 002 with the range switch in position 1 (see **fig. 6**) and no connection at the C_x input. An unsteady count ranging from 000 to about 060 indicates that the meter is picking up stray 60 Hz. If this happens, try redressing or rerouting the wiring between the circuit board, range switch, and C_x input jacks. K4ZKU found that reversing the ac line cord at the wall outlet would help with such a situation. A simple test of U5, the display, and

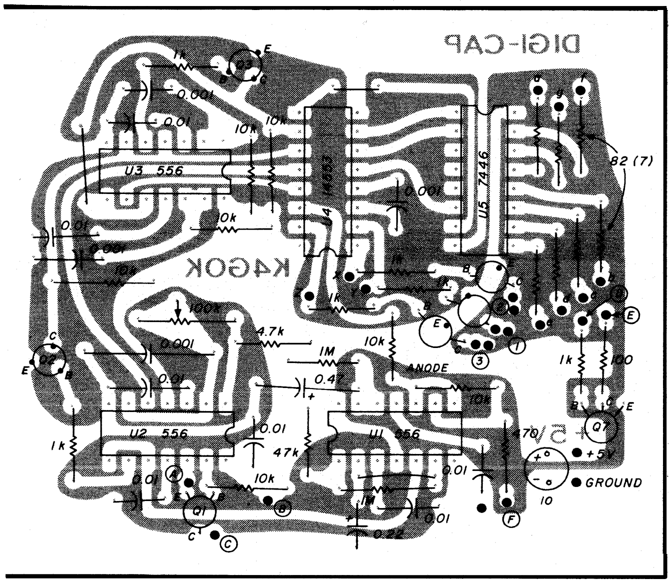


fig. 4. Parts placement diagram for the printed circuit board.

the wiring between can be made by temporarily grounding pin 3 of U5; the display should show 888.

The unit must be calibrated before use. Capacitors of known value are required. Surplus computer and audio boards are a good source for precision capacitors. I found 1 per cent capacitors from 0.001 to 2.5 μF at local hamfests. The meter should be allowed to warm up for about 20 minutes before calibration. If precision measurements in the 10s and 100s of microfarads ranges are not required, the 2000- and 200-ohm pots at positions 5 and 6 of the range switch can be replaced with 1000- and 100-ohm fixed resistors. To calibrate the meter, connect a 0.1- to 0.3-μF

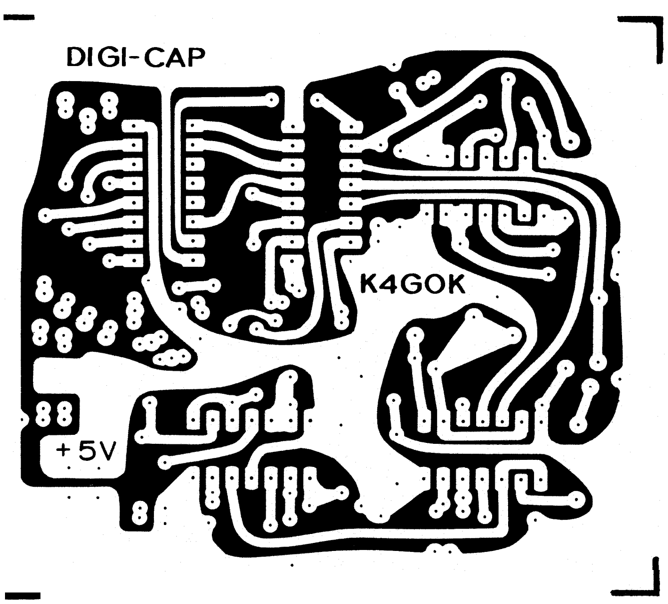


fig. 5. Foil layout pattern for the digital capacitance meter.

capacitor of known value, and with the range switch in position 3, adjust the 100-kilohm reference oscillator pot on the circuit board so that the display indicates the correct capacitor value. This calibrates the 100k-pF range (switch position 3) as well as the 10k-pF (position 2) and 1-μF (position 4) ranges. The 1k-pF is range calibrated by the 1-megohm pot at switch position 1; the 10-μF and 100-μF ranges are calibrated by the 2000- and 200-ohm pots at positions 5 and 6.

using the meter

Operation of the meter is simple. Observing proper polarity, connect the capacitor to be measured, select the largest range that does not cause an overflow, and read the capacitor value shown on the display. **Table 1** shows examples of how the display indicates various capacitor values for each of the range switch positions. The first three ranges measure in thousands of pF and the last three ranges measure in μF. The decimal point is properly positioned. Note that if a 22-μF capacitor is being meas-

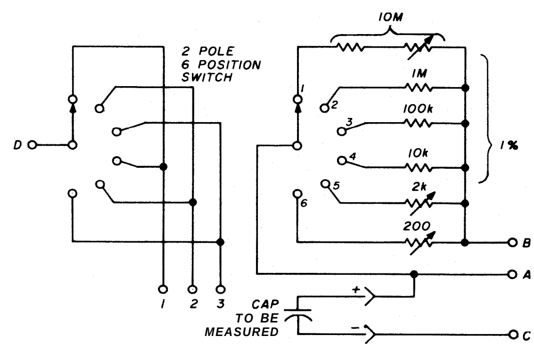


fig. 6. Switch connections for the range switch of the capacitance meter. The points specified are connected to the appropriate location on the circuit board (see fig. 4).

ured the range switch should be in position 5 and the display will show 22.0. A 0.047-μF capacitor is 47k-pF, and it will be measured with the range switch in position 2. The display will show 47.0. Labeling the first three positions of the range switch as kpF (or nF for nanoFarads if preferred), and the last three positions as μF will make the meter very easy to read.

An open capacitor will cause a 000 to 001 to be displayed. A shorted capacitor will cause the overflow indicator to flash and the display to indicate a fixed number that is independent of the range switch position.

Lead lengths should be kept short when measuring small value capacitors. The photographs show a plug-in device made from banana plugs, a small piece of copper clad board, and sheet brass.

conclusion

The digital capacitor meter has been a fun project

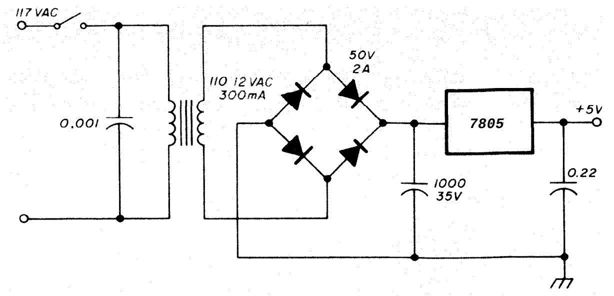


fig. 7. Schematic diagram of a small power supply suitable for home station use of the capacitance meter.

to build and it has been a time- (and agony-) saver around the ham shack. I hope that others who enjoy building and experimenting will find it to be the same. I will offer film negatives (or positives) so that builders can make their own circuit boards. Correspondence regarding the meter will be answered if an SASE is included.

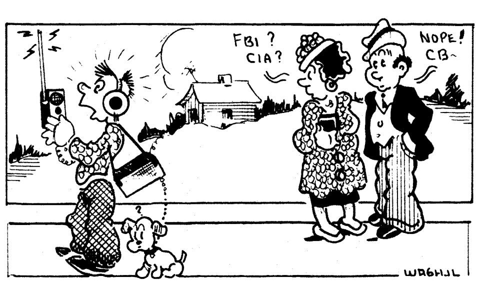
table 1. Switch positions for various measurement ranges showing display and associated capacitance value. In switch position **1,** a display of **1.50** indicates a capacitance of **0.015 μF (1500 pF),** a reading of 2.20 indicates a capacitance of **0.002 μF** (2200 **pF),** etc.

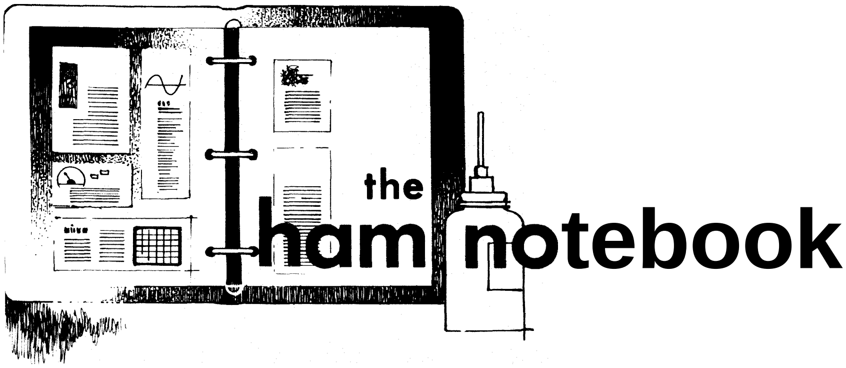
switch position	display	capacitance	range
1	**1.00**	**0.001**	**1000** pF (1 nF)
2	**10.00**	**0.010**	10k pF (10 nF)
3	**100.00**	**0.100**	100k pF **(100**nF)
4	**1.00**	1.000	1 μF
5	**10.00**	**10.000**	10 μF
6	**100.00**	**100.000**	**100** μF

acknowledgments

Several hams have been of great assistance in developing the digital capacitor meter, in particular WA4RVN, K4ZKU, and W4PVA. K4ZKU provided valuable information on driving the display to full brightness, and W4PVA helped with the information on the 14553 counter chip without which the project could not have been undertaken. WA4RVN built his meter according to this article to verify the construction and checkout notes.

ham radio





a fresh look at linear tuning

Most Amateurs expect the benefits of a linear-spaced tuning dial when they purchase new equipment. The expression "linear tuning" refers to the ability to rotate the tuning dial or knob, knowing that a certain number of kilohertz will be traversed with each turn of the knob, say **25** or **50** kilohertz. Amateurs expect to find this feature in professionally designed equipment, but rarely is it found in home-built Amateur gear. Why does this situation exist, and what can the Amateur who perfers to build his own do about it? Let's take a look at how commercial manufacturers handle the problem.

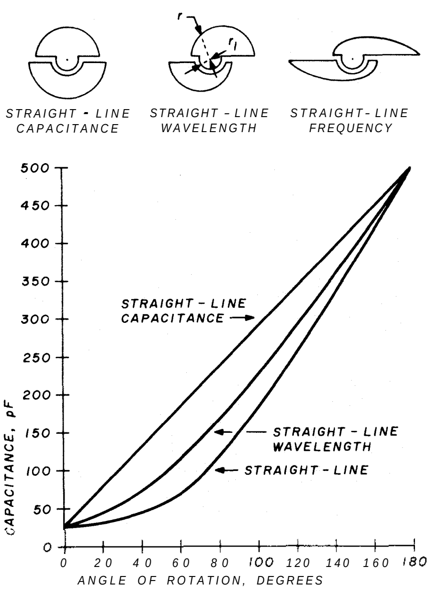


fig. 1. Characteristics of SLC, SLW, and SLF variable capacitors showing capacitance as a function of angle of rotation in typical cases, together with approximate plate shapes. (From *Radio Engineers' Handbook* by F.E. Terman, McGraw-Hill, Inc., 1943.)

Frequency change requires that we square the product of inductance, L, and capacitance, C, to effect a 2:1 frequency change. For example, the product of LC is approximately **520** for **40** meters. The product of LC amounts to about **2080** at **3.5** MHz. Since it would be mechanically unwieldy to alter both inductance and capacitance, the accepted method is to vary either the capacitance or inductance in a typical circuit. (The foregoing remarks apply to high-frequency circuitry in this discussion.)

Most manufacturers handle this situation by limiting the excursion of their oscillator circuit to, say, **500** kHz and by using a VFO coil with windings spaced nonlinearly. A tuning slug moves into the coil form and causes an inductance change. Collins refers to this method as "permeability tuning." It's a good system; unfortunately it's not suitable for easy duplication by the home builder. Another method would be to use capacitor plates with special shaping. This also poses a problem for the home builder.

variable-capacitor plate shapes

Fortunately there's a way of "making it" without having a large machine shop at your disposal. The solution to the problem came to me while watching my wife making some designs on a quilt with a mix-and-match pattern.

A look at most transmitting capacitors shows that they use half-round plates in the rotor section, whereas most capacitors for broadcast reception use different shapes. The first shape is called straight-line capacitance (SLC), while the second is called midline, or straight-line frequency (SLF). See **fig. 1.** In short, a variable capacitor with the proper arrangement of SLC and SLF plates should satisfy the need for truly linear tuning.[1]

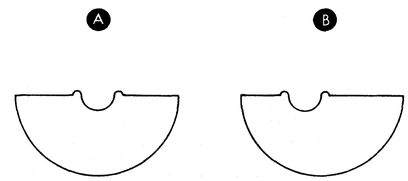


fig. 2. Comparison of rotor plates in the popular "Command" transmitter. A shows approximate shape of an SLC plate; B an SLF plate.

modifying transmitting variable capacitors

Some variable capacitors, which use aluminum plates spaced with washers or metai spacers, can be modified easily in the rotor section to accomplish this objective. In my case, I removed half-round plates from the middle capacitor in a Command transmitter and re-installed them on the rotor shaft of the master oscillator tuning capacitor, which had been altered by lifting out several of the SLF rotor plates.

With the correct amount of fixed L and C, linear tuning will result. If you're willing to settle for a limited frequency excursion, exceptionally high accuracy can be achieved. The

---

[1]Still another shape for variable-capacitor plates is called straight-line wavelength (SLW) in which the plates are shaped so that, when used to tune an inductance to resonance, the wavelength at resonance is a linear function of the angle of rotation. Practical capacitors use intermediate characteristics or a combination of these basic types (see fig. 2). **Editor.**

calibration chart **(table 1)** shows this to the last hertz. Note that this is not a one-of-a-kind experiment. Equally satisfying results have been accomplished in a half-dozen instances. There's no reason why this technique can't be applied to other ranges, such as the popular 5.0-5.5 MHz range used in many VFOs.

My original intentions were satisfied, as shown by a 3.5-3.6-MHz curve. However, high-accuracy linear readout continued throughout at least a 200-kHz span between 3.45-3.65 MHz. All of the above was achieved without trimming or bending of the master-oscillator variable capacitor plates; thus it's fair to say that similar results could be obtained by any careful experimenter or builder.

Neil Johnson, W2OLU

solid-state amplifier switching

Forget about carrier-operated relay (COR) circuits and other mechanical antenna and power switching arrangements by going solid-state. This diode-switching circuit provides maintenance-free, reliable switching without fuss or bother. Best of all, it's simple.

Simply use two quarter-wavelength sections of RG-174/U coax and appropriate diodes **(fig. 1)**. The 1N4148 diodes are adequate for moderate power levels commonly used on 2-meter f-m.

David D. Holtz, WB2HTH

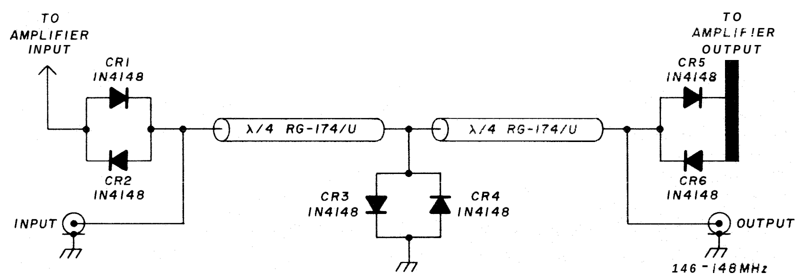


fig. 1. Solid-state switch for 2-meter f-m. Circuit was adapted from an article appearing in the April, 1973, issue of *ham radio*.

de-icing the quad

Probably more quad antennas have come to grief because of ice than from all other reasons combined. At least that seems to have been my experience. It seems that something may be lacking in our planning. A simple means of de-icing the quad should be a real boon to those who usually have a couple such examples of nature's contempt for us each winter.

The quad driven element is usually fed at bottom center through coaxial line. If, for the driven element, we use a wire having a higher resistance at dc than at rf, such as galvanized electric fence wire or smaller size copperweld, 60-HZ power, fed through the coaxial line should provide enough heat to prevent the formation of ice, or if it has already formed, to melt it. After all, ice usually forms at temperatures quite close to freezing, and this idea wouldn't require a temperature increase of more than a few degrees to thwart Jack Frost.

The average quad has at least two elements, and it wouldn't do to leave the parasitic elements out in the cold. By going to the top of the quad, opposite the feed point, one finds a voltage node. A capacitor of suitable power-handling capability may be inserted here without affecting array performance. A value of 0.01 or 0.02 μF should be enough capacitance. The same thing can be done with the reflector (and the director if you have more than two elements). A pair of wires that connect all the elements in series for dc, running parallel to the boom, should permit you to apply enough current through the coaxial line to keep the ice away. An ordinary filament transformer should supply enough power for most applications.

The diamond configuration might be preferable for this application, as more support would be provided for the capacitors and connecting wires; the square configuration makes a clearer illustration **(fig. 1)**.

Henry S. Keen. W5TRS

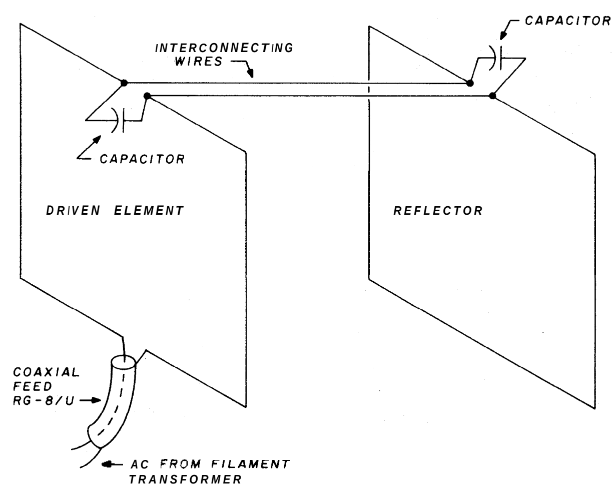


fig. 1. Power applied to a quad from an ordinary filament transformer will generate enough heat to prevent ice formation, or if already formed, to melt it. The capacitors are inserted at the voltage nodes of the elements, and don't affect array performance.

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contents

Observations & Comments

While browsing through **RSGB's** ***Radio Communications*** (August, 1979, page **751**) **I** came across an interesting letter from H. Herzer, DL7DO,* which had been subsequently picked up by ***QST*** in their July, 1980, issue (page **73**). The subject of Herr Herzer's letter was the antiquated **RST** signal-reporting system and what might be done about it.

Herr Herzer presents a pretty good case for eliminating the **RST** system in terms of today's standards of equipment sophistication, band conditions, and operating methods.

During the early years of Amateur Radio, as Herzer points out, the **RST** reporting system was a valuable aid to operators, as most (if not all) equipment was homemade. Amplitude levels were low, power-supply filter systems were primitive or nonexistent, and measurement equipment was crude. So the **RST** system was a means of evaluating on-the-air signals. If your rig used, say, a type 45 tube with 90 volts on the plate, a report of S7 meant that the signal was "fairly strong." A tone report of T8 meant that something had to be done in the power-supply filter department to bring the signal to **T9**: "pure dc note." If you received a readability report of R3, this could mean just about anything, from poor propagation conditions to faulty adjustment of receiving equipment at the other end of the circuit.

Herzer recommends a "*Q* System" to replace the antiquated **RST** reporting system for both CW and radiotelephone communications. In Herzer's system "*Q*" stands for ***transmission quality,*** which accounts for the one and only parameter relevant to successful information transfer by Amateur Radio stations. It all boils down to: "Has the message been received and understood?" The Q reporting system used only three variables:

Q1 ***At no time has there been sufficient transmission quality; that is, no copy has been received.***

Q2 ***Sufficient transmission quality has occurred some of the time; that is, partial copy has been received.***

Q3 ***Sufficient transmission quality has occurred at all times; that is, full copy has been received.***

Intermediate stages of reporting, such as Q1/2 or Q2/3, might be used. You can always ask the other operator for an explanation.

Herzer feels that such a system will result in a considerable reduction in **QRM**, contest reporting, and logging. I agree. Today's **RST** reporting system is not only redundant but meaningless.

Of course, this means changing your QSL cards to show the new system, and it will probably take a long time to implement on a worldwide basis. But the system has real merit in reducing "on-the-air" pollution in today's Amateur bands.

The other night I worked several stations, foreign and domestic, on the low end of the 14-MHz CW band. All reports were **RST** 579. During contest operation, all stations reported **RST** 599. Utterly meaningless! Why bother with such a signal-reporting system? In fact, why is any signal-reporting system useful today?

What do you think? We'd like your opinion.

Alf Wilson, **W6NIF**
technical editor

'*Radio Communications,* **Journal of the Radio Society of Great Britain, August, 1979, page 751.**

comments

microphones

Dear HR:

Readers should be cautioned to avoid using the type of audio cables suggested by W1OLP in his article on microphones and speech processing in the March issue of *ham radio.* Only the audio line should be inside the shield. Having the audio line plus the PTT line and/or battery line inside the shield will, in many cases, result in unwanted noises in the transmitted audio: hum, switching noises, and stray rf. It is preferable to use microphone cable which has one shielded line for audio, with the PTT and other control lines outside the shield.

Buddy Massa, **W5VSR**
New Orleans, Louisiana

Hallicrafters story

Dear HR:

The fine story by W6SAI about Bill Halligan's HT-4 (BC-610) in the November, 1979, issue of *ham radio* surely brought back a flood of memories – Utah Beach, Ste. Mere Eglise, Carentan, Isigny.

We had five mobile units in ADSEC (Advanced Section Communications Zone), 3rd Army, and only one of them was an SCR-299, the others were SCR-399s. I would like to call your attention to the incorrect caption with the photograph on page 24. The mobile unit is an SCR-299, not 399; the 399 differed mainly from the 299 in that it was not housed in a panel truck; it was provided with an HO-17 plywood shelter, designed to fit the equally famous 6x6 International truck. The shelter could be lifted off, complete with its equipment, and operated on the ground as a fixed station.

Our first team's unit was installed in a "Duck" amphibious version of the 6x6, with two little PE-75 gas generators connected in parallel on the aft deck. They were let down from the LST *before* H-hour on D-day and made an attempt to scramble ashore, but they were met by severe mortar fire and forced to withdraw. Later they gained an exposed position on the beach and made contact with our station in England that had been set up in late May.

I was the Platoon Sergeant of the fifth team and some days later we had all units dispersed several miles apart and well camouflaged in the Normandy apple orchards. They were tied with field wire keying lines (duplexed for telephone) to a radio center in the loft of an old French barn. In the stable below were the TC-10 telephone boards and the Message Center. All operation was manual and long press dispatches were cleared between items of military traffic. Later three more SCR-399s arrived and eight were operated for a few days. Then one-by-one the ADSEC units began to leave to follow the action. Some months later one was in Namur, Belgium, in contact with besieged Bastogne.

I should add that the operators who accompanied these units were highly specialized, having worked in the signal center at the Pentagon before leaving the States. Likewise, many of the technicians were specially trained on SSB multi-channel (AFSK) high-power transmitters (40 kW). They formed the nucleus of the Paris communications center in the "Block House" about a block from the Arc de Triomphe de L'Etoile on Rue Wagram. Some followed the action; some had very important missions elsewhere.

We experienced only one real trouble with our BC-610Es. The high-voltage in the modulation transformer would break down to ground, killing the rig. We found that we could set the transformer up on four short standoff insulators supplied in the spares chest and be back on the air in half an hour. Information was sent up through channels on this fix and apparently others had experienced a similar failure because a Field Change Bulletin was put out by the Signal Corps directing that this modification be installed in all BC-610Es.

After VE-day, returning to France and to the Signal Depot at Mohn near Meziers, SCR-399s were stashed in the fields around the buildings as if we were operating a trailer park. I wonder now what happened to all of them. Some were trans-shipped to the Pacific Theater, but most were left behind.

Clifford O. Field, **WA2JVD**
Fair Haven, New York

more Hellschreiber

Dear HR:

E.H. Conklin, K6KA, is not quite right in describing the Hellschreiber as a wideband system (Comment, March, 1980). Admittedly, the bandwidth of any keyed system is a function of the keyed element rise time, but with proper pulse shaping as practiced by the majority of the PA0 and German Amateurs the amount of spectrum space occupied by a Hellschreiber signal is only marginally greater than that of 45.5-baud RTTY.

The bandwidth necessary for Hellschreiber may be quite easily computed by reference to CCIR Recommendations, which in Appendix 5 of Radio Regulations state that this is

(Continued on Page 66)

Gunn oscillator design

for the 10-GHz band

Gunn oscillators are an attractive alternative to reflex klystrons as a signal source in the Amateur microwave bands. However, to some, the design of stable Gunn oscillators is considered to be somewhat empirical. In practice a particular design approach is tried then perfected through cut-and-try. Much research has been done with Gunn devices as described in the literature,[1,2] leading to high-quality commercial units.

In an effort to dispel some of the mystery surrounding Gunn oscillator design, I'll set down some ground rules for a certain design approach that produced good results. I wish to emphasize at this point that what follows is only *one* of several approaches that will give results. The oscillators were designed for 10 GHz, but this design could be used and modified for any microwave band between 5 and 90 GHz. Gunn devices can be made to oscillate in a number of configurations. These can be in the form of coaxial resonators, waveguide cavities, and microstrip circuitry.[3] They can be tuned mechanically or electrically. The theory of how Gunn devices oscillate is covered in other literature.[4,5,6]

The type of Gunn oscillator presented in this article is a waveguide cavity oscillator. If you wish to tune a waveguide cavity oscillator over a wide range, the cavity volume formed between the diode mount and a movable back wall, in the form of an rf choke, is changed. D. Evans,[7] and Tsai, Rosenbaum, and Mac Kenzie[8] describe wideband mechanically tunable waveguide cavity oscillators. However, the other approach to waveguide cavity oscillator design is to use an iris-coupled waveguide cavity. Here the resonant cavity is formed by the iris and the diode mount. The backwall is adjusted close to the diode mount for optimum operation. The design is inherently narrowband, tunable over 100-300 MHz. Since operation is permitted between 10.0 and 10.5 GHz, this type of oscillator presents a practical approach. Other features of the oscillator are that it is fairly straightforward in design and can be easily reproduced.

Following is a presentation of some of the ground rules for iris-coupled waveguide oscillator design. From these, a design procedure is presented followed by testing techniques and precautions to be taken when putting the oscillator into operation.

The iris-coupled waveguide cavity oscillator is shown in **fig. 1.** The diode is mounted across the center of the broad dimension of the waveguide. Dc bias is coupled to the diode through an rf choke. The backwall, which can be moved, is usually close to the diode mount and can be adjusted for stable operation. The function of the iris is to complete the waveguide cavity while providing coupling between oscillator and load. The tuning screw located between the diode post and the iris provides a limited tuning mechanism for the oscillator.

Since the Gunn oscillator is a negative-resistance device, it will have a tendency to oscillate at more than one frequency. The idea is to reduce the number of spurious resonances and make it oscillate where you want it to. The first step in this direction is to examine what makes up the main oscillating cavity of the oscillator.

Waveguide cavity. The fundamental resonant mode of the cavity occurs when the distance between the iris and the effective backwall is one-half the guide wavelength. If the diode is mounted near a sidewall of the cavity, the effective backwall is somewhere between the diode post and the backwall. If, however, the diode is mounted in the cavity center, the effective backwall is in the plane of the diode mount.[9] This is verified by slotted-line measurements looking at the normalized impedance in the plane of the diode mount.

One end of the cavity is now defined. The other end is defined by the iris. The iris reactance can be either inductive or capacitive. **Fig. 2** shows an equivalent circuit of an iris-coupled cavity. If the iris is capacitive, the physical length of the cavity is longer than $\frac{\lambda_g}{2}$ as shown in **fig. 2(a).** If the iris is inductive as in **fig. 2(b),** the actual cavity length is shorter than $\frac{\lambda_g}{2}$. The relationship between cavity length and iris susceptance is:[10]

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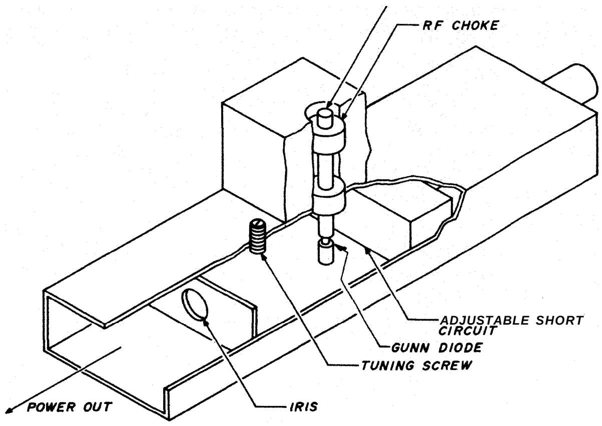


fig. 1. Simple waveguide iris-coupled oscillator. With the diode mounted across the guide center, the distance between iris and diode post is approximately one-half guide wavelength. Iris size is optimized for maximum power output while providing isolation from load mismatches. The rf choke minimizes power loss through the bias line. Backwall is adjusted to provide a stable operating point.

$$\ell = \frac{\lambda_g}{2\pi}\left[n\pi + \tfrac{1}{2}\tan^{-1}\frac{2}{|b_e|}\right] \text{(cap.)} \quad \textbf{(1)}$$

$$\ell = \frac{\lambda_g}{2\pi}\left[n\pi - \tfrac{1}{2}\tan^{-1}\frac{2}{|b_e|}\right] \text{(ind.)} \quad \textbf{(2)}$$

where = cavity length

λ_g = guide wavelength

$$= \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{2a}\right)^2}}$$

n = integral number

$|b_e|$ = absolute value of normalized iris susceptance

λ = operating wavelength

a = wide dimension of waveguide

In most cases $n = 1$ to reduce the possibility of the oscillator operating at a lower frequency. Once the value for b_e is found, the appropriate expressions (**eqs. 1** or **2**) give a value for the cavity length, ℓ.

The estimated value of loaded Q for an iris-coupled cavity is given **by:**[11]

$$Q_L \simeq b_e^2 \frac{\pi}{2} \frac{1}{\left(1-\frac{f_c}{f}\right)^2} \quad \textbf{(3)}$$

where b_e = normalized value of iris susceptance

f_c = guide cutoff frequency, TE_{10} mode

f = operating frequency

From the previous discussion, the cavity length, ℓ, and the cavity loaded Q depend on the value given to the normalized iris susceptance.

Iris. An iris is an obstruction placed into a waveguide system that electrically has a value of reactance or its inverse, susceptance. Irises can take various shapes **(fig. 3)** and can be inductive, capacitive, or resonant. From the standpoint of ease of construction, a centered circular aperture is the simplest form to make. The circular iris is inductive and its value can be calculated; however, it's easier to use the **graph**[12] in **fig.** 4.

The graph gives a good approximation for the value of normalized susceptance, $\frac{B}{Y_0}$ (that is, b_e) as a function of the ratio of iris diameter to the broad dimension of the waveguide. This is done for various waveguide aspect ratios and various ratios of operating-wavelength-to-guide width. The values of b_e from the chart are in good agreement with measured values. Now that the iris susceptance is calculated using **fig.** 4, cavity length can be found.

The question is raised as to how large a hole should be made in an iris plate. Critical coupling, the point where power output is maximum, occurs when the iris area is about 25 per cent of the waveguide cross-sectional **area.**[13,14] If the hole is further enlarged, the oscillator cavity will be overcoupled with a resultant drop in output power and poor stability. On the other hand, by making the hole smaller, the cavity will become undercoupled resulting in a drop in oscillator output power.

So in determining cavity length, choose an iris area between 20 and 25 per cent of the waveguide cross-sectional area. Since the circular iris area is $\pi d^2/4$, the diameter can be found. The next step is to design the cavity to operate at the highest frequency of interest. The cavity can always be tuned downward by a dielectric screw tuner. The operating frequency determines the wavelength. Knowing the a and b dimensions of the guide, the value of the normalized susceptance, b_e, can be found from **fig.** 4. This value is then substituted into **eq.** 2 to find cavity length ℓ.

Rf bias-choke system. Erratic operation, spurious responses, and power loss can be caused by an improper bias-choke design. The diode must be operated with a dc bias while decoupled from the

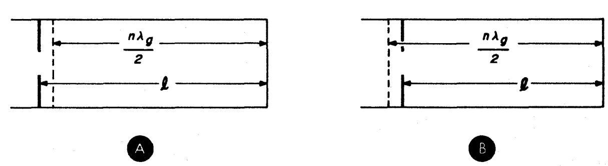


fig. 2. Equivalent circuit of an iris-coupled cavity. When the iris is capacitive, A, the actual cavity length is longer than multiples of one-half guide wavelength; when inductive, B, the actual cavity legnth is shorter than multiples of one-half guide wavelength.

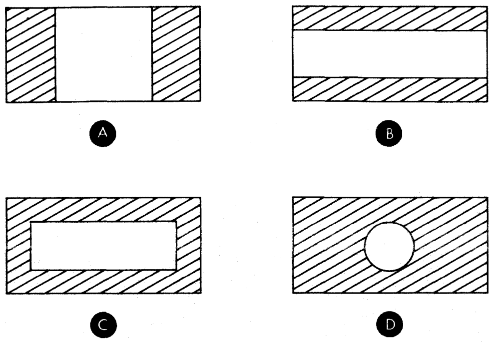


fig. 3. Various iris configurations. A and D are inductive; B and C are capacitive and resonant respectively. The circular iris in D is easiest to make.

bias supply. Early designs operated erratically and even had diode failure because of poor bias-choke design. Two common types of chokes are the radial-line choke and "dumbbell" choke. The radial-line choke requires a large circular plate parallel to the broad side of the waveguide with small separation. The radius of the radial line is approximately $\lambda/4$, with a configuration of an open-ended quarter-wave transmission line. The feedpoint impedance at the diode end becomes very low.

Although this type of bias-choke system is easy to build, some radiation occurs from the open end. A more popular choke arrangement is the "dumbbell" choke, **fig. 5.** As the name implies, sections A, B, and C resemble a dumbbell. This choke design is basically a series of quarter-wave-long coaxial line transformers, alternating between low and high Z_0 sections. The design transforms a wide variety of impedances at the feed point to an extremely low impedance at the Gunn-diode mounting point. This is necessary since the waveguide wall is a current-carrying surface.

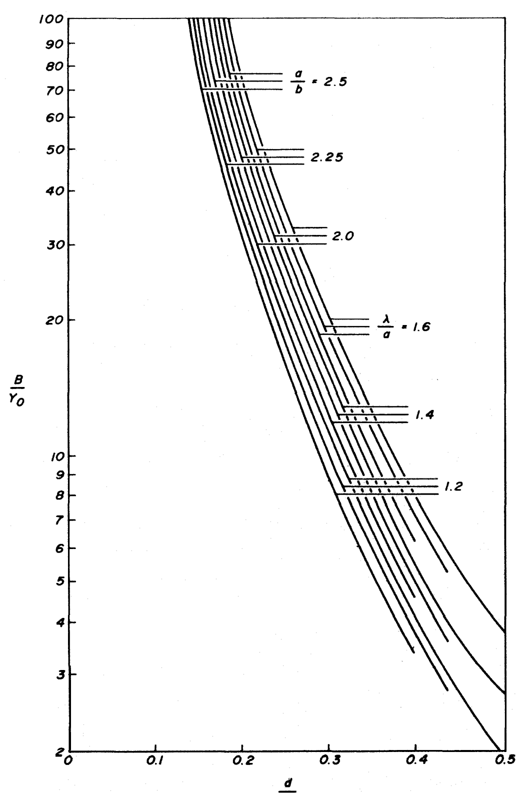


fig. 4. Relative susceptance of a centered circular aperture as a function of the ratio of iris diameter to waveguide broad dimension.

The characteristic impedance of a coaxial line is:

$$Z_0 \simeq \frac{60}{\sqrt{\epsilon_r}} \ln \frac{r_2}{r_1} \tag{4}$$

where r_2 = inside radius, outer conductor
r_1 = outside radius, inner conductor
ϵ_r = relative dielectric constant

Since sections A and C must have close spacing with respect to the wall, a dielectric material such as Mylar tape can be used to prevent the choke from shorting the bias supply. Sections A and C will be shorter in length than section B to account for the difference in phase velocity caused by the different dielectric constant of the tape. The velocity that the wave propagates is given as:

$$v = \frac{c}{\sqrt{\epsilon_r}} \tag{5}$$

where v = velocity of propagation
c = speed of light = $3(10^8)$ meters/second
ϵ_r = relative dielectric constant

To calculate the $\lambda/4$ sections, operating wavelength, X, is found from:

$$\lambda = \frac{v}{f} \tag{6}$$

where λ = operating wavelength
v = propagation velocity
f = operating frequency

Eqs. 5 and **6** will allow you to calculate the lengths of sections A, B, and C, while **eq. 4** gives the characteristic impedance of each section. The diameter of the cylindrical hole into which the choke section slides can be between **114** inch **(6.35mm)** to **112** inch **(12.7 mm)** for **X** band. **A** compromise is to use **318** inch **(9.5mm).**

Diode-mounting configurations. The diode can be mounted on a post centered in the waveguide. The post and mounted diode can excite TEM modes in the vicinity of the post, resulting in spurious responses that can cause oscillator turn-on problems. Post reactance can be eliminated if guide height is reduced to that of the diode package. However, results have shown that the tuning range is narrowed and power output drops.[15] If it's desirable to broadband the oscillator, a tapered or stepped post can be used. Again, because the device is a negative-resistance oscillator, broadbanding in this manner can lead to oscillation in unwanted modes and turn-on problems.

The resonant frequency of the TEM modes can be equated to post and diode height, being approximately a half-wavelength long. Mode frequency can be doubled by centering the diode on the post. The diode will cause a null in the fields at the point that makes the original post-height sections halved. However output power will drop, since mounting the diode in the post center decouples the diode. In many cases diode location on the post is left up to the experimenter.

Optimum post diameter is discussed in the literature.[16] Theoretical calculations verified by experimental results show that a post 0.125 inch (3 mm) in diameter gives the oscillator the broadest bandwidth. Post diameters greater than 0.150 inch (3.8 mm), reduce bandwidth over which the oscillator can operate. The power output, however, increases with post diameters greater than 0.150 inch (3.8 mm).

Cavity backwall. In some Gunn-oscillator designs the cavity backwall is fixed and located approximately one-half guide wavelength behind the diode post. For this configuration, some form of matching network is ahead of the diode. In the iris-coupled waveguide oscillator the backwall can also be fixed; however, this must be done by experimentation to obtain optimum results.

Once the backwall position is determined for optimum power output and stability, the position can then be fixed. From a flexibility standpoint, I found that a movable backwall permitted greater freedom of adjustment. This is particularly evident when different Gunn diodes are used in the same cavity.

Fig. 6 illustrates different movable backwall designs. **Figs. 6A** and **B** are quarter-wave choke sections, while **fig. 6C** is a close-fitting block that can be secured by locking screws once optimum operation is established. In **fig. 6A,** the quarter-wavelength sections are separated from the wall by nylon bearings. The mechanism is spring loaded, so that a constant pressure is exerted onto the choke assembly. An adjusting knob, riding on a threaded shaft, moves the choke assembly in and out.

Fig. 6B is identical to that of **fig. 6A,** except a dielectric is used around the entire sections that come in close contact with the waveguide walls. Note that these sections are narrower than the corresponding sections in **fig. 6A** because the insulated sections form a dielectric loaded guide.

Fig. 6C shows a simple block that is close-fitted in the guide. No attempt is made to insulate here; wall contact is desired with this design. In many cases, the block is cut to a depth of a quarter guide wavelength. A threaded rod acts as a handle for block adjustment.

The choke-section depth is based upon the idea that the space between guide sidewall and choke can be considered as a guide beyond cutoff (assuming TE_{10} mode). The space between the guide broad wall and choke is basically a reduced-height guide propagating in the TE_{10} mode. The relationship for λ_g is the same regardless of guide height. The choke depth is one-quarter guide wavelength. When air is in the dielectric, $\lambda_g/4$ is computed from:

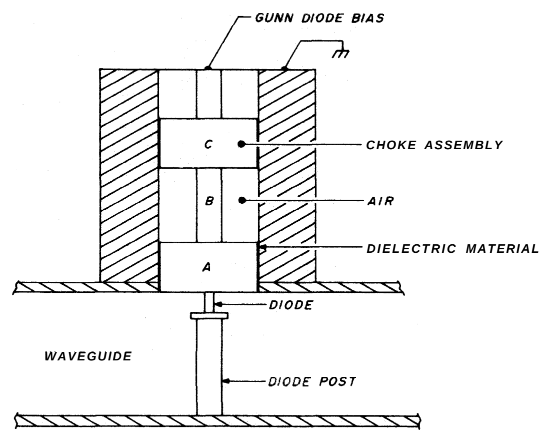


fig. 5. Cross section of the dumbbell rf choke. Sections A and C are identical. The dielectric insulates the center section from the wall. Sections A, B, and C form a coaxial transmission-line system of quarter-wavelength transformers with different characteristic impedances.

$$\lambda_g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{2a}\right)^2}} \text{ for } TE_{10} \text{ mode} \qquad (7)$$

where λ_g = guide wavelength
λ = operating wavelength
a = broad guide dimension

From **eq. 7** the value for λ_g is divided by four, and each section in **fig. 6A** is $\lambda_{g/4}$ long.

In **fig. 6B,** the value of λ_g for the two sections having dielectric tape wrapped around them is:

$$\lambda_{gd} = \frac{\lambda}{\sqrt{\epsilon_r-\left(\frac{\lambda}{2a}\right)^2}} \text{ for } TE_{10} \text{ mode} \qquad (8)$$

where λ_{gd} = guide wavelength in dielectric guide
ϵ_r = relative dielectric constant

Here the first and third choke sections are $\lambda_{gd}/4$,

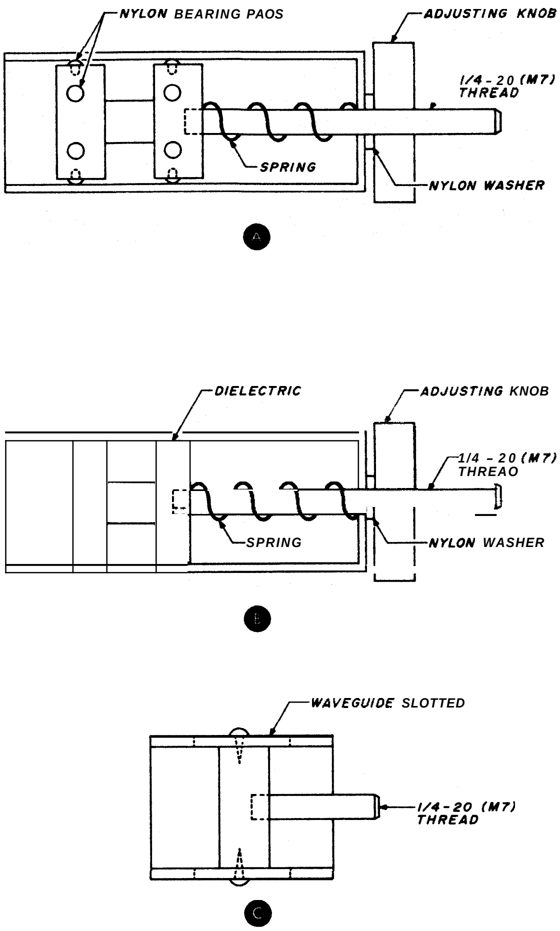


fig. 6. Different moveable backwall designs. An adjustable backwall choke assembly, using nylon pads for isolation, is shown in A. B shows a design using dielectric tape around the two sections. Wall contact is desired in the block design in C. The close-fitting adjustable backwall is locked into place by locking screws on each side of the waveguide.

while the middle section is $\lambda_g/4$. **Eq.** 8 is an approximate solution, assuming that a dielectric with a low-loss tangent is used.

Matching section. This device is an adjunct to the oscillator in that, if there's a considerable mismatch between oscillator and load, some sort of matching device is needed. Several techniques are used in impedance matching. Devices such as E-H tuners and slide-screw tuners, have been fairly common. However, they've been replaced in many applications by a ferrite isolator. This device exhibits low attenuation in the forward direction and high attenuation in the reverse direction. They can be made broadband and present a constant load to the oscillator despite wide variation of load mismatch. However, they're rather expensive, so a simple approach is used.

While it's narrow band, a three-screw tuner will match over a limited range of mismatch conditions. **Fig. 7** is an example of a three-screw tuner. The screws are one-quarter guide wavelength apart and are placed along the center line of the broadwall of the guide. If a wider range of impedances is to be matched, **3/8** and 518 guide wavelength separations can be used if space permits. The distance between the iris and the first screw of the tuner can be made **3/8** guide wavelength.

oscillator design and assembly

Since an iris-coupled waveguide cavity oscillator can tune only over a portion of the 10-GHz band, several factors affect the choice of frequency. First, the frequency will tune downward with increasing bias for most diodes. Diodes are available where frequency increases with bias; these are employed where temperature compensation is required. Second, the frequency will tune downward as the tuning screw penetrates the cavity. Third, the frequency decreases with increasing temperature.

Because the effects of bias, temperature, and mechanical tuning lower the frequency, choose a frequency of, say, 50 MHz or so above the desired operating frequency. With an AFC system, the oscillator can stay locked onto either the incoming received signal or a reference signal.

Design example. In the following example an operating frequency of 10.350 GHz was chosen. To compensate for the effects mentioned previously 10.400 GHz became the design frequency. An iris diameter of 0.25 inch (6.35 mm) was used. Before the cavity length can be calculated from **eq. 2,** λ_g and b_e must be found:

$$\lambda = \frac{c}{f} = \frac{3 \times 10^8 m/s}{10.4 \times 10^9 Hz} = 1.13 \text{ inch}$$

$$(0.02885 \text{ meters or } 28.85 \text{ mm})$$

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}} - \frac{28.85 \text{ mm}}{\sqrt{1 - \left(\frac{28.85 \text{ mm}}{2 \times 22.86 \text{ mm}}\right)^2}}$$

$$= 1.464 \text{ inch } (37.18 \text{ mm})$$

From **fig.** 4, the value for b_e can be found, providing the following ratios are calculated:

$$\frac{d}{a} = \frac{6.35 \text{ mm}}{22.86 \text{ mm}} = 2.8$$

$$\frac{a}{b} = \frac{22.86 \text{ mm}}{10.16 \text{ mm}} = 2.25$$

$$\frac{\lambda}{a} = \frac{28.85 \text{ mm}}{22.86 \text{ mm}} =$$

Using these numbers in **fig.** 4, b_e is 14. Next substitute the values for b_e and λ_g in **eq.** 2 and calculate cavity length:

$$\ell = \frac{\lambda_g}{2\pi}\left[\pi - \frac{1}{2}\tan^{-1}\frac{2}{|b_e|}\right]$$

$$\ell = \frac{37.18\ mm}{2\pi}\left[\pi - \frac{1}{2}\tan^{-1}\frac{2}{14}\right]$$

$$0.717\ inch\ (18.2\ mm)$$

The tuning screw can be placed anywhere between the diode post and the iris; however, a position of 0.22 inch (5.6 mm) from the iris was chosen, since placing it too close to the diode post would cause unstable operation.

Bias choke and diode post. Referring to **fig.** 5, the lengths of sections A, B, and C are calculated. Since sections A and C are sections of coaxial line with Mylar insulation, propagation velocity will be altered by $\sqrt{\epsilon_r}$ **of** the Mylar. Using **eqs.** 5 and **6:**

$$v = \frac{c}{\sqrt{\epsilon_r}} = \frac{3 \times 10^8\ m/s}{\sqrt{2.8}}$$
$$= 5.9 \times 10^8\ feet/second$$
$$(1.79 \times 10^8\ meters/second)$$

$$\lambda = - = \frac{1.79 \times 10^8\ m/s}{10.25 \times 10^9 Hz}$$
$$= 0.688\ inch\ (17.5\ mm)$$

In the above instance, the frequency for the middle of the 10-GHz band was chosen, i.e., 10.25 GHz. The length of sections A and C are $\lambda/4$, therefore:

$$\frac{\lambda}{4} = \frac{17.5\ mm}{4} = 0.17\ inch\ (4.4\ mm)$$

Section B **(fig. 5)** has air dielectric, therefore $\sqrt{\epsilon_r} = 1$ and its length becomes:

$$\lambda = \frac{c}{f} = \frac{3 \times 10^8\ m/s}{10.25 \times 10^9\ Hz} = 1.15\ inches\ (29.27\ mm)$$

$$\frac{\lambda}{4} = \frac{29.27\ mm}{4} = 0.28\ inch\ (7.3\ mm)$$

The characteristic impedance must be low for section A and C and high for section B. Based on choke design appearing in D. Evan's articles,[17,18,19] the outer diameter of the choke cylinder is 0.375 inch (9.525 mm). Sections A and C are chosen so that the spacing accommodates one or two layers of Mylar tape. The diameters of A and C are then 0.360 inch (9.144 mm). The diameter of B is approximately 1/3 of that of A and C. Section B diameter is chosen as 0.125 inch (3.175 mm). Using these diameters in **eq.** 4, the characteristic impedances of the sections are:

Sections A and C:

$$Z_0 = \frac{60}{\sqrt{\epsilon_r}}\ \ell n\ \frac{r_2}{r_1}$$

$$Z_0 = \frac{60}{\sqrt{2.8}}\ \ell n\ \frac{9.525\ mm/2}{9.144\ mm/2} = 1.46\ ohms$$

Section B:

$$Z_0 = \frac{60}{\sqrt{1}}\ \ell n\ \frac{9.525\ mm/2}{3.175\ mm/2} = 65.9\ ohms$$

The section above C maintains the same diameter as in section B and is made long enough to connect to a BNC connector. The Gunn diode connects to section A. **Fig.** 8 is an outline dimension for a typical low-to-medium power Gunn diode package. Each end is 0.061 inch (1.56 mm) in diameter. The mating hole in section A is made slightly larger; i.e. 0.0625 inch (1.59 mm) in diameter and at least as deep. The diode post is made from a 10-32 (M5) screw. The end opposite the head has a hole drilled out to 0.0781 inch (1.98 mm) diameter. The depth of the hole is at least 0.0625 inch (1.59 mm). The larger hole is needed to avoid fracture of the diode caused by misalignment errors.

The quarter-wave dimensions for movable back-wall choke assembly are based on the design shown in **fig. 6B.** Here the first and third sections are loaded with Mylar tape. The inner guide dimension is 0.9 × 0.4 inch (22.86 mm x 10.16 mm). The choke

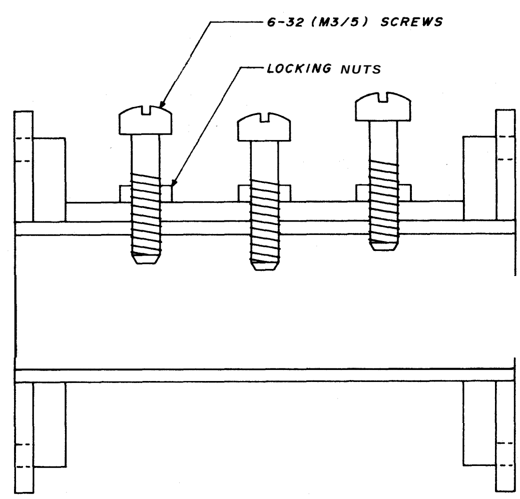


fig. 7. Cross section of the triple-screw matching section. Screws are one-quarter guide wavelength apart and are placed along the centerline on the guide broad wall. Spacings of 318 and 518 guide wavelengths can be used if space permits, since such spacings match a wider range of impedances.

block must be made smaller by the amount of the Mylar tape. If 5-mil-thick tape is used, the choke block is made with at least a 7-mil clearance on each side. These blocks then become 0.886 x 0.386 inch (22.50 mm x 9.80 mm) in their cross-sectional dimensions. Their length is computed from **eq. 8:**

$$\lambda_{gd} = \frac{\lambda}{\sqrt{\epsilon_r - \left(\frac{\lambda}{2a}\right)^2}}$$

$$= \frac{28.85\ mm}{\sqrt{2.8 - \left(\frac{28.85\ mm}{2 \times 22.86\ mm}\right)^2}}$$

$$= 0.733\ inch\ (18.62\ mm)$$

$$\frac{\lambda_{gd}}{4} = \frac{18.62\ mm}{4} = 0\ 183\ inch\ (4.66\ mm)$$

This value is the length of the first and third sections. The middle section length is computed from **eq. 7:**

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}}$$

$$= \frac{28.85\ mm}{\sqrt{1 - \left(\frac{28.85\ mm}{2 \times 22.86\ mm}\right)^2}}$$

$$= 1.464\ inch\ (37.18\ mm)$$

$$\frac{\lambda_g}{4} = \frac{37.18\ mm}{4} = 0.37\ inch\ (9.30\ mm)$$

The remaining calculation is that of the matching transformer or tuner. The first screw of the three-screw tuner shown in **fig. 7** is approximately 3/8 λ_g away from the iris. The distance between the screws is $\lambda_g/4$. Using 10.25 GHz, a band center, $\lambda_g = 1.5$ *inches* (38.10 mm). The distance between screws is 0.375 inch (9.5 mm), and between screw and iris is 0.563 inch (14.3 mm). When the sections were built, this dimension was closer to 0.590 inch (15 mm). The discrepancy produced no apparent problem.

construction

Now that all critical dimensions have been calculated, **figs. 9A** and **B** show the oscillator assemblies. The oscillators operate around 10.4 GHz and 10.3 GHz respectively. **Fig. 9B** differs from **A** mainly in backwall design. The construction of the oscillator follows a sequence of steps.

Cavity. Referring to **fig. 9A**, the cavity assembly is made from a piece of WR-90 waveguide about 2.5 inches (63.5 mm) long. If you have a drill press, milling machine, and lathe, the job becomes easier. However, much of the construction can be done if you have some hand tools and a drill press.

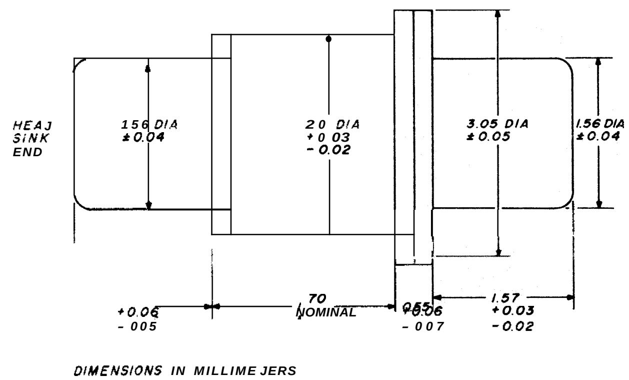


fig. 8. Outline dimensions of a typical low-power Gunn diode encapsulation. Diodes with output power ratings greater than 50 mW (such as the Alpha Industries DGB-6835C) have the cathode on the heatsink end. Those with less than 50 mW output power ratings (such as the Microwave Associates MA-49508) have the anode on the heatsink end.

Square off both ends of the guide and deburr. Obtain a piece of brass stock 0.75 x 1 x 1 inch (19 x 25.4 x 25.4 mm) and another 0.125 x 0.75 x 1 inch (3.175 x 19 x 25.4 mm). Fit a UG-39/U cover flange over one end of the guide and clamp the two brass pieces on each side. Make a waveguide cap for the other end of the guide out of a piece of I x 0.5 inch (25.4 x 12.7 mm) flat stock *60* mils thick. Place this assembly vertically on a fire brick. Rest the waveguide cap over the open waveguide. The flange end should be flush with the brick. Solder the assembly using a torch and let cool. File the flange end smooth; or if a milling machine is available, mill it smooth.

Measure a point 0.717 inch (18.2 mm) back from the flange and center it on the broad dimension of the guide on the 1 inch (25.4 mm) high block behind the flange. Drill a small pilot hole through the entire assembly using a 3132-inch (2.38 mm) diameter drill. This ensures the alignment of the bias choke with the diode post. On the bottom side, drill and tap for a 10-32 (M5) thread. On the choke block, drill a 3/8-inch (9.525 mm) diameter hole.

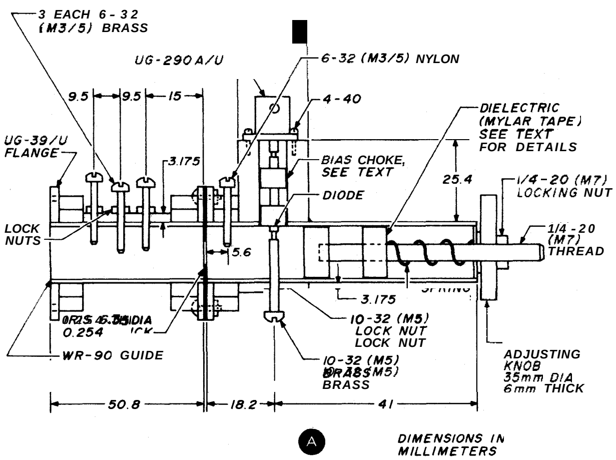
A word of caution here — start by drilling progressively larger holes until the required diameter is reached. Use a slow drill speed and oil. The drill has a tendency to grab, and you may have to anchor the assembly to the work table on the drill press.

Next, locate the holes of a UG-290/A BNC connector on the top of the bias-choke housing. Drill and tap four 4-40 (M3) screws 0.5 inch (12.7 mm) deep. Drill and tap a 6-32 (M3/5) hole 0.22 inch (5.6 mm) back from the flange face and centered on the broadwall in front of the bias block for a 6-32 (M3/5) nylon tuning screw. Drill a 1/4-20 (M7) clearance hole centered on the waveguide cap to accommodate the shorting choke assembly screw mechanism.

Iris. The iris is made from thin copper foil at least 10 mils thick. One of the UG-39/U flanges can be used as a pattern as an outline to locate the center. Cut out the iris and scribe a mark locating the center of the iris hole. Carefully drill a 0.25-inch (6.35 mm) diameter hole. Then drill out the four corner holes to clear 8-32 (M4) screws. (Later, the iris plate will be clamped between the oscillator assembly and the tuner assembly.)

Biaschoke. To make the bias choke, a lathe is handy since it makes the job easier but isn't absolutely necessary. Obtain a 318-inch (9.525 mm) diameter brass rod about 4 inches (101.6 mm) long and square off one end. Referring to **fig. 5,** mark off sections A, B, and C. These will be 0.172 inch (4.4 mm) 0.288 inch (7.32 mm), and 0.172 inch (4.4 mm) respectively. Above section C, a length of 0.236 inch (6 mm) can be cut back and attached to the bias connector.

Sections A and C are reduced in diameter 0.360 inch (9.144 mm) and B and the section above C to 0.125 inch (3.175 mm) in diameter. A 0.0625 inch (1.59 mm) diameter hole is drilled on center in the end of section A. Wrap a layer of Mylar tape around A and C. Fit the bias choke into the bias block and set it flush with the inside top of the waveguide. Cut the section of the choke above section C to mate up with the center conductor on the BNC connector and solder. The assembly is secured into place by four 4-40 (M3) screws.



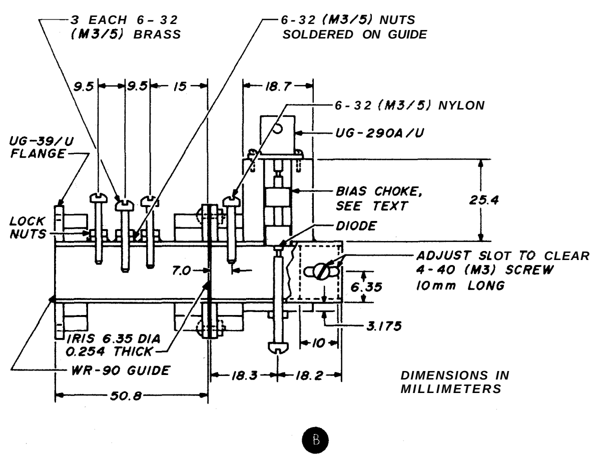


fig. 9. Assembly drawings of two Gunn oscillators. A design using an adjustable choke plunger backshort is shown in A. The oscillator in B has an adjustable contacting backshort that can be locked into place.

Diode post. The diode post is made from a 10-32 (M5) screw with a 0.078 inch (1.98 mm) diameter hole drilled into the center of the screw opposite the head. When the diode post is put into position, a 10-32 (M5) locking nut secures it.

Backwall. The movable backwall is made from two brass blocks 0.886 x 0.386 inch (22.50 x 9.80 mm) in cross section by 0.183 inch (4.66 mm) long, and a 2.5 inch (63.55 mm) long 1/4-20 (M7) threaded rod. Drill a hole, centered on the face of the block, and tap it for 1/4-20 (M7) thread through one block and half way through the other. Screw the rod into one block. Screw down the other block until it's about 0.366 inch (9.30 mm) from the first block. Wrap a layer of Mylar tap around each block. Slide a spring over the free end of the 1/4-20 (M7) rod.

The assembly is backed into the flange end of the oscillator cavity, and the rod goes through the clearance hole in the cap end of the waveguide. An adjusting knob, threaded for 1/4-20 (M7) is screwed on and used to move the choke assembly. A 1/4-20 (M7) is screwed on and used to move the choke assembly. A 1/4-20 (M7) locking nut is secured when the optimum position for the choke is found.

Place the diode, with heat-sinking compound, into the diode post screw. Carefully insert the screw and diode and lock into place. Do not exert too much pressure (see "precautions"). Lock into place with the locking nut. Insert the 6-32 (M3/5) nylon tuning screw at this point.

Tuner. The triple-screw tuner is made from a 2-inch (50.8 mm) section of WR-90 guide. The ends should be squared off and deburred. Solder a brass section 1.125 inch (28.58 mm) long by 1 inch (25.4 mm) wide by 0.125 inch (3.175 mm) thick onto the broad side of the waveguide between the two flanges. Clamp the plate to the waveguide, slide the flanges over each end of the waveguide and solder the assembly. Drill and tap three 6-32 (M3/5) holes centrally along the broad waveguide face through the block according to the locations in **fig. 9A**. Deburr the holes and insert three 6-32 (M3/5) screws 0.75 inch (19.05 mm) long with locking nuts. Connect the tuner to the oscillator section. Place the iris plate as shown and use four 8-32 (M4) screws and nuts to hold the assembly together.

This completes the oscillator assembly. The unit in **fig. 9B** is constructed in the same manner except for the backwall. Here a block closely fitting the inside of the guide is clamped into place on each side by a **4-40** (M3) screw.

test techniques and results

The type of tests that can be made depends on what equipment is available. If you've been able to obtain a fair amount of X-band test equipment, including a frequency meter and a slotted line from surplus dealers or flea markets, you're in good shape. Otherwise, a minimum of test equipment can be borrowed or made. For making some of your X-band test equipment consult reference 20.

Reflectometer test. For those who have access to an X-band reflectometer, looking at the reflected power from the cavity will show where the oscillator cavity resonates and where any spurious resonances occur within the 8.5-12.4-GHz band. Moving the backwall and tuning screw gives an indication of the tuning range. However, when power is applied to the oscillator, these frequencies will shift slightly. If a coaxial-to-waveguide transition is available, a transmission test is made to determine 10-GHz leakage out of the bias choke. The results on both oscillators of **fig. 9** indicate the leakage is down by 60 dB or more.

Slotted line measurements. If a slotted line is available, the reactance of the iris can be measured. I built several irises and the differences between calculated and measured reactance were within 5 per cent. If the values of normalized reactance are calculated carefully from **fig. 4,** they should be within 5 per cent of the actual values.

Swept-bias tests. Probably the most important testing technique used on Gunn oscillator design is

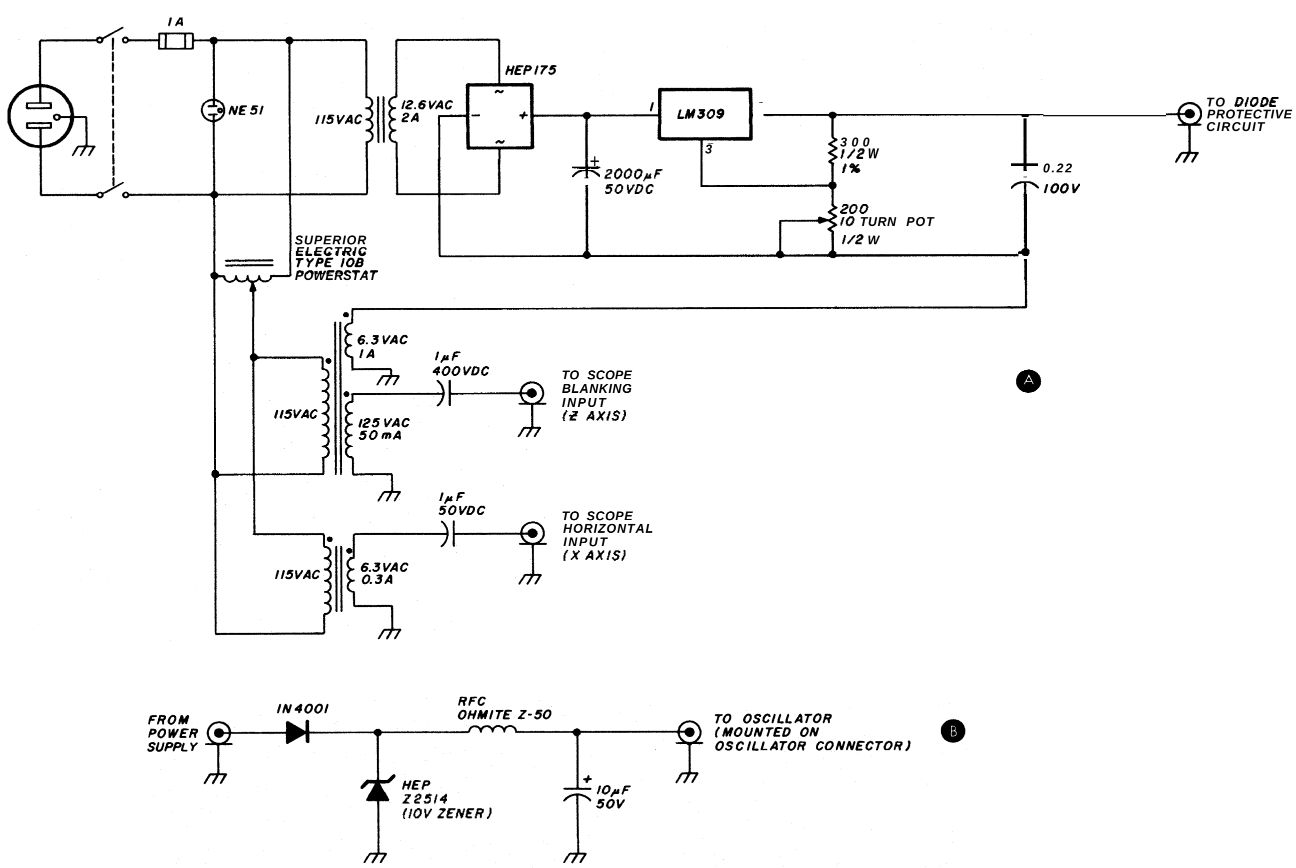


fig. 10. Schematic of the swept bias supply for testing Gunn-diode oscillators. In (A), a dc supply provides regulated dc voltages up to 500 mA. The 6.3 Vac winding of a transformer provides a variable ac supply, which is in series with the dc supply. Both supplies are connected to the Gunn oscillator through the protective circuit in (B).

the swept-bias test setup. A variable bias voltage is applied to the oscillator while its output is monitored with a detector and frequency meter. The swept response is displayed on an oscilloscope. Using such a test setup, the oscillator can be made to look into various loads while making different adjustments toward a stable operating point.

Connect the oscillator and tuner to a load through a directional coupler. Connect a frequency meter and detector to the directional-coupler coupling arm.

The swept-bias supply circuit shown in fig. **10A** is made of several supplies. A dc supply provides a regulated 5-10 volts dc. The supply can also provide up to 500 mA of current. The 6.3 Vac winding of a transformer, providing a variable ac supply, is in series with the dc supply. Both are connected across the Gunn diode oscillator through the protective circuit, fig. **10B**. Provisions are made for synchronized sweep voltage and a blanking voltage. Correct winding sense must be observed on the transformer windings to ensure proper sweep direction and blanking.

The protective circuit has a diode in series to protect the Gunn diode from reverse bias voltages and a zener diode to protect it from overvoltage in the forward-bias direction. The Z-50 choke and electrolytic capacitor ensure that a low impedance is presented to the oscillator and prevents buildup of dangerous voltage levels from parasitic resonances.

The protective circuit is connected to the swept source through a coaxial cable. A dummy load is then connected to the protective circuit. The dc voltage is set to 5 volts, and the sweep voltage is adjusted for **6** volts peak-to-peak if the MA-49508 diode is used, or 10 volts peak-to-peak if the DGB-6835C diode is used.

Note the powerstat settings for these ac voltages. Set the ac voltage to zero, turn off the supply, and remove the dummy load. Note: The protective circuit parts mount onto the Gunn oscillator. A matched load is used on the Gunn oscillator setup, and the nylon screw is backed out so that it's even with the inside top wall of the waveguide. The protective circuit box is connected to the oscillator. Turn on the power and advance the swept voltage to the preset position. Move the backshort until a stable swept response, shown in fig. **11D**, is obtained. (Figs. **11A** through **11C** show various stages of *misadjustment*.) The screws on the tuner can be adjusted for optimum output power while maintaining oscillation. If a particular antenna is to be connected, turn off the supply, connect the antenna, turn on the supply and make adjustment.

Fig. **11D** shows that, under proper operation, the onset of oscillations starts abruptly once the bias voltage exceeds the Gunn-diode threshold voltage.

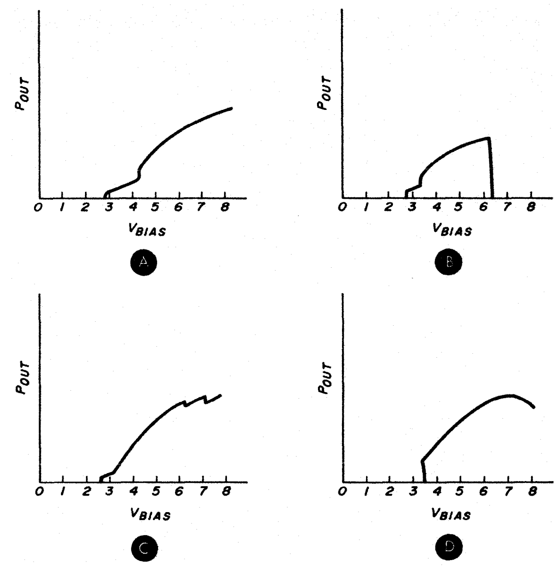


fig. 11. What to expect while running tests. Graphs A through C show respectively poor starting characteristics. oscillator-to-load mismatch, and backwall partially under the bias choke. Graph D shows the proper relationship between power output and bias.

As the bias voltage increases, output power increases to a peak and drops suddenly. Frequency usually decreases with increased bias voltage.

Modulation sensitivity tests were performed on the two types of Gunn diodes. The MA-49508 diode modulation sensitivity measurement was 15.6 MHz/volt, while the DGB-6835C modulation sensitivity measurement was 11.7 MHz/volt (both at room temperature).

For comparison, the published data on these diodes indicate an average modulation sensitivity around 7 MHz/volt at room temperature. The current through the diode remained fairly constant during these measurements.

I made a rough check of temperature effect on the oscillation frequency. The MA-49508 diode frequency drift with temperature was measured at −290 kHz/degree C compared to the maximum of −350 kHz/degree C on the data sheet. The DGB-6835C diode measured −0.87 MHz/degree C to the maximum −1 MHz/degree C on its data sheet.

some precautions

The Gunn diode, like any semiconductor device, can be damaged by an electrostatic discharge. Therefore, use care when handling the diodes. Mechanically, the diodes are fairly rugged upon compression. However, they can be damaged by shear fracture. This particular failure mode occurs when placing the diode into the oscillator cavity and tightening the diode post screw too much, especially if some axial misalignment exists in the diode-socket holes.

Check diode polarity before power is applied. In many cases positive bias with respect to ground is used. The medium- and higher-power diodes, such as the Alpha DGB-6835C, are heat sinked at the cathode end. The diode cathode end makes firm contact with the diode post, which is usually a good heatsink.

On the other hand, a lower-power diode, such as the MA-49508, is constructed with the heatsink on the anode end. It's important to note that, with this type of diode configuration, the diode is physically reversed inside the package. Therefore, if the cathode end is the grounded electrode, the package must be physically ***reversed*** before mounting it into the oscillator. If this is not done, the diode will be reversed-biased and will be damaged.

Another cause of diode damage is parasitic oscillation. Since the Gunn oscillator is a negative resistance device, oscillations will occur at any spurious resonance that exists from hf to the microwave freauencies. Oscillations can exceed the maximum bias voltage on the diode. If not by-passed at the bias choke cold end, any length of bias line can form a resonant system, and the oscillator may put out power at that particular frequency.

For this reason, it's desirable to mount any modulation/bias circuitry close to the oscillator. Any bias modulation circuit should present a low impedance to the Gunn oscillator. Check with the Gunn diode manufacturer for their recommended protection circuit.

One very important precaution must be mentioned regarding Gunn oscillators (or any microwave oscillator). ***Do not*** look into the open end of the waveguide while power is applied to the oscillator. Close up, rf power density can exceed OSHA's *10mW/cm²* safety limit. Fortunately, the rf power density falls off to a safe level a short distance from the oscillator. Your eyes are especially susceptible to damage from rf power radiation, so never look into an open waveguide ***or*** stand in front of a microwave antenna!

conclusion

I've presented some of the details in the design of a Gunn oscillator. Some of the parameters and conditions for building a working oscillator are given. Also presented is a detailed design and assembly description of the oscillator. Test results are given, and some precautions are stated since it's possible to damage these devices if not properly treated.

Detailed circuits to drive and modulate the diode were not presented, since this is a different subject. Further information can be found in references 21-27. The main purpose of this article was to present how a Gunn oscillator can be built and tested.

acknowledgments

I wish to thank Mr. Mark Crandell of Hughes Aircraft Corp. for his helpful advice and information discussions. I also want to thank Mr. Richard Wade of Microwave Associates, and Mr. Richard Gerrish of Alpha Industries, for their advice and for providing the diode engineering samples. My appreciation also goes to Mr. George Cutsogeorge, W2VJN, for his advice on the circuit design used in the test equipment.

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ham radio

measuring capacitance of electrolytic capacitors

Some simple math and a volt-ohmmeter make it possible to evaluate your electrolytics

You can build or buy many instruments that will tell you, with reasonable accuracy, the value of a capacitor. The capacitance of electrolytics is not easily determined, however, because they are polarized and tend to have sizeable leakage current. The following material provides a method for evaluating electrolytic caps.

the electrolytic capacitor

Electrolytic capacitors are made of either aluminum or tantalum electrodes and an electrolyte. Boric acid is the usual electrolyte used with aluminum, and sulfuric acid is frequently used with tantalum. The active electrode may be etched; or in some cases, tantalum is formed as a porous plug. In either case, a thin insulating layer is formed on the active electrode. This layer acts as the dielectric in which the capacitive energy is stored. In a well-behaved electrolytic cap, this thin layer looks a lot like a rectifier. That is, current flow in the polarized direction is small, while in the reverse direction it can become quite large. It is this polarized state which makes bridge-type capacitance measurements difficult.

uses

Because it's possible to obtain a large quantity of microfarads in a small package, electrolytics are used as filter elements in power supplies. Here they function as energy-storage devices, charging when the voltage is high then supplying current when the voltage starts to drop. Electrolytics are also used as coupling elements. The large capacitance acts as a low alternating-frequency impedance between two points in a circuit where a difference in direct-current potential exists. In effect, this application also uses the capability of the electrolytic to store energy.

Nothing lasts forever, and this is particularly true of electrolytic capacitors. Loss of liquid, degeneration of the dielectric film, and abuse may drastically change the basic capacitance of the unit. The question then is, How do you measure the capacitance of an electrolytic? One answer, which goes back to fundamentals, is to measure its ***energy storage ability.***

mathematical derivation

Almost every Radio Amateurs' handbook gives the relationship between supply voltage, voltage across a capacitor, and time as a function of the charging resistance. This is an exponential relationship, but need not cause concern for the nonmathematically inclined. If you're interested in the basic math, finish this section; if not, just jump to the next.

First, the resistance and capacitance (RC) involved in a charge or discharge circuit, when multiplied together, give what is called the "time constant." If you have a large value of resistance, the time to charge or discharge will be long. The same may be said for capacitance. If the RC time constant is large, the time to charge or discharge is large.

We'll concern ourselves with discharge, in which the voltage across a capacitor with a shunt resistor is expressed as follows:

$$V = \frac{e_0}{exp\,(t/RC)} \qquad (1)$$

The term e_0 is the voltage to which the capacitor was charged at ***timezero. Exponent*** (t/RC) is **2.718** raised to the value of t/RC. Here t is the elapsed time in seconds since the resistance and capacitance started to discharge. Note that if t is just equal to RC, we have **2.718** raised to unity power, or simply **2.718.**

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With the technical details out of the way we can now attack the problem. If we measure the time required for a capacitor to reach 112.718 of the voltage to which it was originally charged, then we know the value of the product of the resistance and capacitance. Divide the time by the resistance and you end up with the capacitance. Simple, isn't it?

measurement example

Now for a practical example. Suppose you use a volt ohmmeter with a 20,000-ohm/volt dc sensitivity, and a capacitor has a rated voltage of 35 volts. You'll use the 50-volt scale, so the effective resistance of the meter across the capacitor will be 50 times 20,000 or 10^6 ohms. This is a rather large resistance if the capacitor is several thousand microfarads, since the time constant will be several thousand seconds. Use a 47-k resistor in parallel with the meter so that the ***R*** in the circuit is

$$R = \frac{1}{\left(\frac{1}{10^6} + \frac{1}{4.7 \times 10^4}\right)} \text{ ohms}$$

$$= 4.49 \times 10^4 \text{ or } 44{,}900 \text{ ohms}$$

(round off to *45 kilohms*)

We'll assume a) that the capacitor to be tested has been charged long enough so that it is reformed and up to capacitance, and b) we'll start with 35 volts. The voltage at which timing will be stopped will be 3512.718, or 12.88 volts.

Connect the VOM across the capacitor and the 47k resistor, charge the capacitor to 35 volts, then note the time at which you remove the power from the circuit. Note that you must remove one lead from the power supply. Simply turning off the supply is not enough. Now watch the voltage. When it reaches 12.88 volts, again note the time. The length of time elapsed, in seconds, is the time constant. Suppose it was 83 seconds. This value divided by the resistance, 45 kilohms, is **1844** microfarads, the effective capacitance of the unit being measured.

closing remarks

This method of measurement doesn't take into account any leakage effects, so the capacitance you measure may be a bit less than the actual dynamic capacitance. Also recognize that, when dealing with high-voltage capacitors, there is danger of lethal shock and you should be extra careful of any voltage over, say, 15 volts. Another precaution concerns direct shorting a capacitor. Most units will survive the current surge from a screwdriver short; however some may not and it's better to use a resistor than to burn up a screwdriver.

ham radio

More Details? CHECK—OFF Page 94

appreciating the
L matching network

Theory and application of the L network in Amateur circuits

Often one looks at transmitter schematics and notices the simple two-element networks in the shape of the letter L, which are used to match a source impedance to a load impedance. These seemingly simple devices, which display high efficiencies, often baffle many Amateurs. The thought of using two completely reactive components to transform the value of a resistor by itself seems intriguing. Such simple circuits find their way into almost every piece of communications equipment; matching transistor input and output, matching a reactive antenna to a flat line, and matching a driver stage to a final amplifier. One drawback of the L network is that it is an exact match at only one frequency.

Let's look into the L network to try to understand it better. But we'll try to use less mathematics and more intuition to derive the equations. Once we've done this, we'll then be able to confidently use the networks not only to transform values of purely resistive loads to match a source resistance, but we'll also be able to match any complex impedance as well. The first step in understanding the L network is to examine the series-to-parallel impedance conversions.

series-parallel impedance conversions

A series combination of a resistor and an inductor or capacitor may be transformed into an equivalent parallel circuit. If the impedance of two components inside a black box were measured at a single frequency, you would have two answers: a series circuit and its equivalent parallel combination, **fig. 1.**

To understand the series equivalent of a parallel circuit or vice versa, begin by writing the simple expression for the impedance across two parallel elements, Z_p, and later try to shape it into a series circuit.

$$Z_p = \frac{(jX_p)(R_p)}{R_p + jX_p} \tag{1}$$

where **j** indicates the imaginary or reactive component $(j = \sqrt{-1})$.

To obtain a real denominator, multiply both numerator and denominator of **eq. 1** by the complex conjugate $R_p - jX_p$. The resultant is

$$Z_p = \frac{(jX_pR_p)(R_p - jX_p)}{R_p^2 + X_p^2} \tag{2}$$

where $j^2 = -1$.

Next, separate the equation into the real (resistive)

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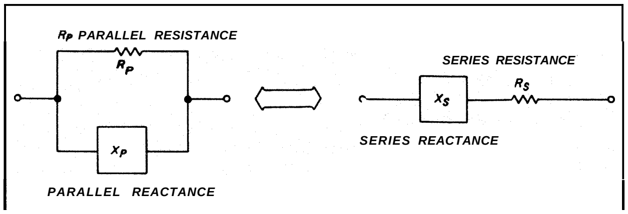

fig. 1. A series combination of resistance and reactance can always be found that exhibits the equivalent impedance of any given parallel combination of resistance and reactance.

and imaginary (reactive) components by grouping all terms with js by themselves.

$$Z_p = \left(\frac{X_p^2 R_p}{R_p^2 + X_p^2}\right) + j\left(\frac{X_p R_p^2}{R_p^2 + X_p^2}\right) \qquad \textbf{(3)}$$

Thus an equation in the form of a series combination of a real (resistive) component and an imaginary (reactive) component results:

$$Z_s = R_s + jX_s = \text{series equivalent impedance} \qquad \textbf{(4)}$$

$$R_s = \frac{X_p^2 R_p}{R_p^2 + X_p^2} = \frac{R_p}{1 + (R_p^2/X_p^2)} \qquad \textbf{(5)}$$

$$X_s = \frac{X_p R_p^2}{R_p^2 + X_p^2} = \frac{X_p}{1 + (X_p^2/R_p^2)} \qquad \textbf{(6)}$$

If circuit Q is introduced, the expressions are simplified. The Q for a parallel circuit is R_p/X_p, whereas that for a series circuit is X_s/R_s. Thus, the series equivalent elements are

$$R_s = R_p/(1 + Q^2) \qquad \textbf{(7)}$$

$$X_s = X_p/(1 + 1/Q^2) \qquad \textbf{(8)}$$

A transformation from a series to the equivalent parallel circuit can be made similarly:

$$R_p = R_s[1 + (X_s^2/R_s^2)] = R_s(1 + Q^2) \qquad \textbf{(9)}$$

$$X_p = X_s[1 + (R_s^2/X_s^2)] = X_s(1 + 1/Q^2) \qquad \textbf{(10)}$$

The sign of the reactive component doesn't change; the series equivalent of a parallel capacitance is still a capacitor.

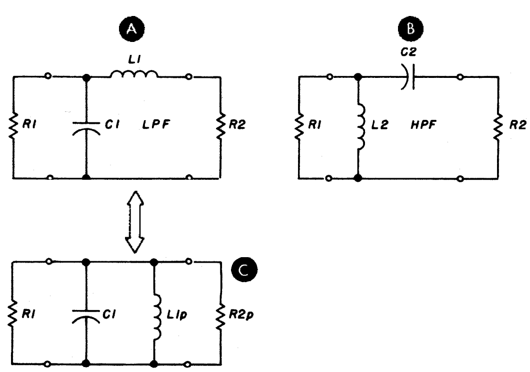

fig. 2. The L network may assume the form of either (A) lowpass or (B) highpass filter. A parallel equivalent circuit is shown in (C).

deriving the L network

Two types of matching L networks are shown in fig. 2, A and B. One resistance, $R1$, is matched to a smaller resistance, $R2$. The shunt element is across the larger resistor, and the series element is connected to the smaller resistance. The method of deriving the equations will be to transform all the elements into the same form and make the reactances cancel.

For the circuit in fig. 2A, begin by transforming the series combination of $L1$ and $R2$ into a parallel equivalent circuit, which then forms the parallel circuit of $R1$, $C1$, $L1_p$, $R2_p$ (fig. **2C**). The parallel equivalent of $R2$ is:

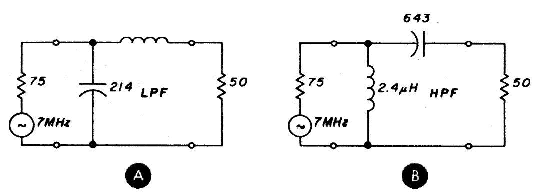

fig. 3. Components for matching networks using a (A) lowpass filter do not match those of the highpass arrangement, (B).

$$R_p = R_s[1 + (X_s/R_x)^2] \qquad \textbf{(11)}$$

$$R2_p = R2\,[1 + (X_{L1}/R2)^2] \qquad \textbf{(12)}$$

This transformed value of $R2$ must equal $R1$ for impedance matching. If **eq.** 12 is solved for X_{L1} (the inductive reactance of $L1$), substituting $R1$ for $R2_p$ in the equation yields

$$X_{L1} = R2\sqrt{(R1/R2)-1}\,;\; L1 = X_L/2\pi f \qquad \textbf{(13)}$$

Thus the value of $L1$ is based on the resistor ratios. Next transform inductance $L1$ into its equivalent parallel value:

$$X_p = X_s[1 + (R_s/X_s)^2] \qquad \textbf{(14)}$$

$$X_{LP} = X_L[1 + (R2/X_L)^2] \qquad \textbf{(15)}$$

$$X_{LP} = R1/\sqrt{(R1/R2)-1} \qquad \textbf{(16)}$$

Now notice that we have a parallel circuit (fig. **2C**). The transformed parallel inductance, X_{LP}, can be cancelled by an equal and opposite parallel capacitive reactance:

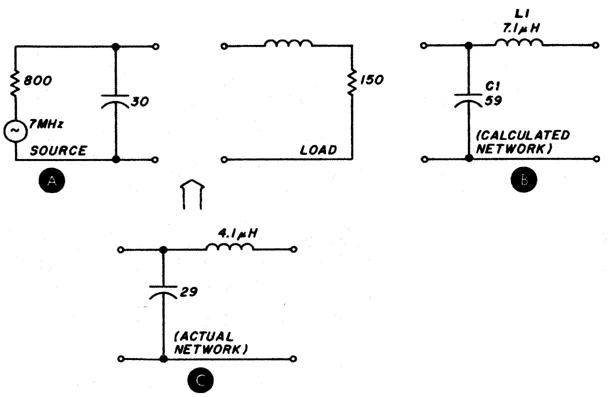

fig. 4. At 40 meters the source and load impedance, (A), partially absorb the L calculated network reactances, (B), resulting in a new network (C).

$$X_{C1} = X_{LP} = R1/\sqrt{(R1/R2)-1}\,; \qquad C1 = 1/2\pi f X_C \tag{17}$$

Thus series inductor $L1$ increased $R2$ resistance until it matched $R1$, then shunt capacitor $C1$ cancelled the reactance of the inductor, as in a parallel tank circuit. For the network in **fig. 26,** simply transform the series arrangement of $C2$ and $R2$ into a parallel-equivalent circuit, with the capacitance value determined by the necessary resistance transformation and with inductance $L2$ of sufficient value to cancel the transformed capacitive reactance.

For each of these networks we could have alternatively transformed the parallel combination on the left into a series-equivalent circuit to match $R2$ impedance. The value of the other reactance would cancel the series equivalent.

resistive matching

Now we'll match a resistive generator to a resistive load. An example might be matching a 75-ohm generator, such as the output impedance of an oscillator, to a 50-ohm load (input impedance of a grounded-grid amplifier) on 40 meters. To accomplish this use either configuration **fig.** 2A or **26.** The larger resistance is always designated $R1$.

$$R1 = 75\ ohms,\ R2 = 50\ ohms,\ f = 7\ MHz$$

Network 2A:

$$X_{L1} = R2\sqrt{(R1/R2)-1} = 50\sqrt{(75/50)-1} = 35.3\ ohms$$
$$L1 = X_{L1}/2\pi f = 0.8\ \mu H$$
$$X_{C1} = R1/\sqrt{(R1/R2)-1} = 75/\sqrt{(75/50)-1} = 106\ ohms$$
$$C1 = 1/2\pi f X_{C1} = 214\ pF$$

Network 2B:

$$X_{L2} = R1/\sqrt{(R1/R2)-1} = 75/\sqrt{(75/50)-1} = 106\ ohms$$
$$L_2 = X_{L2}/2\pi f = 2.41\ \mu H$$
$$X_{C2} = R2\sqrt{(R1/R2)-1} = 50\sqrt{(75/50)-1} = 35.3\ ohms$$
$$C2 = 1/2\pi f X_{C2} = 643\ pF$$

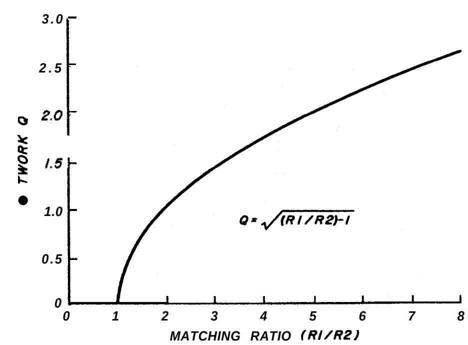

fig. 6. Circuit Q is determined by the ratio of the resistors to be matched.

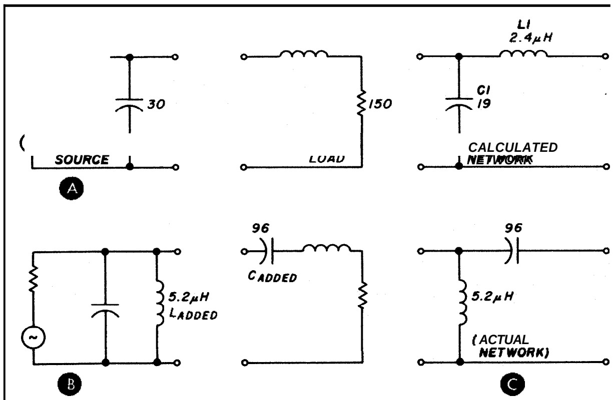

fig. 5. At 15 meters the computed L network (A) reactances must be completely absorbed by the source and load impedance. Additional compensation is required (B) to result in a totally new network (C).

Thus the same impedances can be matched with two different sets of components, **fig.** 3.

The principal difference between networks **2A** and **26 (fig. 2)** is that network **2A** is a **lowpass** filter and network **2B** is a **highpass** filter. For most applications it's desirable to suppress harmonics; thus network **2A** is chosen. **It** turns out that network **2A** is also easier to construct, because one end of a variable capacitor can be physically grounded to decrease lead inductance.

complex impedance matching

Now that we can match one resistor to another, we'll try to match complex impedances. Begin by designing an L network for matching the resistive part of a source to the resistive part of the load, then

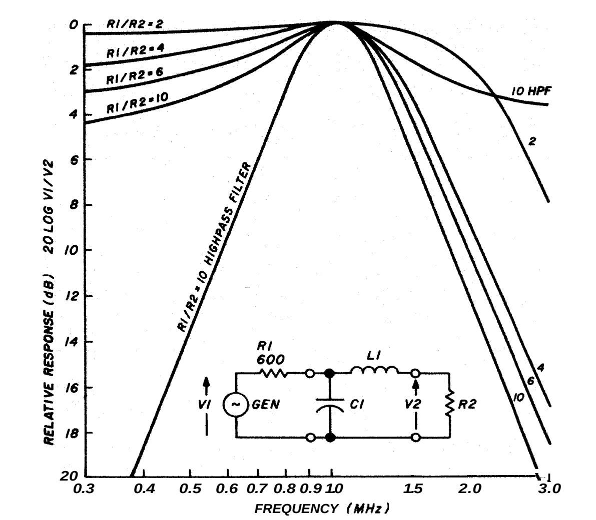

fig. 7. Several L networks were measured using various matching ratios.

modify the component values of the L network to take into account the reactive parts of source and load. Here's an example using network **2A**, because we want to suppress harmonics.

40-meter-band example. Decide whether the sources resistance is higher or lower than the load resistance. The higher resistance is labelled $R1$. Let's match a 40-meter final amplifier, which has a plate resistance of 800 ohms with a shunt capacitance of 30 **pF,** to a 150-ohm load with a series inductance of **3 μH, fig.** 4A.

$$R1 = 800\ ohms, R2 = 150\ ohms, f = 7\ MHz$$

Use the equations for network **2A**, which states:

$$X_{L1} = R2\sqrt{(R1/R2)-1} = 312\ ohms$$
$$L1 = X_{L1}/2\pi f = 7.1\ \mu H$$
$$X_{C1} = R1/\sqrt{(R1/R2)-1} = 384\ ohms$$
$$C1 = 1/2\pi f X_{C1} = 59\ pF$$

If this network, **fig.** 4B, is inserted between the two resistances, we would have matched the resistances at 7 MHz. Note, however, that we must also compensate for the source shunt capacitance and the load series inductance. First, capacitance $C1$ of the matching network is in parallel with the source shunt capacitance. Therefore, $C1$ must be equal to 29 **pF.** The remaining 30 **pF** required for matching is already present in the source shunt capacitance. The same thing can be done with the load impedance. Instead of adding a 7.1-pH inductor as calculated, use the 3-pH inductance present in the load and add 4.1 pH to make up the difference, **fig.** 4C.

15-meter-band example. The previous example was easy in that the source and **load** impedances turned out to be directly absorbed into the matching network. What if we wanted to match the same load and source on a different band, say 15 meters?

Using the same equations at the new frequency, **fig.** 5A, $f = 21\ MHz,\ C1 = 19\ pF,\ L1 = 2.4\ \mu H,$ the value of $C1$ will be totally absorbed into the source shunt capacitance with 11 **pF** left over, which must be cancelled. An inductor could be inserted across the source shunt capacitance to cancel only 11 **pF** of capacitance, **fig.** 5B. The inductive reactance of the added coil must equal the capacitive reactance of the excess shunt capacitance:

$$\begin{aligned} X_{L(added)} &= X_{C(excess)} = 1/(2\pi f_{excess}) \\ &= 1/[2\pi f(30pF - 19\ pF)] \\ &= 1/(6.28)(21\ MHz)(11\ pF) \\ &= 688\ ohms \end{aligned}$$

$$\begin{aligned} L\ added &= [X_{L(added)}/2\pi f] \\ &= 688/2\pi f \\ &= 5.2\ \mu H \end{aligned}$$

The amazing thing is that, when we build the matching L network, instead of a shunt capacitance, $C1$, we'll actually have a shunt inductance, L_{added}, **fig.** 5C.

Now for matching the load impedance. The load already has 3 pH of inductance, which is greater than the 2.4-pH series inductance required, leaving an excess of 0.6 pH to be cancelled. A series-capacitive reactance, just large enough to cancel the 0.6 pH of excess inductance, is needed **(fig. 5B).**

$$X_L(excess) = 2\pi f L(excess) = 2\pi f(0.6\ \mu H) = 79\ ohms$$
$$C(added) = 1/2\pi f X_L(excess) = 96\ pF$$

The L network, **fig.** 5C, looks more like the high-

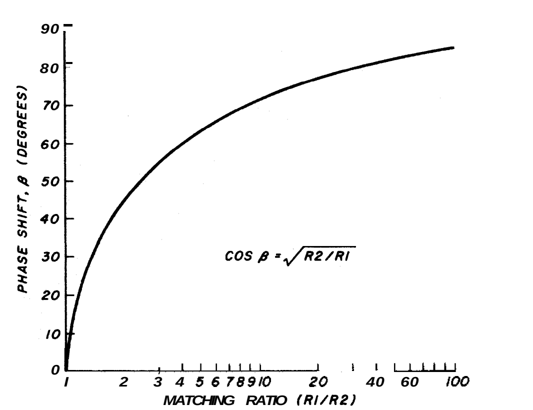

fig. 8. Phase shift of the L network depends also on the matching ratio, ($R1/R2$).

pass filter shown in **fig. 2B** than what we started out to make. Our new filter merely subtracts enough reactance so that the inherent reactance remaining in the source and load perform the actual matching. The total effect is, however, basically that of a lowpass filter.

The only problem now remaining in applying the L network to any impedance is that source and load might not be in the right form. What is needed is for the impedance on the shunt side of the L network to be described in a parallel or shunt form, and that on the series side to be described in a series configuration. If this is not the case, the impedances must be converted using the series-parallel conversion rules discussed earlier. Once this is done source and load reactances can be easily cancelled.

bandwidth

The Q or quality factor of the matching network determines the bandwidth between the upper (f_1) and lower (f_2) 3-dB frequencies.

$$Q = \frac{f_0}{f_1 - f_2} = \frac{\text{operating frequency}}{\text{bandwidth}} \qquad (18)$$

The Q for an L matching network is:

$Q = \sqrt{(R1/R2) - 1}$ as shown in **fig. 6.**

Thus Q, or selectivity, is determined by the ratio of the impedances to be matched and cannot be selected independently, as in the more complex π and T matching networks. As the matching ratio increases, so does the circuit Q. The efficiency of an L network is usually greater than 95 per cent:

$$efficiency = [R2/(R2 + R_{coil})]\,[100] \qquad (19)$$

some experiments

I put a few L networks together and measured them to see what they could do, **fig. 7.** I decided to match a 600-ohm generator to various loads at 1 MHz. How the Q affects attenuation at the second and third harmonics is apparent. For a matching ratio greater than 4 (Q of 1.7), the slope of the attenuation quickly approaches 12-dB/octave. Every time the frequency is doubled, attenuation increases by 12 dB. The series inductor contributes 6-dB/octave, while the shunt capacitor completes the additional 6 dB. Thus you can expect about 12 dB of attenuation at the second harmonic and 19 dB at the third. I also compared the response of an L network for a matching ratio of 10:1 using either the highpass or lowpass version of the L network. They appear symmetrical about the center frequency.

The ability of an L network to match impedances is also limited by the characteristics of the components. As the frequency is sufficiently increased, the series inductance in the leads of a capacitor predominate, making it look more like an inductor than a capacitor. The same thing happens to inductors at high frequencies, with the capacitance between coil windings tending to shunt the effects of inductance.

The phase shift, β, for an L network as shown in **fig. 8A** is:

$$\cos\beta = \sqrt{R2/R1} \qquad (20)$$

when β is measured in degrees, as shown in **fig. 8.** The lagging phase shift approaches 90 per cent as the matching ratio exceeds 10:1. As the matching ratio is increased, the necessary shunt capacitance, $C1$, and series inductance, L1, decrease for a constant input impedance. Therefore, as the matching ratio increases, the network behaves more like a capacitor. This phase delay must be taken into account when matching into different elements of a phased array antenna.

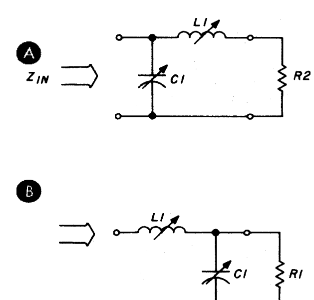

fig. 9. Input impedance varies as the matching elements are adjusted. In A a high impedance is matched to a lower impedance. *B* shows matching to a higher impedance.

Finally, let's take a brief look at the input impedance of an L network as series inductor, $L1$, and shunt capacitor, $C1$, are varied. For the case of matching a high impedance downward, **fig. 9A,** note that the real part of the transformation is controlled mostly by varying series inductor $L1$. Remember that matching circuits have been designed for the resistance ratio, then modified slightly to remove reactances. Increasing $L1$ increases the real part of the input impedance. The value of shunt capacitance $C1$ controls the reactive part of the impedance presented to the input. Increasing capacitance increases the $-j$ term.

For the case of matching to a higher impedance, **fig. 9B,** decreasing $C1$ causes the real part of the input impedance to increase. Increasing $L1$ makes the input impedance appear more inductive.

ham radio

the half-wave balun: theory and application

Some notes on this versatile tool for impedance matching

The half-wave balun, in which the inverted portion of the balanced output signal is obtained by using a half-wave section of transmission line, is perhaps the simplest and most economical balun available; because of this simplicity and low cost, it's one of the most popular. Its design is straightforward — so simple, in fact, that equations are hardly necessary. Just measure a half wavelength of transmission line and you have a balun.

Despite the popularity of this type of balun, equations that predict its performance are generally unavailable. The only previous work on this type of balun of which I am aware was done by Woodward.[1,2]

The usual schematic of the half-wave balun is given in **fig.** 1. For analysis it's convenient to redraw **fig.** 1 as shown in **fig. 2.**

The balanced load is assumed to be perfectly balanced; admittedly this isn't always true in practice, but it would not be fair to blame the balun for imperfections caused by the load. The impedance seen by the generator at the unbalanced port is $Z_L/2$ in parallel with $Z_L/2$, or $Z_L/4$, as expected.

In most baluns, the two most important characteristics are the impedance match and the degree of balance, both versus frequency.

impedance balance

One desirable characteristic of a balun is that it presents an essentially constant impedance over a wide bandwidth. Since the half-wave balun will be exactly one-half wavelength at only one frequency, it's useful to examine the impedance presented to the generator in **fig.** 2. The input impedance in **fig. 2** is composed of one-half the load impedance, Z_1, in parallel with the input impedance of the half-wave balancing line. This line, in turn, is terminated in the other half of the load impedance, Z_2. The impedances Z_1 and Z_2 will both be assumed to be independent of frequency. The input impedance of the balancing line, however, will depend on the characteristic impedance and the electrical length of the line, and hence on frequency.

The impedance presented at the input of the half-wave balancing line is given by the familiar transmission line equation:

$$Z_{in} = Z_c \left[\frac{Z_2 \cos\theta + jZ_c \sin\theta}{Z_c \cos\theta + jZ_2 \sin\theta} \right] \qquad \textbf{(1)}$$

where Z_{in} = input impedance of half-wave balancing line

Z_c = characteristic impedance of balancing line

Z_2 = load impedance of the balun; equal to one-half balanced load

After some arithmetical manipulation, which I'll explain in the appendix, and setting $k = Z_c/Z_2$, we obtain the ratio of input impedance to load impedance for the resistive (real) component:

$$\frac{Z_{in}}{Z_c} = \frac{k^2}{(k\cos\theta)^2 + (\sin\theta)^2} \qquad \textbf{(2)}$$

Eq. 2 is plotted for various values of k over a range of θ from 90 to 270 degrees; see **fig.** 3. This represents a frequency range from one-half to one and one-half times the design frequency. The ordinate gives the input impedance in terms of the load impedance.

As shown, the ordinate is matched ($Z_{in} = Z_c$) for all values of characteristic impedance only when $k = 1$, **i.e.,** when the characteristic impedance equals the load impedance. If you stop and think about it, both these results should have been expected.

For values of $\theta = 90$ and 270 degrees, the balanc-

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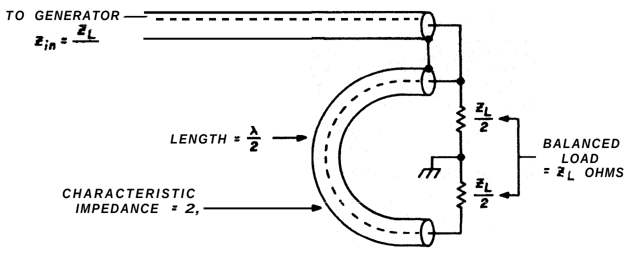


fig. 1. Basic half-wave balun circuit using lengths of transmission line.

ing line is an odd number of quarter-wavelengths; thus the balancing line inverts the load impedance about the characteristic impedance squared. For these line lengths, the ratio Z_{in}/Z_2 is equal to k^2.

Fig. 4 shows the reactive (imaginary) component of balancing line input impedance. In this case the reactance is zero for all frequencies only when $k = 1$, again the matched case.

The balun input impedance is, of course, the impedance given from **figs.** 3 and **4** in parallel with $Z_1 = (Z_L/2)$. (This latter impedance has been assumed to be independent of frequency, but in the real world it may also vary with frequency.)

balance bandwidth

In addition to the impedance bandwidth, it's desirable to know balance bandwidth. Balance bandwidth means that the frequency range over which the currents through the two halves of the balanced load are equal in magnitude and opposite in phase with a specified tolerance. Following Woodward,[1,2] we use as the criteria of balance:

$$balance = \left| \frac{I_1 - I_2}{I_1 + I_2} \right| \quad \textbf{(3)}$$

From **fig.** 2 current I_1 is given by $\frac{e_0}{Z_1}$. Current I_2 is given in terms of the load resistance and characteristic impedance by the well-known transmission line equation3

$$e_0 = e_2 \cos\theta - jI_2 Z_c \sin\theta \quad \textbf{(4)}$$

After arithmetical manipulations as shown in the appendix, we obtain:

$$I_2 = \frac{e_0}{Z_2 (\cos\theta \mp jk \sin\theta)} \quad \textbf{(5)}$$

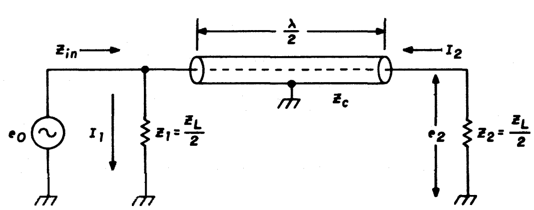


fig. 2. Equivalent of half-wave balun shown in *fig. 1*.

from which $$\left| \frac{I_1 - I_2}{I_1 + I_2} \right| = \left[\frac{(\cos\theta + 1) + jk \sin 6}{(\cos 6 - 1) \mp jk \sin\theta} \right] \quad \textbf{(6)}$$

where k $= \frac{Z_c}{Z_2}$ as before.

After rationalizing, separating into real and imaginary parts, and finding the magnitude (described in the appendix), a plot of this equation is given in fig. 5; the system is balanced when the plot equals zero. As can be seen, a perfect balance occurs only when $\theta = 180$ degrees as might be expected. The balance bandwidth improves for small values of $k = Z_c/Z_2$. Thus, for best balance bandwidth, the load impedance should be large compared with the balancing-line characteristic impedance, This is contrary to impedance bandwidth considerations, in which the widest impedance bandwidth is obtained for the matched case.

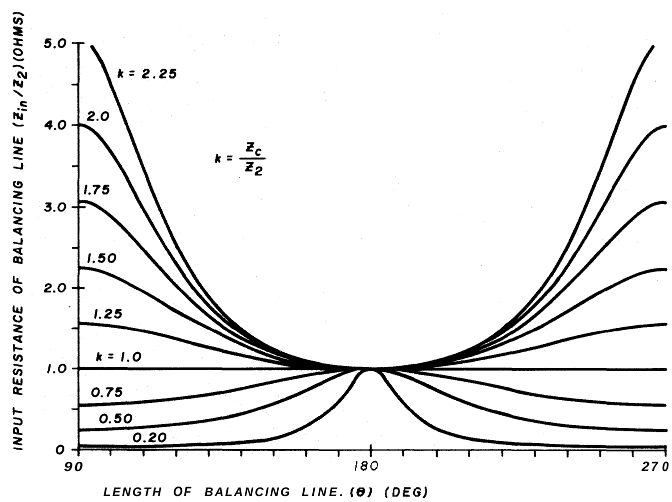


fig. 3. Input resistance of the half-wave balancing line as a function of line length for various values of characteristic impedance.

Increasing the load impedance to values greater than about ten times the characteristic impedance $(k \leq 0.1)$ gives little additional improvement in balance bandwidth. For small values of k, eq. **6** approaches $(\cos\theta + 1)/(\cos \ell - 1)$, which is independent of k.

phase shift

One final factor to consider is the change in phase with frequency caused by propagation delay through the balancing line. At the design frequency, the relative phase at the output of the line is 180 degrees with respect to the input. This phase shift will vary directly with frequency. Assuming that the amplitude (current or voltage) is constant with frequency, the phase change will cause reduction in the in-phase component (referenced to 180-degree phase-shift

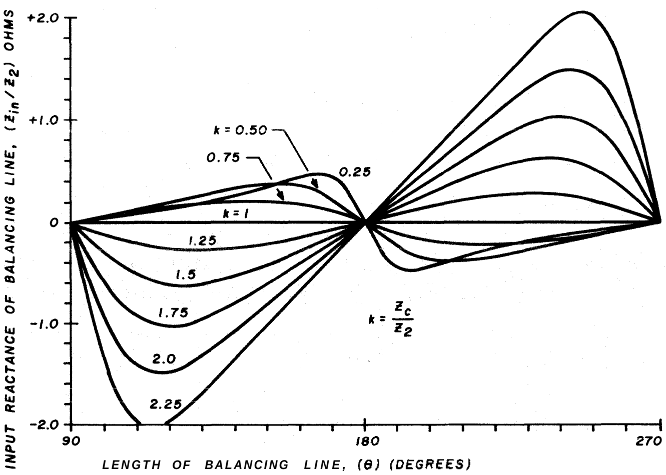


fig. 4. Input reactance of the half-wave balun as a function of line length for various values of characteristic impedance.

from the balancing-cable input) and the generation of a quadrature component.

The in-phase and quadrature components will be proportional to

$$\cos\theta \text{ (in-phase) and } \sin\theta \text{ (quadrature)} \qquad (7)$$

respectively, where 8 is the deviation from 180 degrees with respect to the input. From **eq. 7,** it can be seen that a deviation of 26 degrees (14.5 per cent) of design frequency will cause a I-dB reduction in output amplitude through one-half of the balanced load caused by phase change. Similarly, a 6.24 degree deviation (3.4 per cent of design frequency) will cause a 1-dB quadrature component to be generated.

The effect of the reduction in the in-phase component is an unbalance through one-half the load. The effect of the quadrature component will depend upon the balanced-load characteristics. If the balanced load is an antenna, two possibilities are **a)** generation of unexpected side lobes, or **b)** generation of a circular component to the antenna response.

In other situations the quadrature component may or may not be important. The effect of the quadrature component should be considered for each application.

example

A typical application of the half-wave balun is to couple a balanced folded dipole with a nominal impedance of 300 ohms to a 75-ohm coaxial cable. This represents a **4:1** impedance transformation. For convenience, the same 75-ohm coax is generally also used for the balancing section. In this case, $Z_2 = Z_1 = 150$ *ohms* equals one-half the balanced load, so that with 75-ohm coax, $k = 75/100 = 0.5$. From the $k = 0.5$ curve, it can be seen that the frequency range is not particularly good — only about ±25 degrees or about 12 per cent for an impedance change to the 70 per cent point. Use of higher-impedance coax for the balancing section would give a wider bandwidth.

It can also be seen that the half-wave balun will give better performance when used to couple a circuit to a higher impedance balanced load, such as a folded dipole, rather than to a lower impedance antenna, such as a Yagi antenna.

conclusions

I have discussed the performance of the half-wave coaxial balun. As often occurs in engineering problems, the performance characteristics of a device can have very subtle implications, even when the device itself is extremely simple. It has been shown that the best impedance characteristic is obtained when the impedance of the balancing line is equal to one-half the balanced load impedance. On the other hand, the best balance versus frequency characteristic is obtained when the balanced load impedance is high compared with the balancing line impedance. There is no requirement that the characteristic impedance of the transmission line feeding the **balun/load** be the same as that of the balancing half-wave section.

Whether or not the bandwidth limitations of the half-wave balun are the limiting factor in overall system performance depends on other factors. In particular, if the balanced load is an antenna, as it often is, its performance may deteriorate faster than that of the balun so that it may be the limiting factor.

Despite the bandwidth limitations described, I believe the half-wave balun will continue to be widely used because of its simplicity and economy.

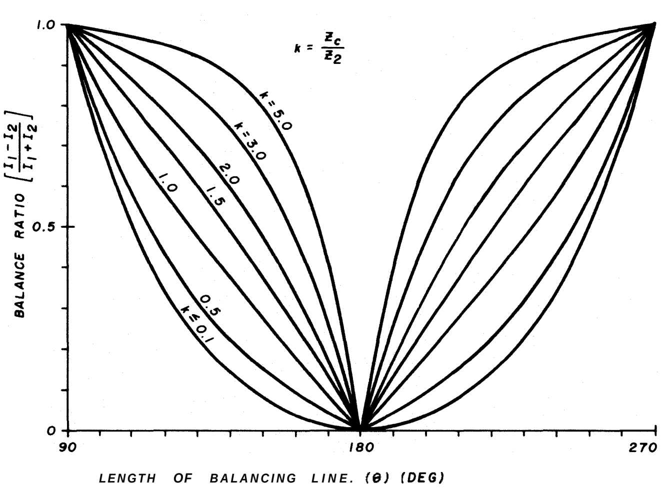


fig. 5. Balance ratio as a function of balancing-line electrical length. Perfect balance is obtained when the balance ratio is zero.

references

1. O.M. Woodward, Jr., "Balance Measurements on Balun Transformers," RCA *License Bulletin* LB-872, July 22, 1952. RCA Laboratories Division, Princeton, New Jersey.
2. O.M. Woodward, Jr., "Balance Measurements on Balun Transformers," *Electronics*, Vol. 26, No. 9, September, 1953, pages 188-191.
3. Terman, *Radio Engineering*, 3rd Edition, McGraw-Hill Book Company, Inc., 1947, figure 4-1, page 75, and Eq. (4-32a), page 92.

appendix

Because many of ham radio's readers are mathematically inclined, I'll present a brief derivation of the more important equations used in this article.

Beginning with the basic transmission line equation, **eq. 1,** divide the numerator and denominator of the right-hand side by Z_2; also divide both sides of the equation by Z_2. This gives:

$$\frac{Z_{in}}{Z_2} = \frac{Z_c}{Z_2}\left[\frac{\cos\theta + j\frac{Z_c}{Z_2}\sin\theta}{\frac{Z_c}{Z_2}\cos\theta + j\sin\theta}\right] \quad \textbf{(A-1)}$$

$$= k\left[\frac{\cos\theta + jk\sin\theta}{k\cos\theta + j\sin\theta}\right] \quad \textbf{(A-2)}$$

Since we defined $\frac{Z_c}{Z_2} = k$

Now rationalize **eq. A-2.**

$$\frac{Z_{in}}{Z_2} = k\left[\frac{\cos\theta + jk\sin\theta}{k\cos\theta + j\sin\theta}\right]\left[\frac{k\cos\theta - j\sin\theta}{k\cos\theta - j\sin\theta}\right] \quad \textbf{(A-3)}$$

Separate **eq. A-3** into real and imaginary parts:

$$\frac{Z_{in}}{Z_2}(real) = k\left[\frac{k\cos^2\theta + k\sin^2\theta}{(k\cos\theta)^2 + (\sin\theta)^2}\right]$$

$$-\frac{k^2}{(k\cos\theta)^2 + (\sin\theta)^2} \quad \textbf{(A-4)}$$

which is plotted in fig. 3.

Note that when $k = 1$, the matched case, $\frac{Z_{in}}{Z_2}$ (real) is independent of θ and equal to unity.

For the imaginary part,

$$\frac{Z_{in}}{Z_2}(imag.) = k\left[\frac{k^2\cos\theta\sin\theta - \cos\theta\sin\theta}{(k\cos\theta)^2 + (\sin\theta)^2}\right]$$

Factoring,

$$\frac{Z_{in}}{Z_2}(imag.) = k\left[\frac{(k^2-1)\cos\theta\sin\theta}{(k\cos\theta)^2 + (\sin\theta)^2}\right] \quad \textbf{(A-5)}$$

which is plotted in fig. 4.

In this case, when $k = 1$, the imag. part of $\frac{Z_{in}}{Z_2}$, which represents the reactive component, is zero regardless of θ.

To derive the balance equation, **eq. 6,** refer to fig. **A-1.** From fig. **A-1,** the well-known transmission line equation can be obtained:[3]

$$E = E_r\cos\theta + jI_r Z_0\sin\theta \quad \textbf{(A-6)}$$

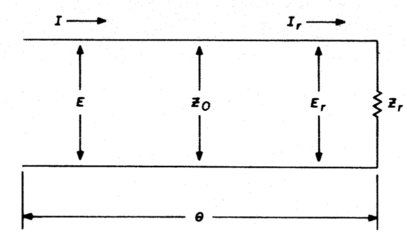
Eq. A-6 gives voltage at the input of the transmission line in terms of the load current and characteristic impedance of the line.

As $E_r = I_r Z_r$, **eq. A-6** can be rewritten as

$$E = I_r Z_r\cos\theta + jI_r Z_0\sin\theta \quad \textbf{(A-7)}$$

Solving for I_r,

$$I_r = \frac{E}{Z_r\cos\theta + jZ_0\sin\theta} \quad \textbf{(A-8)}$$



$E = E_r\cos\theta + jI_r Z_0\sin\theta$

fig. **A-1.** Basic transmission-line circuit.

Rewrite **eq. A-8** using the nomenclature of fig. 2:

$$I_2 = \frac{-e_0}{Z_2\cos\theta + jZ_c\sin\theta} \quad \textbf{(A-9)}$$

The minus sign is obtained since I_2 of fig. 2 flows in the opposite direction from I_r.

Since $k = \frac{Z_c}{Z_2}$,

$$I_2 = \frac{-e_0}{Z_2(\cos\theta + jk\sin\theta)}$$

which is **eq. 5** in text.

Current I_1 in fig. **2** is given by $I_1 = \frac{e_0}{Z_1}$. But since $Z_1 = Z_2$, we can write $I_1 = \frac{e_0}{Z_2}$. Hence the ratio $\left|\frac{I_1}{I_2}\right|$ is given by

$$|ratio| = \left|\frac{\frac{e_0}{Z_2} - \frac{-e_0}{Z_2(\cos\theta + jk\sin\theta)}}{\frac{e_0}{Z_2} + \frac{-e_0}{Z_2(\cos\theta + jk\sin\theta)}}\right|$$

$$|ratio| = \left|\frac{(\cos\theta + 1) + jk\sin\theta}{(\cos\theta - 1) + jk\sin\theta}\right| \quad \textbf{(A-10)}$$

Rationalize **eq. A-10** and obtain

$$|ratio| = \frac{[(\cos\theta + 1) + jk\sin\theta][(\cos\theta - 1) - jk\sin\theta]}{(\cos\theta - 1)^2 + (k\sin\theta)^2}$$

from which the real and imaginary parts are given by

$$real = \frac{\cos^2\theta - 1 + (k\sin\theta)^2}{(\cos\theta - 1)^2 + (k\sin\theta)^2}$$

$$imaginary = \frac{-2k\sin\theta}{(\cos\theta - 1)^2 + (k\sin\theta)^2}$$

The magnitude is given by the square root of the sum of the real and imaginary parts squared, or

$$magnitude = \frac{\sqrt{[\cos^2\theta - 1 + k^2\sin^2\theta]^2 + [-2k\sin\theta]^2}}{(\cos\theta - 1)^2 + (k\sin\theta)^2} \quad \textbf{(A-11)}$$

This equation is plotted in fig. 5. It can be seen from **eq. A-11** that when $k \leq 0.1$, the magnitude approaches

$$magnitude = \left|\frac{\cos\theta + 1}{\cos\theta - 1}\right|$$

and is independent of k.

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Yagi antenna design: quads and quagis

A discussion
of the
one wavelength
in circumference loop
including examination
of gain and loop shape

Up to this point in this series on Yagi antennas, we have considered only linear cylindrical elements about one-half wavelength long with a small diameter compared to wavelength; there is in common use, however, a radiating "element" consisting of a loop of wire about one wavelength in circumference. There are many such loop configurations: a triangular loop (commonly known as a delta loop), a square loop with two sides parallel to the earth (known as the quad), and a square loop oriented so that two diagonal corners are perpendicular to the earth (known as a diamond loop). Loops can also be made that are not equilateral. Triangles can be isosceles or even with three different sides; four-sided loops can be rectangles, and, indeed, loops can have more than four sides or can even be round.

The one-wavelength loop can be used either as a driven radiator or a parasite. To drive the loop it is opened at some point on its circumference where it is excited with a suitable voltage or current. As a parasite the loop is left closed; current will flow depending upon the induced voltages from other loops or elements and the self-impedance of the loop. To help understand the behavior of these loops it is useful to model the loop in terms of dipole elements.

Let's first consider a model for a square quad driven loop. The model starts with two driven half-wavelength dipoles in free space as shown in **fig. 1.** The dipoles are voltage driven at their centers (at xx and *yy*) and in this model are separated (vertically) by a quarter wavelength **($\lambda/4$).** If the excitation is equal at xx and yy (the same voltage and phase), the dipole currents will also be equal and, therefore, the voltages at the ends of the dipoles will be equal. The outer $\lambda/8$ sections of each dipole may now be bent as shown by the dotted line, forming a square. Since the voltages at dipole ends are still equal, the bent dipoles can be connected together without changing any currents in the square.

Since for each dipole there is a unique relationship between voltage at its ends to current at its center, the end connection of both dipoles in the square configuration insures that the center dipole currents will be equal even though excitation voltages at xx and yy may be different! Indeed, we can remove voltage excitation at yy (shorting the terminals) and still be sure that whatever dipole currents flow from excitation at xx, they will be equal in the two dipoles. One can look at the excitation in this case as a dipole ***current*** feed at xx; the dipole centered at yy is ***voltage*** fed at its ends. Note that to realize the same loop current the sum of the voltages supplied to xx and yy must be constant. This leads to the well-known fact that the driving point impedance at xx and yy shorted will be just twice the impedance at xx when yy and xx are excited equally.

Fig. 2 shows the square quad excited at the center of the bottom section and also shows the relative magnitudes and directions of the currents which flow. Note that the horizontal sections of the square show currents in the same direction as those of the original dipoles of **fig. 1;** therefore these sections will provide radiation fields at long distances; since they are shorter than the original half-wavelength dipoles,

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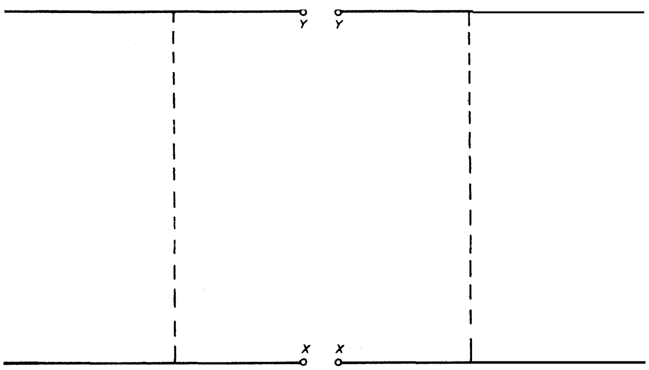

fig. 1. Diagram of two driven dipoles with $\lambda/4$ vertical spacing. Bending the ends of the dipoles together, as shown by the dotted lines, forms the square quad loop.

however, they require more current to produce a given radiation field, *i.e.*, the radiation resistance will drop. Moreover, the directivity and gain of each segment will be somewhat smaller than that of a full half-wavelength dipole. Both of these effects will be discussed shortly.

The vertical portions of the square quad shown in **fig. 2** are quite different. Note that for each of these vertical segments the top and bottom portions have identical currents but in opposite directions. This symmetry insures complete cancellation of the far (radiation) field. In other words, the vertical sections do not radiate; they simply act as a capacitive "top hat" loading (or tuning) for the horizontal radiating segments. Thus the radiating properties of the quad square of **fig. 2** will be identical to the (capacity loaded) resonant shortened or truncated horizontal segments alone. These will, of course, produce only horizontally polarized radiation.

Take another look at the two half-wavelength dipoles of **fig. 1.** The two (broadside) dipoles separated by a distance of $\lambda/4$ will produce a gain increase over one dipole. This separation or ***stacking*** gain can be easily calculated by using known self and mutual impedances; the result is shown in **table 1** and also in **fig.** 3 where the overall gain in **dBi** is plotted against the separation, S, in units of wavelength, A. Note that the separation gain improvement peaks when S is about $5\lambda/8$, leading to the often-quoted (but generally incorrect for other than single elements) "optimum stacking separation."

Fig. 3 can be understood qualitatively by remembering that the effective capture cross section area for a single half-wavelength dipole is about $0.13\lambda^2$. Thus, when the separation between the dipoles is large enough, the capture areas are basically independent (and thus additive), leading to an **effective gain improvement** of **a factor of** 2 or **3 dB**. For smaller separations, however, the capture areas overlap. In this overlap region both constructive and destructive interference can occur; for very small separations the combined capture areas reduce to that of **a** single dipole but for some intermediate regions (such as $S = 5\lambda/8$) constructive interference occurs, making the gain improvement more than a factor of 2. This constructive interference is due to a favorable reduction in the driving point resistance of the dipole, which is a direct result of the behavior of the real part of the mutual impedance of the two dipoles.

Another qualitative way of understanding this entire phenomenon is to view the (transmitting) dipoles as an excitation of a vertical aperture. Broadening this aperture by separating the dipoles is tantamount to narrowing the H-plane pattern, which will increase the gain. When the dipole separation becomes large enough, however, the quasi uniform illumination disappears and the vertical aperture acts like two independent (small) apertures giving rise to a diffraction pattern in the H-plane with maxima corresponding to a gain improvement of exactly a factor of 2. In any case, the actual calculated values of gain are shown in **table 1.** Note that for the quarter-wavelength separation of **fig. 1** the gain of the stacked dipoles is 3.236 **dBi**, or just **1.085** dB above that for a single half-wavelength dipole. This gain increase is all due to ***beam*** pattern narrowing in the H-plane; the E-plane pattern beamwidth remains the same as that for a single half-wavelength dipole.

The square quad loop of **fig. 2** is very similar to the stacked dipoles of **fig. 1,** but there are two significant changes. Because of the truncated or shortened elements, the driving-point impedances of the elements are reduced and the E-plane pattern width is somewhat increased, resulting in a somewhat reduced gain. To calculate these two results I will use the same method outlined by Kraus[1] (page 139-143).

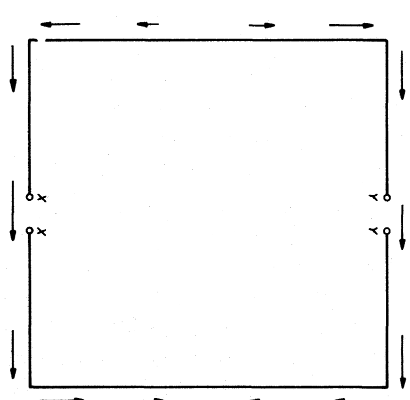

fig. 2. Outline of the square quad loop showing the current distribution in the horizontal and vertical members. Horizontal polarization results with the feedpoints as shown.

Kraus's calculation applies to a thin radiating element which is not capacitance loaded; to make the calculation apply to a single truncated element fed at the maximum current point, the limits of integration must be suitably altered; **fig. 4** shows the essential geometry.

The retarded value of the current at point z on the antenna referred to a point at distance, S, is

$$I = I_o \left[\cos \frac{2\pi}{\lambda} \cdot z\right] e^{jw(t - \frac{S}{c})} \tag{1}$$

where the quantity in brackets is the form factor for the current on the antenna. Following the Kraus development, the antenna can be viewed as a string of infinitesimal dipoles of length dz. The far fields, dE_θ and dH_ϕ, at a distance, S, from an infinitesimal dipole, dz are

$$dH_\phi = j\frac{I \sin\theta \, dz}{S\lambda} \tag{2}$$

$$dE_\theta = 120 \cdot \pi \cdot dH_\phi \tag{3}$$

where ϕ is the azimuth angle around the z axis. The total field from the entire antenna is then

$$H_\phi = \int_{-L/2}^{+L/2} dH_\phi \tag{4}$$

From eq. **1,** eq. **2,** and eq. **4,** **(5)**

$$H_\phi = \frac{j I_o \sin\theta \, e^{jwt}}{2\lambda} \int_{-L/2}^{+L/2} \frac{1}{S} \cos \frac{2\pi}{\lambda} \cdot z \cdot e^{-j\frac{wS}{c}} dz$$

Note that $S = r - z\cos\theta$ **(6)**
and also note that at long distances the amplitude of S is the same as the amplitude of r, so that we may write:

$$H_\phi = \frac{j I_o \sin\theta \cdot e^{j\omega(t - \frac{r}{c})}}{2\lambda r} \int_{-L/2}^{+L/2} \cos \frac{2\pi}{\lambda} \cdot z \cdot e^{j\frac{\omega\cos\theta}{c} \cdot z} dz \tag{7}$$

Let $\beta = \omega/c = 2\pi/\lambda$; eq. **7** may be rewritten:

$$H_\phi = \frac{j \beta I_o \sin\theta \cdot e^{j\omega(t - \frac{r}{c})}}{4\pi r} \int_{-L/2}^{+L/2} e^{j\beta z\cos\theta} \cos(\beta z) \, dz \tag{8}$$

Since $\int e^{ax} \cdot \cos(bx) \, dx = \dfrac{e^{ax}[a\cos(bx) + b\sin(bx)]}{a^2 + b^2}$

and if $a = j\beta\cos\theta$ and $b = \beta$, then eq. **8** becomes:

$$H_\phi = \frac{j\beta I_o \sin\theta \cdot e^{jolt - \frac{r}{c})}}{4\pi r} \tag{9}$$

$$\left| \frac{e^{j\cos\theta\cdot\beta z}}{\beta^2(\sin 2\, 0)} (j\beta\cos\theta \cdot \cos\beta z + \beta \cdot \sin\beta z) \right|$$

Evaluating this expression at both limits and collecting terms:

$$H_\phi = \frac{j[I_o]}{2\pi r} \cdot F(\theta)$$

where $[I_o] = I_o e^{j\omega(t - \frac{r}{c})}$

and $F(\theta) =$ **(10)**

$$\frac{\cos\left(\frac{\beta L}{2} \cdot \cos\theta\right) \sin\frac{\beta L}{2} - \cos\theta \cdot \sin\left(\frac{\beta L}{2} \cdot \cos\theta\right)\cos\frac{\beta L}{2}}{\sin 0}$$

$F(\theta)$ is often referred to as the field pattern factor. Thus the far fields of the truncated element can be written:

$$H_\phi = \frac{j[I_o]}{2\pi r} \cdot F(\theta) \tag{11}$$

and $E_\theta = \dfrac{j\,60\,[I_o]}{r} \cdot F(\theta)$ **(12)**

Note that if there is no truncation $(L = \frac{\lambda}{2})$ eq. **10** reduces to the well-known expression for a $\lambda/2$ dipole:

$$F(\theta) = \frac{\cos\left(\frac{\pi}{2} \cdot \cos 0\right)}{\sin\theta} \tag{13}$$

Kraus has shown that the self-radiation resistance, R_{11}, of such a linear element can be computed by equating the integral of the Poynting vector over a large sphere (total power radiated) to the driving point (current maximum) total power supplied. The result is (see Kraus,[1] 5-90, page 143):

$$R_{11} = 60\int_o^\pi F^2(\theta) \cdot \sin\theta \, d\theta \tag{14}$$

For the truncated element it is now a simple matter to insert the value of $F(\theta)$ from eq. **10** into eq. **14** and integrate. The integration is quite easily done numerically with a simple computer program. The result of such an integration is shown in table **2** where the radiation resistance of a truncated dipole element of length L (in terms of wavelength λ) is shown.

The directivity or gain can also be easily computed. Once $F(\theta)$ is given, the directivity is simply the ratio of the ***maximum*** value of $F^2(\theta)$ to the ***average*** value of $F^2(\theta)$ over the entire 4π solid angle. That is, the directivity $\boldsymbol{D}$ is:

$$D = \frac{F^2\left(\theta = \frac{\pi}{2}\right) \cdot 4\pi}{2\pi \int_o^\pi F^2(\theta) \sin\theta \, d\theta} \tag{15}$$

or from eq. **14**

$$D = \frac{120}{R_{11}} F^2\left(\theta = \frac{\pi}{2}\right) = \frac{120}{R_{11}} \cdot \sin^2\left(\frac{\beta L}{2}\right) \tag{16}$$

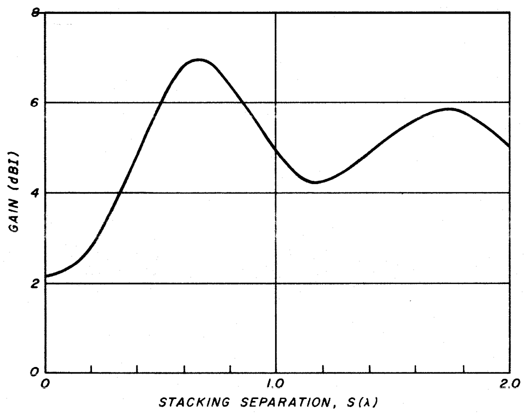

fig. 3. "Stacking" gain *vs* separation for two side-by-side dipoles.

The calculated directivity, D, and the related gain, expressed in dBi, are also listed in table 2.

Note that for small truncations where L is only slightly shorter than $\lambda/2$ there is not much reduction in self-radiation resistance and not much reduction in directivity; this is to be expected because the small ends of the dipole which are truncated carry little current, so do not contribute greatly to element performance. For heavy truncation, however, both self-radiation resistance and directivity decrease significantly; the limiting case where the length goes to zero is the well-known infinitesimal dipole whose directivity is just 1.500 and gain is 1.761 dBi.

To compute the driving-point resistance of the full quad square loop it is necessary to know the mutual resistance between two (truncated) elements separated by $\lambda/4$. From quite fundamental considerations Hurwitz[2] has shown a mathematical expression for the real part of the complex mutual impedance between two elements, R_{21}:

$$R_{21} = 60 \int_0^{\pi} F^2(\theta) \cdot \sin\theta \cdot J_o\,(\beta S \cdot \sin\theta)\, d\theta \qquad \textbf{(17)}$$

where $F(\theta)$ is the pattern function and J_o is the Bessel function whose argument is the product of element separation S and $\beta \cdot sin\ 8$. If S is measured in units of X the argument is simply $(2\pi S \cdot sin\ 8)$. Note that for very small separations J_o approaches unity; for very close separations $R_{21} \cong R_{11}$.

Shown in table 3 are values of mutual resistance vs separation (λ) of truncated elements. Note that the mutual resistance behaves very similarly to the values for full half-wavelength dipoles3 but the magnitudes are much smaller; a careful comparison shows that the reduction factor varies somewhat with separation.

The properties of the square quad loop can now be computed. Table 2 shows that the gain of a *single* (truncated) $\lambda/4$ element is 1.922 dBi and that R_{11} = 38.547 *ohms*. For the full quad loop of fig. 2 there are two "elements;" if they are both equally driven, they will produce the same far-field strength and power density as a single driven "element" carrying twice the current. However, the driving-point resistance of each one of the driven elements, when both elements are equally excited, is $R_{11} + R_{12}$. Therefore, total input power (both elements) is:

$$W_2 = 2 \cdot I^2 (R_{11} + R_{12}) \qquad \textbf{(18)}$$

While for the single element alone (same far field) is:

$$W_1 = (2I)^2 \cdot R_{11} \qquad \textbf{(19)}$$

If the directivity of a single element is designated as D_1 and the directivity of the full quad loop as D_2, then

$$D_2 = D_1 \frac{W_1}{W_2} = 2D_1/(1 + R_{12}/R_{11}) \qquad \textbf{(20)}$$

From tables 2 and 3 values for the square quad loop are: $R_1 = 38.547$ *ohms*; $R_2 = 21.729$ *ohms*; D_2/D_1 becomes 1.279 or 1.069 dB. Note that this ***stacking*** gain of the two truncated elements is nearly identical to the stacking gain of half-wavelength dipoles at a separation of $\lambda/4$ (1.085 dB, see table **1**); the difference is due only to the details of mutual resistance. The driving-point resistance, R, of the total loop, of course, is twice the value for a single element (when both are driven):

$$R = 2(R_{11} + R_{12}) = 120.6 \text{ ohms} \qquad \textbf{(21)}$$

Since D_1 (see table **2**) is 1.557,

$$D_2 = 1.991 \text{ and} \qquad \textbf{(22)}$$

$$Gain = 2.992 \text{ dBi} \qquad \textbf{(23)}$$

These properties of the square quad loop have been obtained rather rigorously; the main assumption in the model is total neglect of far-field radiation from the vertical sections, which are assumed to act as capacitance sinks for the current at the ends of the horizontal radiating segments. Moreover, the assumption is made that the current distribution on

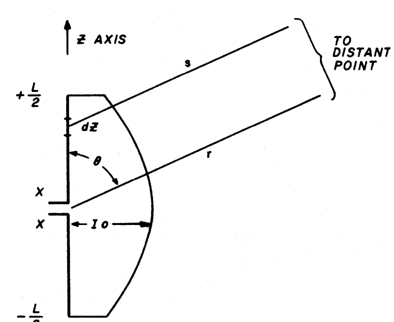

fig. 4. Current relationships for the symmetrical, thin center-fed truncated antenna of length L. This is the relationship necessary to determine driving-point impedance and H- and E-plane patterns of the square quad loop.

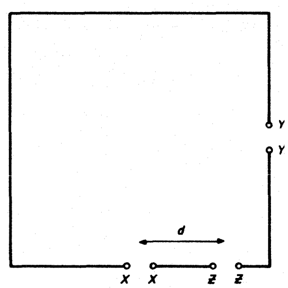

fig. 5. Example of a square loop with the excitation applied at one of three different places.

the horizontal segments is strictly sinusoidal; this is valid for very thin elements. Most quad loops are built with wire, which is thin compared to a wavelength, so one can be quite confident of the model.

It is now easy to understand qualitatively the radiation pattern of the loop. The H-plane profile does narrow, compared with a single linear element, because of the "illumination" of a wider vertical aperture; quantitatively, this profile narrowing is almost the same for truncated quad "elements" and the equivalently separated half-wavelength dipoles. The E-plane profile, however, is not as narrow for the truncated element as for a half-wavelength dipole; this factor accounts for the somewhat reduced directivity (see **table 2**). The loss in gain is about 0.23 dB for the $\lambda/4$ truncated element.

other driven loops

Before I consider a multiloop quad system I will briefly discuss other forms of driven loops. First, I will determine the performance of the square quad loop when it is driven or excited at a point other than the center of the bottom segment. **Fig. 5** shows the square loop with three different feed points: the center of the bottom leg xx, the center of the right vertical leg yy, and at an arbitrary point, *zz*, placed at a counter-clockwise distance *d* (from *xx*) around the loop. It has already been shown that if excitation is applied at xx (with yy and *zz* shorted), the loop will produce a totally horizontally polarized far field; its gain is 2.992 dBi. Similarly, if excitation is applied at yy (with xx and *zz* shorted), only vertically polarized far-field radiation will occur; gain is again just 2.992 dBi. Note that excitation at xx and yy are basically independent, i.e., unit current flow due to excitation at *xx* has a null current at yy and vice versa.

If we now excite the loop at *zz* with the same unit current, $I_O \cos wt$, it is easy to see that current flow at xx is just $I_O \cos(\beta d) \cos wt$ and at *yy* is $I,\ \sin(\beta d) \cos wt$. These currents will produce orthogonal far fields which must be added vectorially to obtain the total far field; the total far field has exactly the same magnitude as that which is produced by the same unit current at xx alone. In other words, the square loop gain and drive-point resistance are totally independent of the feed point; only the polarization changes (from totally horizontal if the feed is at xx to totally vertical if the feed is at *yy*). This simple theorem can be easily proved by the same type of argument for ***any*** equilateral one-wavelength loop.

It is now interesting to consider moving the feedpoint *zz* to the lower right-hand corner of the square loop. Since we are considering free space loop gain for which rotation is unimportant, it is clear that this configuration produces ***exactly*** the same result as the familiar square ***diamond*** loop fed at the bottom or top corner. Thus we now know that the quad square and the diamond square have ***exactly*** the same gain and ***exactly*** the same drive-point resistance. Similarly, the gain and drive-point resistance of ***any*** equilateral (one-wavelength) loop is ***totally*** independent of the position of the drive point on the loop.

I shall now return to the horizontally polarized square of **fig. 2.** We have shown that a square loop (relative to a half-wave dipole) has a somewhat enhanced gain (+0.84 dB), made up of an increase (+1.07 dB) due to the vertical separation of the radiating segments (*H*-plane narrowed somewhat) and a decrease (−0.23 dB) due to the shortened or truncated radiating segments (*E*-plane broadening). Let's now explore the performance of a ***rectangular*** one-wavelength loop; **fig. 6** shows some examples.

Rectangle **A** is a wide but low loop which is recognized as a folded dipole loop; B is a narrower and higher loop than the square, and **C** is a high but very narrow loop. For all of these loops the sum of the width, W, and height, H, measured in wavelengths is constrained to be just 0.5. We are now in a position

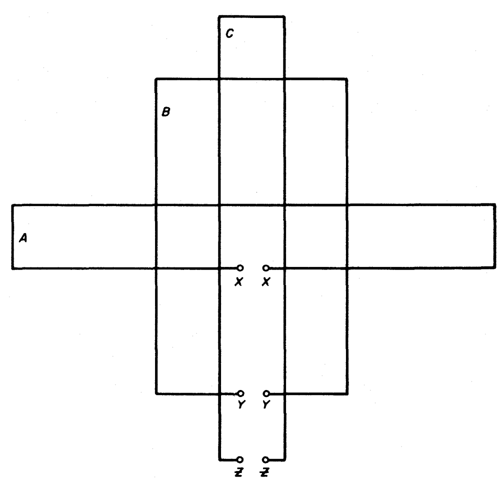

fig. 6. Diagram of three different rectangular configurations. In each case the feed point is located for horizontally polarized radiation.

table 1. Stacked half-wavelength dipoles showing gain vs. dipole separation.

separation S(λ)	gain (dBi)	stacking gain (dB)
0.000	2.151	0.000
0.125	2.425	0.274
0.250	3.236	1.085
0.375	4.514	2.363
0.500	5.978	3.827
0.625	6.938	4.787
0.750	6.758	4.607
0.875	5.831	3.680
1.000	4.929	2.778
1.250	4.373	2.222
1.500	5.275	3.124
1.750	5.844	3.693
2.000	5.097	2.946

to compute loop performance. Table 4 shows the results of calculations for rectangular loops by the same methods as used for the square. D_2/D_1 is the H-plane directivity increase due to the vertical separation of radiating segments (see eq. 20), while D_1 is the directivity of a single horizontal radiating segment (see table 1). R is the loop driving point resistance (see eq. 22). Note that the limiting case of the folded dipole (W = 0.5, H = 0) shows a gain of 2.151 dBi (identical to a single half-wave dipole) and a driving-point resistance of 292.5 ohms (just four times that of a single half-wave dipole). Since reactance effects of the full loop are double that of a single "element," the Q of the folded dipole will be just one-half the Q of a single half-wave dipole, leading automatically to a bandwidth twice as large. As one reduces the loop width from this limiting case the gain increases monotonically. However, this favorable increase in gain is automatically accompanied by a significant reduction in driving point resistance; since reactance effects are essentially identical for all loops, the circuit Q increases commensurately. Thus the potential gain obtainable from high narrow loops is always offset by unfavorably high values of Q and corresponding narrow bandwidths!

table 2. Self-resistance and gain for one truncated element.

element length (λ)	R_{11} (ohms)	directivity	gain dBi
0.05	1.955	1.502	1.768
0.10	7.590	1.510	1.789
0.15	16.255	1.522	1.823
0.20	26.966	1.537	1.868
0.25	38.547	1.557	1.922
0.30	49.780	1.578	1.980
0.35	59.562	1.599	2.040
0.40	67.023	1.619	2.094
0.45	71.612	1.635	2.134
0.50	73.130	1.641	2.151

Let us now consider other loop shapes. For equilateral shapes, we have already seen that gain and driving-point resistance are independent of feed-point resistance on the loop. Moreover, a reasonably rigorous solution has been obtained for a square loop. Because of the independence of properties on feed-point position, which is equivalent to independence on loop rotation with feed point fixed at the bottom, it seems reasonable to assume that gain from all equilateral loops is approximately equal to that of a circular loop having the same enclosed area. This in turn can be equated to an equivalent square of the same area. Such an intuitive model, iogether with a model of how E-plane directivity varies with element length (see table 2) and how H-plane directivity varies with element separation (see table 1), allows a reasonable estimate of equilateral loop gain. Table 5 lists these values for several equilateral loops. The values for the square are the ones already computed (see table 4)l; all others are estimated by

table 3. Mutual resistance, $R_{1,}$, for λ/4 truncated elements vs. separation in λ.

separation (λ)	mutual resistance (ohms)	separation (λ)	mutual resistance (ohms)
0.00	38.547	0.80	−9.783
0.05	37.782	0.85	−7.165
0.10	35.536	0.90	−4.194
0.15	31.951	0.95	−1.142
0.20	27.251	1.00	1.730
0.25	21.729	1.10	6.083
0.30	15.721	1.20	7.713
0.35	9.589	1.30	6.481
0.40	3.686	1.40	3.190
0.45	−1.659	1.50	−0.787
0.50	−6.170	1.60	−4.008
0.55	−9.636	1.70	−5.450
0.60	−11.930	1.80	−4.806
0.65	−13.010	1.90	−2.518
0.70	−12.919	2.00	0.446
0.75	−11.780		

this simple model. It is quite easy to see that the popular triangle or delta loop is slightly lower in gain (by 0.3 dB) than the square (quad or diamond) and that the loop with the highest gain is a circle. However, note that the gain of the circular loop is estimated as 3.28 dBi which is 1.13 dB larger than that of a half-wave dipole. This is not quite as large an increase as the approximately 2 dB which was quoted by Lindsay,[4] but I believe this discrepancy is probably not outside of the experimental accuracy range of Lindsay's measurements combined with the estimation accuracy range of the simple model I have used.

I shall now consider a multiloop square quad array where not only driven elements but parasitic loops as well are used. From what we have just seen, the table 4. Properties of the rectangular loop.

width $w(\lambda)$	height $H(\lambda)$	D_1	D_2/D_1	D_2	gain dBi	R ohms
0.05	0.45	1.502	2.073	3.114	4.933	3.77
0.1	0.4	1.510	1.815	2.741	4.379	16.7
0.15	0.35	1.522	1.596	2.430	3.856	40.8
0.2	0.3	1.537	1.419	2.181	3.386	76.0
0.25	0.25	1.557	1.279	1.991	2.990	120.5
0.3	0.2	1.578	1.172	1.850	2.672	169.8
0.35	0.15	1.599	1.094	1.750	2.430	217.7
0.4	0.1	1.619	1.041	1.686	2.267	257.5
0.45	0.05	1.635	1.010	1.652	2.179	283.6
0.5	0.0	1.641	1.000	1.641	2.151	292.5

quad array should behave just about like an equivalent stacked Yagi array separated (vertically) by 0.25λ and having elements bent together to form the individual square quad loops. However, to proceed with a computation of the properties of the parasitic quad array, we must know all complex self and mutual impedances of the truncated elements. Up to this point it has been possible to rather rigorously describe the properties of a driven loop because only the real part of self and mutual impedances were required to obtain gain, driving-point resistance, and pattern information (both in the E- and H-planes).

When parasitic elements are involved, the imaginary impedance terms are required. Computation of the imaginary impedances is a non-trivial exercise, and I am unaware of any rigorous procedure for carrying out such a calculation for truncated elements. As far as self-impedance is concerned, the reactive or imaginary value is controlled entirely by the "tuning" of the loop; that is, the relationship of wavelength to loop circumference and the effective loop Q. The complex impedance has been calculated for a linear nearly half-wave thin cylindrical element.[5] The method involves treating the metallic cylinder as a boundary-value problem (tangential components of electric field are made to vanish at every point on the conductor surface), from which an integral equation is derived. Approximate solutions of this integral equation yield the current distribution on the cylindrical element from which the input impedance, including the imaginary component, was derived.

I have been unable to find an equivalent rigorous boundary-value calculation for the square quad loop; thus we do not yet have the basis for calculating the precise reactance for a nearly one-wavelength square loop. However, it is possible to at least estimate the loop Q (but not its precise resonant frequency) by remembering that reactance changes with frequency are due to the effective inductance and capacitance of the antenna; that is, near-field stored energies, whereas the resistance (radiation) has to do with far field. Truncating the half-wave dipole changes the geometry of the element but hardly affects its (central) inductance or (end) capacitance. It would be reasonable, therefore, to assume the quad loop contains essentially the same total reactive impedance changes as two half-wave dipoles. Thus, following the argument made in a previous article,[3] we may write for the loop self-impedance:

$$Z_{11} = R_{11} + jX_{11} = R_{11}\,[1 + j \cdot 2Q(F/FR - 1)] \qquad (24)$$

but (empirically)

$$R_{11}Q = (430.3 \log_{10} K - 320) \qquad \textbf{(25)}$$

so that

$$Z_{11} = 120.5 + j\,(860.6 \log_{10} K - 640)\,(F/FR - 1)\ ohms \quad ($$

As in the previous article, F is the normalized frequency and FR the normalized resonant frequency of the loop. Loop Q is readily estimated from **eqs.** 25 and **21.** The only remaining problem is a determination of FR. Although there is no rigorous way of calculating FR from basic principles, it is significant to note that the region where the ends of the two dipoles are joined must have electric (capacitive) fields at right angles to the conducting cylindrical element, exactly like those of an infinitely long straight cylinder near a voltage loop. This observation implies that there should be a negligible "end effect" at the capacitive voltage loop, and, therefore, FR should be very close to the frequency at which the total loop circumference is just one wavelength.

When we consider the imaginary part of the mutual impedance between loop halves or "elements," another computational complication arises. At long distances or separations, the imaginary mutual reactance, X_{21}, must be (except for a phase shift) simply related to the real part, R_{21}, and this relationship should be unaffected by the precise "tuning" of the "elements." However, at very small separations X_{21} must approximate the value of X_{11}, which is fundamentally fixed by circuit Q and resonant frequency and not at all by R_{11}. How to correctly represent X_{21} at all intermediate spacings has not, to my knowledge, been solved quantitatively. For this reason I will model the quad array first as the equivalent Yagi array stacked at a spacing of $\lambda/4$ then, second, apply necessary corrections to direc-

table 5. Estimated equilateral loop properties.

equilateral loop type	sides	equivalent square side (λ)	D_1	D_2/D_1	D_2	gain dBi	oh
triangle	3	0.219	1.545	1.205	1.862	2.70	1(
square	4	0.250	1.557	1.279	1.991	2.99	12
pentagon	5	0.262	1.562	1.309	2.044	3.10	12
hexagon	6	0.269	1.565	1.324	2.071	3.16	12
octagon	8	0.275	1.567	1.338	2.097	3.22	13
circle	∞	0.282	1.570	1.356	2.129	3.28	13

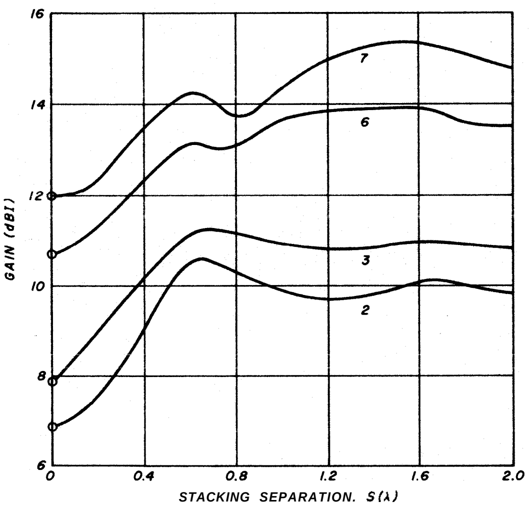

fig. 7. Representative examples of stacking gain for 2, 3, 6, and 7-element Yagis *vs* the stacking separation.

tivity and gain. Compared with stacked ($\lambda/2$ element) Yagis, the square quad will have 0.23 dB ***less*** directivity and gain (all due to E-plane pattern broadening).

It is interesting to calculate the free-space "stacking gain" of Yagis as a function of vertical separation. I chose as representative boom lengths and number of elements the following simplistic Yagis, which were found to have excellent free-space properties:[6] a two-element beam on a 0.15-X boom, a three-element beam on a 0.25-X boom, a six-element beam on a 0.75-X boom, and a seven-element beam on a 1.25-X boom. **Table** 6 lists characteristics of these beams.

Fig. 7 shows the computed gains of these stacked configurations as a function of the stacking separation, S, in units of X. Note that the rise in gain due to S is somewhat different for each case and also somewhat different than the case for dipoles shown in **fig.** 3. Two things seem to be occurring as S is increased. For small separations, the capture area of one Yagi essentially overlaps that of the other Yagi; therefore, the total capture area for both is essentially the same as for one alone. As S increases, the total capture area increases and ultimately doubles if S is large enough. For the larger Yagis (where the original capture area for one Yagi is large), it is easy to see that to realize a given separation, gain S must be relatively larger than for a small Yagi or especially a dipole. In other words, the transition from the gain of one large Yagi to the doubling of gain (3-dB increase) for two large Yagis requires a ***larger*** separation than is required for smaller Yagis. In addition to this rather gradual gain increase due to separation of capture areas, **fig.** 7 suggests that the constructive interference due to mutual impedances noticed for the dipoles in **fig.** 3 also persists in stacked Yagis. An increase in gain is noticeable a? S = 0.6 and alsc at 1.6. Qualitatively, **fig.** 7 shows a combination of these two effects; first, the constructive impedance gain increases at particular separations and second, the capture area separation gain increase which takes place more slowly with large Yagis.

It is interesting to compare the computed gain increase of the seven-element beam with the experimental values reported by **Viezbicke**[7] in his **fig. 11A. Table 7** shows a comparison of my computed results with Viezbicke's published experimental values. The comparison is not totally valid because Viezbicke's seven-element beam is not the same beam I have used in **fig.** 7. The differences cannot be quantified, since Viezbicke failed to include specifications for his seven-element beam. Nevertheless, **table** 7 shows good qualitative agreement and even fair quantitative agreement (close to Viezbicke's stated experimental accuracy of 0.5 **dB).**

table 6. Characteristics of representative beams. Radius of all elements = 0.0005260λ.

number elements	length (λ) reel	driven	dir.	boom length (λ)	element spacing (λ)	fig. 7 curve
2	.4937	.4705	–	0.15	0.150	2
3	.4980	.4896	.4690	0.25	0.125	3
6	.4953	.4803	.4481	0.75	0.150	6
7	.4936	.4762	.4466	1.25	0.208	7

table 7. Stacking gain (dB) of seven-element beams.

spacing S (λ)	(Viezbicke) 7 element	computed 7 element
0.38	0.80	1.3
0.57	1.58	2.15
0.78	1.36	1.8
0.99	1.90	2.35
1.20	2.34	3.0
1.40	2.53	3.25
1.61	1.93	3.3

Computations represented by the data shown in **fig.** 7 cannot really be carried down to very small separations with any confidence because the mathematical model uses mutual impedances of full $\lambda/2$ elements. When reactive parasites get very close together their mutual impedance has an imaginary value quite different from that of $\lambda/2$ dipoles.

I shall now examine the stacking gain increases for a separation of $\lambda/4$. **Table** 8 shows these increases for the four computed cases (dipole data from **fig. 3**). Note that the $\lambda/4$ stacking gain for very large Yagis is not as large as that for dipoles;

this can easily be understood because of the (relatively) smaller increase in capture area. One might expect this gain increase to fall monotonically from the value of **1.09** dB (for dipoles) as the array length is increased, however, **table** 8 shows the actual increase to vary somewhat due to the detailed way in which impedances vary.

We are now in a position to estimate the performance of a quad array. If instead of the dipole elements of the stacked **($\lambda/4$)** Yagi arrays I use (bent) elements connected in square loops, **I** will make a quad array in which all conditions remain about the same ***except*** that the E-plane pattern is broadened and the gain is correspondingly reduced due to the truncated "elements." The gain performance of the quad array is therefore about 0.23 dB lower than the stacked $\lambda/4$ Yagis. These estimations are also shown in **table** 8.

table 8. Quarter wavelength stacking gain for Yagis and estimated quad/single Yagi gain.

number elements	boom length (λ)	Yagi $S = \lambda/4$ stacking gain (dB)	quad/single Yagi gain difference (dB)
1	0.0	1.09	0.86
2	0.15	1.03	0.80
3	0.25	1.38	1.15
6	0.75	0.84	0.61
7	1.25	0.65	0.42

It must be evident by now that a square quad array is very much like an equivalent Yagi. Its overall gain is expected to be somewhat higher than the Yagi because of individual loop gain, but not by very much. E-plane pattern is slightly broader due to the current distribution on the truncated "elements," whereas H-plane pattern is slightly narrower due to the "stacking gain" of the loops. Because of this similarity of Yagi and quad arrays, it is plausible that one can intermix quad loops and Yagi elements to provide a hybrid structure of roughly equal performance. Such a hybrid is known as a "quagi;" if properly constructed, it should provide a pattern and gain intermediate between a similar quad and a similar Yagi. There are obviously an enormous variety of possible quagi configurations; there will remain for some time a challenge to the quagi designer to determine preferred configurations and best dimensions for all radiating elements.

summary

A number of interesting conclusions have been reached regarding antenna arrays constructed with conducting loops roughly one wavelength in circumference.

1. A single (driven) loop will provide a free-space gain somewhat larger than that of a half-wave dipole. The gain increase comes about through a narrower H-plane pattern and a slightly broader E-plane pattern.

2. Loop gain varies significantly with shape. For rectangular loops fed in the center of the lower horizontal segment, gain depends on the ratio of height, H to width, W, varying from **2.15 dBi** (equal to a half-wave dipole) if $H/W = 0$ to about **5 dBi** (3.8 dB above a half-wave dipole) as H/W approaches zero.

3. For equilateral loops, gain depends only on the number of sides (see **table** 5). For the square loop, the free-space gain calculated rather rigorously is **2.99 dBi** and driving-point resistance is **120** ohms. **It** is significant that these properties of the square loop are totally independent of the feed point on the square. As an example, the gain and drive-point resistance for a quad square is ***exactly*** the same as that of the diamond square.

4. The free-space gain of a quad array is estimated to be somewhat higher than that of an equivalent (single) Yagi. Calculations show this difference depends somewhat on the particular array (see **table 8**), but ranges from about 1 dB for short arrays to less than 0.5 dB for long arrays.

5. Quagi configurations are expected to show performance figures between those of an equivalent Yagi and equivalent quad.

6. A rigorous theory does not yet exist for self and mutual quad loop reactances. Consequently, quad (and quagi) parasitic loops must be experimentally adjusted for correct resonant frequency. **It** is unlikely that such experimental adjustments can be made with the same precision and confidence that Yagi element lengths can now be specified by present theory. Therefore, the slight gain advantage of the quad over the Yagi shown in **table** 8 may well disappear in practice.

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ham radio

navigational aid for small-boat operators

An idea for approaching harbor entrances using two simple beacon transmitters and a harmonic phase detector

Many small-boat operators have probably wished for a simple navigational aid during fog, darkness, or other hazards. The system suggested here provides guidance between two simple low-powered transmitters, which may be located on opposite sides of a channel, breakwater entrance, or even on a wharf that may be the goal of the boat. It's a sort of SHORAN system, with simplifications, which should put it well within the technical ability, as well as the pocketbook, of the average Amateur.

theory

Consider **fig. 1,** which shows points **A** and **B,** representing the location of two transmitters, which may be battery powered and in the milliwatt range. Our boat is located at point ***C,*** and our aim is to guide the boat between points **A** and **B** using signals transmitted from these two points. Transmitters at **A** and **B** operate on closely chosen radio frequencies, which differ by some arbitrary audio frequency, perhaps several hundred to one thousand Hz. The frequency separation is well within the passband of a receiver located on the boat at point ***C.*** Receiver output, therefore, is the beat frequency caused by this difference in the transmitted frequencies.

If now, the boat maneuvers so that the path length between it and the higher-frequency transmitter becomes relatively shorter than the path to the lower-frequency transmitter, the phase of the received beat frequency signal will advance. This effect may be generalized by stating that the received beat frequency is phase-variant with the position of point ***C,*** the point of reception.

analysis

A number of paths exist along which the receiving point, ***C,*** can move without changing the phase of the beat note. These paths are a family of hyperbolas, ***all of which pass between* A *and* B.** On the line **AB,** they will be a half-wavelength apart, but as they appear further from line **AB,** they separate. This, in effect will "funnel" the boat between the two target points. The number of these paths will be a function of the separation between **A** and **B** in wavelengths **(fig. 2).**

To use this phase-variant characteristic of the beat frequency, a reference signal is needed with which the beat frequency can be compared. The generation and transmission of this signal is a relatively simple process, for which two alternatives exist.

reference signal

In the antenna feedline of either transmitter, we can insert a directional coupler. While signals from the local transmitter are being radiated from this antenna, signals from the other transmitter will be

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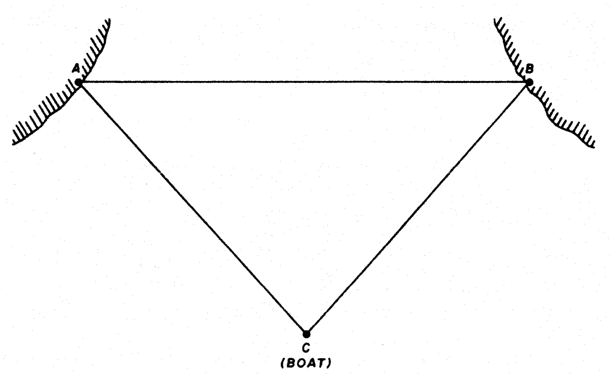

fig. 1. Example of a typical navigational problem. Beacon transmitters are at points A and B at a harbor entrance. Guidance may be provided to a small boat entering the channel using the idea in the text.

picked up. The directivity of this coupler is used to reduce the local signal, while accepting as much as possible of the received signal from the other transmitter.

When detected, a beat note is produced, which is the same frequency as that received at point **C,** but which is completely independent of the position of point **C.** In other words, here is our reference signal.

It could be transmitted to point **C** by using it to modulate a transmitter operating on a third frequency, but this would require a considerable amount of additional equipment both on shore as well as on the boat. Instead, we can double the beat frequency thus produced and use it to amplitude-modulate that transmitter, to a modest percentage, probably below 50 per cent, to avoid distortion. Appropriate filtering built into the modulator system will remove the fundamental.

The second alternative would be to divide the beat frequency by two, and with appropriate filtering, as before, use it to modulate that transmitter. Frequency division by digital methods should produce a symmetrical square wave, which would contain, theoretically, no trace of its second harmonic, which of course is the beat-note fundamental.

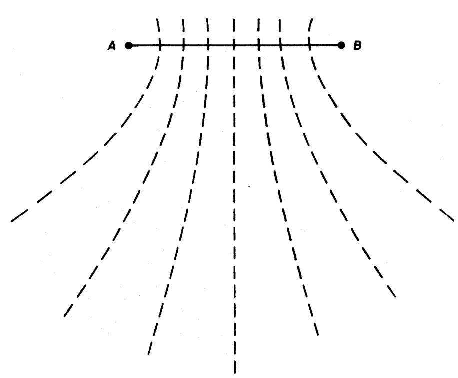

fig. 2. Geometry of equi-phase paths between points A and B of fig. 1. In this example AB = 4 wavelengths. Paths above AB line are the mirror image of those below.

This square wave would have to be reduced to a sine wave by some means, such as a low-pass filter to remove harmonics. A square wave contains only odd harmonics, so this doesn't appear too difficult for several stages of **RC** feedback filtering with a multiple op-amp.

phase detector

Now let us move to the boat, at point **C.** Its receiver will be producing two audio frequencies, having a **2:1** frequency relationship, one of which is phase-variant with the boat's position. This audio-frequency signal is now fed into a harmonic phase detector, **fig.** 3, which responds to two such harmonically related signals, giving **1)** a balanced zero voltage when both signals cross the zero axis simultaneously, and **2)** a positive or negative voltage, according to which signal crosses the zero axis before the other (reference 1).

The differential voltage produced by the phase

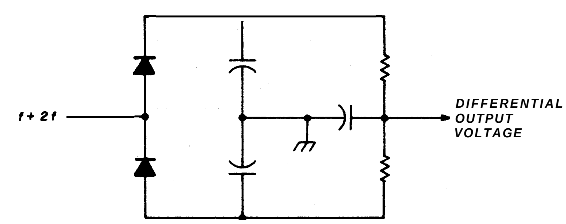

fig. 3. Harmonic phase detector produces a differential voltage that can be used to drive a zero-center meter that will deviate from center in accordance with the boat's deviation from one of the equi-phase paths in fig. 2.

detector is easily amplified to drive a zero-center meter, which will deviate from center in accordance with the boat's deviation from one of the equi-phase paths.

some final thoughts

I suspect that such a system might operate, for example, in one of the 20-kHz slots that appear in the CB spectrum. This would reduce the acqusition of new equipment by the average boat enthusiast to a bare minimum. If he stayed too late at the fishing grounds, he could call the operator of the marina where he keeps his craft and ask him to set the beacons out by the entrance through the breakwater. The marina operator should enhance his popularity by doing so, not only with his own customers, but with the boating fraternity in general.

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ham radio

optimum pi-network design

Methods for optimizing bandwidth without using the Qfactor — design parameters and examples are given for this useful circuit

The familiar pi network shown in **fig.** 1 is both a lowpass filter and a simple impedance-matching network. Stray reactances at source or load may be compensated for by capacitor adjustment. Traditional design methods have used a Q factor to choose correct component values. This article presents design methods for optimizing bandwidth without using the Qfactor.

The main function of the pi network is to provide a resistive match of source resistance, $R1$, with a load resistance, $R2$. An infinite number of component possibilities are available; two examples are shown in **fig.** 2.

This article was rewritten by Leonard H. Anderson, who is a member of the technical staff of Rocketdyne division of Rockwell International, Inc., and is well known for his contributions to ***ham radio*** in the ***Digital Techniques*** series. The ***ham radio*** staff expresses its thanks to Mr. Anderson for his help in interpreting this difficult subject.

impedance transformation diagrams

Fig. 2 is useful for any simple network. Each resistance circle represents the series-form impedance of the parallel combination of end resistance with shunt reactance. This chart is used in the following manner:

1. Calculate series form $R+jX$ of one end, including the shunt reactance.

2. Find the intersection on the resistance circle for both R and X.

3. Move vertically from this intersection to intercept the other resistance circle.

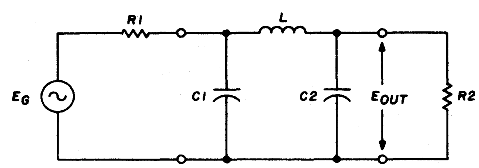

fig. 1. Classic pi network. Circuit is both a lowpass filter and impedance-matching device. Its main function is to provide a resistive match of source resistance, $R1$, with a load resistance, $R2$.

4. Measure the length of the vertical movement; this total reactance is the series-arm value.

5. Determine the reactance at the second intersection as measured from the zero reactance point (R axis passing through reactance axis).

6. Take the opposite reactance sign and calculate the new combination of parallel resistance and reactance.

The pi network first intersections are shown for the series impedance of $R1$ and $C1$. Total vertical movement is the reactance of series inductor L. The

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second intersections on the $R2$ circle represent the impedance presented to the inductor and $R1$, $C1$. To obtain a resistive match, the second intersections must be changed from a positive reactance to negative; this *conjugate* operation moves the Z_2 intersection to the R axis.

As long as each set of intersections falls on each resistance circle, any component combination will yield a perfect impedance match. One particular combination will give the widest bandwidth, and some network synthesis techniques can be used to find that combination.

image impedances

The image impedance of a network is the square root of the product of one input impedance with the other end open and the same input impedance with

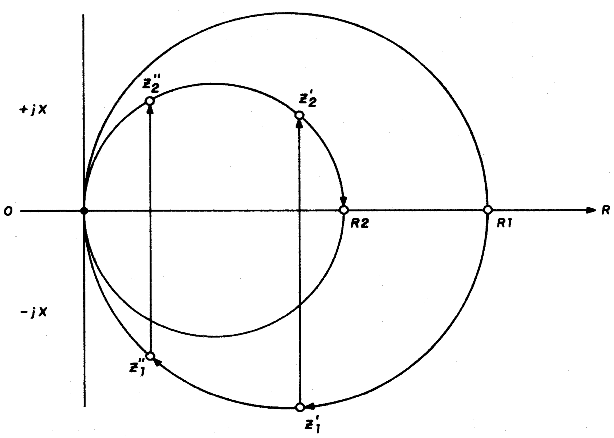

fig. 2. Impedance-transformation diagram. Each resistance circle represents the series-form impedance of the parallel combination of end resistance with shunt reactance.

the other end shorted. An image impedance is not the parameter for direct design but a mathematical tool for achieving the final solution.

The open/shorted opposite-end technique is used in microwave measurements where it's difficult to obtain a pure resistance.' At one particular frequency the image impedance will represent the actual input impedance when the opposite end is loaded with a resistance. An example is an infinite length of transmission line: at any frequency the image impedance is equal to the *characteristic impedance* of the line.

Fig. 3 shows the pi-equivalent circuit composed of pure reactance and susceptances. Using a matrix representation of the network, the following is true:

$$A_{11} = 1 + Z_2Y_3 \qquad A_{12} = Z_2$$
$$A_{21} = Y_1 + Y_3 + Y_1Z_2Y_3 \qquad A_{22} = 1 + Y_1Z_2$$

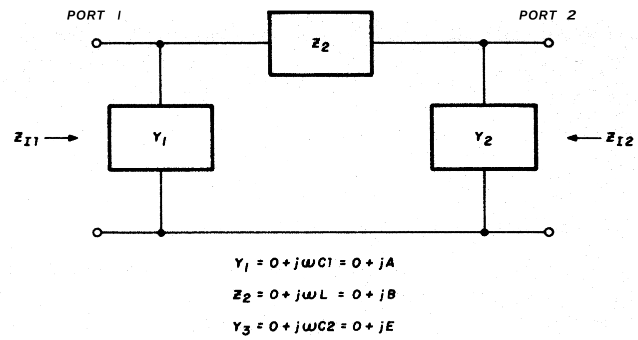

fig. 3. Pi-equivalent circuit composed of pure reactance and susceptances. These parameters at any frequency are derived in the text.

Using these matrix values, image impedance Z_{I1} becomes:

$$Z_{I1}^2 = Z_{o1}Z_{s1} = \frac{A_{11}A_{12}}{A_{21}A_{22}}$$

$$= \frac{B(BE-1)}{(AB-1)(A+E-ABE)} \qquad \textbf{(1)}$$

Letters A, B, and E are the susceptances and reactances at any frequency as given with **fig.** 3. Z_{o1} is the impedance into port 1 with port 2 open; Z_{s1} is the port-1 impedance with port 2 shorted. The image impedance at port 2 is:

$$Z_{I2}^2 = Z_{o2}Z_{s2} = \frac{A_{12}A_{22}}{A_{11}A_{21}} = \frac{B(AB-1)}{(BE-1)(A+E-ABE)} \qquad \textbf{(2)}$$

Z_{o2} and Z_{s2} are the port-2 impedances with port 1 open and shorted respectively. Both right-hand expressions are complex numbers with a zero-value imaginary component.

A broad spectrum plot of both image impedances is shown in **fig. 4** for a typical network. **It** is common to denote frequency in such plots as ω, the product of 2π X ***frequency***. The dash lines indicate a negative-real-part image impedance.

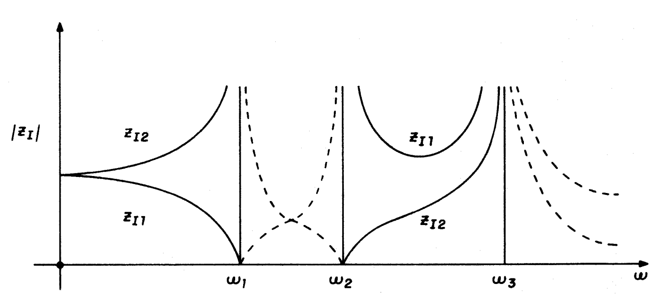

fig. 4. Image impedances at input and output ports as functions of frequency. Frequency is denoted as $\omega = 2\pi f$. Dashed lines indicate a negative-real-part image irnpedance.

Synthesis of component values is not possible between ω_1 and ω_2 nor above ω_3 because of the negative real parts of the impedance. The only choice is below ω_1 or between ω_2 and ω_3.

center frequency impedance transformation

Assuming that $R1$ is greater than $R2$ will exclude any frequency condition below ω_1. The range between ω_2 and ω_3 is left and shown expanded in **fig. 5.**

Frequency ω_0 is the geometric mean of ω_2 and ω_3, expressed as:

$$\omega_0 = \sqrt{\omega_2 \omega_3} \qquad \textbf{(3)}$$

The intersection of ω_0 and each image impedance plot line will give $R1 = Z_{I1}$ and $R2 = Z_{I2}$. Frequency ω_0 is the network center frequency.

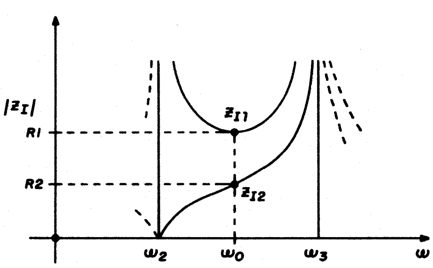

fig. 5. Impedance-transformation relationship to geometric center frequency, ω_0, which is the geometric mean of ω_2 and ω_3. Note that Z_{I1} has the least change at ω_0.

Holding **to** the network center frequency, ω_0, the susceptances and reactances of each pi-network arm are defined as:

$$A_0 = \omega_0 C1 \qquad \textbf{(4)}$$

$$B_0 = \omega_0 L \qquad \textbf{(5)}$$

$$E_0 = \omega_0 C2 \qquad \textbf{(6)}$$

Each end resistance can then be related to all component values by substituting **eqs. 4, 5,** and **6** into **eqs. 1** and **2:**

$$R1^2 = \frac{B_0(B_0E_0 - 1)}{(A_0B_0 - 1)(A_0 + E_0 - A_0B_0E_0)} \qquad \textbf{(7)}$$

$$R2^2 = \frac{B_0(A_0E_0 - 1)}{(B_0E_0 - 1)(A_0 + E_0 - A_0B_0E_0)} \qquad \textbf{(8)}$$

Defining a ratio, m, as $R1$ divided by $R2$, an identity is obtained from **eqs.** 7 and **8:**

$$m = \frac{R1}{R2} = \frac{B_0E_0 - 1}{A_0B_0 - 1} \qquad \textbf{(9)}$$

The relationship of ω_2 and ω_3 frequencies must now be established.

optimizing the design-center frequency

Examination of **fig. 5** shows that Z_{I1} has the least change of value at ω_0. This can be proved by taking the derivative of Z_{I1} as a function of frequency with **eq.** 1. Z_{I2} has a relatively constant slope at ω_0. Choosing ω_0 as the design center frequency results in a minimum impedance change at each network port over a given bandwidth.

To find ω_0 from **eq. 3,** cut-off frequencies ω_2 and ω_3 must be found. Image impedance plots in **figs. 4** and **5** will show that Z_{I1} goes to infinity at each frequency and Z_{I2} is zero at ω_2 and infinity at ω_3. **Examination of eq. 1** shows that the denominator becomes zero if $AB = 1$. A zero denominator will yield a result of infinity. Similarly, the numerator of **eq.** 2 will be **zero if $AB = 1$. A relationship to ω_2 now exists, and** from the expressions in **fig. 3:**

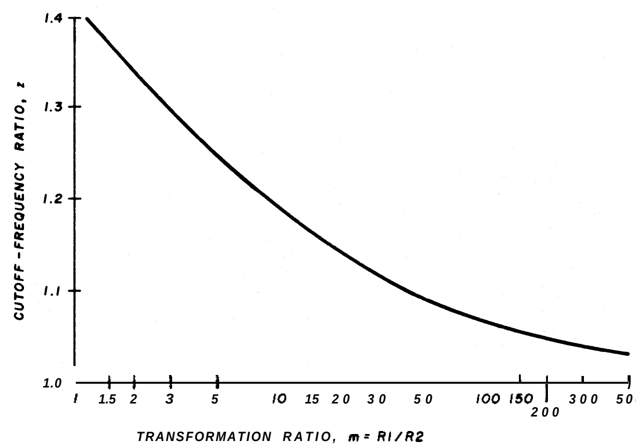

fig. 6. Cutoff frequency ratio, z, as a function of transformation ratio, m. Note that z is an inverse function of m. A lower resistance ratio will provide greater matching bandwidth; a higher resistance ratio gives a narrow bandwidth.

$$\omega_2^2 = \frac{1}{C1L} \qquad \textbf{(10)}$$

Both image impedances are infinity at ω_3. This **condition will** result if the term $(A + E - ABE)$ is zero or $(A + E) = ABE$. This term is common to **eqs. 1** and 2. Using the expressions in **fig. 3,**

$$\omega_3^2 = \frac{C1 + C2}{C1\,LC2} \qquad \textbf{(11)}$$

Eqs. 10 and **11** are a bit clumsy to handle directly. It

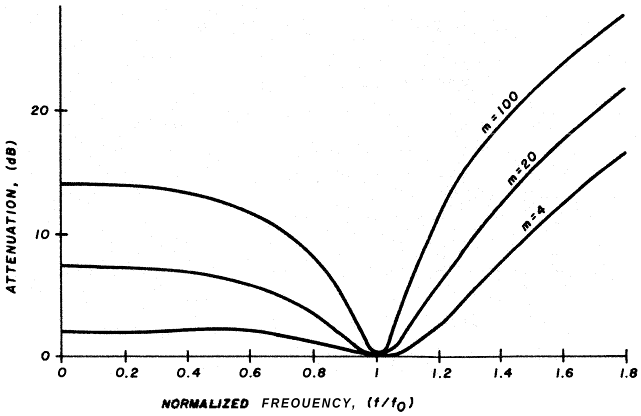

fig. 7. Normalized attenuation characteristic for various transformation ratios. Curves illustrate the difference in response caused by transformation ratio *m*.

will be easier to handle if another ratio is used:

$$z = \frac{\omega_3}{\omega_2} \qquad \textbf{(12)}$$

Some algebraic manipulation of eqs. **3** through **6** and **12** result in the identities

$$z = A_0 B_0 = \sqrt{\frac{A_0 + E_0}{E_0}} \qquad \textbf{(13)}$$

normalizing the component values

More algebraic manipulation [redacted]s. 4, 5, 6; 9 through 11; and 13 give the follo[redacted]tionships:

$$A_0 = \frac{\sqrt{m(z+1)}}{R1} \quad \text{and } C1 = \frac{}{\omega_0 R1} \qquad \textbf{(14)}$$

$$B_0 = \frac{z\,R1}{\sqrt{m(z+1)}} \quad \text{and } L = \frac{z\,R1}{\omega_0\sqrt{m(z+1)}} \qquad \textbf{(15)}$$

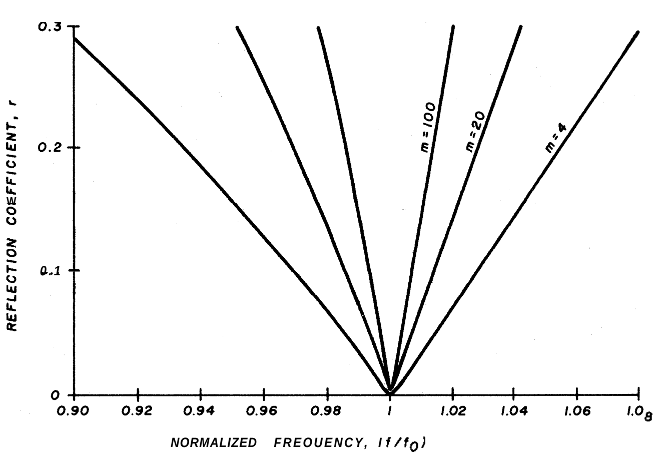

fig. 8. Normalized reflection characteristic for various transformation ratios. Best operation occurs when reflection coefficient $r < 0.1$.

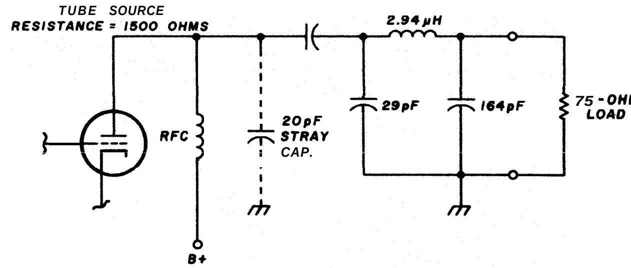

fig. 9. Pi net example for a tube amplifier. Frequency range is between 13.9-14.4 MHz. The reflection coefficient can be checked against fig. 8 for the band edges.

$$E_0 = \frac{1}{R2(z-1)\sqrt{m(z+1)}} \qquad \textbf{(16)}$$

$$\text{and } C2 = \frac{1}{\omega_0 R2(z-1)\sqrt{m(z+1)}}$$

Cutoff frequency ratio, z, can't be selected arbitrarily. It has a definite relationship to end-resistance/transformation ratio m. Using **eqs.** *9* and **13** through **16**, this is:

$$z^3 - z^2[(m-1)/m] - z[(m+1)/m] + [(m-1)/m] = 0 \qquad \textbf{(17)}$$

Fig. 6 is a plot of cutoff frequency ratio, z, versus transformation ratio, m. Eq. **17** is valid only for m greater than unity and $R1$ greater than $R2$.

Note that z is inversely proportional to m. A lower resistance ratio will give a greater matching bandwidth; a higher resistance ratio gives a narrow bandwidth. A check on bandwidth is possible by direct analysis of a network versus frequency, but it may be easier to test by mathematical means.

transfer functions and properties

A transfer function is a mathematical expression of a network that allows comparison of input to output *vs.* frequency. It can also be used to find the *reflection coefficient* at the input port as a check of matching bandwidth. This reflection coefficient is the same as that of a loaded transmission line VSWR expression.

The common transfer function, $F(\Omega)$, representing available generator power versus output power, may be expressed in the normalized form for the pi network as:

$$F(\Omega) = \left(\frac{1}{2\sqrt{m}}\right)\left[A_{11} + \frac{A_{12}}{R2} + A_{21}R_1 + A_{22}m\right]$$

$$\equiv G + jU = \sqrt{\frac{P_o}{P_2}} = \frac{E_g}{2E_{out}\sqrt{m}} \qquad \textbf{(18)}$$

where A_{11}, A_{12}, A_{21}, A_{22} are the matrix terms given previously and

Ω = ω/ω_0, ratio of calculation versus center frequency

G = real part of $F(\Omega)$

U = imaginary part of $F(\Omega)$

P_o = available power of generator, R1

P_2 = power absorbed by R2

The transfer function, $F(\Omega)$, may be expressed in terms of Ω, z, and m by:

$$F(\Omega) = \frac{1}{2\sqrt{m}}\left[\frac{m(z^2-1)(1-\Omega^2 z)+z^2-\Omega^2 z-1}{z^2-1} + j\,\frac{\Omega z(2z-\Omega^2-1)\sqrt{m(z+1)}}{z^2-1}\right] \quad (19)$$

The term $\frac{1}{2\sqrt{m}}$ is the normalizing factor. Any combination of z and m will result in unity *magnitude* of $F(\Omega)$ at ω_0; this allows a response comparison without scale shifting.

Eq. 19 may be programmed on a calculator but must be converted to polar form.* Taken directly, the term in brackets of the normalized transfer function is the generator voltage of **fig.** 1 divided by the complex output voltage across $R2$. The ratio of complex-output-voltage-to-complex-input-voltage of the network is the inverse of $F(\Omega)$, with the result multiplied by $1/2$.

Normalized attenuation around design center frequency can be expressed by:

$$e^{2\alpha} = |F(\Omega)|^2 = G^2 + U^2 \quad (20)$$

This response is plotted in **fig.** 7 showing attenuation as a function of design center frequency. Attenuation in dB is $20\log_{10}(e^{\alpha})$. This plot illustrates clearly the difference in response caused by transformation ratio m.

Reflection coefficient, r, is the reflected voltage

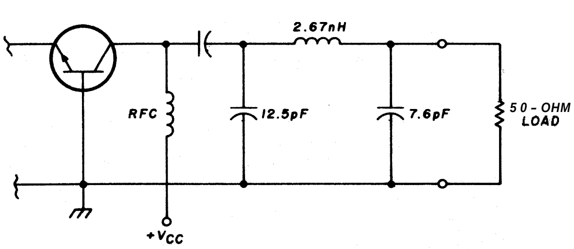

fig. 10. Example showing a transistor amplifier in the range of 12151300 MHz. Transistor output impedance is resistive at 12.5 ohms across the frequency band; $R1$ is now the load and $R2$ is the generator.

*A rectangular-form expression is usually easier to show in texts. The problem with such expressions is that the complex real pan becomes **negative!** Conversion to polar form gives a positive magnitude with the correct phase angle.

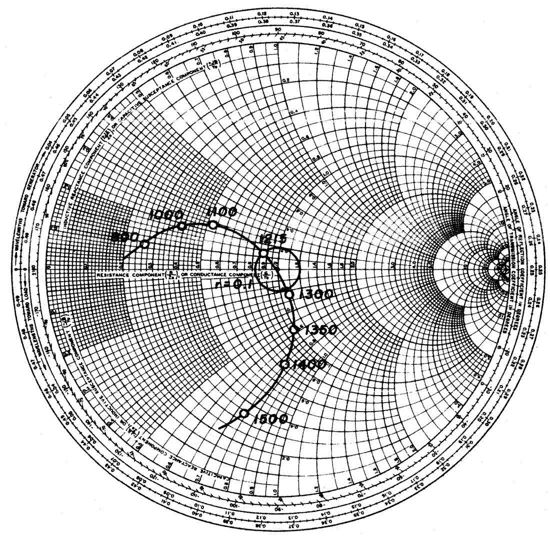

fig. 11. Smith chart showing network output impedance of a transistor amplifier and pi network. Output VSWR is less than 1.25 across the desired bandwidth.

divided by the forward voltage at the generator. Normalized reflection coefficient can be derived from **eqs.** 19 and 20:

$$r = \sqrt{\frac{e^{2\alpha}-1}{e^{2\alpha}}} \quad (21)$$

This expression is plotted in **fig.** 8 for **m** values of 4, 20, and 100. Best operation occurs when r is less than 0.1. Bandwidth again depends on transformation ratio.

some examples

A **vacuum-tube** amplifier has an output impedance of **1500** ohms with stray capacitance of 20 **pF. It** is to be matched to a 75-ohm line over the range of **13.9**-14.4 MHz. The geometric center of the range will be the design center frequency:

$$f_0 = \sqrt{13.9 \times 14.4} = 14.15\,MHz$$

$R1$ is **1500,** R2 is 75; so the transformation ratio, m, equals 20. From **fig.** 6, **m** $= 20$; $z = 1.14$. **Eqs.** 14, 15 and 16 give the component values after calculation of some common terms:

$$\omega_0 = 2\pi f_0 = 6.283 \times 14.15 \times 10^6 = 88.91 \times 10^6$$
$$\sqrt{m(z+1)} = \sqrt{20 \times 2.14} = 6.542$$

From **eq.** 14:

$$C1 = \frac{6.542}{88.91 \times 10^6 \times 1500} = \frac{6.542}{133.4 \times 10^9} = 49.06\,pF$$

Subtracting the **20-pF** stray capacitance gives a component value of 29.06 **pF.** From **eq.** 15,

$$L = \frac{1.14 \times 1500}{88.91 \times 10^6 \times 6.542} = \frac{1.710 \times 10^3}{581.6 \times 10^6} = 2.940\,\mu H$$

From **eq. 16:**

$$C2 = \frac{1}{88.91 \times 10^6 \times 75 \times 0.14 \times 6.542}$$

$$= \frac{1}{6.107 \times 10^9} = 163.7\ pF$$

The tube and network circuit is shown in **fig. 9.** The cutoff frequencies, ω_2 and ω_3, can be checked with **eqs.** 10 and **11** as 13.25 MHz and 15.11 MHz respectively.

The reflection coefficient can be checked against **fig.** 8 for the band edges. The high end is *14.4/14.15* or *1.018* relative to center; low end is *13.9/14.15*, or *0.982*. The reflection coefficient is approximately 0.12, or 12 per cent.

Another example is a microwave amplifier output circuit **(fig. 10).** The desired range is 1215-1300 MHz. The transistor must match a 50-ohm load. The transistor output impedance is resistive at 12.5 ohms across the band; *R1* is now the load and *R2* the generator end. The design center frequency is:

$$f_0 = \sqrt{1215 \times 1300} = 1257\ MHz$$

Transformation ratio is **50/12.5** or 4, and z is 1.27 (from **fig. 6**). Remembering that *C2* is next to the generator and *C1* next to the load, **eqs. 14,** 15, and **16** give the following values:

$$C1 = 7.63\ pF,\ L = 2.67\ nH,\ C2 = 12.45\ pF$$

A Smith-chart plot of the network output impedance, normalized to **50** ohms, is given in **fig. 11.** Output **VSWR** is less than 1.25 across the desired bandwidth.

group delay and element dissipation

Group delay is the differential phase delay divided by the differential frequency across the desired band. **It** is the time delay of a signal through the network and is important for wideband modulation transmission and determination of network dissipation loss. Normalized group delay, $\omega_0 T_g$, can be determined from the complex transfer function of **eq. 19** as:

$$\omega_0 T_g = \frac{\left(\frac{dU}{d\Omega}\right) G - \left(\frac{dG}{d\Omega}\right) U}{G^2 + U^2} \qquad \textbf{(22)}$$

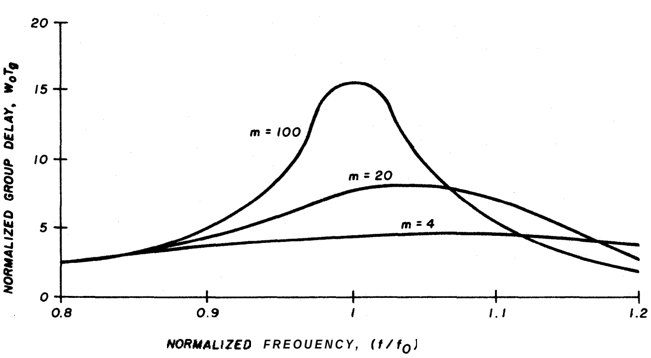

fig. 12. Normalized group delay for three transformation ratios, *m*.

Normalized group delay for three values of ***m*** is plotted in **fig. 12.** In general, maximum group delay occurs at rapid attenuation versus frequency.* The pi-network maximums are slightly higher than design center frequency.

Network attenuation other than the transformation ratio is determined by the unloaded Q of each reactive element. Knowing the element *Q* **allows** determination of loss through the normalized group-delay expression:

$$a, = loss(in\ dB) = \frac{4.343\omega_0 T_g}{Q_u} \qquad \textbf{(23)}$$

Eq. 23 is assisted by the design-center-frequency normalized group delay plotted in **fig. 13.** Assuming

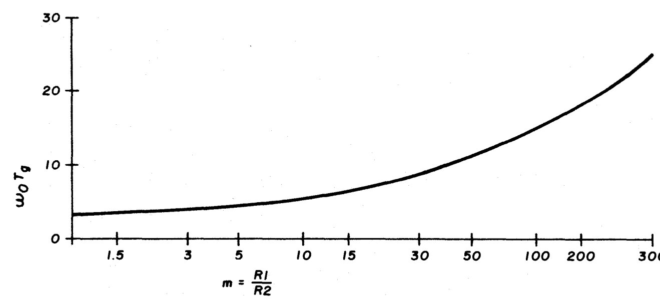

fig. 13. Normalized group delay at center frequency for various transformation ratios.

very high unloaded Qs of the capacitors, an unloaded Q of 160 for the inductor, and an ***m*** of 20 (from the first example), additional network loss will be 0.20 dB, a negligible amount.

summary

The optimum design of a pi network depends on the transformation ratio. Bandwidth is inversely proportional to this ratio. Simple calculation of components is possible with the aid of a few graphs. Reflection coefficient is proportional to transformation ratio and may be used to determine if a network must be retuned for a particular bandwidth.

***Phase shift through a filter is responsible; all filters have rapid phase changes as amplitude response moves from passband to stopband regions.**

a note on the appendix

Appendix A was included in Mr. Leonard H. Anderson's rewritten version of this article. It discusses the "image impedance" method of network design with respect to matrix notation. Also included are normalized component values and their relationships with respect to eqs. **14, 15,** and **16** as well as an explanation of transfer functions with regard to the derivation of eqs. **19, 20,** and **22.**

Author **DL9LX,** in his original version of this article, furnished other appended material. This includes a listing of computed pi-network elements as a function of impedance-transforming ratios for various center frequencies (Appendix **B**); computed values of $z = f(m)$ from eq. **17** (Appendix **C**); and a table of normalized network elements for various transformation ratios useful in general network design (Appendix **D**).

Interested readers may obtain a copy of author **DL9LX's** appendices from ***ham radio*** upon receipt of a large self-addressed, stamped envelope with 28 cents postage. The material in these appendices is in the author's original notation.

Editor.

references

1. George L. Matthaei, Leo Young, and E.M.T. Jones, ***Microwave Filters, Impedance-Matching Networks, and Coupling Structures,*** McGraw-Hill Book Company, **1964**.

2. S.B. Cohn, "Dissipation Loss in Multiple-Coupled-Resonator Design Filters," ***Proceedings of the IRE,*** August, **1959.**

3. H.F. Mayer, "Über die Dampfung von Siebketten im Durchlässigkeitsbereich," E.N.T., **1925,** Band **2.**

4. William I. Orr, W6SAI, "Linear Amplifier Design, Part II," ***ham radio,*** July, **1979.**

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Temes and Mitra, ***Modern Filter Theory and Design.*** John Wiley and Sons, **1973.**

appendix A

image impedance

Modern network theory tends to ignore the "image method" of design. While image methods may be disregarded for complicated structures, they are valid for simple networks and quite useful at frequencies where it's difficult to obtain a purely resistive load.

Many readers will be unfamiliar with network matrix notation. Those who are familiar may be more acquainted with "A, B, *C*, D" notation instead of the subscripted form. Open and short-circuit impedances are given below, referred to **fig.** 3.

Port 1 input impedance with port 2 open:

$$Z_{o1} = \frac{Z_2Y_3 + 1}{Y_1Z_2Y_3 + Y_1 + Y_3}$$

Port 1 input impedance with port 2 shorted:

$$Z_{s1} = \frac{Z_2}{Y_1Z_2 + 1}$$

Port 2 input impedance with port 1 open:

$$Z_{o2} = \frac{Y_1Z_2 + 1}{Y_1Z_2Y_3 + Y_1 + Y_3}$$

Port 2 input impedance with port 1 shorted:

$$Z_{s2} = \frac{Z_2}{Z_2Y_3 + 1}$$

Multiplication of Z_{o1} and Z_{s1} or Z_{o2} and Z_{s2} will still give the same result as with matrix notation. The fact that each image impedance expression, while complex, results in an imaginary part of zero comes about by ***completing*** the complex division; this can be verified by completing all steps. This also applies to eqs. **(7)** and **(8)**. Real-part-only complex expressions are common to purely reactive networks.

Image impedances show the individual resonances within the network. They are synthesis tools — not an actual input impedance ***when loaded.*** The "missing" expression for network resonance at ω_1 of **fig. 4** is because of the $(BE - 1)$ term of eqs. **(1)** ***and*** **(2)**, where:

$$\omega_1^2 = 1/(C_2L)$$

normalized component values and relationships

Eqs. **(14)**, **(15)**, and **(16)** could have been expressed without the frequency ratio, z. In fact, z could have been omitted, but at a price: the usable bandwidth would not be optimized, since the design center frequency could not be located for minimum reflection coefficient, or "goodness of match."

Many readers are under the false assumption that a pi network is a resonant circuit with a quality factor, Q. It is simply an impedance transformation network with the ***appearance*** of resonance due to the sharp cutoff above design frequency. Low-side response behaves more like a conventional asymmetrical lowpass with varying ***passband*** response. Since many power amplifiers are still tube types with pi networks, transformation ratios will be high and the network will ***appear*** to peak at center frequency. As high-frequency, high-power semiconductor technology improves, transformation ratios will decrease, and the pi network will be treated as the simple lowpass filter it really is.

transfer functions

Eq. **(18)** may be found in reference 1, page 37, equations (2.10-1) and (2.10-5), using ABDC matrix descriptors. The transfer function is the generator-voltage-to-load-voltage ratio. The normalizing term will yield "available voltage" at the load; that is, all power from the generator is assumed dissipated in the load, Rl; none in source resistance, Rl.

Eq. **(19)** is obtained by substitution of the network arm reactances into eq. **(18)**. The steps of substitution and simplification are too long to be included here; they have been checked independently.

Input/output complex voltage ratio is obtained by deletion of (1/2) from the normalizing term. This yields a condition in which the generator is a constant-current source with a source conductance always present at the network input.

More detail on the attenuation and delay functions may be found in reference 1, sections 3.02 and 3.03. These use the ***image propagation function,*** γ, expressed in general terms as:

$$\gamma = a + j\beta = \ln(\sqrt{A_{11}A_{22}} + \sqrt{A_{12}A_{21}})$$

where a = image attenuation in nepers
β = image phase in radians

Eq. **(20)** is derived by manipulation of this basic expression in terms of eq. **(19).** Normalized group delay, eq. **(22)**, is derived from the basic group delay expression

$$T_g = \frac{d\beta}{d\omega} = \left(\frac{d\beta}{d\Omega}\right)\left(\frac{d\Omega}{d\omega}\right)$$

With the partial differential $d\Omega/d\omega = 1/\omega_0$,

$$\omega_0T_g = \frac{d\beta}{d\Omega}$$

which yields eq. **(22)** in terms of eq. **(19)**.

ham radio

regulated power supply for VHF transceivers

Many of today's 2-meter fm transceivers feature 25 watts (or less) of rf power. When used in the home station, these radios generally require 13.8 Vdc at 7-8 amperes. Since I didn't have a power supply of this capacity, the only solution was to build one!

Simplicity is reliability, so I decided that the circuit couldn't be complicated; but good regulation was a requirement. The circuit shown in fig. 1 features no-load-to-full-load (8.0 amperes) regulation of 0.2 Vdc. Also featured in the circuit is "fold back" current limiting and overvoltage protection.

circuit description

The power supply consists of a full-wave bridge rectifier with capacitor input. Any transformer-capacitor combination that produces 28 volts dc at 7.5 amperes at Q1 collector will work. Voltage at Q1 collector should not be greater than 40 Vdc, otherwise damage to U1 may result. Q1 and Q2 form a Darlington pair. The output voltage (in this case 13.8 Vdc) is set by potentiometer R1. This sampled voltage is applied to U1, a UA723HM voltage regulator, which contains a voltage reference amplifier and an error amplifier. U1 output is applied to Q2 base to adjust the voltage at R1 to its proper value. The low-value resistor, R2, is the current "fold-back limiter." If the power-supply output should exceed 10 amperes (i.e., a short circuit), regulator U1 will bias the transistors to cutoff; thus the output voltage will drop to near zero until the short circuit condition is corrected.

Capacitor C1 and OVP form an over-voltage-protection circuit. The OVP limits the maximum output voltage to +15 Vdc. (The OVP unit is available from VHF Engineering.) The voltmeter and ammeter are optional, depending on your junk-box supply. The

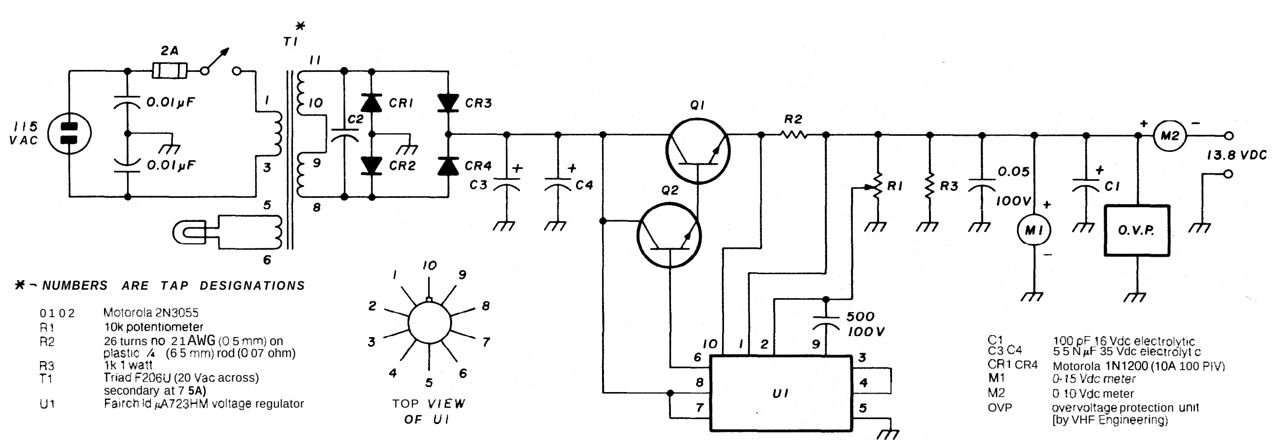

fig. 1. Schematic of the base-station power supply. Current-limiting resistor R2 is hand wound and designed for 10 amperes. If other than 10 amperes is desired, a different value must be used (see text). The overvoltage protector (OVP) is available from VHF Engineering. Q1, Q2 must be mounted on a heat sink, which must be insulated from the chassis.

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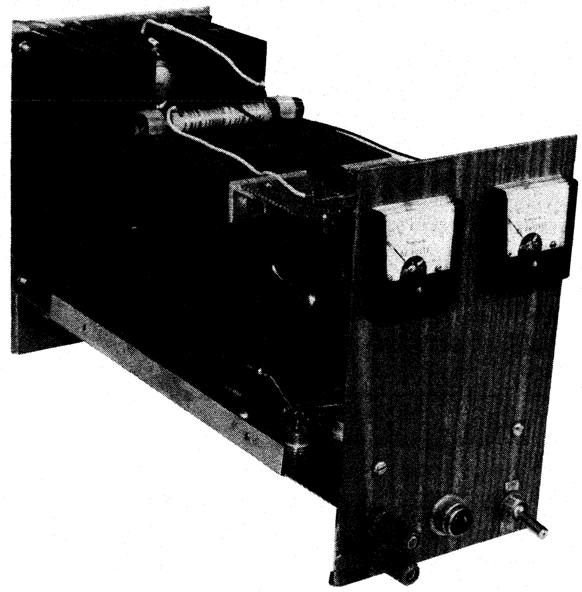
Inside view of the vhf transceiver regulated power supply.

output voltage (determined by R1) is adjustable from about 7-14 Vdc.

construction details

Component layout isn't critical; there's room for wide variation in this regard. Both Q1 and Q2 must be mounted on a suitable heat sink, which must be insulated from the chassis. "Current foldback" resistor, R2, should be wound on a plastic or Teflon rod of about ¼ inch **(6.5 mm)** diameter. Regulator **U1** and potentiometer **R1** are mounted on a piece of Vector board. If you wish to have the current limited to other than 10 amperes, a different resistor for R2 will have to be wound. To determine the new resistance:

$$R_{limit} = \frac{0.7}{I_{limit}} \qquad (1)$$

Where R_{limit} is the new value resistor (ohms), and I_{limit} is the maximum desired current (amperes).

closing comments

Operating results with the power supply have been excellent. If you have a good junk box, or are a good trader at the Hamvention circuits, this supply should cost less than half that of a similar commercial model. Circuit and component layout aren't critical so you have a weekend of fun in constructing your own base-station power supply.

ham radio

More Details? CHECK–OFF Page 94

counter mixer

for the Kenwood TS-520-SE transceiver

This interface circuit between your transceiver and counter provides high accuracy

Most of the digital readout Amateur equipment available today is accuracy-specified to the nearest 100 Hz. This may mean plus or minus 100 Hz, plus or minus the the-base accuracy, and it may include programmed beat frequency oscillator (BFO) allowances. In this case the BFO output is not actually counted, and accuracy will suffer from tolerances of the BFO crystals. Because of linearity problems with the mechanical dial, the digital system will probably be more accurate across the dial, but just barely.

I've used the system described here in various forms and with different gear for the past eight years. It involves mixing the transceiver's three oscillator outputs to produce the operating frequency.[1] To measure the frequency of an incoming carrier, the low i-f is amplified and limited, then substituted for the BFO oscillator output in the mixing scheme. This option is useful for frequency-measuring tests and to calibrate the frequency counter used as a readout by checking WWV.

This particular unit (**fig. 1**) is for use with a Kenwood TS-520-SE. This transceiver has the oscillator outputs as well as a dc-supply connection available on the back panel. To provide an i-f output, another phono jack is installed on the back panel and an emitter follower is used to bring out a tap to the i-f board. See **fig.** 2.

description

Use of this unit with other gear would require providing the proper oscillator, i-f, and power-supply connections. If the rig uses a different mixing scheme, the bandpass circuit between mixers will have to be changed to the new high i-f. Also the input and output coils on the low i-f amplifier and limiter stage will need to be resonated to the different low i-f. The Heath SB-102, for instance, uses the same mixing scheme, and the same coils and capacitors may be used. The six output circuits remain the same, one for each band.

Doubly balanced mixers attenuate unwanted outputs. Separate bandpass circuits cover each frequency range. The 14-15.6-MHz range is covered by one filter as is the entire 10-meter band.

After the bandpass circuits have selected the desired bands, a two-stage broadband amplifier brings up the level to operate a frequency counter.[2] The bandpass circuits are calculated for an R of 10,000 ohms and were designed after an article by Anderson.[3] The requirements in this application are not strict. The filter caps were changed to the nearest standard value.

construction

Use good shielding to avoid feedback of the output signal to the receiver input. When used in the signal-measuring mode, feedback is more likely to occur. To operate in this mode, the two isolation amplifiers are necessary to prevent signals from entering the receiver on the heterodyne- and tuning-oscillator cables. Another source of feedback is the frequency counter. The counter should be enclosed in a metal case to prevent radiation of the high-level

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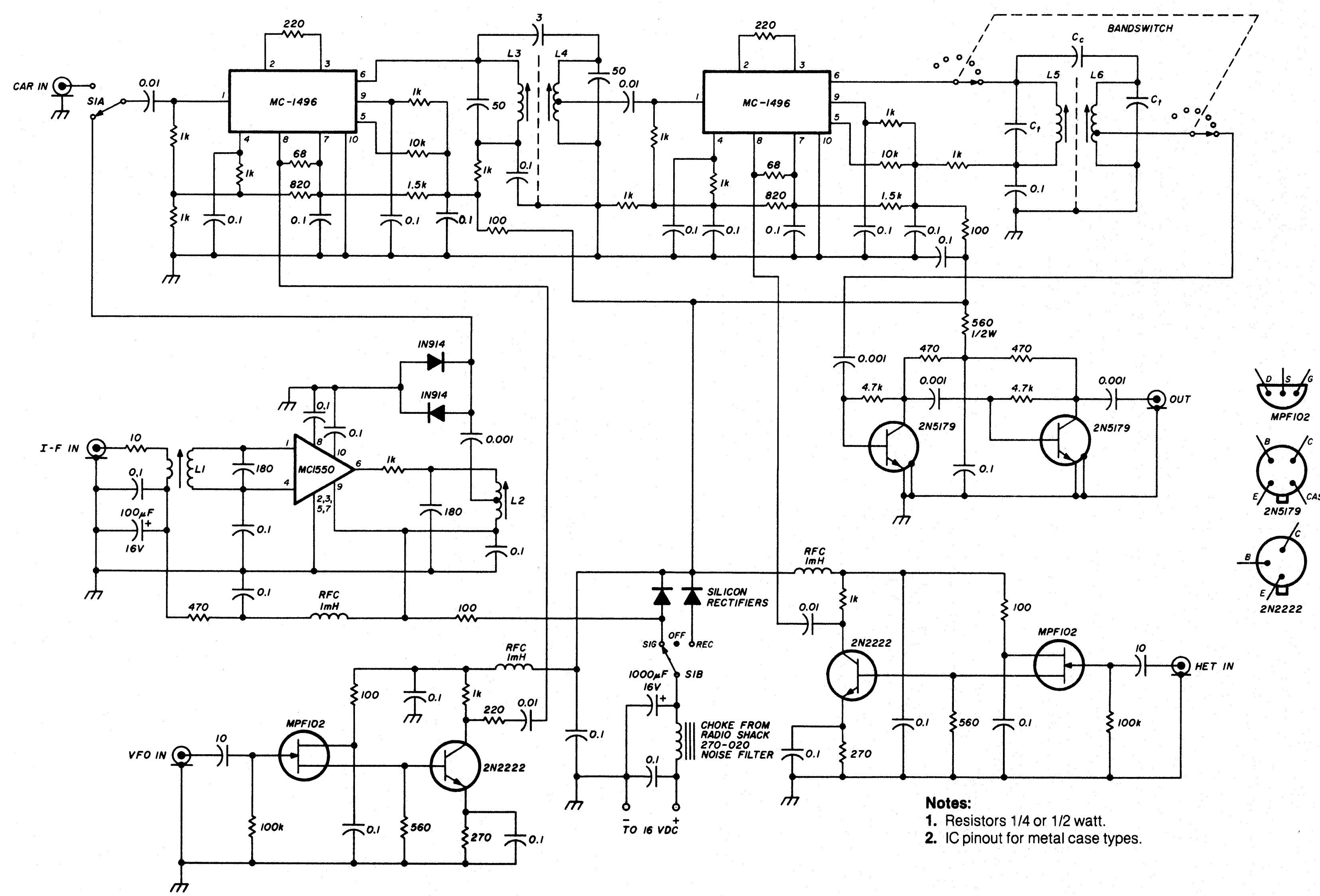


Notes:
1. Resistors 1/4 or 1/2 watt.
2. IC pinout for metal case types.

fig. 1. Schematic of the counter mixer circuit designed for the Kenwood TS-520-SE Amateur-band transceiver. Circuit is easily adaptable to other transceivers.

signals in its input circuits. A short cable between the counter and the mixing unit is helpful, as is a short direct chassis-to-chassis connection between all units.

If you need the unit only as a digital readout of the operating frequency, the isolation amplifiers, i-f amplifier, mode switch, and the i-f connection to the transceiver may be omitted.

operation

When in the operating-frequency mode, the readout changes as the dial is moved. in the signal mode, the readout indicates the frequency of a carrier as long as it is in the passband and of sufficient strength. An S-1 signal will register in the absence of interfering signals. As an example, tuning across WWV results in an unchanging 15 000 000 on the counter. Of course, fading and multipath distortion will alter the reading for some count periods. The counter will count what it sees, and the receiver must provide a countable signal. To help ensure a clean count the receiver should have a CW filter to narrow the passband and separate modulation from the carrier. However, the system will work well with an SSB filter, especially on WWV.

tune up

The following procedure is for use with the Kenwood TS-520-SE transceiver.

1. Set mode switch to RECEIVE.

2. Adjust L3, L4 **(fig. 1)** for maximum signal at pin 1 of the second mixer (MC1496) with the transceiver tuned to about band center.

3. Adjust L5, L6 on each band for proper counter operation. Make small adjustments to allow coverage of the entire band. Observation of the counter is probably the best indication of proper tuning.

4. Set mode switch to SIGNAL.

5. Tune in a steady signal, such as that from the calibrator. Adjust L1, L2 for maximum output at the tap on L2. Reduce signal strength as needed to allow peak tuning.

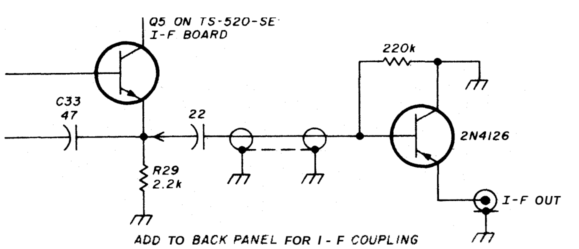


fig. 2. Emitter-follower circuit added to the back panel of the Kenwood TS-520-SE transceiver for i-f signal coupling.

table 1. Coil data for the counter mixer circuit in *fig.* 1

band (MHz)	L5-L6	L6 tap from bottom (turns)	C_t (pF)	C_c (pF)	approximate inductance (μH)
1.8-2.4	no. 32 (0.2 mm) 124t	10	39	12	109.0
3.5-4.1	no. 32 (0.2 mm) 68t	6	50	6	32.75
7.0-7.6	no. 32 (0.2 mm) 36t	3	50	3	8.9
14.0-15.6	no. 30 (0.25 mm) 28 t	3	20	2	5.75
21.0-21.6	no. 30 (0.25 mm) 10 t	1	50	1	1.04
28.0-29.7	no. 30 (0.25 mm) 13t	1	20	1	1.6
high i-f (MHz)	**L3-L4**	**L4 tap**			
8.295-8.895	no. 32 (0.2 mm) 30 t	7 turns from bottom			
low i-f	**L1-L2**	**L2 tap**			
3.395	no. 32 (0.2 mm) 42 t	21 turns			

Note: All coils are wound on slug-tuned forms 7132 inch 15.5 mm) diameter with 1/2 inch (12.5 mm) winding space. Coil is confined to 3/8 inch (9.5 mm) nearest terminals. Windings are layer or scramble wound.

6. As a final check, observe the counter and adjust for readout of the weakest possible signal.

performance

To give an example of the capabilities of this system, I made sixteen consecutive readings of the WWV carrier at 15 MHz. A count gate time of 100 seconds was used. The readings showed a slow drift above and below the 15-MHz target frequency, with a maximum error of 0.35 Hz. Most likely, the major portion of the error was the result of the oven control, with some error due to propagation delay. The counter is controlled by a 1-MHz crystal in a proportionately controlled oven, which has been on for over five years except for power failures.

Of course I've not approached the point where I can begin to look to WWV as a source of error. But obtaining consistent readings with an error of ***less than one part in fifteen million*** is quite satisfying.

references

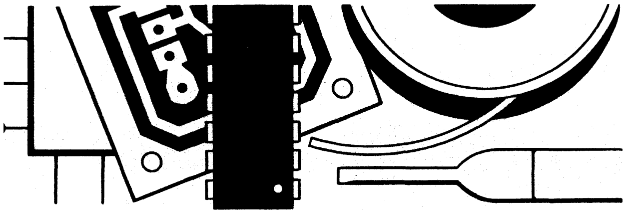
1. Mac Leish, "A Frequency Counter for the Amateur Station," QST, October, 1970.

2. Randall Rhea, WB4KSS, "General-Purpose Wideband rf Amplifier," ham radio, April, 1975. pages 58-61.

3. Leonard H. Anderson, "Top-Coupled Bandpass Filter — a Chebyshev Design," *ham* radio, June, 1977, pages 34-40.

ham radio

the weekender



A simple 40-meter receiver

This article is for those who like to build their own equipment. **It** is a summary of a solid-state receiver that has performed very well. The receiver is the result of my experience trying to find circuits that work. It's about as simple as you can find. The receiver uses an **rf** stage, which certainly helps at night when foreign broadcast stations come through on 40 meters.

The most time-consuming part of the project was making the PC boards. **I** used black PC drafting tape to lay out the boards, which were etched in ferric chloride. Others will probably come up with a better method.

brief circuit description

Many of the receivers shown in the handbooks don't include an rf stage ahead of the mixer. This receiver was first tried using a double-tuned circuit directly into the mixer. However, the circuit was mounted in the chassis and capacitive coupling wasn't satisfactory. At night the shortwave broadcast stations came through. By using this simple rf stage, the selectivity problem was resolved. You can use whatever toroid forms you can find. The T-80-2 forms are probably too large, but these are what **I** have used. They have a red core and are about 1 inch (25.5 mm) in diameter. Resonance can be checked by holding a grid-dip oscillator to the hot end or by placing a turn or two of wire around the core and checking with the grid-dip oscillator. The input coil was mounted on the chassis topside and the output coil was mounted underneath. Fig. 1 shows the circuit.

mixer

The mixer output coil was tuned to 5.5 MHz. **It** is wound on a 318-inch (9.5-mm) ceramic slug-tuned coil using a fixed 100 **pF** cap for tuning. L5 (about 13 turns) on the bottom of the coil feeds directly into the Swan crystal filter.

By Ed Marriner, W6XM, 528 Colima Street, La Jolla, California 92037

variable-frequency oscillator

The **VFO** tunes 12,500-12.800 kHz to cover the 40-meter band. A small cap with two rotary and two fixed plates came out to about 35 **pF,** which just covers the band. You can pull plates out after the set is going to obtain desired bandspread. Coil L6 was wound on a 3/8-inch (9.5-mm) ceramic slug-tuned coii form. **It** was wound with silver-plated wire, nylon covered. The two 500-pF caps from the MPF-102 gate to ground are silver micas. A 9-volt zener stabilizes the MPF-102 drain. The two 2N2222 stages are buffers to reduce pulling effect on the **VFO** and to obtain the 1.5 V rms to feed the mixer.

i-f stage

Only one i-f stage was necessary for this receiver. L7 is another 318-inch (9.5-mm) diameter slug-tuned coil adjusted to 5.5 MHz. It's coupled into the product detector with a 0.001 μF cap.

beat-frequency oscillator

The **BFO** can be varied about 1 kHz with the 0-30 **pF** variable cap to adjust the **SSB** tone. The output coil, L8, is another slug-tuned coil. The 10 **pF** coupling cap to the product detector should be sufficient, but you can try other values if there isn't enough signal injection.

construction

I built the receiver starting from the audio stage and worked backward to the front end. **I** used 2-inch (50-mm) square PC board. **I** laid out the circuits using black drafting tape and etched the boards with ferric chloride. Parts were mounted on a 7 x II-inch (178.5 x 280.5-mm) aluminum chassis." The **VFO** was mounted in a partition topside, which was about 3-inches (76.5-mm) square. **I** used a **VTVM** and rf probe for tune up.

performance

I'm amazed at the performance of this little receiver. **I** frequently operate it from a battery supply during park picnics and wonder how **I** ever got along without it. Don't ask me how to make a transceiver for CW; so far I've not been able to make a mixer to drive a transmitter section. t

"Pans that may be useful for construction are available from Radiokit, Box 411H, Greenville, New Hampshire 03048.

†See John Keith's article, "40-Meter Transceiver for Low-Power Operation," ***ham radio.*** **April, 1980, page 12.**

L1 35T NO.26 (0.3mm) TAPPED 10T BOTTOM ON MICROMETALS T-80-2 FORM (RED CORE)
L2 3T TAPPED FROM GROUND END OF L1
L3 SAME AS L1
L4 25T NO.30 (0.25mm) ON 3/8 IN. (9.5mm) CERAMIC FORM
L5 13T NO.30 (0.25mm) ON BOTTOM OF L4 (FILTER INPUT)

L6 7T NO.26 (0.3mm) ON 3/8 IN. (9.5mm) CERAMIC FORM
L7 25T NO.30 (0.25mm) ON 3/8 IN. (9.5mm) CERAMIC FORM
L8 SAME AS L7

NOTES:
1. ALL RESISTORS ARE 1/2 WATT
2. ALL CAPS 16-25 WVDC

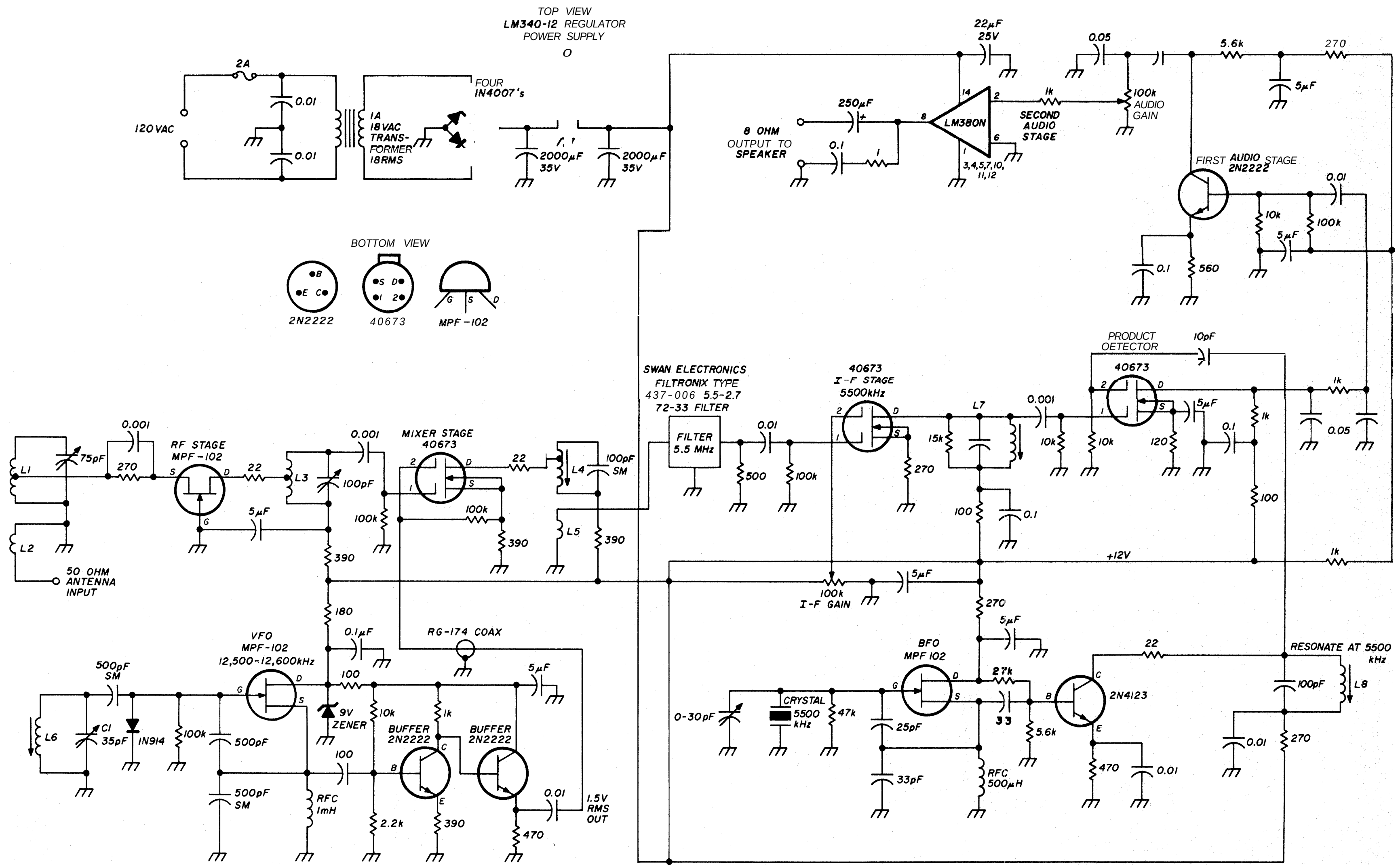


fig. 1. Schematic of the solid-state 40-meter receiver. The MPF-102 rf stage helps selectivity, especially when foreign broadcast stations appear at night. PC boards are used, which were made using the tape-and-etching method. Low-frequency circuits could be mounted on perf boards. The 5.5-MHz filter was picked up at local flea market.

comments

(Continued from Page 6)

he product of baud speed times K, K eing a factor depending on the goodness" of a circuit. The baud peed of Amateur Hellschreiber is 22.5 which multiplied by a K factor f 3 gives a bandwidth of 367.5 Hz. he K factor of 3 comes into the picure because it has long been recognized that a square wave and its third armonic is perfectly acceptable for ormal communications. The bandvidth of a 45.5-baud, 170-Hz shift TTY computed according to CCIR is 45 Hz.

K6KA is correct in his criticism of he Chinese Hell-Fax signal, lately on 4140 and believed still to be working n the Region 2, 80-meter band. But his is a different system with a baud peed of somewhere in the region of 00, and observedly with little or no ttempt at pulse shaping; some chanels are even FSK with 800 Hz shift! 'hey are certainly wide band and not o be compared with the Amateur hell' in Europe.

Finally, I hold no brief for the Hellchreiber system as such but, as I vorked with the system throughout nost of its active life and am fully onversant with its advantages and hortcomings, I thought I'd like to put he matter straight.

Stanley A.G. Cook, G5XB
Radio Society of Great Britain
Reading RG4 9BP, England

PCB "threat"

Dear HR:

I noted in "Presstop" in May, 1980, ssue your warning regarding the 'potentially deadly threat" existing in he form of PCBs or polychlorinated iphenyls, and should like to thank ou for bringing the attention of the raternity to this material.

However, I should like to point out hat the PCB hazard has been, like nany others, vastly overrated by nedia exposure. PCB in massive loses fed to lab test animals has been shown to produce malignant tumors, and repeated applications to the skin of mice has indicated some potential as a dermal carcinogen.

PCB came to the attention of health authorities through two major instances. One was in Japan, where, by error, it was substituted for fish oil in food packaging. The second instance occurred in the U.S., where, in error, it was added in place of vegetable oil, to cattle feed. In both cases severe illness resulted from the consumption of the PCB-contaminated food.

Occasional handling of PCB has shown no deleterious effects on humans. In fact, many Amateurs who are also Industrial Electricians will testify that they have had their hands in it innumerable times, and in big transformer work have literally been immersed in it, with no visible short- or long-term effects.

The properties of PCB, which make it such an excellent electrical insulating fluid, are the qualities that cause the physical and ecological problems. It is heavier than water, non-conductive, and will not break down or decompose at temperatures under 2000°F. In fact it requires the full 3500°F heat of a cement kiln to break it down. Under normal conditions, it is not bio-degradable. This is its biggest hazard. Once spilled, it remains in the ground indefinitely, being propagated by natural ground waters, absorbed unchanged by plants, which are then eaten by animals.

Incidentally, if you have a tube-type television set or refrigerator more than ten years old, fluorescent lights, or a car with brake fluid or hydraulic fluid more than ten years old, you probably have another source of PCB.

Amateurs, building or buying dummy loads without transformer oil, and having gone to their local utility for a gallon of "good, hi-temperature transformer oil" have received a gallon of PCB. All the above is presented to show that PCBs have been around and done a good job for years, and pose no "potentially deadly threat" in the quantities hams use.

PCBs can be differentiated from mineral or vegetable transformer oils by the following means:

1. The smell of PCBs is somewhat similar to that of moth balls. Ordinary vegetable or mineral transformer oils smell like oil.

2. Pure PCB is heavier than water, and a drop dropped into a bottle of water will sink. Ordinary transformer oil will float on water.

If you have a PCB-filled dummy load that has a leak or a filter capacitor filled with PCBs that shows a leak around the bushings these leaks can be easily repaired using "Weldfast 220 or equivalent epoxy. First clean off all PCB seepage with a good solvent; "Xylene" will do fine. Wear rubber gloves to protect you from both the Xylene and the PCBs, and store contaminated wipers in a sealable can. Mix the epoxy, smear over and around the leak, and let it set. Job done.

Clean up any spilled PCBs well with Xylene and rags. Store rags, rubber gloves, and all contaminated materials in a sealed can. The whole object of the game is to keep the PCB from getting directly into your food and from getting into the food chain via the earth and ground water. A call to your public utility will provide a safe method of disposing of your PCB wastes.

Above all, remember PCBs are a hazardous substance, not a "deadly threat." Inspect your capacitors and ensure they are not leaking PCBs. If they are, repair the leaks and clean up the spills properly, or remove the bad component and clean up the spill properly. Put all contaminated materials in a sealed can, wash your hands well, call your public utility and make the necessary disposal arrangements. Don't panic and throw them in the garbage. If you do, you can be sure of getting your share of them back through the food chain.

Tom Ruynon, VE5UK
Saskatoon, Saskatchewan

speed of light

Dear HR:

In his amusing exposé, W7ITB has drawn before our eyes the frightening picture of all our emissions eventually coming back upon us because of the speed of light becoming negative by the 2100th century.

May I draw your attention to a hint at a much earlier date of this reversal? In the May, 1976, issue of *ham radio* on page 31 VK2ZTB cites the speed of light to be 290500 km s^{-1}. Should we all possibly have overlooked this dramatic 3 per cent decrease or could it be a specific development of our fellow Australian hams working towards a "light boomerang?"

Gunter Hoch, **KL6WU**
Darmstadt-Eberstadt

surplus tubes

Dear HR:

I agree with Bill Orr's suggestion in December *ham radio* that one should test surplus tubes as soon as possible after their receipt. However, I object to his blanket condemnation of mail-order surplus houses, that customers have a "fat chance" of getting a refund or replacement for a defective surplus tube purchased via mail order.

We at Fair Radio Sales have been selling used and unused surplus tubes to Radio Amateurs for many years. As a matter of policy, we replace an unsatisfactory tube or refund its price, provided the customer's claim is made within ten days or so of the tube's receipt.

Like any reputable business, we feel that we have the responsibility to make every reasonable effort to satisfy our customers. Undoubtedly there are other surplus dealers who share this commitment to their customers.

Orr's remark was a disservice to the conscientious surplus mail order companies — many of which advertise in *ham radio.*

George Sellati
Fair Radio Sales
Lima. Ohio

short circuits

Yagi antenna design: performance calculations

The caption for **Table 1** on page 25 of the January, 1980, issue of *ham radio* should read: Element reactance for different wavelength-to-radius ratios, K. The caption for **fig. 2** should read: Graph showing the relationship between the wavelength-to-radius ratio, K . . . .

coaxial-line transformers

W6TC reports that on page 17 of the February, 1980, issue of *ham radio,* eight lines below the heading "50/200 ohm transformers," the text should read: ". . . two pairs of RG-58 A/U [not **RG-59** A/U] cable."

Touch-Tone decoder

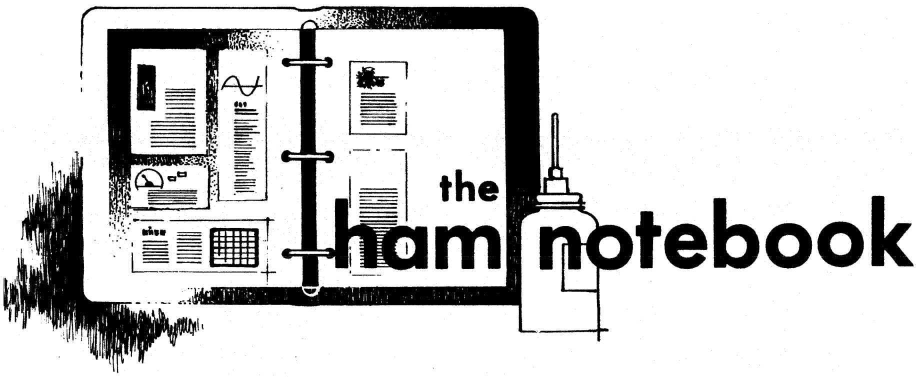
The schematic of the *Touch-Tone* decoder that appeared in the February, 1980, issue of *ham radio* (page 37) should show a crystal, not a resistor, at pin 2 of U3. Pins 3 and 5 of U10 are tied to +12 volts, with pin 4 tied to D1 of the sequential control outputs. The price of the complete kit is $140, assembled and tested $160, from James Wyma, WA7DPX, 12952 Osborne St., Arleta, California 91331.

capacitance measurement

Fig. 1 of the *Capacitance Measurement* article that appeared on page 44 of the April, 1980, issue of *ham radio* appeared with the **HI** and **LO** positions of **S1**B inadvertently reversed. The open contact, which should be marked **HI,** is grounded.

experimental high-gain phased array

In KL7IEH's high-gain phased array article, which appeared on page 44 of the May, 1980, issue of *ham radio,* the reflector elements in **figs. 1** and **5** should be broken in the middle, as should the first director in **fig. 1.** In **fig. 4,** elements D_1 and D_2 are reversed.



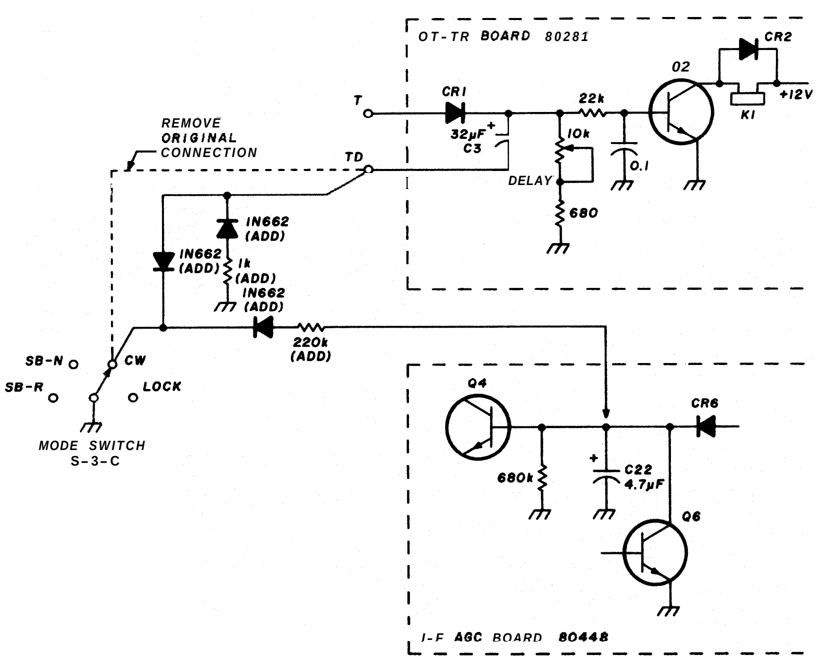
nother improvement 'or the Ten-Tec Omni-D CW agc

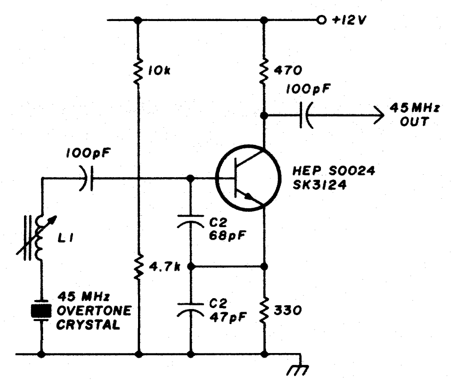
The improved CW **agc** for the Ten-'ec Omni-D transceiver by Doug /lcDougall *(ham radio,* January, 980, page 88) adds a needed fea-ure. However, **McDougall's** approach ısing a new switch deck didn't ıppeal to me because it requires a ;ooling-off delay and material not in ny junk box.

An examination of the circuit suggested that the CW-agc change could be made using diodes instead of a new switch deck. Fig. 1 shows the circuit installed in my radio. The original mute relay circuit seems to be unaffected by the change, and the improved CW **agc** works as described by **McDougall's** modification. The **1-k** resistor and diodes allow C3 to charge and discharge only in the CW mode.

My thanks to **McDougall** and *ham radio* for publishing the article, which led to my circuit.

John Bunting, **W4NET**





overtone crystal oscillator

I found **WB2EGZ's** article, "Quartz Crystals — Gems for Frequency Control," *ham radio,* February, 1979, page 37, very interesting. I had a transmitter on my bench with a crystal fault. The transmitter had been modified so that it would operate on 144 MHz; however, the oscillator wouldn't operate in the overtone mode of the crystal. While the article by **WB2EGZ** was first-class, this is one type of oscillator he didn't cover, although he had a variation of it in his fig. 4. So it was a case of finding out what was happening in my transmitter.

I found that the critical part in the circuit (fig. **2**) was inductance **L1** in series with the crystal. With the crystal shorted out L1 will resonate with **C1,** C2, C3 at or near the crystal overtone frequency when the oscillator is operating as an overtone circuit. When the short circuit across the crystal is removed, the crystal will operate in its overtone mode. If **L1** is tuned to a much lower frequency, the crystal will operate in its fundamental mode.

The overtone circuit appears to operate as a crystal-locked Clapp oscillator; with the crystal shorted, it becomes a Clapp oscillator.

B.E.G. Goodger, **ZL2RP**

impedance of a random-length antenna

Several parameters must be established to provide input for this calculation. There are the wire size and its height above ground, from which the characteristic impedance of the wire as a length of transmission line parallel to the ground is computed. The height above ground must also be expressed in terms of wavelength to establish the radiation resistance of a typical half-wave dipole. This information is found in many texts from Terman, (Radio Engineers' Handbook) to publications such as the ARRL Antenna Book or Handbook.

A change from one band to another will have a major effect on this last parameter. So as a starting point, to illustrate the method, I decided to consider a No. **12 (2.1-mm)** wire, **37.5** feet (**11.4** meters) above ground, which yields the convenient characteristic impedance of **600** ohms. The 20-meter band was chosen, where the height of **0.57** wavelength results in a radiation resistance to a half-wave dipole of 68 ohms. This number was divided by two, assigning **34** ohms to each quarter wavelength.

After all these preliminaries we have two numbers: 600 ohms and 34 ohms as the wire characteristic impedance; and we have the radiation resistance of a quarter wavelength (68 ohms). We now shift our attention to the Smith chart (fig. **3**). If the length of our wire is zero, its impedance must be infinite. This is plotted on the Smith chart as point **0**. The 34-ohms radiation resistance of a quarter wavelength, when normalized to 600 ohms, is **0.05666,** which was rounded off to **0.057** and plotted on the real axis of the Smith chart as point **1.** Points between zero and a quarter wavelength lie on a spiral connecting these two points, which, for simplicity, was approximated by a semicircle centered on the real axis and passing through those two points lying in the left-hand, or capacitive, side of the chart.

The VSWR of point **1** is the inverse of **0.05666,** or **17.65.** From this number, the reflection coefficient, Γ, is computed as **0.893.***

The wire length is now increased to two quarter wavelengths. The reflection coefficient is now the second power of **0.893** or **0.797,** corresponding to a VSWR of **8.85.** This is plotted as point 2, and is connected to point 1 by a semicircle, centered on the real axis as before, but this time lying in the right-hand half of the chart because it is inductive.

In a like manner, successive values for Γ are computed, as the length of wire is increased by successive quarter wavelength additions, and connected by semicircles, as before.

Although this method is only an approximation, it does afford considerable insight into the characteristics of a long- or random-length wire antenna. For example, suppose you're considering erecting a full-wave antenna fed a quarter wavelength from one end. The quarter-wave end section will have a radiation resistance of about 34 ohms, while the three-quarter-wave end section will present a radiation resistance of about **100** ohms. Your chances of balancing your feed system to prevent **feedline** radiation have just gone out the window! It will still radiate effectively, but the opportunity for complications is enhanced.

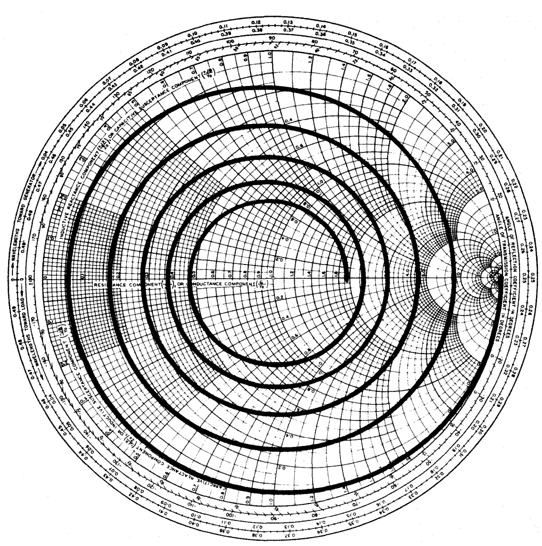
Using a chart of this type is simple enough. You might become confused with the markings of wavelength on the circumference of the Smith chart if you're not careful. Suppose, for example that you want to make an educated guess about this wire at a length of, say, **1.2** wavelengths. This point would lie between four and five quarter wavelengths and would be located by a radius from the center of the chart to where the inner scale reads **0.05** wavelength. Unfortunately, our starting point (0) is marked **0.25** wavelength, rather than **0,** and you must be aware of the possible foul-up. A straight edge marking out this radius intersects the spiral at about **0.296-j0.296.** Multiplying these values by **600** ohms gives **177.6-j177.6** as our estimate of what we would have to match to load such an antenna. All of this may seem merely academic, but it should put us in the ball park when it comes to designing a matching network.

Henry S. Keen, W5TRS

$$\Gamma = \frac{VSWR - 1}{VSWR + 1} \qquad (1)$$

$$Z_0 = 138 \log \frac{4h}{d}$$

where h = height above ground
d = diameter of wire



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contents

Observations & Comments

We **receive many** articles dealing with microprocessors and microcomputers. Only a few of these articles, however, are published in ***ham radio.*** Here's the explanation.

We are certainly interested in the digital world. Like it or not, digital technology is here to stay. A recent article in ***Ham Radio HORIZONS*** by Doug Blakeslee, N1RM, pretty well sums it up (*HRH*, October, 1980).

For ***ham radio*** magazine, our policy is to publish computer articles only if they are Amateur-Radio oriented. We don't publish fun-and-games articles. We're interested in articles that put the computer to work in the ***Amateur station.*** These are the kinds of things we're interested in:

1. Dedicated or single-purpose microprocessor-based circuits such as keyers, displays, decoders, and data bases, to name a few.

2. Interface of the popular microcomputer systems, such as the TRS-80, APPLE II, or PET with Amateur Radio stations.

3. Use of newer, more sophisticated programmable calculators to solve Amateur Radio problems.

These criteria are not a change in our policy at ***ham radio*** but rather a clearer statement of our requirements.

Now for the bad news. If your computer or calculator article contains long program or output listings, these listings must be suitable for ***direct reproduction.*** This means that the printouts must be clean and clear enough so that photographic reproduction can be accomplished directly from your printouts. The task of setting your printouts in type is a formidable one; it is expensive, prone to typesetter errors, and requires extensive proofreading. With a long listing (and we have received quite a few), this problem can be really devastating from a publishing standpoint.

A case in point: suppose your article has a listing or output from a thermal printer. Most of these printouts, especially those from portable calculators, are worse than useless for direct photography. The print paper is fugitive; that is, the copy fades with time and exposure to light. If you must use this kind of printout, immediately get a photocopy before the original fades to oblivion. Send the clean, high-contrast copy with your article. ***Do not*** send the thermal printout.

A few tips for handling thermal printouts:

1. Do not allow the printout to be exposed to sunlight.

2. Especially avoid exposure to fluorescent lights.

3. Most paper is treated with polyvinyl alcohol. Do not store the tape in or near other vinyl material.

4. Do not mount printouts with ***any*** adhesive tape. A chemical action causes disastrous results. A recommended adhesive is "Glu Stic"™ from Faber Castell.

5. Try to use printing tape that prints out in black, not blue (publishing cameras have difficulty with blue).

A last tip: ***Please*** make certain you have copied the final program or output. A "bug" can remain hidden for months and become embarrasing for all. Ask a friend to do your program to see if it is "bomb" proof.

These recommendations were suggested by Dave Buren, N2GE, and associate editor Len Anderson.

Authors who keep these tips in mind will stand a better chance of having their work published in ***ham radio.*** There are all sorts of fascinating applications. Send 'em in!

Alf Wilson, **W6NIF**
editor

comments

amateur band intruders

Dear HR:

I view with great alarm the ever increasing intrusions upon the Amateur 20-meter CW band by Russian and Soviet bloc military CW radio stations, in direct violation of the ITU rules and regulations. I have monitored them on many occasions. These stations use CW with the additional Russian characters, and their traffic is transmitted in 5-character random groups normally associated with cyphering. In addition, the operators use international Q and Z signals reserved for military use.

The transmitters I have monitored exhibit the typical chirp/drift signals usually associated with Russian transmitters. I have found the intruding signals originate on a true bearing between 010 and 030 degrees from my station, signal strength is between S-5 and S-7, and the frequencies are usually between 14,060 and 14,095 kHz.

I hope this letter will make more Radio Amateurs aware of this important problem, and that the FCC and our ITU representatives will be successful in preventing further illegal use of the Amateur bands.

Carl Spikes, W5SAD
Gulfport, Mississippi

sealing coaxial connectors

Dear HR:

Regarding the short article, "Sealing Coaxial Connectors," on page 64 of the March, 1980, issue of ***ham radio,*** I agree with Mr. Wheaton that silicone seal doesn't work well, but PVC electrical tape is by no means adequate either.

Here in Oregon, where weather is quite wet during the winter and cold temperatures with snow and ice prevail in the mountains (where I live), better methods must be found. Also, coaxial cable tends to "breathe" from warm to cold weather and draws air into itself, including any moisture in the air.

The best way I have seen of preventing this is to use rubber, self-vulcanizing ***Electrical Splicing Tape.*** A good seal is provided with one coat of tape; over this, to protect it from sunlight, should be a layer of PVC electrical tape (the rubber tape will decay if exposed to sunlight). Cracking takes one to two years, so allows plenty of time for annual antenna maintenance (which should be done anyway).

Before installing electrical splicing tape, stretch it to 1½ to 2 times its original length. Then wrap the ***entire*** coaxial fitting, leaving no gaps or open spaces. In winter, cover the wrapped connector with your hand for three to four minutes to warm it and initiate vulcanizing action (not necessary during summer). Then cover the rubber tape with one layer of PVC electrical tape.

Working as a radioman here in Oregon, I have radio base stations in some of the worst places for weather this side of Alaska! At one site in northern California, I have two antennas treated with this tape. Conditions are such in winter that winds as high as 70-80 mph prevail, with ice as much as six inches thick on the tower and coaxial feedlines. Inspection during the summer shows only minor contamination of the coaxial fittings, and this can be quickly cleaned out with ***Print Coat Solvent.***

Jim Foster, K7ZFG
Klamath Falls, Oregon

auto-product detection

Dear HR:

I was very interested in K4UD's auto-product detection article ***(ham radio,*** March 1980). Some six or seven years ago I supervised a student project on DSB at Southall College of Technology. We used an MC1496 as the squarer and regenerated the double frequency carrier with a 567 phase locked loop. We obtained excellent results. One operating hint for anyone who is prepared to transmit DSB is not to suppress the carrier too well. If you only reduce it to about 20-25 dB below peak envelope power it will help keep a PLL receiver in lock during modulation pauses.

DSB certainly simplifies the design of transmitters — both in the areas of frequency stability and complexity — and, I think could have considerable application in VHF/UHF hand portables. (NBFM is wasteful of transmitter battery power as there is a full drain on the battery as soon as the press-to-talk switch is operated.

By the way, it isn't too difficult to modify old-style AM transmitters with parallel output tubes (like the DX100) for DSB and perhaps give them a new lease of life!

Joe Hill, G3JIP
Gerrards Cross SL9 8NS, England

long transmission lines for optimum antenna location

This article is **for the Amateur** who has located the ideai antenna site, but finds that it is too far from the transmitter to be reached in a technically acceptable fashion with coax-cable transmission line. An ideal site, of course, is that part of your property that slopes downward in all directions.

What is "technically acceptable?" Let's assume you have a three-element Yagi with traps that permit operation on 20, 15, and 10 meters. The Yagi has a nominal gain of 8 dB on these bands. As the coaxial transmission line is made longer, the antenna-system gain (antenna plus line) becomes lower. At 30 MHz, RG-8/U line, for example, has a l-dB loss[1] for every 100 feet (30.5 meters) neglecting losses caused by standing waves (that is, standing-wave ratios greater than 1). If the ideal site is 500 feet (153 meters) from the operating position, a transmission loss of at least 5 dB car: be expected. **This** leaves an antenna-system gain of 3 dB.* A 5-dB loss would be technically unacceptable.

open-wire line

The solution would be either to locate the transmitter at the antenna site or to reduce the transmission line losses substantially by using an open-wire line, which has an attenuation of 0.1 dB per 100 feet (30.5 meters). Thus the antenna could be removed 1000 feet (305 meters) from the transmitter with the same loss as one fed by coax cable located 100 feet (30.5 meters) from the transmitter.

When discussing open-wire lines, one immediately thinks of a two-wire line that can be constructed with 2-inch (51-mm) to 6-inch (152-mm) spreaders using wire sizes of No. 8 to 22.[3] With the various combinations permitted, line characteristics of 325-800 ohms can be constructed. However, 325 ohms impedance is higher than desired because, ultimately, the line must match a 50-ohm output impedance from the transmitter and probably a 50-ohm input impedance to the antenna.

the four-wire line

Although commercial high-frequency communicators have used four-wire transmission lines extensively, little use of them has been made by Amateurs when open wire lines are needed. Their use in transmission-line runs, however, provides considerably lower characteristic impedances. A 200-ohm line using four No. 14 (1.6-mm) wires on a 0.9-inch (23-mm) diameter can be easily made.4 This type of balanced feeder has been extensively applied where feeder lengths exceed one-half mile (0.8 km). Its relatively low impedance makes this type less susceptible to the irregularities introduced by insulators and switching arrangements. It has high-power transmission capacity for the amount of copper used, and its attenuation can be less than that of two wire feeders.

Four-wire lines may be either side connected or cross-connected, and such connections are made at both ends of the line. A common arrangement of the four wires provides a square when looking at the cross section of the line (**fig. 1**). Side-connection is shown in **fig. 1A**, where the two side wires are connected together vertically at each end of the line. Cross-connection is shown in **fig. 1B**.

Cross-connected lines have a smaller external field than the equivalent side-connected line and therefore have lower pickup when used for receiving.5 This type of line was used extensively at the RCA overseas receiving station, Riverhead, Long Island. In a private communication with Marshall Etter, W2ER, chief engineer of that now inactive installation, I learned that a four-wire line, handled properly, can out perform coaxial lines in terms of reduction of unwanted pickup.

Transmitting loss is not as low in four-wire line as in a two-wire line with large copper conductors, but the loss is probably negligible in relatively short lines, of say 500 feet (153 meters).

By Henry G. Elwell, Jr., N4UH, Route 2, Box 20G, Cleveland, North Carolina 27013

"Reference 2 shows that 3 dB doesn't mean very much under actual operating conditions. Editor.

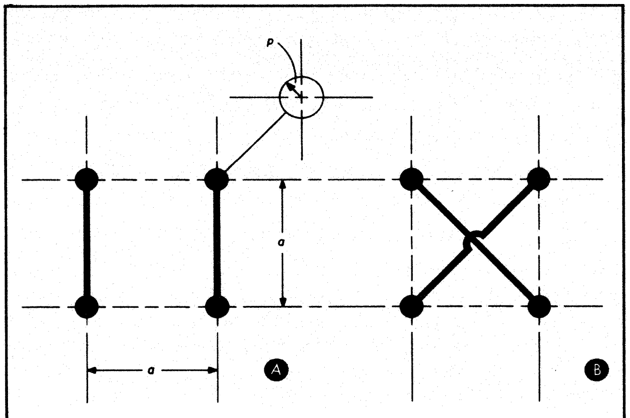


fig. 1. End view of a four-wire transmission line. Side connection is shown in (A) where the two side wires are connected together vertically. Cross connection is shown in (B). The (A) configuration was used in this project for reasons discussed in the text.

The insulation loss in a cross-connected, four-line would be about proportional to the relative characteristic impedances; but since more insulators are required in parallel in its construction than in the side-connected line, overall insulator losses are usually greater. It's therefore not as desirable for transmitting purposes as is the four-wire, side-connected line using the same amount of copper. Its principal use is for receiving, in which its performance is outstanding.

When a square cross section feeder with side-connections is used, the characteristic impedance is equal to that of a pair of two-wire feeders in parallel, each having a spacing equal to the diagonal of the four-wire line. Each diagonal pair is in the neutral plane of the other with no intercoupling. Double power rating is therefore obtained on one set of supports and insulators, and the characteristic impedance is one-half that of one pair.

It's interesting to see the difference in impedance and attenuation between a side-connected and cross-connected line using the same insulators. Consider a four-wire line using the cross section of fig. 1. For the side-connected line, characteristic impedance, Z_0 is calculated from:

$$Z_0 = 138 \log_{10} \left(\frac{a\sqrt{2}}{p} \right) \quad (1)$$

where a is the distance between wires (inches), and p is the radius of the wire (inches).

For the cross-connected line, the characteristic impedance is

$$Z_0 = 138 \log_{10} \left(\frac{a}{p\sqrt{2}} \right) \quad (2)$$

For those wishing to design their own four-wire system, the equation for spacing for cross-connection would be:

$$a = (\sqrt{2})\,(p)\left(10^{\frac{Z_0}{138}}\right) \quad (3)$$

Table 1 shows the characteristic impedances for the two configurations using a spacing, a, of 1.28 inches (32.5 mm) and No. 14 (1.6-mm) bare copper wire, with a radius, p, of 0.032 inch (0.8 mm). These were the constants used in the construction of the four-wire line for this article.

table 1. Comparison of side and cross-connected four-wire transmission line.

configuration	input impedance (ohms)	attenuation: dB/1000 feet (305 meters) copper only	total
side connected	242	1.48	2.04
cross connected	200	1.79	2.48

The losses in a feeder are the sum of copper loss, earth return loss, insulation loss, and loss caused by direct radiation. Radiation loss from a matched feeder is usually so small as to be negligible for carefully designed systems; the loss is always very small with respect to all other losses for almost any type of feeder. Insulation loss in a well-designed system is also a minor quantity, except in long feeders in the high-frequency range. Measurements made on two- and four-wire balanced lines[5] show that, for a 550-ohm, two-wire line, insulation and other losses are about 70 per cent of the copper loss; for a 320-ohm, four-wire line, that number is about 22 per cent at 20 MHz.

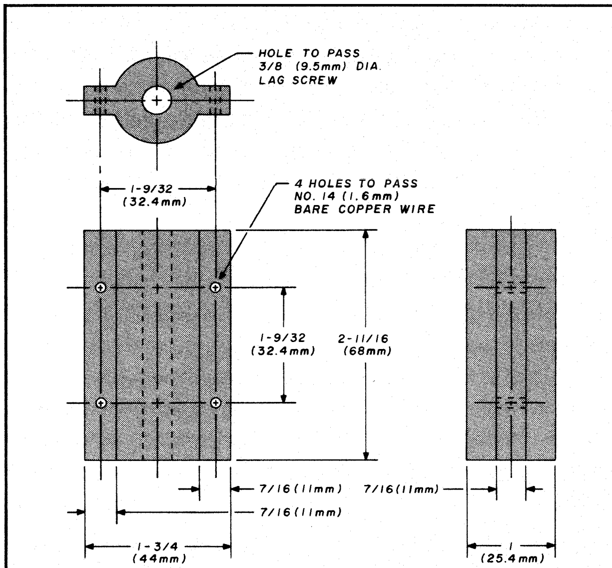


fig. 2. Details of the four-wire insulator, which is made of ceramic material and is secured by a lag bolt.

The attenuation of a four-wire balanced line is:

$$(copper\ loss)\ att = \frac{2.17\ (\sqrt{f\ in\ MHz})}{p\ in\ inches\ (Z_0)} \qquad (4)$$

and the approximate total attenuation for typical construction is:

$$(total\ loss)\ att \cong \frac{3(\sqrt{f\ in\ MHz})}{p\ in\ inches\ (Z_0)} \qquad (5)$$

If p dimensions are in millimeters instead of inches, the constants preceding the radical signs in the numerators of **eqs.** 4 and **5** should be replaced by 53.9 and 74.9 respectively. The attenuation for the two constructions is also shown in **table 1** for 28 MHz.

The side-connected, four-wire transmission line was used in this project because I thought the transmitting characteristics were of greater importance than the receiving ones.

The dimensions of the four-wire insulator used in the project are shown in **fig. 2.*** It's made of a ceramic material called Isolantite and is secured by a galvanized lag bolt. The insulator could also be made from Micarta or Lucite.

selecting the optimum site

In some cases, an Amateur can look at his property and say, "The beam goes on top of the hill." In most cases, the subtlety of the terrain requires that a survey be made of property elevations. By a survey, I mean studying the maps of your area available from the United States Department of the Interior Geological Survey.

using USGA charts

The Geological Survey has a series of standard topographic maps that cover the United States, Puerto Rico, Guam, American Samoa, and the Virgin Islands. Under the plan adopted, the unit of survey is a quadrangle bounded by parallels of latitude and meridans of longitude. Quadrangles covering 7½ minutes of latitude and longitude are published at the scale of 1:24,000. Quadrangles covering 15 minutes of latitude and longitude are published at the scale of 1:62,500.

Each quadrangle is designated by the name of a city, town, or prominent natural feature within it. On the margins of the map are the names of adjoining published quadrangle maps. The maps are printed in three colors. Features such as roads, railroads, cities, and towns (as well as all lettering) are in black ink; water features are in blue, and features of relief, such as hills, mountains, and valleys, are shown by brown contour lines.

The contour interval varies with the scale of the map and the characteristics of the country. On maps that contain supplemental information, additional colors are used, such as green for woodland areas and red for highway classification, urban areas, and U.S. land lines. A booklet describing topographic maps and symbols is available free upon request.

The extent of coverage ot each map is shown on the index map. All quadrangles for which published maps are available have a quadrangle name, publishing agency (if other than the Geologocial Survey), and the date or dates of survey, also printed in black. Further information concerning maps may be obtained from the Map Information Office, Geological Survey, Washington, D.C. 20244.

An inquiry to the above address might request an "Index to Topographic Maps of (name your state)." You'll receive a folder which contains a chart of your state overprinted with all the available quadrangles, identified by name. The folder will also contain a description of other special charts you can purchase. Order the quadrangle featuring your property and perhaps those charts adjoining.

On the Cool Springs, North Carolina, quadrangle that I used, terrain elevations every 10 feet are shown.† All roads and buildings, including individual homes are shown, making it simple to identify the desired property and its boundaries; the scale is 1 inch = 1,000 feet. **Fig.** 3 shows a portion of the Cool Springs, North Carolina, quadrangle in the vicinity of my property.

It's desirable to enlarge the area of interest as much as possible. By making successive enlargements of the chart, I obtained a copy in which 1 inch represented 500 feet **(fig. 4)**. A commercial reproduction office should be able to supply the same service.

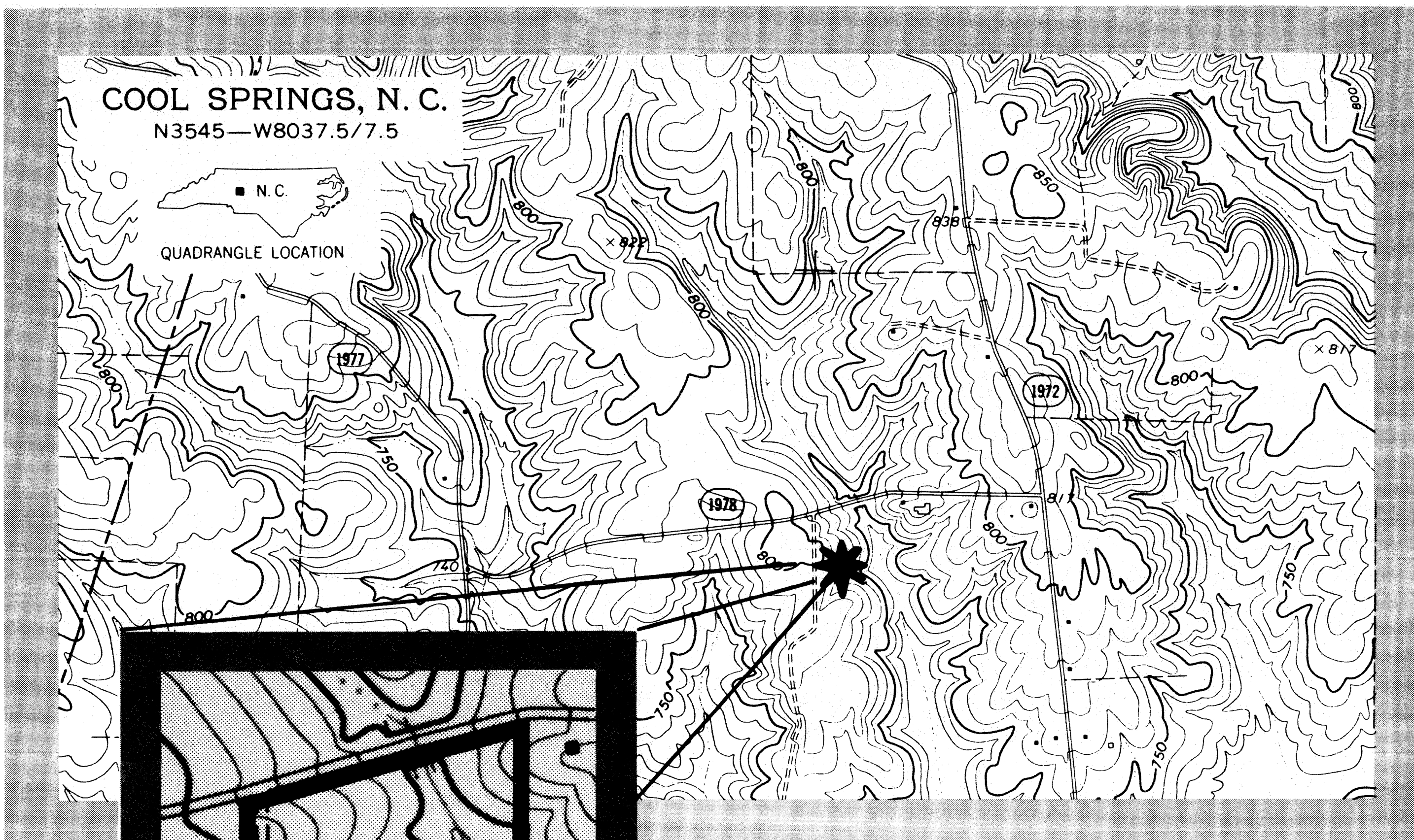
My property boundaries, including the transmitter location, are placed on the topographical map and the best location for the antenna is then determined. It would appear that the best location for my tower would be directly to the northwest at the 800-foot (244-meter) level. Unfortunately a 2,200-volt line on 30-foot (9-meter)-high poles follows the course of the farm road, shown dotted, all the way to the house. I decided to place two towers to the rear of the house on or near the 790-foot (241-meter) elevation. This provided my optimum location so far as terrain as well as distance between antenna and power lines was concerned.

transmission line poles

With the location of the towers established on the topographical map, a property survey drawing on a *1-inch = 200-foot (1 cm = 61 meters)* scale was

*Marshall Etter, W2ER, 16 Fairline Drive, East Quoque, New York 11942, has a limited number of insulators available at $1.00 each plus postage.

†USGS hasn't yet provided charts with metric equivalents. But that will probably change. Editor.



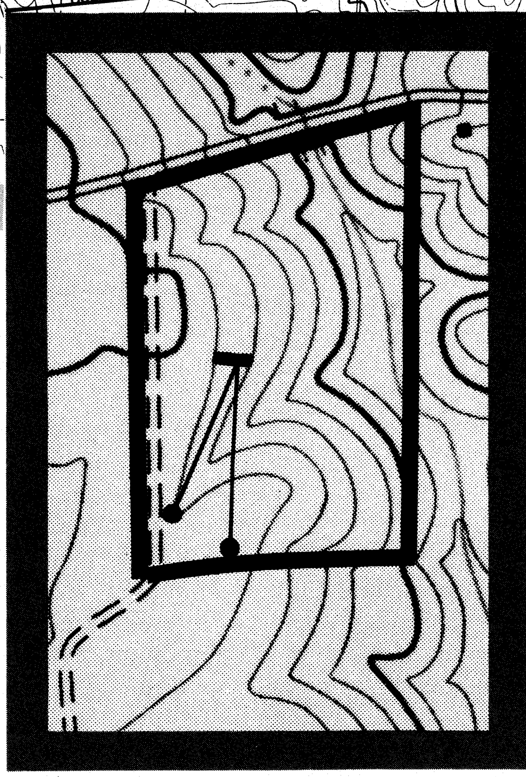


fig. 3 (above). Topographical chart of the author's location (shown by an asterisk). Such charts are available from the U.S. Department of the Interior Geological Survey at nominal cost.

fig. 4 (left). Enlargement of fig. 3 showing author's property boundaries, house location, and tower location. This enlargement helped to determine the best antenna location with respect to local geological factors.

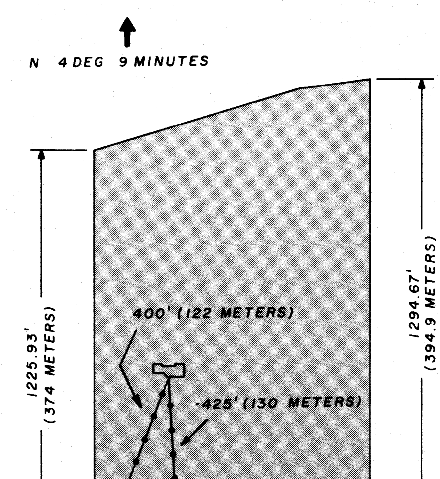


fig. 5 (right). Property survey of author's location showing the relationship between house, transmission line runs, and antenna tower locations. The USGS topographical charts (fig. 3) also include elevation contour lines, which are helpful in determining best location for the antenna towers.

made to obtain distances from house to towers (fig. **5).** Then the number and spacing of transmission line supporting poles were determined. I found that the four-wire transmission lines would have to be 425 feet (130 meters) to one tower and 400 feet (122 meters) to the other tower.

W2ER states that, in early lines, poles were all about 25 feet (7.6 meters) apart, but in later years a staggered spacing from 20-30 feet (6-9 meters) was used to prevent recurring discontinuity at poles from resonating at some certain discrete frequency. To minimize the discontinuities, a copper shield, shown in fig. **6,** was placed to cover three sides of the line at each insulator. (These shields were not used in this project.)

Pole spacing. I decided on 70-foot (21-meter) pole spacing for economy. While not a mistake it required greater wire tension than originally used to prevent twisting under high wind conditions. The additional tension requirement was noted after I traced an antenna system malfunction to a tangle on the four-wire transmission line after a wind storm. The wider separation also produced a fluctuation of SWR readings with heavy wind. From an operational standpoint I've had no noticeable additional problem.

Height. Insulator height above ground was selected

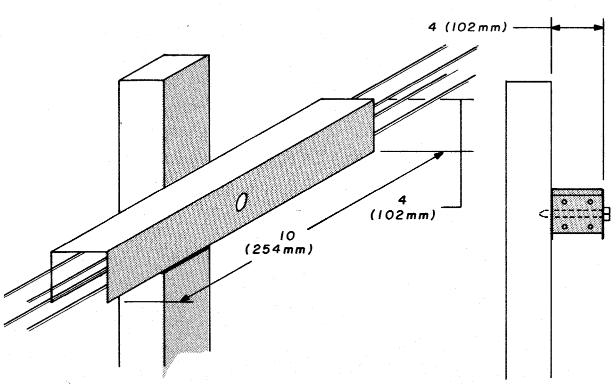
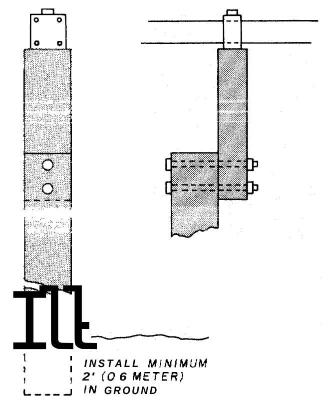


fig. 6. A copper shield may be used at each pole insulator to minimize impedance discontinuities caused by the insulator.

to be 12 feet (3.7 meters) to permit farm equipment to pass underneath. Twelve-foot-high (3.7 meters) treated poles were used (fig. **7**).

Stress. Although there are no stresses on the intermediate poles, other than from wind, there are high stresses on the end pole. This pole should be a single

fig. 7. Typical transmission-line support showing how to increase pole height if required.



quantity	description
1	3/8 in. (9.5 mm) diameter lag bolt
1	four-wire insulator
1	2″ x 4″ x 3' (51 x 102 mm x 0.9 meter) lumber
1	4″ x 4″ x 12' (102 x 102 mm x 3.7 meters) lumber
2	3/8 in. (9.5 mmt machine bolts
4	3/8 in. (9.5 mm) flat washers
2	3/8 in. (9.5 mm) nuts

Install minimum 2 feet (0.6meter) in ground

piece if possible. To maintain the 12-foot (3.7-meter) height, it was necessary to use a 3-foot (0.9-meter) length of lumber cut as shown in fig. **8.** The end pole must be guyed as shown and from the level at which the transmission iine terminates. The objective is to minimize stresses on the pole and have the guy counteract the pull of the transmission line.

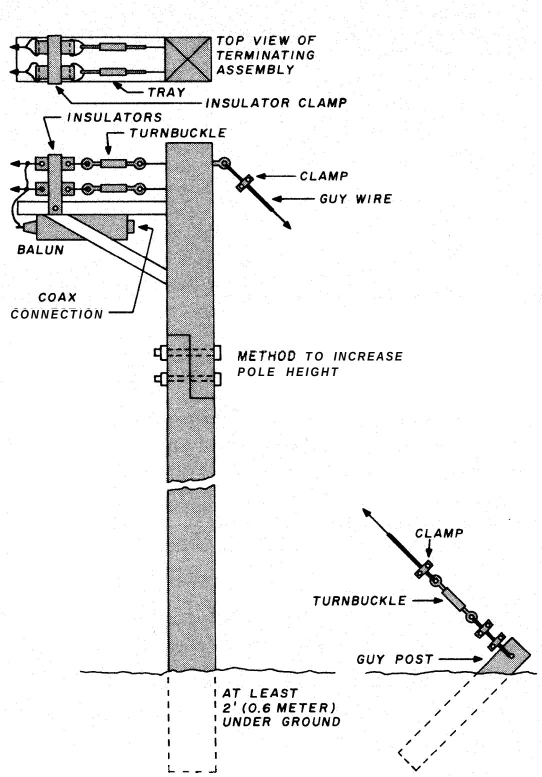
far-end physical termination

Physical termination of the transmission lines is by individual turnbuckles for each wire. Fig. **8** shows the details. Each wire actually terminates on an insulator attached to the turnbuckle by wire. The turnbuckle is connected to a hook on the pole.

The four lines are connected to eye bolts at the transmitter end and threaded through the many four-wire insulators to the terminating post. With the turnbuckle at maximum length, each transmission line is threaded through its end insulator and pulled as tightly as possible **by** hand. When all wires are installed, the turnbuckles are tightened for uniform tension on all wires.

house termination

A %-inch (12.5-mm) lucite panel was used to tie the transmitting end of the line (fig. **9**). Eye bolts were used to terminate the wires. The panel was secured to the house structure by lag bolts. It's very



quantity	description
4	insulators
5	turnbuckles
5	hooks
4	cable clamps
1	14″ x 4″ (36 x 10 cm) wooden tray
1	insulator clamp, homemade
2	sets bolt, flat washer, nut
1	pole to suit application

fig. 8. Construction details of hardware used to terminate the antenna end of the transmission line. The end pole must be guyed as shown to minimize stress on the pole caused by the line.

important to use ***flat*** washers on the panel bolts to distribute the load and to be sure the house structure selected is a primary member; the pulling force at this point is very high.

To maintain proper spacing it's desirable for the wires to leave the termination at *90°* to the panel; this is also the easiest way. If the wires must leave at a smaller angle because of the location of the house and antenna site, some trigonometry must be employed.

Tie-post displacement. Look at **fig. 10A**, which shows a transmission line leaving at *90°* and *45°*. It can be seen that if the wires are connected against the house, distance **A** becomes less as the angle approaches *0°*; that is, at *0°* the wires are touching. It's necessary to move the left support out with respect to the right support to maintain the proper spacing. How do you determine how much to move the one support?

Moving the support "out" can mean in either of two directions. The left wire shown in **fig. 10B** will maintain the l-inch (25.4-mm) spacing at *45°* if the support point is placed at point **A** or point **B.** However, if point **A** is used, the left side of the transmission line will be slightly longer than the right wire. Since it's just as easy to use point **B,** this is the one to be pursued. It's necessary to solve the equation $y = mx + b$, where y is the vertical distance from 0, m is the slope of the wire, x is the distance along the horizontal from 0, and b is the point on the y axis where the left wire crosses it; that is, point **B.**

To simplify the calculation, find x where $y = 0$. Dimension m is known from trig tables since m is the tangent of the angle between the wire and the sup-

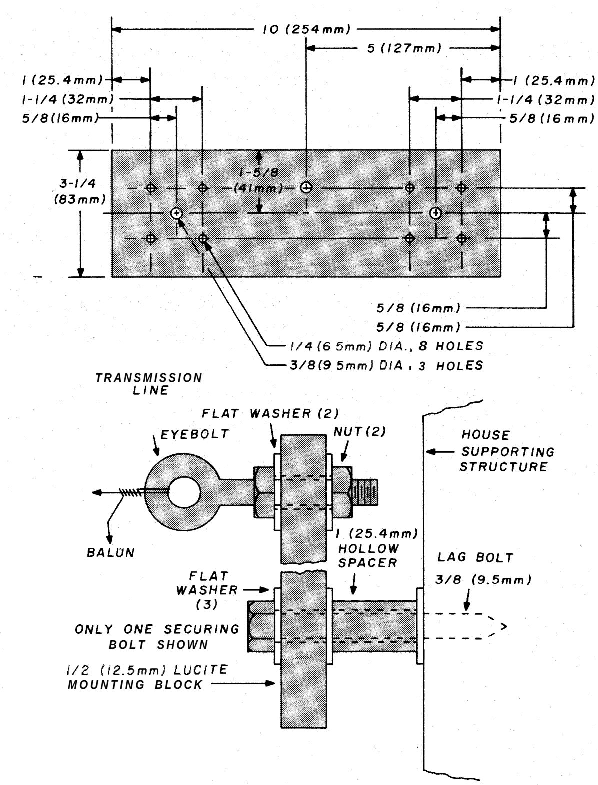


fig. 9. Terminating block at the end of the transmission line made from a ½-inch (12.5-mm) thick piece of Lucite. Upper drawing shows drilling layout for two four-wire lines.

port structure. For the left wire, x is the distance OA. We know the distance OC, which is the desired spacing, and we find CA by solving a right triangle CDA:

$$\sin \angle A = \frac{DC}{AC}$$

Thus $AC = \dfrac{DC}{\sin \angle A}$, and $x = OC - AC$

Therefore we can say that

$$\left(x = OC - \frac{DC}{\sin \angle A}\right)$$

Going back to $y = mx + b$ and substituting m and x at $y = 0$,

$$0 = \tan \angle A \left(OC - \frac{DC}{\sin \angle A}\right) + b, \text{ or}$$

finally: $b = B = -\tan \angle A \; OC - \dfrac{DC}{\sin \angle A}$

Example. Let's say the angle to the antenna from the house is 70°, and a spacing of 1.25 inches (31.8 mm) is required,

$$B = -\tan 70^\circ \left(1.25 - \frac{1.25}{\sin 70^\circ}\right)$$

$$= 0.22 \text{ inch} (5.6 \text{ mm})$$

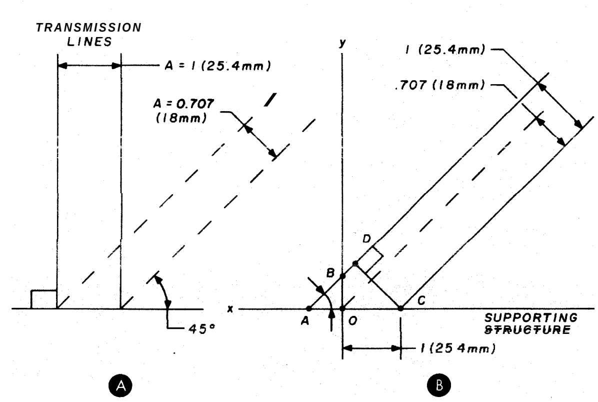


fig. 10. Geometry for determining proper displacement of one tie post when angles less than 90 degrees are used for departure of the line from the transmitting and supporting structure.

To calculate dimension B directly in millimeters, merely substitute the metric equivalent of 1.25 inches, or 31.8 mm, into the above equation. Therefore the eyebolts for the left pair of wires must be displaced 0.22 inch (5.6 mm) farther from the terminating panel than the right pair of wires to maintain proper wire spacing.

To complete the construction of the four-wire transmission line, solder each vertical pair of wires together at both ends to produce the arrangement shown in **fig. 1A**. If the design is for a cross-connected line, diagonal wires would be connected. I used the side-connected arrangement even though the transmission line impedance is 242 ohms for the reason mentioned earlier. Some day I'll revert to the cross-connected arrangement to provide the proper match but am accepting the 1.2:1 mismatch at this time.

impedance transformer — receiving and sending ends

It's safe to say that all popular modern transceivers and linear amplifiers have an output impedance of 50 ohms. Thus a 50-ohm-unbalanced-to-200-ohm balanced balun transformer is necessary: a 4:1 ratio.

These are exciting times for those wanting to make their own baluns. After I constructed the system shown here, Joe Reisert, W1JR,[6] and George Badger, W6TC,[7] described highly efficient baluns in great detail that would be applicable for the design described in this article.

Depending on your degree of purism in such matters, you may settle for a commercial balun; references 6 and 7 spell out the consequences. Their work post dates my effort; I'd already gone commercial, with modifications.

I purchased two kW 4:1 baluns from Caddell Coil Corporation, Putney, Vermont. Each balun coil is wound on two stacked toroids with a total dimension of 2 inches (51 mm) diameter, and 2¼ inches (57 mm) in length. The two windings of the coil have fifteen turns of No. 12 (2.1 mm) wire.

This was my first application of 4:1 baluns, and their characteristics were unknown to me. A test setup as in **fig. 11** was used to measure some of the characteristics of the baluns.

Balun measurements. The station transceiver, a Kenwood TS520, was used for the signal generator, and an Allied Radio model A2516 all-band receiver was used as the null detector. The bridge was a

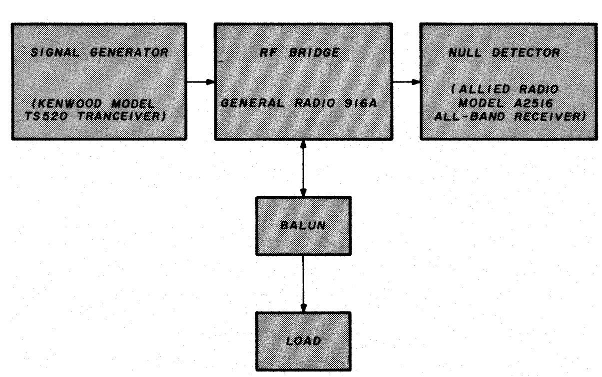


fig. 11. Test setup for measuring balun impedances.

General Radio Model 916A rf bridge. A carbon composition resistor, measured on the bridge at 28.2 MHz, was 250 + j2.8 ohms. This was used as the load.

Impedance measurements were made on five frequencies, from 3.8-28.2 MHz on two of these baluns. Table 2 shows measurement results.

table 2. Balun characteristic versus frequency.

frequency (MHz)	balun 1	balun 2	Hygain balun	Antennabal balun
3.8	56/20°	59.1/18.6°	54.2/21.1°	55.4/23°
7.2	55.8/15°	57/10.8°	60/22.4°	60.6/18.6°
14.2	50.8/18°	55/15°	69.8/17.6°	68/14°
21.2	42.8/39.7°	43.9/38°	79.9/15.6°	74/16°
28.2	55/73°	64/57°	88.3/9.7°	97.0/16°

Based on the 4:1 ratio and a 250-ohm load resistor, the reflected impedance should be 62 ohms,

Having no real feel for the merit of the above readings, I measured a Hy-Gain 1:1 balun with a 50-ohm load, as well as a 1:1 Antennabal device that was on hand. These measurements are also shown in table 2.

The phase angle, indicative of balun reactance, seemed high, even in the commercial units, but far less than the 4:1 baluns to be used. I wasn't interested in the 80- and 40-meter bands for the transmission line, so I removed turns from the existing baluns. A final value of ten turns produced the optimum impedance for the 20-, 15-, and 10-meter bands (table 3).

table 3. Final balun characteristics.

frequency (MHz)	balun 1	balun 2	baluns back-to-back	complete line transmitter to antenna
3.8	60.2/38.7°	56/44.4°	62.6/65°	-----
7.2	69.3/23.6°	69.5/28.6°	61.6/46.6°	-----
14.2	70.2/14.5°	73.5/17.8°	89.7/28.3°	83.4/20.8°
21.2	63.9/17.2°	68/19.8°	88.4/10.2°	39.8/12.7°
28.2	60.4/26.6°	63.7/26.5°	46/0°	94.3/47°

As expected, removing additional turns beyond five improved the 28-MHz readings but caused deterioration of the 14-MHz impedance.

The next step was to connect the two baluns back-to-back and use a 50-ohm resistor for a load. Measurements on the input of the two baluns produced the readings shown in table 3.

I connected a 50-foot (15-meter) length of RG-8/U coax between the transmitter location and the first balun. Then I connected a 100-foot (30.5-meter) piece of RG-8/U coax between the end balun and antenna. Using a 50-ohm resistor in place of the antenna produced readings as shown in table 3. Therefore, the complete 600-foot (183-meter) transmission line provided standing-wave ratios varying between 1.25 and 1.9.

Attenuation characteristics. More important was the attenuation that such a line would produce. Attenuation equations were used[8] based on coaxial cable characteristics. These equations were used, although a hybrid transmission line was being measured. My rationale was that the rf bridge was measuring the short-circuited and open-circuited resistances and reactances of what may be considered a "black box," and didn't care what is in the box. The attenuation of the total line, from transmitter to antenna, is as shown in table 4.

table 4. Transmission-line attenuation.

frequency (MHz)	total line attenuation (dB)	attenuation using four-wire line plus two baluns (dB)
14.2	2.39	1.73
21.2	2.74	1.91
28.2	3.95	2.48

The difference in attenuation between the total line and the four-wire line plus balun is caused by the 150 feet (46 meters) of RG-8/U. W2ER had estimated a I-dB loss per balun without knowledge of the actual loss, and his estimate is good. The baluns most likely account for the loss in the four-wire line. Use of baluns suggested by references 6 and 7 would undoubtedly minimize this loss.

Minimum line loss. Minimum loss would be realized if a tuned line were used and eliminating the baluns altogether. This would require an antenna tuner and would lose the advantage of being able to change bands with minimum tuning. The open-wire could end up in a two-wire polyethylene 200-ohm line for the loop to the rotary antenna.

Another suggestion by W2ER would be to run the four-wire line to the base of the tower, with an exponential line running up the tower, either a two- or four-wire line. The output impedance could be 50 ohms, balanced, and would directly connect to an antenna such as the balanced-input TH6DXX type. With a tower height greater than 30 feet (9 meters) the exponential line would be the minimum permitted length of a half wave on 14 MHz, and of course greater on the 21- and 28-MHz bands; see reference 5.

The use of four-wire lines for the reduction of line losses is wide open for experimenting in the Amateur bands, and the subject is especially pertinent to vhf, where the cost of super coax cable is getting out of hand.

A second tower and transmission line were constructed, as discussed for the first arrangement. Results were almost identical.

results of transmission-line use

The antenna on each tower is a Hy-Gain THGDXX. The antennas were adjusted using the rf bridge with the affects of the total transmission line subtracted from the measurements of the total system. The final SWR readings, as measured at the transmitter end, are acceptable for the three bands: less than 2:1.

In review, two towers separated by 150 feet (46 meters) were located approximately 425 feet (129.6 meters) from the transmitter. The location was selected to provide an antenna site from which elevations in all 360° compass directions decreased and no other structures were close by. I used a four-wire transmission to minimize line loss with 4:1 baluns at each end to accept 50-ohm input and output matching. The two towers are 60 and 50 feet (18 and 15 meters) high.

Results exceeded all my expectations, with no dead spots in any directions. Perhaps a review of my contest activities since erecting the systems is indicative of its success:

1978 ARRL sweepstakes	**first place in North Carolina**
1978 ARRL ten-meter tests	**first place in North Carolina**
1979 ARRL DX contest	**winner, Roanoke Division, high band**
1979 ARRL Radiosport championship	**highest score, North Carolina, phone only**

The ARRL DXCC country total has changed from 285 to 320 confirmed during the period from October 1978 to June 1980.

final comments

The expenditure of time and money has been well worth the effort to get the antenna out where it can do the most good for my signal. It's time more Amateurs with ideal antenna sites not currently in use take advantage of them with a low loss four-wire transmission line.

references

1. *The ARRL Antenna Book,* Table 3-I, *Commonly Used Tranmission Lines,* and Table 3-11, *Standard Coaxial Cables,* American Radio Relay League, Newington, Connecticut, 1964 edition, pages 84 and 85.

2. A. Wilson, W6NIF, "Trap Yagi Antennas," *Ham Radio Horizons,* June, *1980,* page 13.

3. *The ARRL Antenna Book,* Fig. 3-25, *Characteristic Impedance vs. Conductor Size and Spacing for Parallel-Conductor Line,* American Radio Relay League, Newington, Connecticut, 1964 edition, page 81.

4. *op cit,* Fig. 3-26, page 82.

5. Edmund A. Laport, *Radio Antenna Engineering,* Chapter 4, *Radio-Frequency Transmission Lines.*

6. Joe Reisert, W1JR, "Simple and Efficient Broadband Balun," *ham radio,* September, 1978.

7. George Badger, W6TC, "New Class of Coaxial Line Transformers," *ham radio,* February and March, 1980.

8. Leonard H. Anderson, "Antenna Bridge Calculations," *ham radio,* May, 1978.

ham radio

versatile CW identifier

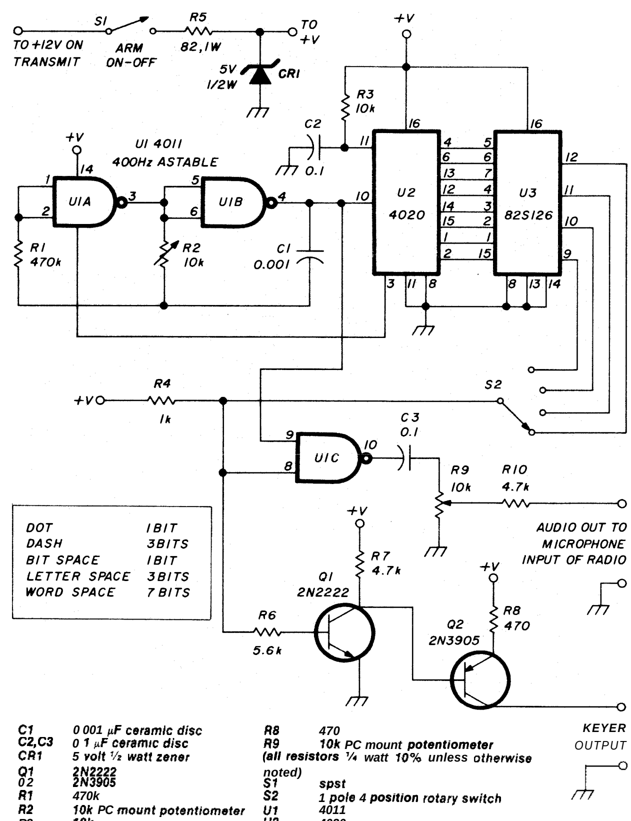
Whether you like CW or not, it's still a useful mode for identifying a station, be it a repeater, RTTY, slow-scan TV, or as an anti-theft device. However, most devices that send preprogrammed CW messages are either bulky, complex, or expensive. I've designed a small, simple, and inexpensive CW identifier that will find many uses in your station.

programming logic

In designing the identifier, I used the standard ratios for sending international Morse code: one bit represents a dot, three bits a dash, one bit an element space, three bits a letter space, and seven bits a word space. As an example, the message "CQ DX" would be represented, in a binary format, as: 11101011101000111011101011100000001110101000 11101010111. My scheme was to program this sequence (or any message) into a PROM by sequentially programming the "ones" at the appropriate addresses. I chose an 828126 PROM, which is a 1K device arranged in four sections of 256 bits. Hence, eight bits are required to address each of the four sections of memory. With 256 bits, an average message will last about 15 seconds at 10 words per minute. **Fig. 1** shows the circuit.

how it works

The identifier plays back the message stored in the PROM by addressing it with eight address lines from a 4020 binary counter driven by an astable oscillator formed from one-half of a 4011. A particular 256-bit section is selected by S2, which serves as a message selector. This output is NANDED with the clock signal, giving an output available at R10 consisting of bursts of tone whenever a "one" is encountered in the PROM memory. Tone level is controlled by R9, so that signal deviation may be adjusted in an fm transmitter. The PROM output also is applied to a keying



C1	0 001 μF ceramic disc	R8	470
C2,C3	0 1 μF ceramic disc	R9	10k PC mount potentiometer
CR1	5 volt ½ watt zener		(all resistors ¼ watt 10% unless otherwise noted)
Q1	2N2222	S1	spst
Q2	2N3905	S2	1 pole 4 position rotary switch
R1	470k	U1	4011
R2	10k PC mount potentiometer	U2	4020
R3	10k	U3	82S126, 82S129, 82S26 or 82S29 (256X 4 PROM — available programmed from author for $6 00 postpaid.)
R4	1k		
R5	82 1 watt		
R6	56k		
R7,R10	4 7k		

Note **Printed-circuit** boards are available from the author for $7 00 postpaid Complete kits with one programmed PROM are also available for $25 00 postpaid

fig. 1. Schematic diagram of the CW identifier. With 256 bits of preprogrammed memory, an average CW message lasts about 15 seconds at a CW speed of ten words per minute.

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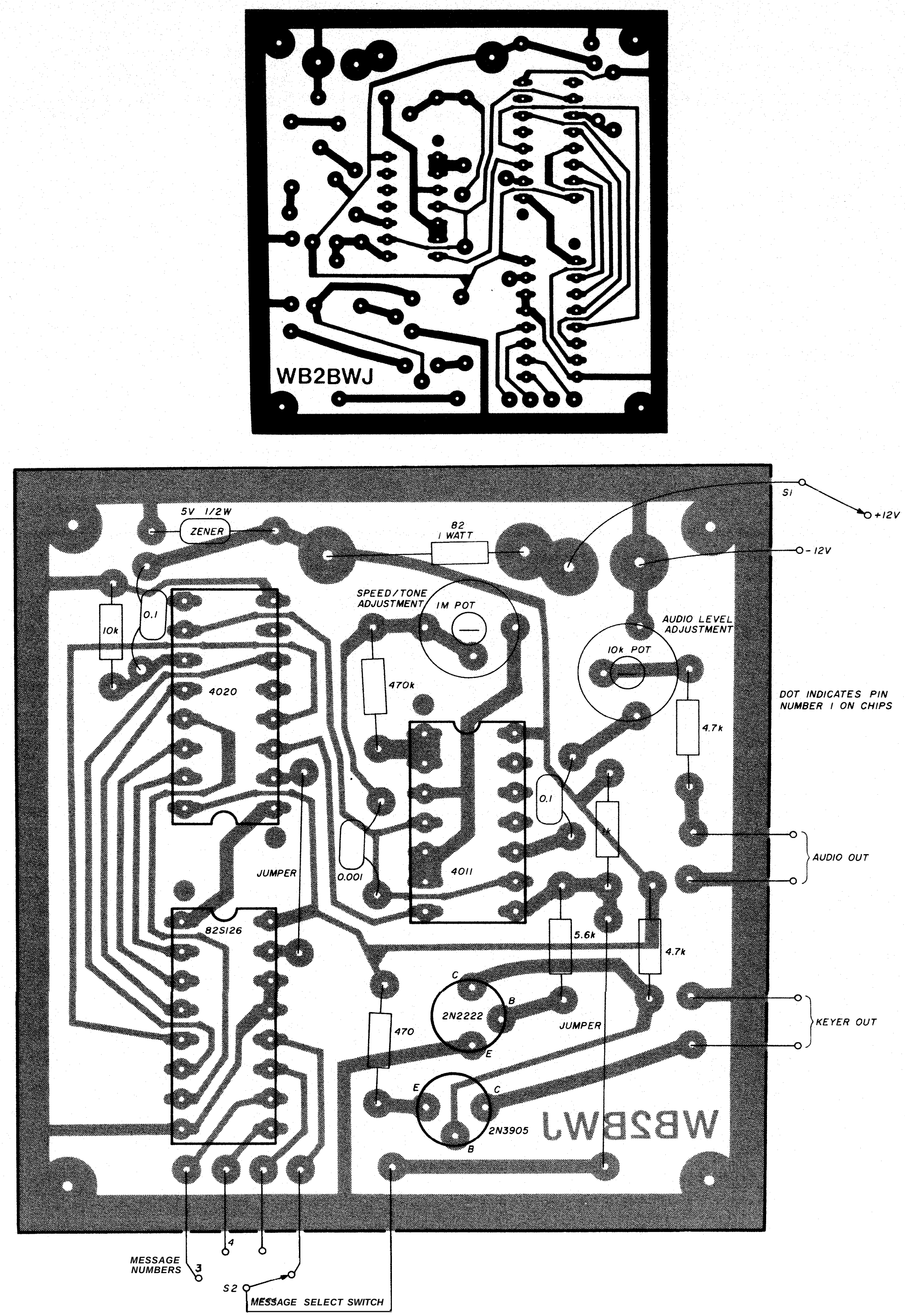

fig. 2. PC board layout and parts placement of the CW identifier. Kits are available (see text).

network composed of Q1, Q2, R6, R7 and R8, which provides an output compatible with the grid-block keying used in most hf transmitters.

Resistor R2 controls the clock rate, which not only determines CW speed but also changes output-frequency tone, since the oscillator does double duty as a clock and a tone generator. The identifier sends the message only once for each S1 closure because the clock is disabled after a count of 256 is reached by the 4020 IC. The 4020 is reset each time power is applied through the differentiator formed by C2 and R3.

The identifier is designed to operate on 12 volts, as

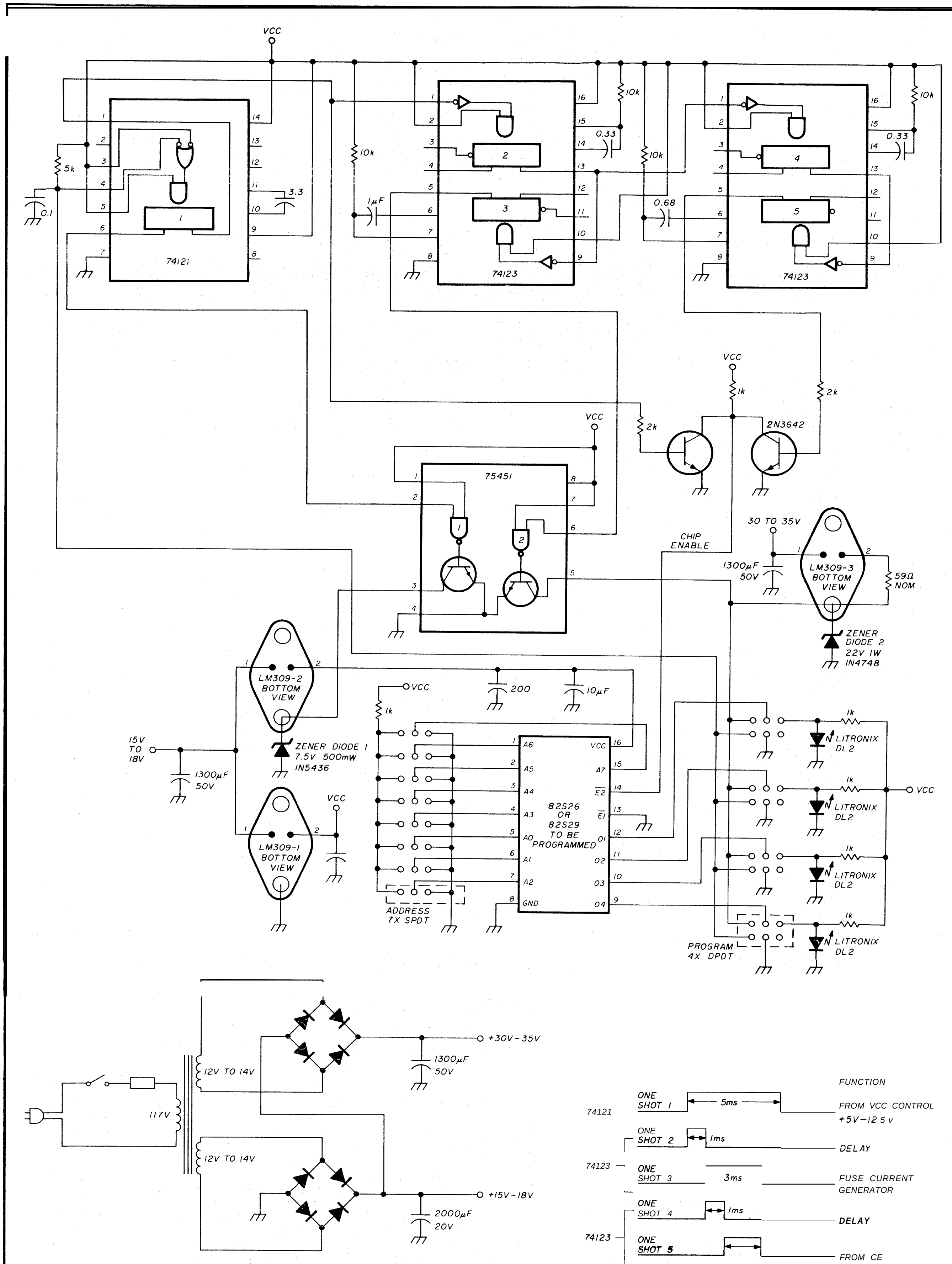

fig. 3. Suggested programmer using the Signetics 82S126 TTL/memory device. (Reprinted courtesy of Signetics, Inc.)

its primary use will be with mobile type equipment. That's why CR1 and R5 are needed to reduce the supply to 5 volts so that the ICs will function properly. If 5 volts are available you can eliminate the regulator network. **Fig. 2** shows the PC-board layout and parts placement.*

Fig, 3 is a schematic diagram of a suggested programmer for the 82S126. (The schematic is reprinted from the Signetics Integrated Circuits Manual.) After the messages have been decided upon, a table of the binary sequence that must be programmed in the PROM should be made. For each "one" in the table, address its location using the eight address switches, then depress the appropriate program switch depending on whether you're programming message 1, **2, 3,** or **4.**

suggested uses and installation

As mentioned, the identifier can be used for repeaters, slow-scan TV, or RTTY identification. Other uses are as a contest keyer, a beacon ID, or an anti-theft device. To use it as an anti-theft device, one of the messages that could be programmed into the identifier to protect your rig, is "STOLEN DE (your call letters)."

Install the unit in your mobile rig, replacing S1 with a magnetic reed switch. Install the switch as closely as possible to the cover of the radio. Connect the +12-volt line to a point of the switched 12 volts within the radio that is available only when in the transmit mode. (This voltage can usually be found on the transmitter strip.) As an alternative, isolate the identifier from ground and connect the ground wire from the identifier to the microphone PTT switch. In this case, connect the +12 volt line to a constant source of 12 volts in the radio. Now, every time the reed switch is closed and you're transmitting, the message will be sent with your voice. The reed switch will close only when a magnet is held close to it. Hence, when you're using the rig, don't put a magnet near the reed switch!

When the radio is unattended leave a magnet on the cover, about where the reed switch is located, and the identifier will activate when the radio transmits, telling all your friends on the repeater that your radio has been stolen (unknown to the felon operating it).

Many of these identifiers are in operation and are performing reliably. The design is simple and effective.

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*As an alternative to building the programmer, I'm offering the benefits of a sophisticated, automated programmer that I've constructed. I'll program any **PROM** that a reader sends me for $1.00 to cover postage and handling. Please be sure that the total number of bits, that is, "ones" and "zeros," in your message doesn't exceed 256. See also my note in fig. 2 regarding PC boards and kits.

three-element switchable quad for 40 meters

Antenna direction can be changed by switching reactances in the parasitic elements

At N8ET I've done quite a lot of serious contest operating. As a result the antenna system became quite extensive — five elements on 10 meters, four on 15, five on 20, dipoles on 40 and 80, and my 30-foot (21-meter) tower was gamma matched for 80 meters with 2000 feet (610 meters) of radials. Forty meters was the weak link as results in the contests verified. To improve my signal on 40, I began to look for better antennas with low-angle radiation. A bobtail was the first antenna to be tried. It took as much space as an 80-meter dipole and was only 33 feet (10 meters) high. Results were encouraging. I almost missed two KL7s because they were so strong that I thought they were locals. The problem was that I still needed something for the European and ZL/VK paths.

the switchable quad

While going through old *QST* articles I came across one about a 40-meter switchable two-element quad.[1] It looked easy enough to erect, and quads are supposed to have excellent low-angle radiation, even when close to the ground. The spacing used in the *QST* article was 22 feet (6.7 meters), with the feedline switched between elements. Since I just happened to have a 40-foot (12-meter) boom, I decided to go with three elements spaced at 20 feet (6 meters) and to switch the parasitic elements only (fig. 1). The results were far better than expected.

As in the *QST* article, I found that the parasitic elements could be tuned by a reactance some distance from the element by using the correct length of feedline between the element and the reactance, The reactance at the end of the feedline is transformed to the proper reactance at the element to tune the element either as a director or a reflector. By using a dpdt relay, I could switch reactances and change from reflector to director and *vice versa.*

construction

All three elements were cut to the same length using the formula for a quad driven element:

$$l = \frac{1005}{f} \tag{1}$$

where l is element length (feet)

f is frequency (MHz)

In metric terms, eq. 1 is

$$l = \frac{306.6}{f}$$

where l is element length (meters)

f is frequency (MHz)

Using a Smith chart, I determined the correct reactance at the end of a 450-ohm open-wire feedline to make a reflector and a director (about ± 150 ohms of reactance is needed at the element). A variable capacitor with a maximum value of 150 pF easily provides the reactance at the end of a 16-foot (4.9-meter) section of the open-wire line. The line can be any multiple of ½ wavelength long plus 16 feet (4.9 meters). This allows the reactances to be placed at the operating position for optimum adjustment of any signal.

adjustment

I erected the quad at a height of about 60 feet (18

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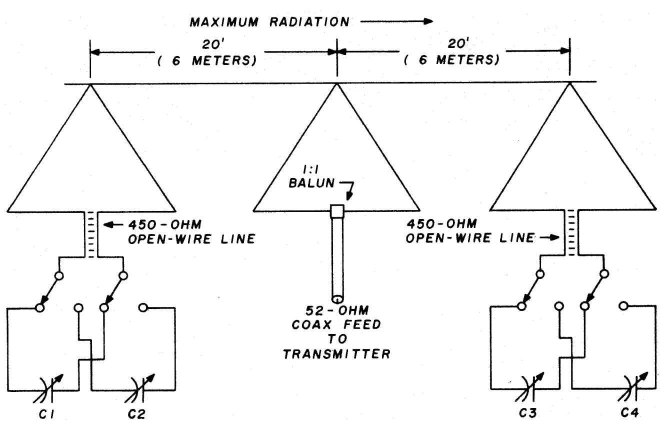
meters). Each element was in the form of a triangle with the peak at the top. The driven element was fed at the bottom center using a 1:1 balun and 52-ohm coax. This element could have been fed directly, or through a wire gamma match, or even with open wire and a tuner. Don't make the same mistake I did by erecting and tuning the driven element and then erecting the parasitic elements! As the books say, they do interact, and I had to resolder all the wire I'd just cut out of the driven element to get a good match. Put up all three elements then adjust the driven-element length for the best match.

Next, adjust the antenna by bringing the receiver out to the antenna base, where you can see the *S*-meter. A distant station in the appropriate direction can be used as a signal source to adjust the parasitic elements. If a nearby station is used, be sure he's using horizontal polarization. (During my initial check with station AD8P, I had the most gain off the back of the quad until AD8P switched from his vertical to his dipole. Then we obtained the expected results.)

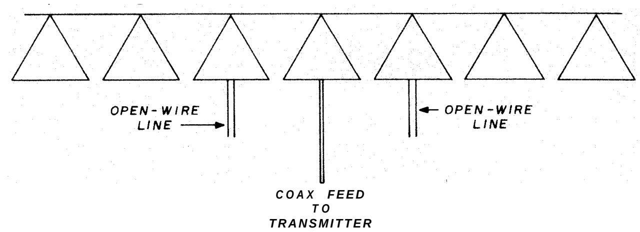
Adjust the capacitors for best front-to-back ratio. This is a fairly critical adjustment (fig. 1) and corresponds very cosely to the point of maximum forward gain. The forward-gain adjustment is quite broad, so it's ***much better*** to adjust for maximum front-to-back ratio and accept the very small loss in forward gain.

The ratio I finally obtained was about 25-30 dB. I was able to get about 50 dB front-to-back ratio using a nearby oscillator as a signal source. This obviously was not right, so be very careful about using signal sources close to the antenna for adjustments. Forward gain over a dipole was not measured, but with the quad I've heard and worked stations I'd never



C1 350 pF, adjust for reflector operation
C2 350 pF, adjust for director operation
C3 350 pF, adjust for director operation
C4 350 pF, adjust for reflector operation

fig. 1. Three-element quad adapted from data in reference 1. The parasitic elements can be switched from director to reflector or vice versa.



Directors:	**elements** 1,2,6,7
Driven Element:	**4**
Switchable director/ **reflector:**	3,5

fig. 2. Proposed switchable quad using seven elements. Mechanical construction would pose a problem for a 40-meter array. Perhaps the antenna could be strung between several high trees. See reference 2 for ideas in erecting long antennas.

heard on 40 meters before with my dipole, which is up 60 feet (18 meters).

alternative supports

For those who don't have a 70-foot (21-meter) high tower and a 40-foot (12-meter) boom, there are other ways to get the quad in the air. All you need are two supports at least 40 feet (12 meters) high, spaced at least 40 feet (12 meters) apart in the right direction. The supports could probably even be less than 40 feet (12 meters) high, The triangular loops can be flattened to accommodate the lack of height. The loops don't have to be a perfect triangle. Of course, the higher the antenna, the better it will perform. Instead of a 40-foot (12-meter) boom, a piece of line between supports can be used to hold up the elements.

future work

I hope to try another configuration, which will give more forward gain, by using more than three elements, feeding the center element, switching only the two closest parasitic elements, and making all the remaining parasitic elements as directors, fig. **2.** The reflector should isolate the appropriate directors so they have a minimum effect on antenna performance.

I would be interested in hearing from anyone who tries this configuration since I am now living in a house with a very small lot and won't be able to try this idea for some time. Perhaps all the elements will have to be switched.

references

1. John E. Kaufmann, WA1CQW and Gary E. Kopec, WA8WNU, "A Convenient Stub-Tuning System for Quad Antennas," ***QST,*** May, **1975,** pages 18-21.

2. G. E. Smith, W4AEO, and Paul A. Scholz. W6PYK, "**Log-Periodic** Fixed-Wire Beams for 40 Meters," ham ***radio,*** April, 1980, pages 26-30.

ham radio

Yagi antenna design: ground or earth effects

A discussion of ground effects and performance of real Yagi antennas above real ground

To this point, the antennas I've discussed were considered in free-space conditions. In the real world, however, the performance of an antenna is profoundly affected by the proximity of the ground or earth. The ground can be beneficial in some circumstances and detrimental in others; it is important to understand ground effects and how to use the presence of ground advantageously.

Unfortunately, the ground or earth near an antenna is not easily characterized; ordinarily, it is complex in shape with highly variable radio-frequency properties. The ideal "ground" will consist of a flat plane of material with high electrical conductivity, usually approximated quite well by the surface of the ocean. Actual ground, however, has a much lower conductivity and lower dielectric constant than salt water and is ordinarily far from flat. Values of electrical conductivity, a, range from about **4** mhoslmeter for salt water to as low as 10^{-3} mhoslmeter for typical residential areas; corresponding values of dielectric constant, ε, range from about 80 for water to **5** for typical residential areas.

The presence of ground alters antenna performance in two ways. First, it changes the normal **free**-space pattern (in the illuminated half sphere) by adding (vectorially) a ***reflected*** pattern. The combination of direct and reflected radiation fields produces regions of enhanced gain, but at the expense of reduced gain in other regions. Second, it alters the antenna itself; currents that flow in the ground surface couple back into the antenna and change all element currents. In other words, an antenna becomes a somewhat different antenna when placed near the ground.

reflections from a plane ground

Let us first consider the reflection from a plane ground surface: the reflected wave can be characterized by a reflected amplitude, K, and a reflected phase, ϕ. For horizontally polarized radiation, K_H remains close to unity and ϕ_H **is** close to a or **180°** for all incoming (and therefore reflected) angles with respect to the surface, β. The phase change is due to the fact that if the conductivity of the surface is high, Maxwell's equations require the tangential E-field at the surface to vanish; this can only occur if K_H is nearly unity and ϕ is nearly ***n.*** The limiting values only occur if the conductivity of the plane surface is infinite; however, for all practical values of ground conductivity, K_H and ϕ_H remain reasonably close to the limiting values.

By contrast, the reflection of vertically polarized radiation is much more complicated. In this case, the limiting value of K_V (for infinite conductivity where the electric field normal to the surface must be continuous) is also unity and that of ϕ_V is zero. However, these values change drastically where ground has a finite conductivity and dielectric constant. Values of K_V and ϕ_V vary radically with the elevation angle relative to the surface, β. As the surface, β, is changed from zero to $\pi/2$ or 90 degrees, K_V from unity down through a minimum at an angle called Brewster's angle and slowly back up to a value usually significantly less than unity. At the same time, ϕ_V **varies** from a or 180 degrees monotonically down to zero. Brewster's angle, β_B, is a function of the dielectric constant of the surface, $\beta_B = \cot^{-1}\sqrt{\epsilon}$. Note that for water $\beta_B = 6$ ***degrees*** (already quite a low angle) and if $\epsilon = 5$ (**poor** ground), β_B rises to only **24** degrees. Thus, it is easy to see that if one models flat real ground as an infinitely conducting plane, the result should be generally trusted for horizontal polarization, but not for vertical polarization. It is significant that far-field radiation for vertical polariation over real

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(finite conductivity) ground vanishes at grazing angles. Only when conductivity is truly infinite does grazing angle reflection add to direct antenna radiation.

Let me return briefly to the observation that the ground surface near an antenna is rarely flat. Even if this real ground surface were perfectly reflecting (infinite conductivity), the pattern of reflected radiation would be exceedingly complicated. It is useful to imagine a small optical model of the ground surface terrain made with a good shiny reflecting surface and illuminated by a light source (replacing the antenna). From this it is easy to understand that the reflected pattern will show bright spots or glints where converging rays are focused from "dished" concave surfaces; in fact, the brightness from such dished areas is potentially much higher than the direct light from the light source itself. if the dished surface is ideal, one can realize enormous brightness gains; the ideal surface is, of course, a parabolic reflector at the right distance from the source.

Excellent optical searchlights are made with such reflectors. A modern radiofrequency example is the large Arecibo dish, which started with a natural, roughly parabolic ground surface and was subsequently improved by high-conductivity, accurately figured reflecting surfaces. To get effective cohesive radiation from the entire surface required the surface to maintain an accuracy of a fraction of a wavelength.

The shiny optical model produces a reflected pattern somewhat hut not exactly, like the real ground. The principal difference is due to the very short wavelength of light compared with the size of the surface features of the model. The brightness gain of any well-configured dished area is theoretically proportional to the reflecting (concave) area, but inversely proportional to λ^2. Thus, the optical glints will tend to be small, bright spots, whereas the real radio glints, occurring at the same angles, will be fuzzy, larger, and less bright. Note, however, that the broader radio glint will approximate the average value of the optical glint over the radio glint angle.

Thus, we see that ground around an antenna can not only reflect, but it can focus (and defocus) radiation as well. To get much of a focusing effect, the concave surface must not only be large (compared with a wavelength) and of the correct focal length, but it must also have a surface accuracy of a fraction of a wavelength. This type of surface is not likely to occur naturally at wavelengths in the 10-to-100 meter range, especially if the antenna is situated in a region where the reflecting surface is quite complicated. Nevertheless, for those users fortunate enough to be able to locate their antennas on a relatively smooth, concave slope of the right curvature, some significant focusing should take place.

ground model

I shall model the ground surface in the conventional way; that is, as an equivalent to a perfectly conducting flat plane. This model should be valid for horizontally polarized radiation at antenna sites in which the actuai ground is reasonably flat out to distances where specular reflection occurs at the lowest elevation angles of interest. For high angles, the reflection takes place nearly under the antenna, and the ground must be flat in that area to a fraction of a wavelength for the model to apply; this is usually the case.

As the elevation angle is reduced, the reflection point recedes from the antenna location until, at very low angles, it is many wavelengths distant. Under this condition the real ground does not have to be very flat to reflect energy with amplitude and phase coherence; it can in fact be quite rough, with variations in height of several wavelengths. This situation is analogous to the well-known optical reflection observed from surfaces that are rough in comparison with a wavelength; one can observe at grazing angles nearly specular reflection. A sheet of paper has roughness variations so large compared with the wavelength of light that, at normal incidence, no reflected images can be seen; nevertheless, at grazing angles, one can easily observe specular reflection effects – and even fair images. For these reasons, the model can be expected to be fairly valid for most horizontally polarized antenna systems.

One more point should be mentioned. Because of the finite conductivity of the real ground, the currents which flow are not strictly at the physical surface of the ground but are distributed throughout the top "skin depth" of the ground. This skin depth is usually quite small; as an example, at a radio frequency of 14 MHz, where the free-space wavelength is 21 meters, this skin depth in salt water is less than 0.1 meter, and even for poor ground, where the conductivity is 10^{-3} mhos/meter, the skin depth increases to a value of less than one meter. Therefore, the infinite conductivity plane model of the ground should give quite acceptable results.

Besides adding a reflected wave to the space pattern of an antenna, the presence of the highly conducting ground plane changes the properties of the antenna itself. Excitation of the antenna produces a current distribution in the nearby ground plane surface, which in turn couples mutual voltages back into all antenna elements; these mutual voltages will obviously affect antenna currents in all of its elements.

To model this interaction, it is useful to replace the ground plane conducting surface by an antenna image. This image is located just as far below the original ground plane as the real antenna is located above. For a horizontally polarized antenna, the image is excited equally with, but exactly out of phase with, the real antenna. Because of the geometrical symmetry of the antenna and its image around the original ground plane surface and the opposite excitation, it is easy to see that, at all points on the original ground plane surface, the tangential electric fields vanish (image field cancels the antenna field).

Thus, by means of the image model, we produce exactly the same tangential field at the ground plane coordinates as would be produced by currents flowing in the real conducting ground plane because of antenna excitation alone. The antenna itself cannot distinguish whether a real ground plane conductor or its oppositely excited image exists. Therefore, the ground interaction with the antenna is identical to the image interaction. One can therefore model all antenna properties over the real ground plane by using the antenna and its oppositely excited image in free space.

Note, however, that there is one significant difference between the image model in free space and the real ground plane. In the real ground plane case only a hemisphere is **actually** irradiated, while in the image model a full sphere is irradiated. Although all fields in the (common) half space will be exactly the same, it is obvious that the total radiated power of the image model will be just twice that of the real ground plane. Therefore, even though antenna element impedances are the same for both situations, gain calculations for the image model must be multiplied by two to get gain for the real antenna over earth.

antenna over earth.

Before I evaluate detailed antenna properties over earth, I would like to briefly discuss the elevation angles that should be of paramount interest. It is well known that long-distance radio communications take place primarily by ionospheric F_2 layer reflection (or, more properly, refraction). While the F_2 layer can vary in (virtual) altitude over the earth (250 to 400 km), it is instructive to make a very simple model of this layer as a reflecting shell at an altitude or height, H, of 300 km. A radiated wave at an elevation angle, β, will bounce from this shell and return to earth at the same elevation angle, β, but at reat circle range, R. The relationship of R to β depends only on simple geometry. For a single hop the maximum range on the earth is limited; thus, communications at very long distances will involve several hops.

Fig. 1 shows a plot of β versus R for different numbers of hops (up to 6), n. This diagram shows clearly that a given range can be reached with different discrete elevation angles, or that a given elevation angle arrives back at the earth at discrete ranges. To cover all values of range requires a continuous spread in elevation angle β, but the limits of this spread in β can be narrowed somewhat by taking advantage of different numbers of hops. As an example, ail ranges beyond $R = 1600\ km$ can be accommodated by the heavy line in **fig. 1**; that is, β varying over a range of only 3 to 17 degrees.

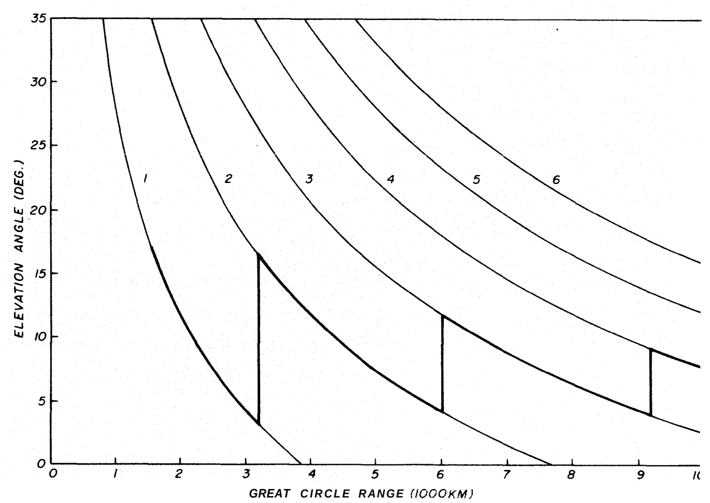

fig. 1. Diagram of the elevation angle required for propagation over various great circle distances and number of hops.

It is generally desirable at the higher frequencies to use low angles to minimize attenuation resulting from multiple hops and reflection losses and to ensure ionospheric refraction at the highest frequency. For such frequencies (say 14 to 28 MHz) the range of elevation angles shown in **fig. 1** seems quite appropriate. At lower frequencies, (≤ 7 MHz), however, if the propagation path at either end is in daylight (where absorption is high), a higher range of angles (using a greater number of hops) may give a lower overall absorption. The reasoning derived from this oversimplified model gives expected results not inconsistent with observations reported in the ARRL Antenna *Book*[1] on page 18. However, real propagation is clearly more complicated than is shown in **fig. 1**.

Kift[2] has shown in an elegant way that long distance propagation involves many propagation modes. He has shown, in measurements made between Ascension Island and Slough, England, that measured arrival angles, when complete path ionospheric soundings are known, correlate well with ray-tracing expectations. (Ray tracing can identify actual propagation modes.) His results indicate that eleva-

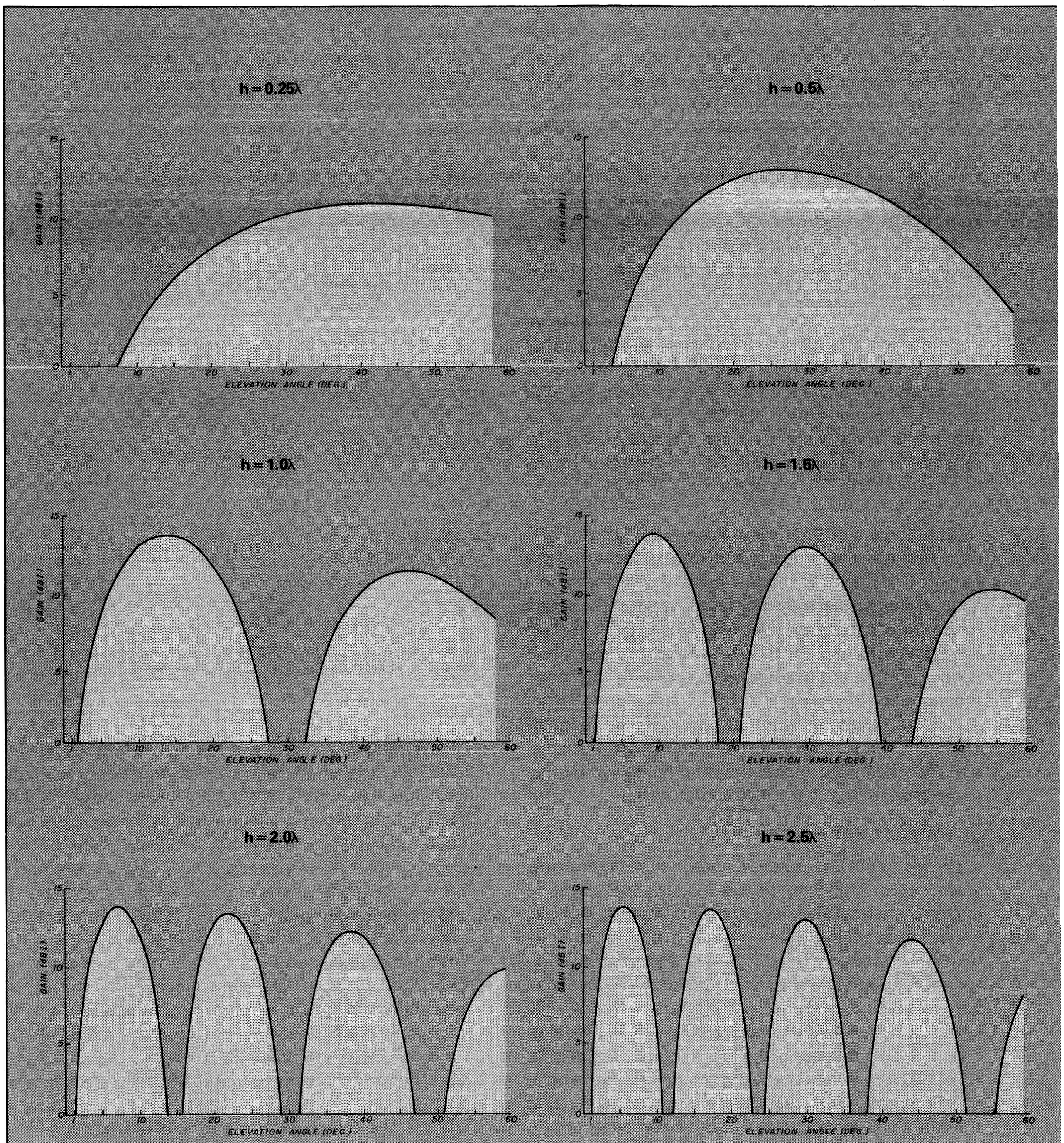

fig. 2. Angle of radiation for three-element Yagis at heights from 0.25λ to 2.5λ, above. At right, height of 3λ.

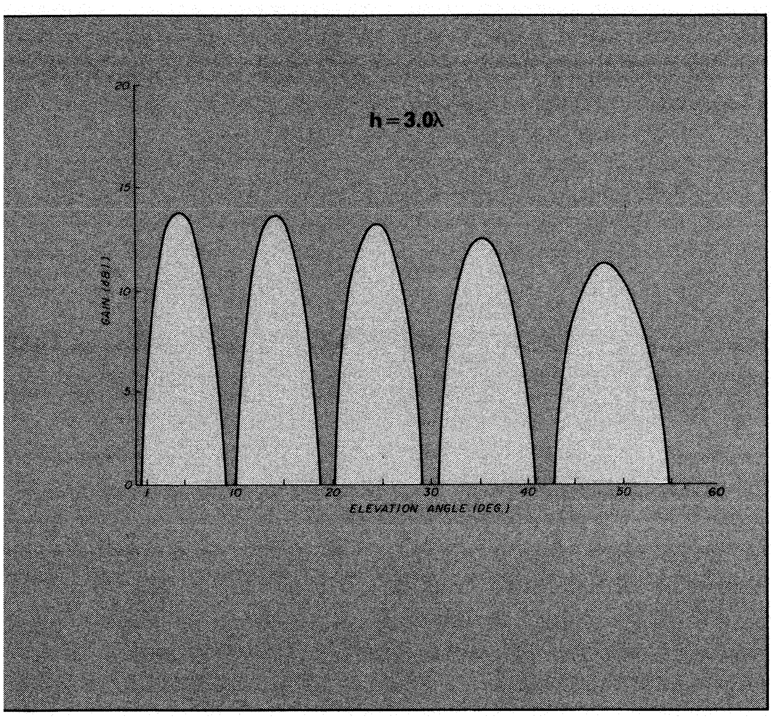

table 1. Representative Yagi beams. All elements are cylindrical with radius ρ = 0.0005260(λ).

element	three-element boom length 1/4λ length (λ)	boom position (λ)	six-element boom length 3/4λ length (λ)	boom position (λ)
reflector	0.49801	0.000	0.49528	0.000
driven	0.48963	0.150	0.48028	0.150
D1	0.46900	0.300	0.44811	0.300
D2			0.44811	0.450
D3			0.44811	0.600
D4			0.44811	0.750

table 2. Three element beam (from table 1) over ground.

height over ground (λ)	gain (dBi)	F/B (dB)	angle (deg.)	driver impedance (ohms) R	X
0.1	9.63	4.90	59	43.05	13.23
0.25	10.88	13.39	42	14.21	−4.25
0.5	12.96	42.94	27	14.76	1.49
0.75	13.46	17.13	19	16.18	− 1.25
1.0	13.68	29.42	14	15.15	0.70
1.25	13.69	19.40	11	16.07	−0.66
1.5	13.78	27.22	10	15.32	0.45
1.75	13.77	20.52	8	15.98	−0.45
2.0	13.85	26.24	7	15.42	0.33
2.5	13.84	25.70	6	15.47	0.26
3.0	13.85	25.34	5	15.51	0.22
Free space	7.86	23.60	0	15.70	0.00

tion angles from 3 to 20 degrees are indeed quite important. He also shows focusing effects of a given mode and the great variety of results which can occur in practice. Thus, in evaluating an antenna system over ground, it is most important to ensure good gain over ***all*** lower angles (say 3 to 17 degrees) and for the lower frequencies ($\leq$7 MHz) over even higher angles (up to say 30 degrees). I shall therefore show, in all cases to be presented, a plot of H-plane gain as a function of elevation angle, β.

antenna performance over ground

To illustrate typical Yagi antenna performances over ideal ground, I shall use as representative Yagi beams a three-element Yagi on a 0.25λ boom and a six-element Yagi on a 0.75λ boom; the basic characteristics are given in **Table 1.** I have made calculations for each beam at a number of different elevations over ground. The results for the three-element beam are shown in **Table 2** and those for the six-element beam in **Table** 3. As a reference, the free-space performance for each beam is also listed in these tables. **Figs. 2** and **3** show the H-plane, or vertical, pattern of each of these cases at certain chosen elevations. It is apparent from these H-plane patterns that maximum gain in the forward direction occurs at an elevation angle which is an inverse function of the antenna height; this relationship is tabulated quantitatively in **Tables 2** and **3.**

These figures also show that the antenna gain has a number of lobes, the biggest lobe is the first one (lowest elevation angle). For each succeeding lobe, the peak gain is somewhat lower. This reduction in gain is caused by the natural free-space directivity of the antenna. The overall pattern is a series of lobes (produced by interference of the direct and reflected waves) essentially modulated by the inherent free-space pattern of the antenna. Note that the relative gain reduction at high angles is greater for the (more directive) six-element beam than for the three-element beam. Moreover, a careful analysis of the lobes shows that the maximum point on each lobe is slightly altered by the natural beam directivity. This is shown in **Tables 2** and **3** by the slightly lower elevation angles of the main lobe for the six- versus the three-element beam at a given height above ground.

Thus, we see that the main lobe of an antenna occurs at an angle ***primarily*** determined by its height over ground, but ***secondarily*** by the natural antenna directivity. This latter effect is most pronounced at low antenna heights, and it is also responsible for the relatively poorer gain at these heights. One would ordinarily expect the ground reflection to double the radiated field (or to add 6.01 dB to gain), but if it occurs

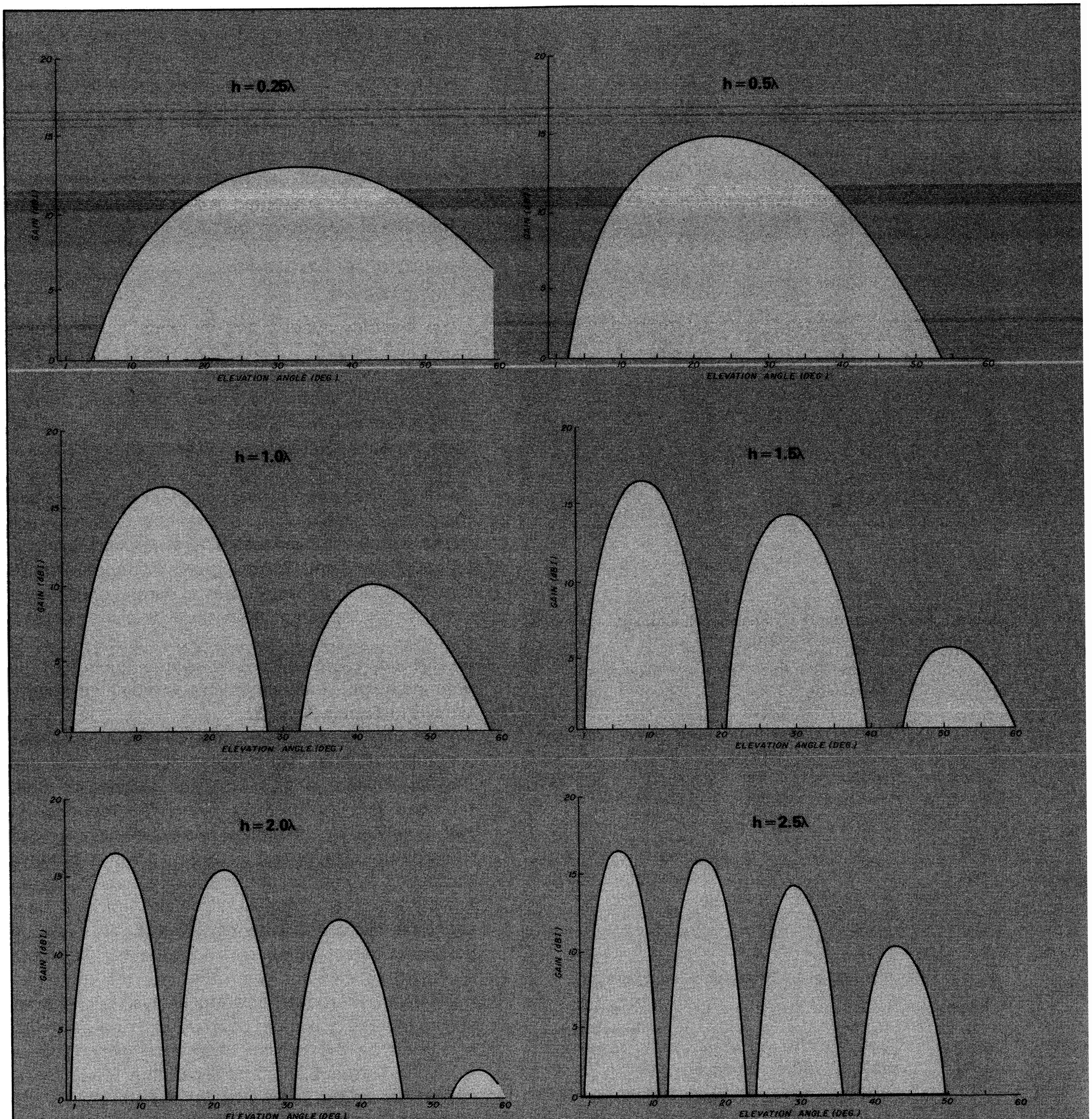

fig. 3. Angle of radiation for six-element Yagis at heights from 0.25λ to 2.5λ, above. At right, height of 3λ.

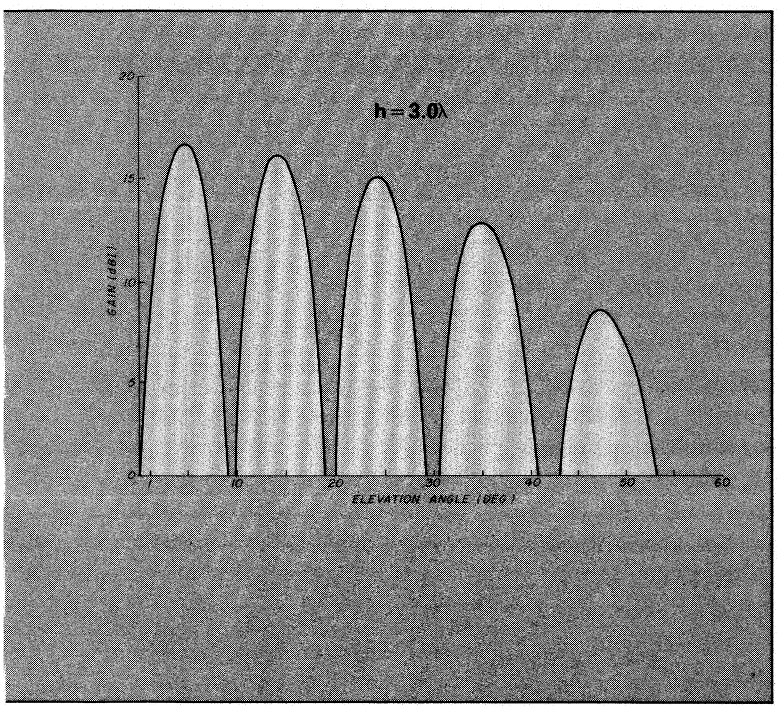

table 3. Six-element beam (from table 1) over ground.

height over ground (λ)	gain (dBi)	F/B (dB)	angle (deg.)	driver impedance (ohms) R	X
0.1	11.89	8.67	46	14.68	8.58
0.25	13.01	13.14	33	23.90	−7.98
0.5	14.86	34.31	24	21.62	−1.73
0.75	15.59	21.52	18	21.24	−0.34
1.00	16.36	35.77	14	22.38	−1.51
1.25	16.23	29.69	11	20.96	−0.95
1.50	16.58	38.61	9	22.23	−1.29
1.75	16.44	35.04	8	21.07	−1.17
2.00	16.67	40.94	7	22.10	−1.23
2.50	16.67	42.86	6	22.01	−1.21
3.00	16.69	44.76	5	21.94	−1.21
Free space	10.70	57.21	0	21.62	−1.28

table 4. Specifications for six-element beam optimized by slight shifts in boom positions for D1 and D3. All element radii are 0.0005260(λ).

element	optimized for free space: length (λ)	optimized for free space: boom position (λ)	optimized for 1.5λ over ground: length (λ)	optimized for 1.5λ over ground: boom position (λ)
reflector	0.49528	0.000	0.49528	0.000
driven	0.48071	0.150	0.48028	0.150
D1	0.44811	0.2991650	0.44811	0.3039068
D2	0.44811	0.450	0.44811	0.450
D3	0.44811	0.5999658	0.44811	0.5958952
D4	0.44811	0.750	0.44811	0.750

at a high elevation angle (low antenna heights) the original antenna gain (free space) is significantly lowered (at the same high angle).

The front-to-back ratio is also shown in Tables 2 and 3. Recall that the definition I use for F/B is the ratio of forward energy flux density at the best elevation angle to the reverse energy flux density ***at the same*** reverse elevation angle. **Tables 2** and **3** show this quantity to fluctuate rather widely with antenna height; the cause of these fluctuations is the altered antenna element complex currents that result from the mutual coupling of antenna and its image. These mutual effects are large when the antenna is low and relatively small when the antenna is high. Note, however, that even when the antenna is three full wavelengths above ground, enough interaction occurs to noticeably alter the free-space value.

Similarly, the antenna driving-point impedance fluctuates with antenna height. When the antenna is very low, for example, at a height of 0.1λ, driving-point resistance and reactance are far from their free-space values. This shows dramatically that if one adjusts an antenna near the ground (say at 0.1λ) for best performance, it certainly will ***not*** be the best adjustment at final operating height.

These ground mutual effects, which alter the antenna element currents, are present to some degree at all antenna heights likely to be used in practice. This is tantamount to saying that the antenna over ground is ***not*** the same as the antenna in free space. An antenna optimized for free space will therefore ***not*** generally be quite optimum over ground. Obviously, one should really optimize the antenna over ground at the desired height.

What is the best antenna height? Recall from **fig. 1** that one should strive for a large gain over a ***range*** of angles, for example, 3 to 17 degrees. An inspection of **figs. 2** and 3 shows that this occurs when the antenna height over ground is about 1.5λ. For 14-MHz radiation, this height would be about 30 meters, or 100 feet. Practical operating experience does verify that such an antenna height gives excellent results. Note also that at a height of 3λ a deep lobe null occurs at an elevation angle of 10 degrees; this angle is sometimes important, such as for a range of 4500 km using two F-layer hops. Such a high antenna, even though excellent as a band opener at very low angles, would not be expected to be a good overall performer. I have tried a large 14-MHz antenna at a height of 2.6λ; from my location in New York State, the average European signals were found to be substantially inferior to those received from an antenna at a height of 1.5λ.

It is fortunate that an antenna at a height of 1.5λ over ground is not seriously degraded from its free-

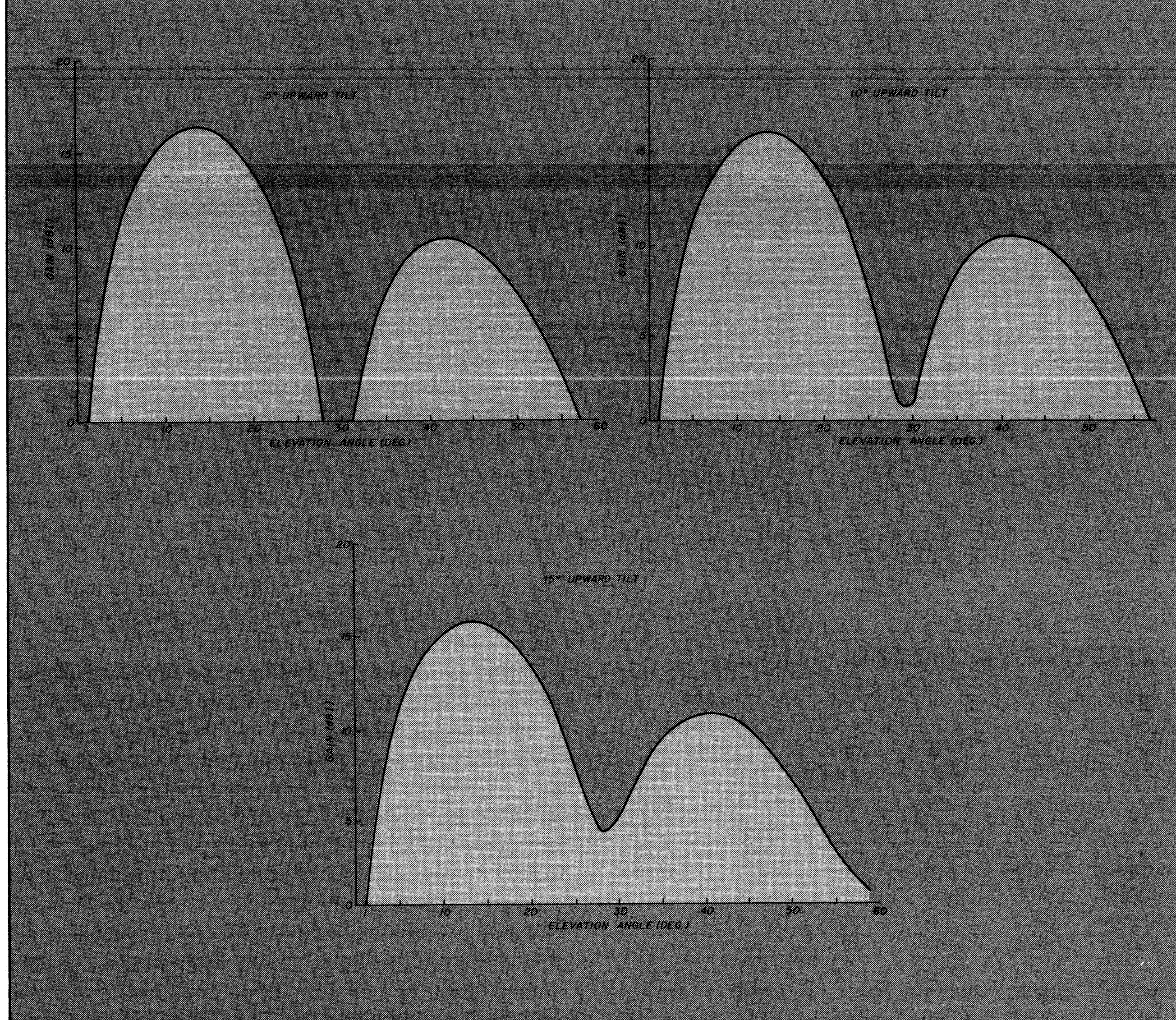

fig. 4. These graphs show the angle of radiation of a six-element beam that has been tilted upward above the horizon. In comparison with fig. 3, the peak gain is reduced, as are the nulls between the lobes.

table 5. Performance of free-space optimized six-element Yagi (specification in table 4).

frequency	gain (dBi)	FIB (dB)	angle (deg.)	driver impedance (ohms) R	X
0.996	10.59	26.59	0	22.42	−6.53
0.998	10.65	32.64	0	22.10	−3.50
1.000	10.70	120.18	0	21.80	−0.42
1.062	10.75	32.69	0	21.51	2.72
1.004	10.79	26.68	0	21.25	5.93

table 6. Six-element optimized Yagi at 1.5λ over ground (specification in table 4).

frequency	gain (dBi)	FIB (dB)	angle (deg.)	driver impedance (ohms) R	X
0.996	16.50	26.45	9	22.40	−6.12
0.998	16.56	32.52	9	22.17	−3.09
1.000	16.61	123.18	9	21.95	−0.00
1.002	16.65	32.59	9	21.75	3.14
1.004	16.69	26.60	9	21.57	6.33

space performance. **Table** 3 shows that the *F/B* ratio (at central design frequency) for the six-element beam is a superb **57** dB in free space and degrades only to a still superb value of 38 dB when the beam is mounted at 1.5λ over ground. In both cases, optimization procedures described in a previous article3 can tune up the *F/B* ratio with only minor effects on other performance features.

I have carried out such an optimization by varying the boom positions of *D1* and *D3* slightly; final beam specifications are shown in **Table 4,** and the performance around the frequency of best *F/B* is shown in **Tables 5** and **6.** Note that optimization requires very delicate boom position adjustments. These boom positions have been adjusted sufficiently well to give a *F/B* well over 100 dB at the central design frequency. It is interesting to note that ***different*** boom positions are needed for the free-space Yagi and for the Yagi to be mounted over ground. This is because they are really slightly different antennas because of ground interaction.

Let me stress, as I did in the previous article on optimization, that although the Yagis are mathematically optimized to give a very high (but narrowband) *F/B* ratio, the basic model cannot really be trusted to this level of accuracy. It should be quite possible in principle to carry out this type of optimization experimentally on a real Yagi; the basic behavior should be similar, but the final boom positions might be slightly different.

Note that when a Yagi is mounted over ground the lowest lobe of radiation has a maximum at an elevation angle usually sufficiently high that the direct wave from the antenna is somewhat reduced from its peak free-space value. It is interesting to see if any improvement could be made by purposely tipping the antenna boom upwards to increase the direct wave.

Unfortunately, tipping the antenna upward automatically tips the image ***downward*** by the same angle; the net result is that, while the direct wave is increased, the reflected wave is decreased, and unfortunately by a greater amount. As an example, consider the six-element Yagi mounted 1.0X over ground. The maximum gain of **16.36dBi** occurs at an angle of **14** degrees as shown in **Table** 3. **Table 7** and **fig. 4** show the result if the antenna is tipped upward at angles of **5, 10,** and **15** degrees. It is easy to see that maximum overall gain is actually best when the antenna is parallel to the ground plane; as one tips the antenna the peak lobe gain is reduced slightly and the deep nulls between lobes tend to become shallower. This is precisely the behavior expected from a consideration of the vectorial addition of direct and reflected waves.

table 7. Six-element Yagi at 1.0 λ over ground. Performance vs upward tipping angle.

tipping angle (deg.)	gain (dBi)	F/B (dB)	angle (deg.)	impedance (ohms) R	X
0	16.36	35.77	14	22.38	– 1.51
5	16.31	32.85	14	22.51	– 1.65
10	16.12	29.58	14	22.47	– 1.84
15	15.77	26.60	14	22.27	– 1.99

summary

1. Although ground is difficult to characterize, there is reason to believe that for horizontal polarization a good model is an ideal, infinitely conducting plane.

2. The H-plane (vertical) pattern consists of a number of lobes caused by the interference of direct and ground-reflected waves. The first (lowest) lobe is the strongest; succeeding lobes are reduced somewhat in gain by the natural free-space directivity of the antenna.

3. Mutual effects between the antenna and ground cause antenna element currents to change; these changes cause significant alterations to the antenna properties. The most noticeable variations occur in *F/B* ratio, but there are also significant variations in gain and driving-point complex impedance.

4. Best overall antenna performance appears to occur if the antenna height is about 1.5λ. This is not a critical figure, but it is believed that **3X** is probably too high.

5. Tuning or adjusting an antenna near the ground for best performance guarantees that the antenna will ***not*** be optimum at operating height.

6. Because of the significant mutual effects with the ground, the antenna should be optimized at its final operational height. Generally, this optimization will not be quite the same as the optimized free-space antenna.

7. Large antennas are more handicapped at low heights than small antennas; this is due to their higher natural free-space directivity.

8. For best gain the boom of the antenna should in principle be parallel to the effective ground plane surface; however, the degree of parallelism required is not critical. Tipping the antenna upward to improve gain will, in actual fact, decrease maximum gain.

references

1. The ARRL Antenna Book, 1974, 13th edition, published by The American Radio Relay League. Newington, Connecticut.
2. F. Kift, "The Propagation of High-Frequency Radio Waves to Long Distances," Proc. IEE, 107 Part B, March 1960, page 127.
3. J.L. Lawson, "Yagi Antenna Design," *ham* radio, July, 1980. page 18.

ham radio

how to determine true north for antenna orientation

Simple procedure for establishing a reference baseline to set up your beam antenna

Whenever a rotary beam antenna is installed a question arises. Which way is north? This article describes a simple, accurate, and little-known method of laying out a true north-south baseline using the sun for orienting a new installation or for checking the accuracy of an existing rotator direction indicator. The only equipment required is an accurate source of time.

description

The procedure is based on the principle that, at local noon, under certain conditions, the sun bears true south, so the shadow of your tower or other structure may be marked to create a permanent reference baseline. Under most conditions, the time when the sun bears true south occurs at times other than noon. However, the exact time may be easily computed with the data included in this article.

Computing the time. The time correction has two parts: the first remains constant and depends on your location (longitude, actually); the second is variable and depends on the day of the year the observation is made. Once the correction has been calculated it's simply added or subtracted, as the case may be, from local noontime to determine the exact time of the observation.

Although the computation takes less time to perform than to explain, an understanding of the factors composing the total correction is helpful in applying the correction.

Celestial dynamics. As is well known, the earth revolves once every 24 hours. However, for ease of understanding, let's assume that the sun rotates around the earth's equator, from east to west, once every 24 hours. The earth's equator, a circle, is divided into 360 degrees with 0 degrees arbitrarily set at a point at the intersection of the equatorial circle and a line drawn due south from the Royal Navy Observatory in Greenwich, England. Thus, the sun moves in an angular arc of 15 degrees in one hour. It is also equivalent to a movement of 1 degree of arc every four minutes of time or an arc of 1 minute (a degree being subdivided into 60 minutes of arc) every four seconds of time.

The location of a point on the earth may be defined in terms of latitude (degrees of arc north or south of the equator) and longtitude (degrees of arc east or west of the line through Greenwich). Longitude in North America is typically stated in degrees and minutes of arc west of Greenwich. For example: Philadelphia is 75 degrees, 0 minutes west; San Francisco is 122 degrees, 27 minutes west longitude. The longitude of your location may be taken from a map or obtained from local civil authorities (U.S. Coast and Geodetic Survey).

Time zones. The earth has been divided into 24 standard time zones based on standard meridians spaced every 15 degrees. Eastern Standard Time is based on the 75th meridian, CST on the 90th, MST on the 105th, PST on the meridian at 120 degrees west, and so on. Each zone is designated by a letter of the alphabet. The Greenwich meridian, for example, is

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designated *Z* (Zulu). Unless you're fortunate enough to live on one of the standard meridians, the sun will not be on the local meridian passing through your location at noon standard time. However, since we now know the sun's rate of rotation in terms of angular displacement vs time elapsed, it's easy to compute the time of local meridian passage.

examples

The following example illustrates an arc-to-time conversion. My location is Greensboro, North Carolina — longitude 79 degrees, 53.43 minutes west. The local time zone is EST (75th meridian time). The lateral offset in longitude is 4 degrees, 53 minutes, derived from subtracting 75 degrees from 79, degrees, 53 minutes (rounded from 79 degrees, 53.43 minutes). As each degree equals four minutes of time and each minute of arc is four seconds of time, the total offset is 16 minutes and 212 seconds or, ex-

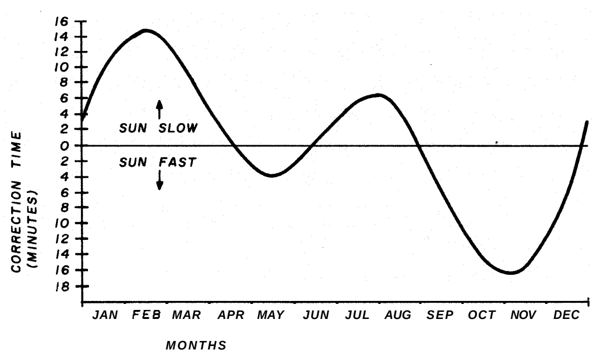

fig. 1. Plot of equation of time (EOT) taken from the *Nautical Almanac,* used to correct bearings for antenna orientation. The phenomenon is caused by the eccentricity of the earth's orbit with respect to the sun. The graph illustrates the difference between solar and local clock time.

pressed more conventionally, 19 minutes and 32 seconds. As the location is west of the standard meridian, the sun will be late and won't arrive on the local meridian (i.e., bear true south) until 12:19:32 clock time, neglecting other factors for the moment.

Another example: Greenville, New Hampshire, longitude 71 degrees, 51 minutes west, zone time, EST. In this case, the offset from the standard time meridian is 3 degrees, 9 minutes. Converting this arc to time gives 12 minutes and 36 seconds. In this case, the location is east of the standard-time meridian, so the sun will arrive early at 11:47:24 AM, neglecting other factors.

the equation of time

The other correction factor, which was neglected in the foregoing calculation, arises from the fact that the sun is like a poorly regulated clock: it runs fast during some periods of the year and slow during others. This is because the earth's orbit is an ellipse rather than a circle. Fortunately, this eccentric motion is highly predictable and forms a pattern that repeats year after year. The difference between solar time and clock time is called the ***Equation of Time*** (EOT) by astronomers and navigators. Rather than resetting clocks every day to agree with solar time, civil time uses an average, or mean, of the overall yearly variation; e.g., Greenwich Mean Time or Universal Coordinated Time (UTC).

For any location, the correction for EOT may be taken directly from **fig. 1,** a plot of data taken from the ***Nautical Almanac,*** a U.S. Government publication. For Amateur antenna alignment purposes, the data is valid through the year 2100.

correction for daylight savings time

Although perhaps obvious it should be mentioned that, when local time is based upon daylight savings time, clocks are arbitrarily advanced one hour, thus making indicated time one hour ahead of "actual" time. Therefore, an additional correction of plus one hour should be added when applicable. The following example illustrates a complete calculation:

Find: Local time when the sun bears true south

Given: Location, Greenville, New Hampshire, longitude 71 degrees, 51 minutes west

Zone time, EDST, 75th meridian

Day, August 1, EOT, 6⅓ minutes or 00:06:20

longitude correction	0:12:36 (−)
EOT correction	0:06:20 (+)
combined correction	6:16 (−)

Subtracting from 12:00:00 noon EST (stated as 11:59:60) gives:

11:59:60 EST
06:16 (−)
11:53:44 EST

However, since EDST is in effect, one hour must be added to the calculated time to agree with indicated time: *11:53:44* + *01:00:00* = *12:53:44* EDST. Therefore, the sun will bear true south (and the shadow of your tower or other structure will point true north) at 44 seconds after 12:53 PM clock time.

Once the shadow has been marked you can, at your leisure, sight along the baseline from tower to mark to identify a more distant terrain feature. Then it's simply a matter of aligning the antenna boom along this line (with the rotator control set at north) to ensure an accurately calibrated direction indicator.

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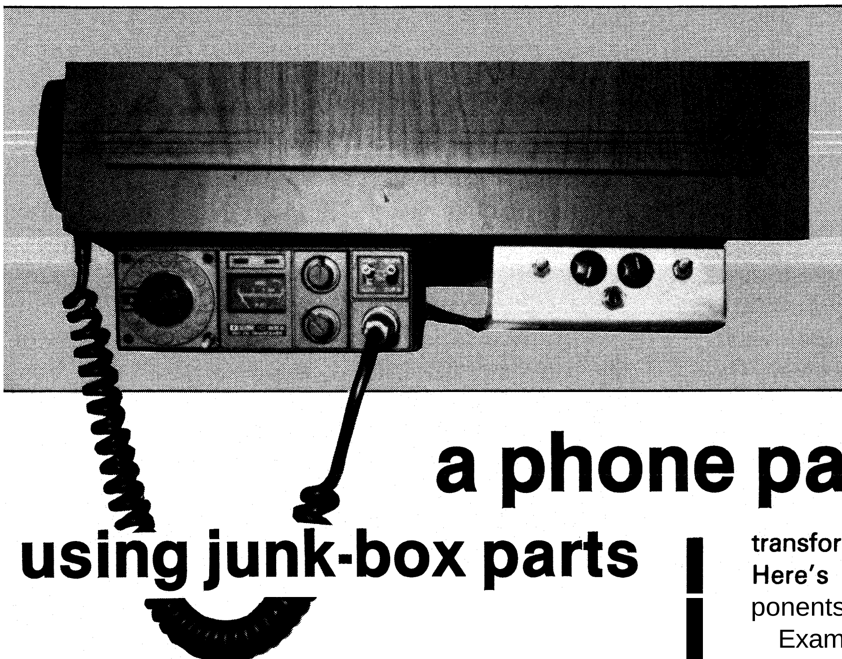

a phone patch using junk-box parts

Old transformers used in tube-type radios are put to use in this easy-to-build phone-patch circuit

Often I get the urge to upgrade my station with one accessory or another. In these days of inflation, however, running out and buying accessories usually results in a severe wallet-ache. This article describes how you can raid your junk box rather than rob a bank to build a reliable phone patch. The best part is that the patch takes only a couple of hours to put together.

theory

The main problems to overcome in constructing the phone patch are impedance matching and converting the balanced condition of the telephone line to the unbalanced audio condition of the Amateur transmitter and receiver. A transformer with proper impedance specifications offers the easiest solution.

Not long ago, before transistors or integrated circuits were available, everyone used "field-effect transistors with heaters in them" — more commonly called tubes. Since the age of the inexpensive transistor has arrived, more and more tube-circuit power transformers have been retired to the junk box. Here's an interesting application of these components.

Examination of the specifications of one of these transformers shows that it may be quite suitable for use in a phone patch. First, since power transformers are designed to operate at 60 Hz, there's little worry of audio distortion because of insufficient core material. Second, a review of the winding information shows good compatability between an Amateur rig and the telephone line.

Eq. (1) shows the relationship of winding voltages to the winding turns ratio.

$$\frac{V_1}{V_2} = \frac{1}{a} \qquad (1)$$

where V_1 = voltage across winding 1 (primary)
V_2 = voltage across winding 2 (secondary)
a = ratio of the number of turns of winding 2 with reference to winding 1

checking out the transformer

After blowing the dust off the old transformer, carefully determine which pair of wires constitutes the primary leads (usually the black pair). By measuring the voltages of the *open* windings while the primary is connected across the 120-volt house current, enough information may be obtained to calculate the turns ratio of the windings. (If your nerve is hardened to the idea of plugging a transformer into 120 Vac house current, a step-down transformer may be used to drive the winding.) I had a transformer with two windings in addition to the primary. The voltages

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measured on the open windings and the turns ratio calculated from **eq. (1)** are listed in **table 1.**

After the turns ratio has been determined, another equation may be used to determine the degree of impedance matching possible with the transformer. **Eq. (2)** relates the impedance that will be observed between the windings of a transformer if the transformer is "ideal." Although no transformer is ideal, most are very good, and the equation will closely predict the impedance relationships.

$$\frac{Z_1}{Z_2} = \frac{1}{a^2} \qquad (2)$$

where Z_1 = impedance across winding 1 (primary)

Z_2 = impedance across winding 2 (secondary)

a = ratio of the number of turns of winding 2 with reference to winding 1

The telephone-line impedance is approximately *600* ohms. By using **eq. (2),** I calculated the impedances that would be present across the secondary windings if the primary were connected across the phone line. The results appear in **table** 2.

As may be observed from **table** 2, if the primary (black-black) is connected across the 600-ohm phone line, the 12-volt filament winding (green-green) will present a reasonable load to an 8-ohm speaker output from an Amateur receiver. Likewise, the high-voltage winding (red-red) presents a reasonable microphone input match to either high- or low-impedance Amateur transmitters.

construction

I constructed the phone patch in a minibox chassis. **Fig. 1** illustrates the final circuit. Transmitter audio leads should be shielded cable.

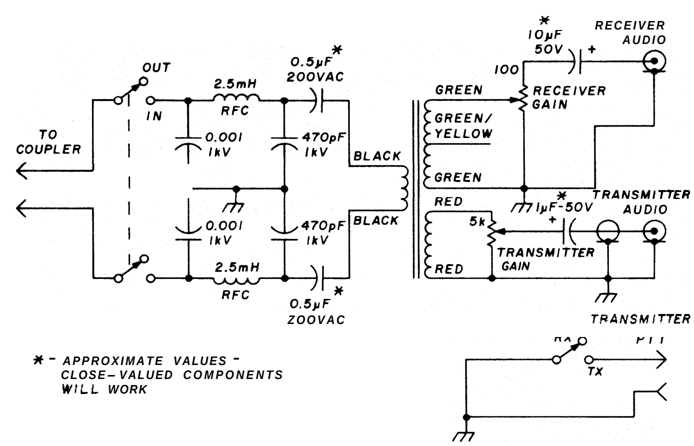

fig. 1. The K7NM phone patch made from junk-box parts. The transformer was salvaged from a tube-circuit radio. The text explains how to check out such transformers for compatibility with the telephone line.

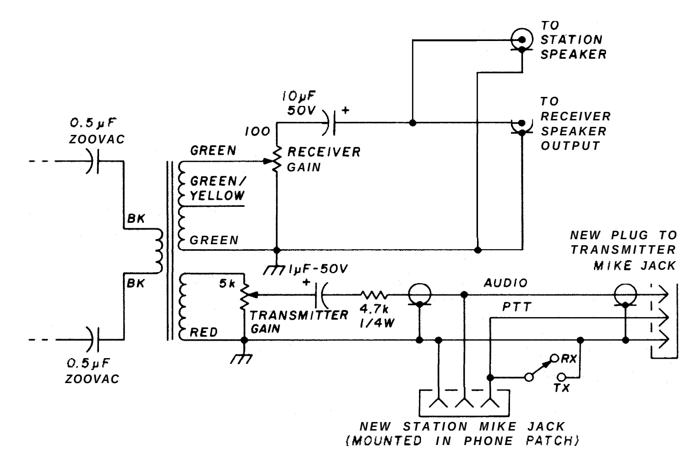

fig. 2. Alternative connection of transmit and receive audio to the transmitter.

table 1. Measured voltages from the power-transformer windings.

winding colors	voltage (Vac)	turns ratio (a)
black-black	120.0	1.000
green-green	12.6	0.105
green-green/yellow (center-tap)	6.3	0.052
red-red	240.0	2.000

table 2. impedances across windings.

winding colors	turns ratio (a)	impedance (ohms)
black-black	1.000	600
green-green	0.105	6.82
green-green/yellow (center-tap)	0.052	1.62
red-red	2.000	2400

An rf filter was placed in the phone line to prevent rf from entering the transmitter audio through the patch from the phone line. This filter is only necessary if high power levels are used.

The coupling capacitors on the primary winding must be nonpolarized and have at least 200-Vac ratings. These capacitors prevent the primary transformer winding from shorting any telephone-line dc voltages that may be present.

The phone patch is connected to the station telephone through an acoustic coupler (to be installed by the phone company). It provides a 1/4 inch (6.4 mm) phone jack for the phone-patch connection. On most desk telephones, a button on the receiver hook must be lifted to activate the coupler. Wall phones will probably have some method of activating the coupler also. I've used only desk phones in the phone patches installed to date.

The receiver audio is connected in parallel with the station receiver speaker, reducing the receiver speaker load to about 4 ohms.

The transmitter input is connected to the phone-patch input jack on the rig. If no such input is provided, **fig. 2** illustrates an alternative method of connection to the transmitter. The phone patch is connected permanently to the transmitter through the microphone input jack. A new jack is mounted in the phone patch chassis for connection to the station microphone. The transmitter patch audio is coupled through a 4.7-k resistor into the audio line.

operation

When operating the phone patch, simply throw the switch located in the transformer primary to the IN position. The telephone must remain off the hook, with the coupler button activated, to maintain the phone connection to the phone patch. I've found it convenient to use the telephone exclusively for modulation and monitoring during a phone-patch contact, The transmit switch is placed in the TX mode during transmit and returned to RX during receive periods, facilitating standard push-to-talk operation When the patch is completed, the primary switch is returned to the OUT position.

Dual level adjustments exist for both receiver and transmitter levels to and from the phone patch. The receiver audio gains on both receiver and patch will affect the line level to the telephone. Also, both the transmitter level adjustment and the gain contol in the phone patch will affect transmitter modulation level. To set levels, I set the receiver audio gain to a comfortable speaker volume and the transmitter level for proper modulation from the station microphone. Next, the patch was switched IN and the telephone removed from the hook with the coupler button activated. A single digit (other than 1) was dialed to remove the dial tone. The receiver patch gain was adjusted for a comfortable audio level in the telephone. Similarly, the patch transmitter level was adjusted for proper modulation of the transmitter while I counted into the telephone. If your city has a time number, this service could be used as a reliable source for the transmitter level adjustment. Once the patch levels are set, there should be no need for additional adjustments.

conclusion

Since the first phone patch, several others have been built and work very well. Even though tables **1** and **2** show that a 6.3-volt filament winding has a very low impedance (about 2 ohms), I've successfully used transformers with this winding as a speaker winding.

Some tube-type transformers have dual 6.3-volt windings, while others may have both a 6.3-volt winding and a 5-volt winding. These transformer windings may be wired in series to improve speaker impedance matching.

Although this phone patch was not designed for VOX operation, some reports have been received that VOX has been used successfully. I've used this phone patch design on both the hf and vhf bands with excellent reports of audio quality. Often Amateurs inquire what kind of patch it is. It's simply a junk box special.

acknowledgement

I would like to thank Rob Yaw (WA7IAL) for his assistance during the testing of these phone patches.

ham radio

a folded end-fire radiator

Introducing the "W8" — an antenna with modest gain and interesting possibilities for the experimenter

This antenna fits into an area approximately 1/10 by **1/6** wavelength and provides a compact, effective radiator for the lower-frequency bands. A slight reshaping converts it into a horizontally polarized, omnidirectional radiator for vhf. The antenna can be considered as a folded, series-connected, two-element, end-fire beam.

evolution

Start with a two-element, bidirectional, end-fire beam. If the feed is omitted, the essential elements are two half waves, closely spaced and excited equally but 180 degrees out of phase **(fig. 1A)**. The arrows in **fig. 1** indicate current flow, with the large arrows at current maxima. Spacing is usually 1120-1/**5** wavelength. Gain can approach 4 dB over that of a **halfwave** antenna.' Radiation is maximum to left and right in the plane of the figure. For illustration, take spacing to be **1/10** wavelength.

Now fold each element into a rectangle 1/10 by **1/5** wavelength, as shown in **fig. 1B**. Note that the wire ends labelled **a** are at the same rf potential; similarly, the ends labelled **b** are also at the same potential, opposite to that at **a.** These respective pairs of ends may therefore be joined without disturbing the current distribution as shown in **fig. 1C**. Note that the array now consists of one wavelength of wire folded into a sort of figure eight, accounting for its name, abbreviated **"W8."** At the center crossover point are two current nodes (voltage maxima). **Fig. 1D** is another possible arrangement with only one center crossover connected, about which more later.

There is nothing sacred about the 1/10 by 115 wavelength rectangle. For a low-frequency, vertically polarized array on a single pole the configuration of **fig.** 2 is appealing. The W8 can be squeezed or stretched, with directivity declining as the shape spreads horizontally **(fig. 1)**.

feed methods and impedance

Of the various possible feeds, I've tried only one, balanced current feed, achieved by opening the **W8** at a current maximum and attaching parallel-wire feeders, as in **fig.** 2. In December 1973 1 measured a 6-meter model built to the **1/10** by 115 wavelength pattern and obtained a driving-point impedence between 400-500 ohms. Measurements were made

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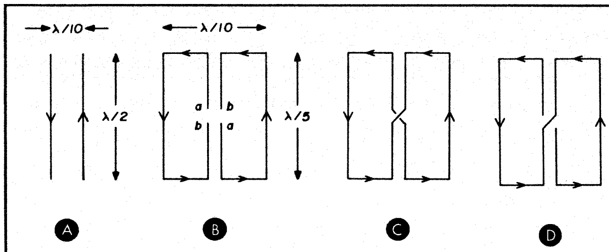


fig. 1. Evolution of the W8 end-fire antenna. The arrows on the elements indicate current flow; the large arrows show current maxima. In A a basic radiator is shown consisting of two half-wave elements, closely spaced, and excited 180 degrees out of phase. Each element is then folded into a rectangle, B. Wire ends may be joined as shown in C or D, as explained in text.

with a crude resistance bridge driven by a grid-dip meter. I hope someone who can make more reliable measurements will look into this. For this feed method 450 ohm TV ladder line should do well.

The W8 could be voltage fed at the center cross-over point, using either tuned feeders or an open-wire matching transformer as in **fig.** 3. This might be convenient for the vertically polarized, one-pole configuration of **fig.** 2, particularly if you wish to run a line along the ground for some distance.

Finally, if the W8 is not broken at a current maximum you should be able to drive it by any of the methods used for plumbers' delights, in particular the gamma match applied at a current maximum. This might be a convenient choice for the omnidirectional W8 mentioned later.

performance and application

A reasonable guess at gain can be made by comparing **fig. 1C** and the basic, two-element endfire, **fig. 1A**. The center section, consisting of the folded ends of the original elements, is just a short section of open-wire line and can be expected to have negligible loss and radiation. Thus the only real waste represented by the configuration of **fig. 1C** is that which results from radiation from the horizontal portions. However, these useless portions are only half the length of the useful radiating sections and, in addition, contain portions of the current distribution farthest from the two current maxima. This argument suggests that gain should still be perhaps 2 dB over a similarly placed halfwave. The 40-meter vertically polarized version was used for a couple of years, but I have no way to make comparisons and can only hope this discussion will arouse some of the more ambitious experimenters.

An interesting comparison can be made with the quarter-wave vertical. If the configuration of **fig.** 2 is used and each radiating section makes a 45-degree angle with the vertical, then the vertical extent of the array is somewhat less than a quarter wavelength. The mechanical disadvantage of the W8 is that guy cords are needed to position the two side corners of the diamond (I use nylon fish line). On the other hand, the W8 requires no radials or ground plane, and its current maxima are elevated to the array center. This should make it practically simpler to achieve low-angle radiation with a W8.

The small gain of the W8, though interesting, will be swamped by the effects of propagation and interference. Its main advantage is relative simplicity and compactness. A 40-meter horizontal W8, for example, could be supported slightly above your roof in about the same space as a full-sized, two-element 8JK for 20 meters.

dimensions and adiustment

The 6-meter model used 20 feet (6 meters) of wire arranged in a 2 by 4 foot (0.6 by 1.2 meters) rectangle, with the center section of transposed line spaced about 3/4 inch (19 mm). With a small loop soldered across an opening at the center of one side, the grid dipper resonated at about 50 MHz.

When the 40-meter model was put together, 136 feet (41.5 meters) of wire were used at first in the antenna proper, with 68 feet (20.7 meters) of open wire line to the house. When this arrangement was grid-dipped at the sending end, the resonant frequency turned out to be 5.8 MHz. The explanation for this remarkable discrepancy is the capacitive loading effect of the W8 center section. On the 40-meter model, the spacing in wavelengths was closer by a factor of 7 than the spacing on the 6-meter model. As noted in the section describing the evolution of the W8, the center crossover point and most of the center section of line are at high rf potential, so capacitance here is very effective in loading to a lower frequency. The major advantage of the W8 is its compactness, however, and this loading effect makes the antenna even more compact. Incidentally, a small lumped capacitance across the center crossover point is a conveni-

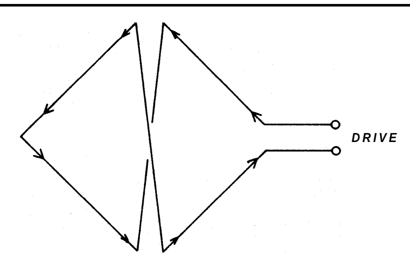


fig. 2. Diamond-configured W8 antenna with current feed. This arrangement may be used for a low-frequency, vertically polarized array mounted on a single support.

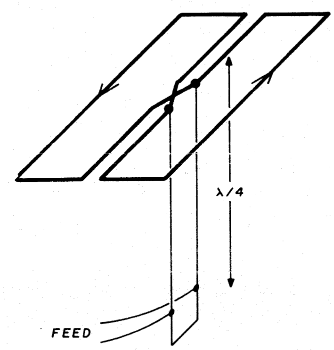


fig. 3. Feeding the W8 at the center crossover point using tuned feeders or open-wire matching transformer.

ent way to tune a vhf W8 to a lower frequency. This was checked in the 6-meter model.

To adjust the 40-meter W8, two steps were taken. The center line section spacing was increased from 314 inch (19 mm) to about 3 inches (76 mm) to reduce the loading effect. Then the lengths of the four vertical wires (two radiating portions and the center line section) were gradually raduced from an original 28 feet (8.5 meters) to a final 21 feet (6.4 meters). At this point the antenna, crudely draped horizontally about 4 feet (1 meter) off the ground, grid-dipped at about 7.1 MHz. The full width across the top is 14 feet (4.3 meters). The vertical extend of this W8 is about 1/6 wavelength at the operating frequency.

A few years later the 40-meter W8 was replaced by a 75-meter version following fig, 2. Its slanting sides were 28 feet (8.5 meters) long, forming a square standing on one point, with the center vertical section of line 40 feet (12 meters) long. These dimensions were only a first guess based on experience with the 40-meter model, although the antenna worked about as expected. Although the main radiation was vertically polarized endfire, there was a significant horizontally polarized radiation broadside. Total length of wire in this version was 192 feet (58.7 meters) or about 0.76 wavelength. For the rectangular 40-meter configuration, the total length of wire was 112 feet (34 meters) or about 0.82 wavelength. The diamond configuration produces more center loading, so the shorter length is reasonable, although resonant frequency is unknown for the 75-meter version.

node forcing

An interesting question about the W8 or any other electrical-full-wavelength continuous conductor, such as a quad loop, is, "What determines the location of the current maxima?" Presumably this is fixed by the feedpoint and by the actual loop resonant frequency. If a damped wave train were excited in an unattached loop at a frequency not exactly resonant, you'd expect the current maxima to chase around the ioop as the wave train died away rather than remain at fixed locations. However, if the loop were opened at one point, this point is compelled to be a current node.

This brings us to the variation of fig. ID, where one of the center crossover points is closed while the other remains open, forcing the array to produce a current node at the open ends. I've used the 40-meter W8 with both crossover points connected and with only one connected. As expected, I found no difference.

As an aside, it seems possible that some of the occasional disappointments with quads whose elements are either mistuned or incorrectly coupled might be partly explained by the current distributions on the parasitic elements having "slipped" around the loop, so that the current maxima are not in line with those of the driven element. This may help to explain why quads occasionally "squint." Although I've never heard of anyone doing so, it should be similarly possible to ensure the current distribution on quad loops, particularly the parasitic elements, by opening one of the vertical sides a quarter wavelength from the driven-element feedpoint and at corresponding points for the parasitic elements.

horizontally polarized omnidirectional radiator

When the current maxima are close together, as in fig. 1C, the W8 should have a directivity in its own plane similar to that of the two-element 8JK beam: a figure-eight pattern with the nulls filled in a bit by the radiation from the short ends of the rectangle. As the shape is pulled out, so that the current maxima are farther apart and the center section of line is shorter, more of the current distribution contributes to radiation over the whole of the plane of the array.

It's clear that, if the shape is pulled out to something like that in the sketches of fig. 4, the pattern in the plane of the array will be close to omnidirectional. In three dimensions the pattern should be roughly doughnut shaped. If the array is mounted in a horizontal plane, the result will be omnidirectional horizontal polarization. Because the current maxima pro-

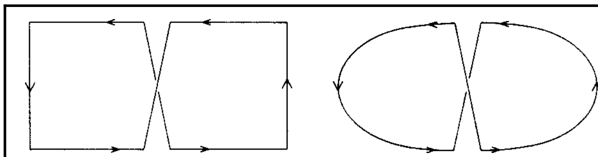


fig. 4. W8 antenna shapes for an omnidirectional radiation pattern.

vide greatest radiation, it's necessary to pull out the shape past the point where it's roughly square or circular. For a start you might try an ellipse with about a 2:1 axial ratio or a rectangle about 1/4 by 1/8 wavelength, as suggested in **fig.** 4.1 hope some vhf experimenter will take a shot at this.

If rod or tubing is used for construction, a quarter-wave transformer can be used for combined support and symmetric center voltage feed, as shown in **fig.** 3. If you wish to force two current nodes, one of the crossover points can be left open and the corresponding end of the quarter-wave transformer also left free, resulting in a Zepp-fed arrangement.

At vhf, horizontal W8s could also be stacked to produce an omnidirectional array with gain and horizontal polarization. The optimum stacking distance for antennas with some gain of their own is greater than a half wavelength. A convenient choice with the W8 would be one wavelength interconnecting feedlines, bent enough to accommodate the desired spacing.

odds and ends

The W8 is essentially a one-band antenna because the capacitive loading of the center section causes its natural modes not to be harmonically related. However, it can be used at harmonics if one employs tuned, open wire feeders, as I do. Such use should be considered makeshift rather than desirable. I have used the 75-meter version on 40 meters, where the pattern should be roughly omnidirectional with a slight bias toward broadside. On the third harmonic I'd expect a slight endfire directivity.

This may also be the place to point out that the W8 is a narrowband antenna compared to a halfwave antenna. This is a property it shares with the 8JK, although the W8 antenna bandwidth can be expected to be somewhat narrower than that of the 8JK because of its smaller size and center loading. The current maxima are about a quarter wavelength apart, so I'd expect the omnidirectional version of the W8 to have somewhat greater bandwidth than that of the 8JK, however.

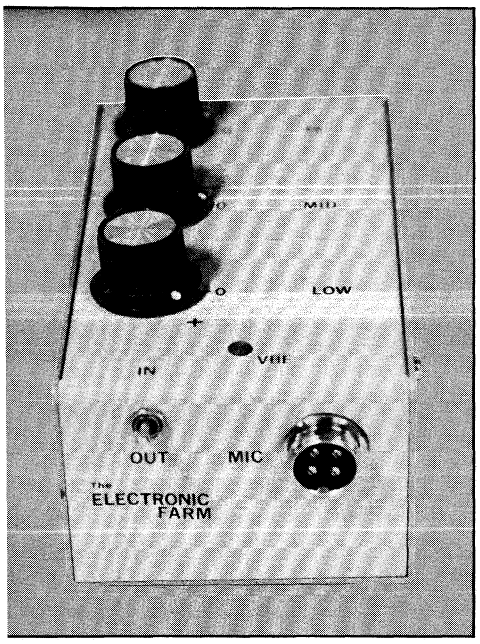
A major aim of this article is to interest people with access to facilities in carrying out gain and impedance measurements and in experimenting with the omnidirectional version. I've tried without success to find any hint of this simple but interesting configuration in the commonly available works on antennas.

reference

1. John D. Kraus, Antennas, McGraw-Hill, 1950, sections 11-3 and 11-5.

ham radio

More Details? CHECK — OFF Page 94

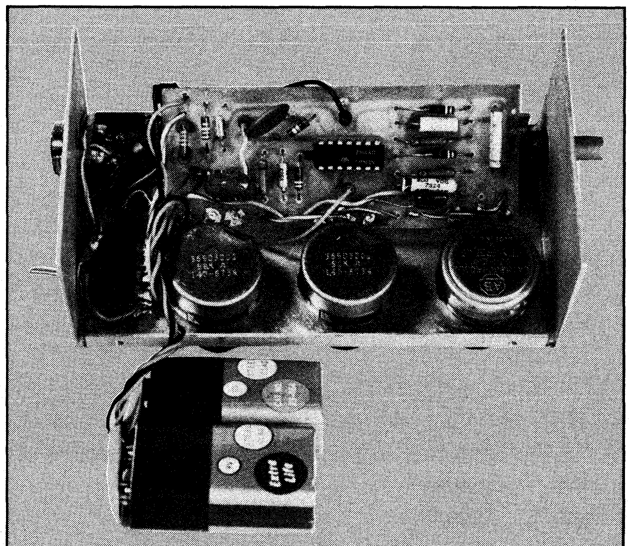


the voice-band equalizer

This addition to your phone station uses an LM324 quad op amp

Here is an equalizer for your microphone that plugs into the microphone jack. The voice-band equalizer (VBE) gives you control of your microphone frequency response. Three adjustments allow you to balance the low, mid-range and high frequencies. The VBE uses three bandpass filters designed to work on the most important part of the vocal spectrum. The passband of your transmitter (2.4 kHz nominal) is the section of audio frequencies I call the "voice band." The VBE divides this important voice band into three parts: low, mid-range and high. By adjusting three controls you can change your audio characteristics. You can boost mid-range and high frequencies to add punch – or accentuate low frequencies for more bass – all without changing your microphone. You can easily build the VBE in a few hours.

This article shows how to work with and apply contemporary active circuits to an Amateur Radio design. Well-known passive tuned-circuit theory is reviewed, and newer tuned-circuit theory and design are introduced. The article also gives information on standard construction practices and how these are applied to build the VBE. Also shown is how to use existing equipment and resources to test the VBE. The voice-band equalizer is a project that will impart a sense of pride and accomplishment to the Amateur who builds one.



Inside the voice-band equalizer showing construction details. A PC board is used for the small components. Control pots are mounted on the minibox as shown.

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circuit design

I built the circuit around an LM324 quad op amp. The LM324 consists of four independent operational amplifiers in a 14-pin DIP package. Each amplifier is internally compensated. Three of the amplifiers operate as tuned circuits; one combines the action of the three separate tuned circuits. Each of the three active tuned circuits acts like the passive tuned circuit of fig. **1.** Using the active rather than passive tuned circuits eliminates the need for bulky and expensive inductors. In fig. **1** L, C, and R are combined to form a series-resonant circuit. At resonance, the electrical energy presented to ***LCR*** is shunted to ground and the inductive reactance, X_L, and capacitive reactance, X_C are equal. Also, at resonance, the impedance of the ***LC*** combination is theoretically zero ($X_{LC} = 0$).

The inductive reactance creates a current lag of 90 degrees, while at the same time the capacitive reactance creates a current lead of 90 degrees. Since the two reactances ($X_L + X_C$) are equal and opposite in phase, they cancel. In practice the impedance at resonance is never actually zero. The Q of this series-resonant ***LC*** circuit is an expression of the ability of the circuit to reach "zero" impedance with respect to frequency:

$$Q = \frac{F_r}{\Delta F(0.707\,F_r)} \qquad (1)$$

where Q = quality factor
F_r = resonant frequency
AF = excursion of frequency from F_r

Also, $Q = \frac{X}{R}$, where X = *reactance* (ohms) and ***R*** = *series resistance* (ohms) for passive series-resonant circuits. The resonant frequency of the circuit is:

$$F_r = \frac{1}{2\pi\sqrt{LC}} \qquad (2)$$

The passive ***RLC*** circuit has low impedance at resonance and high impedance at other frequencies. So energy entering the ***RLC*** circuits through the control pots (fig. **1**) will be shunted to ground at resonant frequencies. As you move the pots toward the noninverting amplifier, input signal is deleted from this input and attenuation occurs within the bandwidth of ***RLC.*** Similarly, when the control pot is turned in the opposite direction (toward the inverting input), frequencies are deleted to the IC, causing gain within the bandwidth of ***RLC.*** Placing the control pot in the center of rotation causes equal effect on both inverting and noninverting inputs. The amplifier output is then flat.

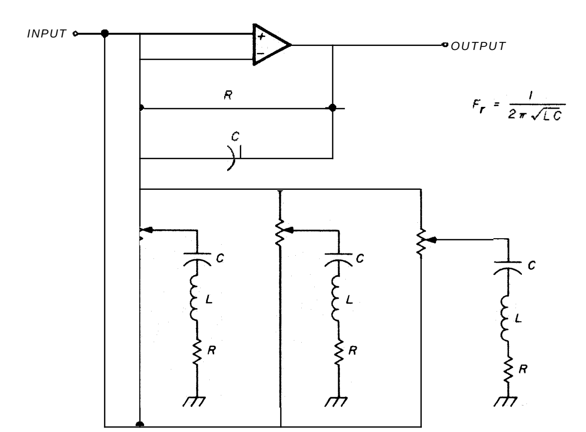


fig. 1. Passive series-resonant circuits using RLC components with an operational-amplifier IC. The voice-band equalizer active circuit (fig. 2) exhibits equivalent characteristics.

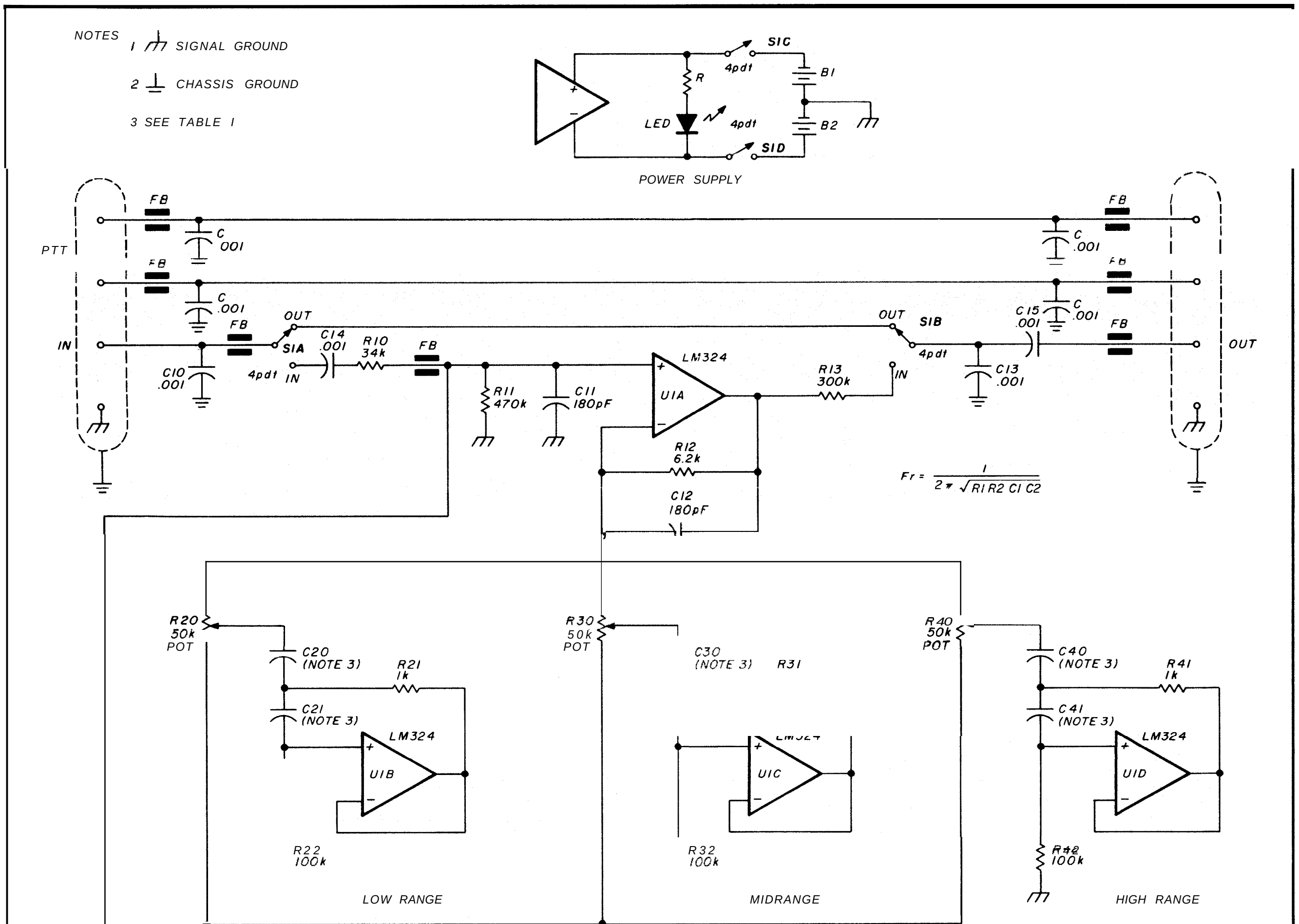
active circuit

The operation of the active circuit for the VBE, fig. **2,** is similar to that of the passive design shown in fig. **1.** Let's look at how the active circuit works. Amplifiers U1B, U1C, and U1D are used in active tuned circuits.2 The active design eliminates the need for inductors.

Now, instead of the center (or resonant) frequency being controlled by ***LC*** values, C20 and 21 determine the center frequency. R21 and 22 also affect the center frequency. I kept their value constant, since they

table 1. Suggested values for the active tuned circuits. Center (resonant) frequencies have been determined using the equation in *fig.* 2.

filter	resonant frequency (Hz)	capacitor (fig. 2)	value (μF)
low frequencies	335	C21	0.2
		C21	0.01
	410	C20	0.15
		C21	0.01
	521	C20	0.15
		C21	0.0062
	1299	C30	0.05
		C31	0.003
mid-range frequencies	1412	C30	0.047
		C31	0.0022
	1959	C40	0.033
		C41	0.002
high frequencies	2905	C40	0.033
		C41	0.001
	3558	C40	0.02
		C41	0.001



C10	0.001	R20	50k pot
C11	180 pF	R21	1k
C12	180 pF	R22	100k
C13	0.001	R30	50k pot
C20,21,		R31	1k
30,31,	see table 1	R32	100k
41		R40	50k pot
FB	ferrite beads	R41	1k
R10	34k	R42	100k
R11	470k	S1	4pdt switch
R12	6.2k	U1	LM324
R13	300k		bypass caps all 0.001

fig. 2. Schematic of the VBE. A single LM324 operational amplifier IC is the active device. The equation is used to determine the center (resonant) frequency of each active tuned circuit. Using active rather than passive tuned circuits eliminates the need for bulky and expensive inductors.

also control the *Q.* The circuits of U1B, U1C, and U1D are identical except for the center frequencies, which are controlled by the capacitor values. The center (resonant) frequency, can be computed by the equation shown in **fig.** 2.

In **table 1** I've provided some suggested component values and center frequencies for **fig.** 2. Use 10 per cent or better tolerance values of C to ensure that the center frequencies arrive at computed values.

calculator input

Table 2 is an algorithm for TI30 or similar calculators. If you decide to experiment with the formula in **fig.** 2, the calculator algorithm will save time and work. Whether you work longhand or use a calculator, remember that the C values in the equation are in *farads;* the typical circuit values are in *microfarads* (μF). Remember also to divide the μF value by 10^6 to convert to farads before substituting into the equation.

bandpass filters

The VBE requires three **bandpass** filters. The problem is to generate a **bandpass** characteristic using only one op amp and a minimum number of related components. At first glance, the circuit of **fig.** 2 might appear to be a highpass and a lowpass section. A bandpass characteristic can be generated by the circuit of **fig.** 2 because, at sufficiently high *Q*, both low and highpass active circuits exhibit bandpass functions.[3] The peaks shown in **fig.** 3 illustrate the similarity to a bandpass function that the lowpass and highpass functions can exhibit. The circuits of U1B, U1C, and U1D (**fig. 2**) each generate a function similar to those of **fig.** 3 and **fig.** 1. To combine the three active circuits, I use each to independently control the response of a master amplifier, U1A. Its output reflects the constant *Q* peak generated by U1B, U1C, and U1D. The characteristics of the three tuned circuits remain constant. Only the response of U1A is modified by the degree to which U1B, U1C, and U1D are inserted. The VBE circuit gives ± 12 dB of adjustment with minimal interaction between controls.

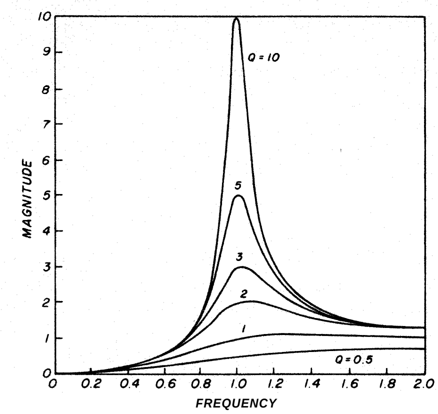


fig. 3. Typical response curve of an active circuit. As *Q* increases, the circuit approaches bandpass-filter characteristics. (From reference 3.)

construction

Construction of the voice-band equalizer is divided between the single PC board and the box that encloses it. Most of the components are mounted on the board. The layout shown in **fig.** 4 is compact and contributes to easy assembly. Ample space is provided for components of many different sizes. The single quad op amp (LM324) greatly simplifies layout and parts placement. You'll have no difficulty building the VBE if you follow the suggested layout. If you choose to experiment with single or dual op amp packages, you'll have to duplicate plus and minus power-supply lines as well as the ground connections.

Follow the general layout in either case to simplify construction and ensure predictable results. Be sure to use good electronic and mechanical techniques. Use a low wattage soldering iron, 60/40 solder, and watch for solder bridges, especially near the IC socket.

table 2. Algorithm for determining center (resonant) frequency of each VBE tuned circuit. The algorithm is for a TI30 (or later) calculator.

enter	press	display	notes
R1	x	R I	R in terms of ohms
R2	=	R1R2	the calculator has automatically converted to scientific notation
	STO	R1R2	place R1R2 in memory
	REC	R1R2	check memory
	CL	0	clear
C1	EE	C1	converting to scientific notation
	+/−		negative exponent
6	x	C1 $\times 10^{-6}$	C in terms of farads. (Don't press = yet.)
C2	EE		converting again
	+/−		
6	=	C1 x C2	C1C2 product in scientific notation
	x		
RCL(R1R2)	RCL		R1R2 from memory
	=	R1R2C1C2	in scientific notation
	$\sqrt{\ }$	$\sqrt{R1R2C1C2}$	square root
	x		times
2	x		two, times. . .
π	=	$2\pi \times \sqrt{\ }$	pi
	=	F_0	F_0 (center frequency) in scientific notation
	x		
1	INV		
	EE		convert back to Hz
	=	F_0 = Hz	your answer

fig. 4. VBE circuit board showing foil side.

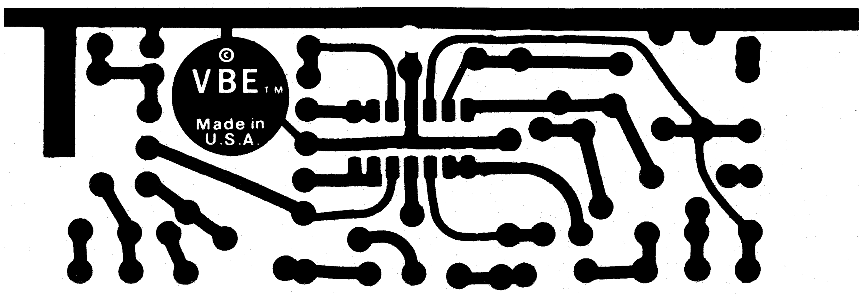
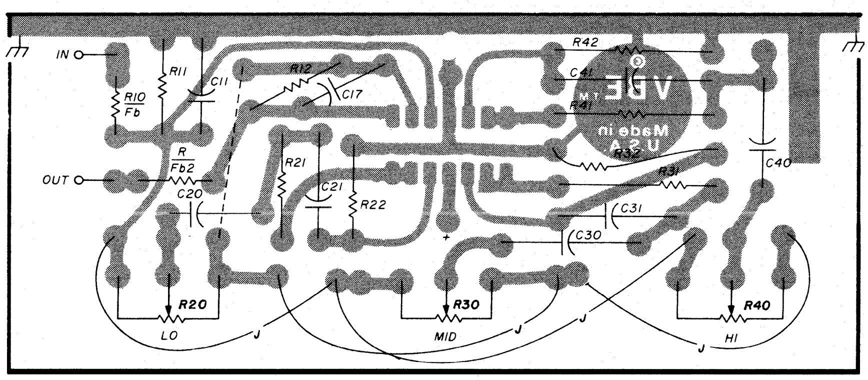


fig. 5. Component placement for the VBE. Control pot jumpers are mounted on the board but are shown off the board for clarity. Switch SW1 mounts onto the chassis as shown.



I started to build the VBE with the printed circuit board.4 Etch and drill the PC board. I drilled the nine control-pot connection points to accept small eyelets. I used the three pots to provide a convenient way to mount the board in the box (the copper foil is not strong enough for this). The eyelets provide a strong mechanical connection for the pots. I found this method convenient, but you can use wires for mounting the pots and other means to mount the PC board. I used a good quality 14-pin IC socket for the LM324 op amp. Molex pins or even direct soldering would work too. Next, I mounted the resistors and the jumper according to the parts placement shown in fig. 5. The jumper runs under C20.

Include ferrite beads on the legs of R10 and R13. I installed the capacitors next, and then the jumpers for the control pots. I did not mount the controls until after the holes were drilled into the box. Since I soldered the controls directly to the eyelets on the circuit board, the holes in the box must line up exactly with the controls that have been soldered to the circuit board.

Once the holes have been drilled for the controls, place the pots loosely into the holes, from the outside of the box. Now solder the main board to the controls and they will line up for a good fit. Mount the foil side of the board toward the knob side of the controls so that clockwise rotation results in gain.

The smallest box I could find to fit the VBE was 5 x 3 x 1-3/4 inch (12.7 x 7.6 x 4 cm) minibox. The small box requires good construction techniques and a fair amount of care to make all the parts fit correctly. You can use a larger box for easier construction. A 5-1/4 x 3 x 2-1/8 inch (13 x 7.6 x 5 cm) minibox is a good size. Before drilling the box, I placed all the parts in their relative positions to see how they fit together. Be sure to leave enough space for the batteries, switch, indicator LED and cables or connectors that you choose. Beware of the screws that hold the box together! Drill the required holes and mount the connectors, switch, and LED. This is a good time to attach labels to the box.

S1 controls the power and the input-output switching. When you turn on the VBE, you're also placing it on line. In a pinch you can use two separate switches (dpdt) instead of one 4 pdt; just remember to use them both together! I covered the tail of the switch with some electrical tape to prevent accidental short circuits. Remember that the input and the output are connected to the switch at the front of the box. I forgot this and cut some of the wires too short before realizing it and making corrections. Also don't forget the wires for the PTT circuit. It's a good idea to keep as much rf energy as possible out of the active circuitry. Note the use of ferrite beads and bypass caps. I put some beads inside the connectors to and from the rig and microphone for good measure.

working with the circuit

Two good-quality 9-volt batteries will operate your VBE for many hours. Current consumption is on the order of 30 mA at 15 volts for full rated output. Typical quiescent current is less than 2 mA. The LED indicator can draw as much as several times the current of the circuit, so I chose a low-current LED to increase battery life. I used two batteries to supply both plus and minus voltages to the LM324 op amp. The bipolar power supply simplifies circuit design and operation; many single-supply designs require additional components to create an artificial ground and to provide a low-impedance dc return for the input circuit. The minibox has enough space to include an ac power supply as an alternative to battery operation.

Some of the newer solid-state transceivers require that the microphone ground return be separate from the chassis ground. Also, some rigs require a separate ground return for the PTT control lines. Be sure to follow your manufacturer's schematic and instructions explicitly. Misconnecting these critical ground returns can often result in erratic operation or oscillation without apparent cause.

An additional number of filters could be added to the VBE circuit by an enterprising builder. Simply add extra controls and extra active tuned circuits. Adjust their frequency centers by substituting *C* values according to the equation shown in **fig. 2.**

I used an audio sweep generator and scope to check the VBE. Even if you don't have this equipment available, all you need is a variable-frequency signal source and an output indicator.

If you don't have a signal generator you can use your SSB receiver as the signal source. Place the receiver in the calibrate mode and use the audio output as your signal source. Disconnect the antenna to eliminate background noise and keep the af gain low. An oscilloscope is the best output indicator. You could use an ac VTVM or even a standard VOM in a pinch. The VOM will probably give only a relative indication. The VOM is a linear voltage indicator and it doesn't have flat frequency response. Your tape recorder RECORD meter is a good indicator. It has logarithmic calibration and is reasonably flat for the narrow voice band. Sweep the frequencies and note the output on the indicator you're using. I found it useful to plot the levels onto a graph to get a visual feel for the way the controls work. If you monitor the output you'll hear the VBE operate.

using the VBE

The voice band equalizer low-, mid-, and high-range controls each have a total range of about 24 dB. Each control can boost to 12 dB. The same control can also attenuate as much as 12 dB. A full 12 dB is a considerable amount of gain in a relatively narrow bandwidth. Often the addition of gain will require readjusting the transmitter microphone gain. I watch the ALC indicator on my rig to avoid overdriving.

I often find it more useful to use attenuation to produce the desired response. I try to balance the boost then attenuate to keep an even average level. For example, I boost the mid and high ranges, but at the same time roll off the lows. Or sometimes I boost the lows but also roll off the highs. Turning up all three controls together offers no real advantage.

I've also found that listening to a tape recording of my station is a good way to hear the VBE work. The next best thing is to let a fellow ham borrow your VBE so you can hear how it sounds on the air. Reports from other hams can be misleading. On SSB the receiver varies the pitch as you tune a signal. This effect makes it difficult for others to describe the sound of the VBE. Keep in mind that the VBE doesn't add any distortion of its own, and most hams have come to expect some distortion from transmitting audio accessories, Using the VBE is most like having several different microphones to choose from. When I heard a tape played of myself, I could hear the sound of the VBE right away. Hearing the VBE myself helped me know when and how to best make adjustments.

speech processors and the VBE

A normal speech pattern consists of highly varied peaks and valleys of rather low average level. Speech processing is a method of creating a more constant pattern with higher average level; this is why a speech processor seems to increase signal strength. Audio limiters, clippers, compressors and rf processors are methods commonly used. Often speech processing will upset your normal tonal balance, imparting exaggerated qualities: too much bass, too many highs, not enough bass. If you already use a processor, the VBE can give you extra control. By using both a processor and the VBE, you can make changes in the sound of the processed audio.

conclusion

The voice-band equalizer offers a new dimension in control and flexibility. Its straightforward design means that the average Amateur can build it successfully. To assist those who may have difficulty finding parts, I can supply many of them at a nominal cost. If you have any questions concerning the VBE or require information or parts, please send me a letter with a self-addressed stamped envelope for a prompt reply. I'll be glad to answer your questions if I can. The basic circuit of the VBE can be used in many applications limited only by your creative ability.

references

1. Radio Handbook, 20th edition, chapter 3, page 3.15.
2. Linear Applications, Volume I, National Semiconductor Corporation, Application Note AN 31-14.
3. Larry Hutchinson, "Practical Photo Fabrication of Printed-Circuit Boards," *ham* radio, September, 1971, pages 6-18.

ham radio

installing radials

for vertical antennas

A novel approach to the problem of installing radial wires in a grass lawn

Antennas and propagation are always prime topics for discussion by hams. Each ham develops his favorite type of antenna and will readily champion it to any who are willing to listen. Most antenna discussions are a combination of myth and reality, success and failure. I'd like to share my experience with a vertical antenna.

the vertical antenna

The mention of vertical antennas brings many thoughts to mind. How many have heard these or similar comments from time to time: "A vertical antenna radiates equally poorly in all directions." "You have to copperplate your backyard to make it work at all." "I just nailed it to my fence post and got DXCC in six months."

I'm neither going to make any fantastic claims, nor tell a tale of failure. I will describe a technique of stringing radials that I used as a solution to my vertical antenna installation problems.

The antenna I used was the Hustler 4BTV. This antenna can be used with or without radials when ground mounted. I preferred to use radials, but how many and how long? To answer these questions I researched the literature on vertical antennas. Most of this material indicated that many radials produced the best performancce. I was limited by my lot size in the length of radials I could run. Two articles were of particular interest to me because of the problems I was attempting to solve.[1,2] Using the information in these articles, I decided to use 14 radials each 25 feet (7.6 meters) long. Now I had to decide how to install them.

installation

Installation of radials is another topic that has many "best ways." One method is to bury the radials, while another is to place the radials on top of the earth. Burying the radials defeated my purpose, and placing them on top presented a safety hazard to neighborhood children. I needed something in between.

I derived my method from two completely different bits of information. The first was from an antenna ar-

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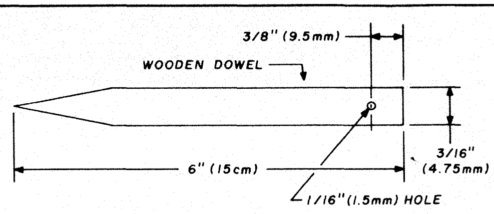

fig. 1. Threading tool made from a piece of wood dowel.

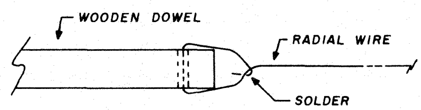

fig. 2. Method for securing radial wire to the wooden threading tool, or "needle."

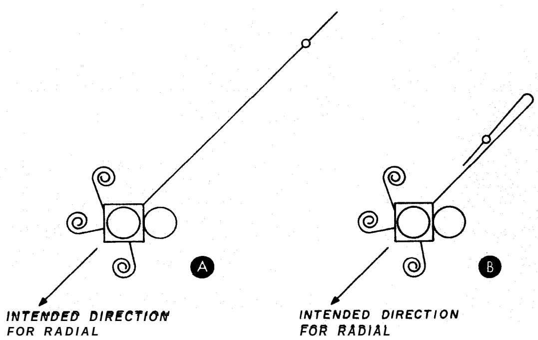

fig. 3. Preparing radials for threading into a grass surface. Each radial is uncoiled and laid on the ground in a direction opposite to that in which it is to be laid along the earth, (A). Sketch (B) shows how the wooden needle is placed underneath the antenna base and pointed in the desired direction. These views are shown looking down on the antenna base.

ticle in which the radials were laid on the ground and grass seed sown over them: when the grass grew, the radials were held under the grass so that you could walk on them, or even mow the lawn, and not disturb the radials. My lawn was in place and I didn't want to strip it out and replant it. However if I could place radial wires along the ground ***under*** the grass, perhaps the safety advantage was attainable.

The second piece of information was my memory of how I lost many arrows while an archery student many years ago. When I missed the target, the arrows struck the ground at a very shallow angle and traveled along the earth beneath the grass, sometimes becoming completely buried. If I could fashion a tool to penetrate the grass like an arrow perhaps it could pull the radial wires behind it under the grass.

The tool I developed is similar to a large sewing needle. I used pieces of 3/16-inch (4.75 mm) diameter hardwood dowel cut 6 inches (15 cm) long. I sharpened one end of each dowel in a pencil sharpener; then I drilled a 1116-inch (1.6 mm) diameter hole through the dowel 318 inch (9.5 mm) from the end (fig. 1). I made 14 of these needles — and was then ready to begin laying the radials.

Each radial consisted of a length of No. 22 (0.6 mm) copper wire, wire obtained from a 50-foot (15-meter) spool of twisted antenna wire (Radio Shack catalog no. 278-1329). This twisted wire, which is made of seven strands of No. 22 (0.6-mm) wire, was cut in half, resulting in two 25-foot (7.6-meter) lengths. I soldered one end of each length to a spade lug for attachment to the antenna base. I then untwisted each length into seven individual strands and coiled them to prevent tangles. I threaded the free end of each strand through the hole in the wooden needle, looped it back upon itself to form a bridle, then soldered, (fig. 2).

When all of the needles were attached, I connected the spade lugs to the antenna base mount. I laid seven radials at a time to avoid congestion at the antenna base. This is the point of this article: threading the radials into the lawn.

laying the radials

I began by uncoiling a strand of wire and laying it on the ground in a direction opposite to that in which it is to be sewn, (fig. 3A). Double the strand back on itself so that the wooden needle is under the base of the antenna, (fig. 3B). Now push the needle through the grass and along the earth, making sure it points in the direction the radial is to lay. Work the wooden needle along under the grass with your fingers, and the wire will slowly follow behind it. (A helper is handy during this operation to prevent kinks from being pulled into the wire as it loops back on itself.) When the entire length of the radial is threaded, simply raise the needle up far enough to turn it so that it points downward. Then push it into the earth, thus anchoring the radial.

some useful hints

A little time spent in preparation of the lawn will make the task of threading radials easier. Cut and rake the lawn before working with the radials. This will reduce the mass of grass through which you must work the needle. After installing the radials, rake the surface ***in line*** with the radials — not across them. My lawn mower is set for a cutting height of 2-112 inches (6.4 cm). None of the radials has been disturbed during mowing.

references

1. Sevick, "The Ground-Image Vertical Antenna," ***The ARRL Antenna Anthology***, pages ***22-24.***
2. Stanley, "Optimum Ground Systems tor Vertical Antennas," ***QST, December, 1976.***

ham radio

A CW keyboard using the APPLE II computer

Program listing and simple interface circuit for using this popular computer with your Amateur station

The APPLE II computer has numerous possibilities for simplified interfacing to external devices. Four 74LS flip-flop outputs are available that are programmed to set or reset. Three inputs sense whether the data is TTL zero or 1. Four other inputs return a number 0-255, depending upon the series resistance of a 150k pot.

As an example of what can be done, I drive an IDS IP-225 printer at 1200 baud from the GAME I/O socket, although the APPLE serial interface won't operate above 600 baud. The AN0 output provides serial data, while SWØ accepts "handshaking" or CTS signals from the printer. With the four flip-flop outputs fed to a 4-bit decoder, up to 16 circuits can be controlled by the computer.

CW keyboard

Converting the APPLE II to a CW keyboard is quite simple. A program for this follows, with a circuit for interfacing to a relay driver (fig. 1). In the following discussion, $ denotes hexadecimal numbers, with decimal equivalents in parentheses (see program listing).

Subroutine SBR 5 forms a dash, SBR 6 forms a dot, SBR 7 provides a short space between dots and/or dashes, while SBR 8 inserts a long space after formation of each numeral or letter.

The keyboard is read in line 300 until a key is pressed. The test for numeral 0-9 or letter A-Z is in lines 376-418. Note that there's a relationship between line numbers and the ASCII code read from the keyboard. For example, A = ASCII 193 (with bit 7 set). The test for "Is it an A?" is in line 393. If it is an A, the program jumps to line 193, where the dot, space, dash SBRs are called. The program then returns to line 300 to read the next key. That's all there is to it.

The program sends numerals or letter only. Obviously, punctuation and special characters for AR, SK, etc., could be included. (But see line 420.) String input, then reading ASC for each string character in turn at line 300, would be a simple modification. (See page 89 of the APPLE Basic Manual for an equivalent of MID$.) These refinements are not included in my program because I use a bug for serious CW work. However, if a good keying relay is used, the CW is quite acceptable. (See notes in fig. 1.)

A machine-language listing, resulting from the POKE statements in lines 10-23 follows the basic listing. The number selected for speed is loaded into $306 (774), although this location is initially loaded with $FF (255). Note that a simple modification of LDA $CØ3Ø, following 0308 DØ FD would provide for sidetone from the APPLE speaker (CØ3Ø is the speaker address, of course). Follow this with 88 DØ F5 AD 58 CØ 60. Conversion to POKE statements is left as an exercise for the user.

interface/relay driver

The interface comprises an emitter follower, mounted directly on the 16-pin plug that plugs into the GAME I/O socket. Pins 1, 16 are toward the front of the computer, so the two-wire output cable is brought straight back from pins 8, 9 through one of the slots on rear wall.

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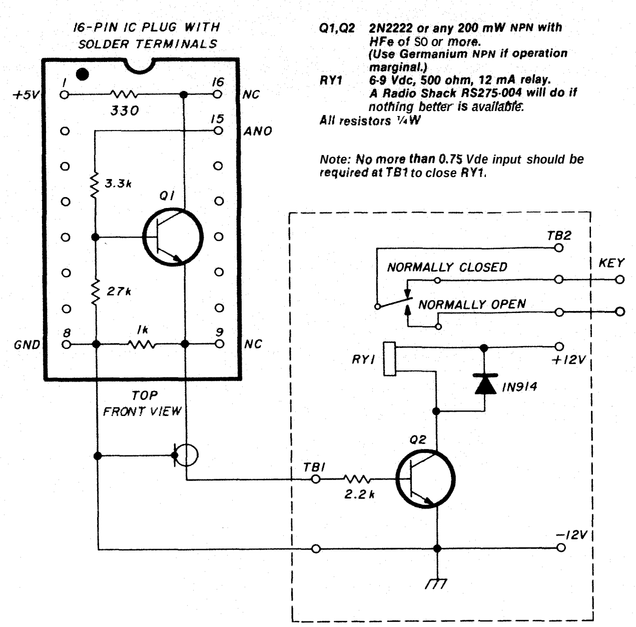


fig. 1. Interface circuit for the APPLE II computer and a CW keyboard.

The emitter-follower provides 0.8-volt output. The relay driver should be tested to confirm that the relay closes with not more than 0.75 volt. If marginal, the use of a Germanium NPN transistor for *Q2* is suggested.

The Radio Shack relay is listed only because of ready availability. If you have a better relay with the same characteristics, by all means use it. The +5 Vdc for the emitter follower is from the APPLE supply (pin 1). The driver uses 12 Vdc. Although this voltage is available at pin 50 of the peripheral sockets — or at the power-supply socket — it's not readily accessible. Inclusion of a small 12-Vdc, 12-mA supply on the relay driver chassis is probably preferable.

For those not familiar with LS chips, *do not* plug anything in with power on. *Do not* reach inside the APPLE cabinet without first grounding yourself by touching the power supply shield.

radio-frequency interference

Judging by what I hear on the ham bands from other computer hobbyists, and by articles in other Amateur publications, there appears to be a real RFI, or "hash" problem with some computers. Fortunately, this is not the case with the APPLE computer. I operate mine within 6 inches (15 cm), or less of the receiver. No hash occurs on 80 or 40 meters except for weak subharmonics of the 14-MHz crystal oscillator. (I also run a color TV set on rabbit ears within 6 feet (2 meters) of the computer. You can't do this with most computers.)

Program listing for the CW keyboard using the APPLE II computer:

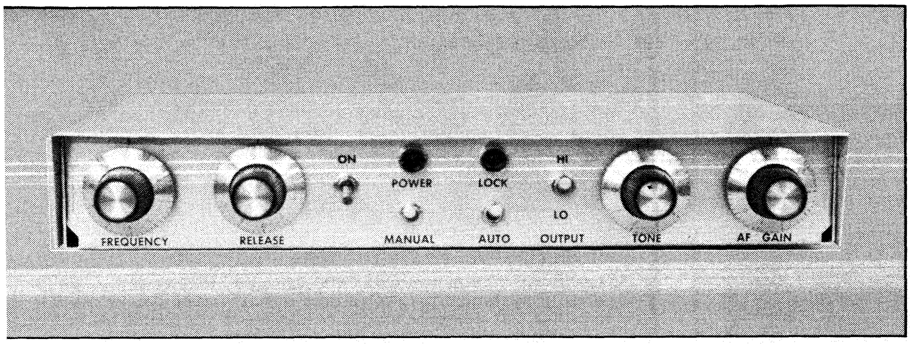
```
>LIST
  1 GOTO 10
  5 POKE 774,SP: CALL 768: RETURN
  6 POKE 774,SP/3: CALL 768: RETURN
  7 FOR X=1 TO SP/4: NEXT X: RETURN
  8 FOR X=1 TO SP/2: NEXT X: RETURN
 10 POKE 768,173: POKE 769,89
 12 POKE 770,192: POKE 7719160
 13 POKE 772,255: POKE 7739162
 14 POKE 774,255: POKE 775,202
    REMX(774) = speedselected.
 16 F'OKE 776,208: POKE 777,253
 18 POKE 778,136: POKE 779,208
 20 F'OKE 780,248: POKE 781,173
 22 POKE 782,88: POKE 7839192
 23 POKE 784,96
 25 GOTO 500
176 GOSUH 51 GOSUH 7: GOSUB 5: GOSUH
    7: GOSUH 5: GOSUB 7: GOSUB
    5: GOSUB 7: GOSUB 5: GOSUB
    8: GOTO 300                    REM0
177 GUSUB 6: GOSUH 7: GOSUB 5: GOSUH
    7: GOSUB 5: GOSUB 7: GOSUH
    5: GOSUB 7: GOSUH 5: GOSUH
    8: GOTO 300
178 GUSUH 6: GUSUB 7: GUSUH 6: GOSUH
    7: GOSUB 5: GOSUB 7: GOSUB
    5: GOSUB 7: GOSUB 5: GOSUH
    8: GOTO 300
179 GOSUB 6: GOSUB 7: GOSUB 6: GOSUB
    7: GOSUB 6: GUSUB 7: GUSUH
    5: GOSUB 7: GOSUB 5: GOSUH
    8: GOTO 300
180 GOSUB 6: GOSUB 7: GOSUB 6: GOSUH
    7: GOSUH 6: GOSUH 7: GOSUB
    6: GOSUB 7: GOSUB 5: GOSUH
    8: GOTO 300
181 GOSUB 6: GOSUB 7: GOSUH 6: GOSUH
    7I GOSUB 6: GUSUB 7: GOSUB
    6: GOSUB 7: GOSUB 6: GOSUH
    8: GOTO 300
182 GOSUB 5: GOSUB 7: GOSUB 6: GOSUB
    7: GOSUB 6: GOSUB 7: GOSUH
    6: GOSUB 7: GOSUB 6: GOSUB
    8: GOTO 300
183 GOSUB 5: GOSUB 7: GOSUH 5: GOSUH
    7. GOSUB 6: GOSUB 7: GOSUH
    6: GOSUB 7: GOSUB 6: GOSUB
    8: GO70 300
184 GOSUB 5: GOSUB 7: GUSUH 5: GOSUB
    7: GOSUB 5: GOSUB 7: GOSUH
    6: GOSUB 7: GOSUB 6: GOSUB
    8: GOTO 360
185 GOSUH 5: GOSUB 7: GOSUB 5: GOSUH
    7: GOSUB 5: GOSUB 7: GOSUB
    5: GOSUB 7: GOSUH 6: GOSUB
    8: GO?O 300: REM  9            REM9
190 REM  LETTERS FOLLOW
193 GOSUB 6: GOSUB 7: GUSUB 5: GOSUH
    8: GOTO 300      :REM A
194 GOSUB 5: GOSUH 7: GOSUB 6: GOSUB
    7: GOSUB 6: GOSUB 7: GOSUB
    6: GOSUB 8: GOTO 300: REM  B
195 GOSUB 5: GOSUB 7: GUSUH 6: GUSUH
    7: GOSUB 5: GOSUH 7: GOSUH
    6: GOSUB 8: GOTO 300: HEM  C
196 GUSUB 5: GOSUB 7: GOSUH 6: GOSUH
    71 GOSUB 6: GOSUH 8: GOTO 300
    : REM  D
197 GOSUH 6: GOSUH 8: GOTO 300
198 GOSUH 6: GOSUB 7: GOSUB 6: GOSUB
    7: GOSUB 5: GOSUB 7: GOSUB
    6: GOSUB 8: GOTO 300: REM  F
```

```
199 GOSUF 5: GOSUB 7: GOSUF 5: GOSUB
    7: GOSUB 6: GOSUB 8: GOTO 300

200 GOSUB 6: GOSUB 7: GOSUB 6: GOSUH
    7: GOSUB 6: GOSUF 7: GOSUB
    6: GOSUB 8: GOTO 300: REM  H
201 GOSUB 6: GOSUB 7: GOSUB 6: GOSUB
    8: GOTO 300
202 GOSUB 6: GOSUH 7: GOSUB 5: GOSUB
    7: GOSUB 5: GOSUP 7: GOSUH
    5: GOSUP 8: GOTO 3003 REM  J
203 GOSUB 5: GOSUB 7: GOSUB 6: GOSUB
    7: GOSUEI 5: GOSUB 8: GOTO 300
    : REM  K
204 GOSUH 6: GOSUH 7: GOSUH 5: GOSUH
    7: GOSUB 6: GOSUB 7: GOSUH
    6: GOSUB 8: GOTO 300: REH  L
205 GOSUB 5: GOSUH 7: GOSUR 5: GOSUR
    8: GOTO 300: REM  M
206 GOSUB 5: GOSUEI 7: GOSUB 6: GOSUR
    8: GOTO 300: KEM  N
207 GOSUH 5: GOSUB 7: GOSUH 5: GOSUB
    7: GOSUB 5: GOSUB 8: GOTO 300
    : REM  O
208 GOSUB 6: GOSUB 7: GOSUH 5: GOSUH
    7: GOSUH 5: GOSUB 7: GOSUB
    6: GOSUB 8: GOTO 300: KEH  P
209 GOSUH 5: GOSUH 7: GOSUB 5: GOSUB
    7: GOSUB 6: GOSUB 7: GOSUB
    5: GOSUB 8: GOTO 300: REM  Q
210 GOSUH 6: GOSUH 7: GOSUB 5: GOSUB
    7: GOSUH 6: GOSUB 8: GOTO 300
    : REH  R
211 GOSUB 6: GOSUB 7: GOSUB 6: GOSUB
    7: GOSUB 6: GOSUB 8: GOTO 300
    : REM  S
212 GOSUB 5: GOSUB 8: GOTO 300:
    REM  T
213 GOSUB 6: GOSUB 7: GOSUH 6: GOSUH
    7: GOSUR 5: GOSUH 8: GOTO 300
    : REM  U
214 GOSUH 6: GOSUH 7: GOSUR 6: GOSUB
    7: GOSUR 6: GOSUB 7: GOSUH
    5: GOSUB 8: GOTO 300: KEM  V
215 GOSUH 6: GOSUH 7: GOSUB 5: GOSUR
    7: GOSUB 5: GOSUB 8: GOTO 300
    : REM  W
216 GOSUH 5: GOSUB 7: GOSUH 6: GOSUB
    7: GOSUH 6: GOSUB 7: GOSUB
    5: GOSUEI 8: GOTO 300: REH  X
217 GOSUH 5: GOSUH 7: GOSUB 6: GOSUB
    7: GOSUB 5: GOSUB 7: GOSUB
    5: GOSUEI 8: GOTO 300: REH  Y
218 GOSUH 5: GOSUB 7: GOSUH 5: GOSUB
    7: GOSUH 6: GOSUB 7: GOSUR
    6: GOSUH 8: GOTO 300: REM  Z
300 KB= PEEK (-16384)
302 IF KB<128 THEN 300
304 POKE -16360,0
310 REM  BY W. S. SKEEN - W6WR FEB 7
    9
376 IF KB=176 THEN GOTO 176 :REM 0
377 IF KB=177 THEN GOTO 177
378 IF KB=178 THEN GOTO 178
379 IF KB=179 THEN GOTO 179
300 IF KB=180 THEN GOTO 180
381 IF KB=181 THEN GOTO 181
382 IF KB=182 THEN GOTO 182
383 IF KB=183 THEN GOTO 183
384 IF KB=184 THEN GOTO 184
385 IF KB=185 THEN GOTO 185 REM 9
393 IF KB=193 THEN GOTO 193: REM  A
374 IF KB=194 THEN GOTO 194
395 IF KB=195 THEN GOTO 195
396 IF KB=196 THEN GOTO 196
397 IT KB=197 THEN GOTO 197
398 IF KB=198 THEN GOTO 198
399 IF KB=199 THEN GOTO 199
400 JF KB=200 THEN GOTO 200: KEM  H
401 IT KB=201 THEN GOTO 201
402 IF KB=202 THEN GOTO 202
403 IF KB=203 THEN GOTO 203
404 IF KB=204 THEN GOTO 204
405 IF KB=205 THEN GOTO 205
406 IF KB=206 THEN GOTO 206
407 IF KB=207 THEN GOTO 207
408 If KB=208 THEN GOTO 208
409 IF KB=209 THEN GOTO 209
410 IF KB=210 THEN GOTO 210: REM  R
411 IF KB=211 THEN GOTO 211
43.2 IF KB=212 THEN GOTO 212
413 IF KB=213 THEN GOTO 213
414 IF KB=214 THEN GOTO 214
415 If KB=215 THEN GOTO 215
416 IF KB=216 THEN GOTO 216
417 IF KB=217 THEN GOTO 217
418 IF KB=218 THEN GOTO 218
420 IF KB<176 OR KB>218 THEN 300

490 END
500 CALL -936
510 FRINT : PRINT "     CW KEYBOARD
     (RELAY DRIVER)"
520 PRINT : FRINT "            BY W6W
    R"
530 FRINT : PRINT "  SELECT SPEED BY
     ENTERING"
540 PRINT : PRINT "  A NUHHER 75 TO
    255."
550 PRINT : PRINT "  100 = APPROX, 2
    0 - 25 WPM"
560 PRINT : PRINT "  255 = VERY SLOW

580 PKINT : PRINT "  75 = FAST"

600 F'RINT : INPUT "  NOW PLEASE SELE
    CT SPEED",SP
610 PRINT : PRINT "  READY TO SEND"

620 POKE 774,SP
630 GOTO 300
```

: **Machine-Language Program.**

```
*300L

0300-  AD 59 C0    LDA  $C059
0303-  A0 FF       LDY  #$FF      REM
0305-  A2 64       LDX  #$64      Set AN0
0307-  CA          DEX            DLY2
0308-  D0 FD       HNE  $0307     DLY/
030A-  88          DEY            DLY/
030B-  D0 F8       HNE  $0305     DLY2
030D-  AD 58 C0    LDA  $C058     Reset AN0
0310-  60          RTS            Return to Basic
```

REMARK: In the example above, $64 (Speed = 100) has been located in $306(774), in preparation for a dash.



CW regenerator

for Amateur receivers

Shake hands
with the "Golden Articulator"
kick back, and enjoy
near-perfect CW reception

Interference eliminated, superb audio, mesmerizing CW regeneration – these words are in the minds of every devoted CW operator. Well brace yourselves fellow Amateurs; these words are about to be realized. The Golden Articulator is here and you can prepare for a thrilling adventure in CW regeneration.

features

No longer will your headphones be filled with multiple interfering signals, because this device contains useful features. These include:

1. A frequency acquisiton adjustment with a range of **400-**1800 Hz.

2. A variable release tie between 1-20 seconds for various incoming code characteristics. The circuit will automatically return to normal audio after the conclusion of incoming CW (very useful when my tube-type transceiver was warming up). This feature can be aborted with the front-panel RELEASE control.

3. Adjustments for pitch and gain of the internal tone oscillator. The pitch control has a range of 280-800 Hz. This feature was included as an alternative to monotonous single-tone copying.

physical description

Additional front-panel controls include a) power ON/OFF switch with LED; b) a LED that operates in agreement with incoming CW when the PLL is locked; and c) an OUTPUT impedance selector giving a choice of either **8** ohms or 1 kilohms.

The rear panel supports the audio input jack and the two audio output connections. The power supply connection uses a twisted pair of insulated wires fed through the factory-produced opening in the chassis corner.

The internal layout can be seen in the photo. Ample space is provided for all components by using the low-profile, cut-down chassis design. Obvious in the photo is the absence of an internal power supply. My unit uses an external power source consisting of two batteries: a 6-volt lantern battery in series with a 1%-volt D cell. (The additional D cell was required to pull in relay K1.) An internal ac supply[1] can be included if desired **(fig. 1)**.

operation

Operation of the Golden Articulator begins with receiver audio connected to its input jack **(fig. 2)**. With headphones or a speaker plugged into the output jack, receiver audio can be monitored even if the unit is off. When the power is switched on, **C1** feeds audio to the activated LM-567 PLL tone decoder; simultaneously audio is connected to the output device through C2 and KIA. By varying frequency control R2, you can select the particular CW signal you wish to regenerate. When the PLL is locked, the lock LED will operate in agreement with the chosen incoming signal. The resultant digital signal produced on pin **8** of the **LM567** is applied to the 555 delay and 7413 in-

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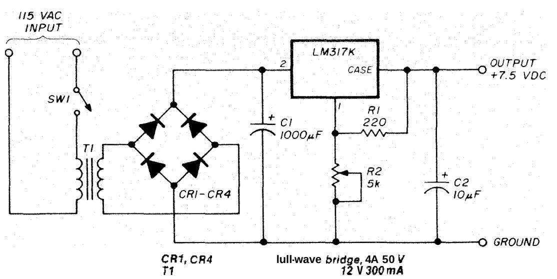


fig. 1. Optional internal regulated ac power supply.

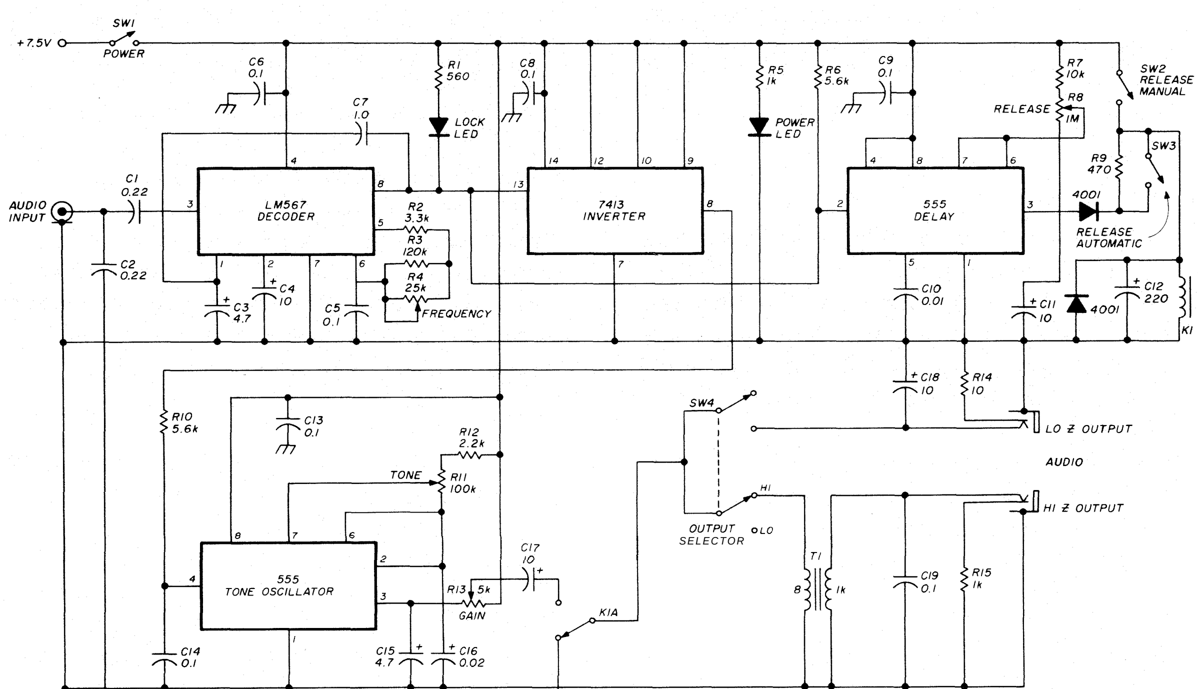
verter circuit inputs. The inverter is required because a positive trigger is needed by the 555 tone oscillator, which at this time is activated and output to the open contact of K1A.

auto release

You now have a decision to make. If auto-release switch SW3 is momentarily depressed, relay K1 will energize and K1A contacts will transfer. This action connects the internal tone oscillator to the output device and voila! — those dream words come true, producing tape-quality interference-free copy. In the auto-release mode, front-panel variable-release control R8 must be adjusted for the incoming CW keying characterisics. This adjustment controls the holding time of K1 and should be set slightly longer than spaces or pauses, whichever is greater. Additionally, at the end of an incoming transmission, the 555 delay will begin its last timing cycle. At its conclusion it will automatically release K1, which returns the circuit to normal receiver audio. This transfer will usually occur during the early seconds of transmission.

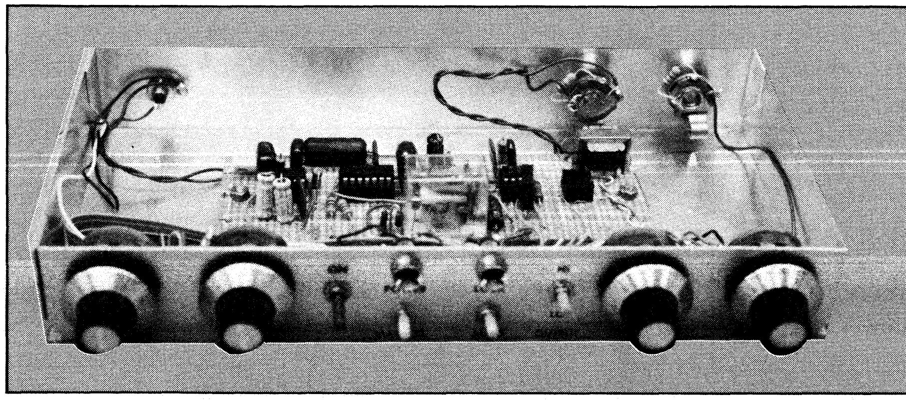
incoming-signal drift

The beauty of this feature lies in the fact that, if the receiver is prone to drift, the signal can be first verified as still on frequency, otherwise the headphones may be filled with emptiness. If the signal has drifted off frequency, as indicated by the extinguished lock LED, adjust the main tuning of the receiver — or better yet, slightly vary frequency control R4 while monitoring the lock LED for acquisition. Then depress the



Notes:
1. SW3 is spring return type
2. K1A shown in normal position.
3. R13 is audio-taper pot; all others are linear
4. T1 is an audio transformer, 8-ohm primary; 1k ohm secondary.
5. K1 is a miniature spdt relay. Coil resistance 500 ohms; current 12 mA

fig. 2. Schematic of the Golden Articulator. Capacitor C7 was chosen to eliminate chatter from the LM567 output at moderate-to-high code speeds. Capacitor C12 across K1 coil eliminates relay dropout during reception of critical information. Caps C15, C18, and C19 tame the 555 oscillator leading-edge spikes to produce a pleasing CW tone.



Internal view of the circuit. Ample space is available for an internal power supply.

auto release and you're back in business. This process sounds time consuming, but in reality it takes only a few seconds.

manual release

As mentioned earlier, you have a decision to make in the selection of release modes. The second choice is manual operation. By switching manual release SW2, K1 is immediately connected to the power supply. To use this feature the incoming signal must be frequency stable. This capability was additionally useful during checkout of the unit, thereby avoiding the repeated dropout of K1. Incidentally, while on the subject of checkout, I used the calibration signal from my transceiver as a very convenient variable audio source.

measurements

The bandwidth of the tone decoder was measured over its input frequency range of 400-1800 Hz. As shown in table 1, the bandwidth varied between 40-120 Hz over the range. The bandwidth can be shifted, if desired, by changing PLL loop filter capacitor C4. A smaller value will widen the bandwidth and vice versa.

Input sensitivity was measured for the same input frequency range. A value of 28 mV rms sine wave input was required for the LM567 to produce a stable lock and transfer its output state high to low.

precautionary notes

The output circuit of the LM567 will develop chatter* when C3 is relatively small. This phenomenon is a result of the output stage moving through its threshold more than once after lock. At moderate-to-high code speeds this chatter severely disrupted the input triggering for the 555 tone oscillator. To remedy this situation, capacitor C7 was connected between output pin 8 and output-filter pin 1 of the LM567. This eliminated the switching transient, thereby cleaning the trigger pulse to the tone oscillator.

One other critical component is capacitor C12 connected across K1 coil. This capacitor's charge holds the coil energized during periods when pin 1 of the 555 delay is positive ***and*** the delay has just timed out. Admittedly, this condition doesn't occur very often. But it becomes a nuisance when K1 drops out during reception of critical information. A value of 220 μF was sufficient to overcome this problem.

The 555 tone oscillator produced a displeasing square wave output, which needed some help to produce a pleasant tone. With the addition of C15, C18, and C19, the annoying leading-edge spikes were removed and sufficient waveform rounding was induced to produce a pleasant tone.

construction

Fabrication of the unit began with the modification of a standard 5 x 10 x 3 inch (12.7 x 25.4 x 7.6 cm) aluminum chassis. The bottom of the chassis was removed, leaving an open box measuring 5 x 15 x 1% inches (12.7 x 38.1 x 3.8 cm). This low-profile design gives a streamlined appearance to the unit, accompanied by the cover and four graduated knobs. The switches are packaged with several different colored

Rear view showing low- and high-impedance output jacks and audio-input jack. Note twisted-pair power-supply wires in lower right corner.

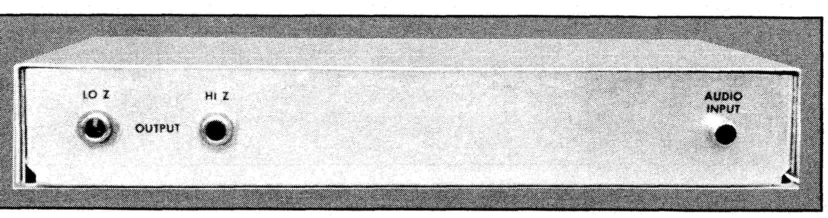


table 1. Tone-decoder bandwidth measurements.

F center (Hz)	F low (Hz)	F high (Hz)	BW (Hz)
400	380	420	40
600	580	630	50
800	775	835	60
1000	970	1040	70
1200	1150	1230	80
1400	1370	1460	90
1600	1540	1650	110
1800	1750	1870	120

Notes:
1. Loop filter C4 10.0 μF.
2. Output filter C3 4.7 μF.
3. F low data taken when lock LED just energizes with increasing input frequency.
4. F high data taken when lock LED just de-energizes with increasing input frequency.

toggle slip-ons. To aid in switch identification I used one of each color.

Board layout, as seen in the lid-off photo, follows the circuit diagram with parts being ordered left to right. The board is spaced from the chassis bottom with ¼ inch (6.4 mm) spacers leaving adequate space above the highest part; that is, K1.

The right side of the board shows an empty socket. The original design used this socket with an audio-amplifier circuit. Experimentation indicated that this circuit was really not required because the 555 tone oscillator could drive either a speaker or headphones with more than enough audio. The only loss was the convenience of gain adjustment for normal receiver audio, but the transceiver gain control served the same purpose.

concluding remarks

Operation has been a pleasant experience in CW copying. With the 7.5-volt battery supply, total current drain is 150 mA. Obviously battery life will depend on use of the equipment, but a conservative estimate would be 3-4 months.

The greatest obstacle I encountered in phasing this unit into my operating habits was the pronounced absence of anything but one signal in my headphones. After all, some of us old timers will find it difficult to accept the fact that there are other ways besides listening to headphones filled with earsplitting interference and noise.

references

1. *Voltage* Regulator Handbook, National Semiconductor, 1977 edition.
2. Howard M. Berlin. Design of Phase-Locked Loop Circuits with Experiments, Howard W. Sams & Co., Inc., Book No. 21545, 1st edition, 1979.

ham radio

geometry of Phase III Spacecraft orbits

Despite the failure of AMSAT Phase III on May 23, **1980,** AMSAT officials have urged publication of this article. Their thinking is that a vast majority of the Amateur fraternity was, at launch time, "too far down the learning curve" to have been effective users of the new satellite. We must begin now to prepare for the next satellite, scheduled for launch in **1982.**

I have just completed a project for AMSAT in which I wrote a BASIC tracking program for the Phase-III Spacecraft elliptical orbits.* While this project is still fresh in mind I'd like to share some facts I've gleaned about the first of these satellites, Phase III-A, which was to be launched in May, **1980.** My source for this data is Rich Zwirko, K1HTV, Vice President of Operations, AMSAT. This satellite, after the kick motor is fired, will have the following orbital characteristics:

a = ***25028 km*** (length of the semi-major axis)
e = ***0.6852*** (Eccentricity)
i = ***57 degrees*** (Inclination)
P = ***655 minutes*** (Period)
ω = ***210 degrees*** (Argument of perigee)

*This program listing is available from ham radio on receipt of a self-addressed, stamped, 8 × 11 envelope. AMSAT volunteers may make available cassette duping for the PET, TRS-80, AIM65, APPLE, and others. Watch the AMSAT newsletter. Orbit.

(The ***argument of perigee*** will increase about 0.08 degree per day.) These are the projected data if everything goes well. Now, for the uninitiated, I'll explain each one individually.

orbit with respect to earth — definitions

The orbit will be an ellipse (fig. 1), which lies in a plane called the ***orbital plane.*** Once and for all let's agree to view the satellite from "above" from where the satellite appears to move counterclockwise. The Earth will be at one focus of the ellipse. The closest approach to the Earth is called ***perigee;*** the furthest approach ***apogee.*** The perigee, the two focii, the geometric center of the ellipse, and the apogee all lie on a line called the ***major axis.*** Exactly one half the distance from perigee to apogee, that is, the distance from the center of the ellipse to perigee (or apogee), is called the ***length*** of ***the semi-major axis, a.***

The ***eccentricity,*** e, is a measure of ellipse elongation. If $e = 0$, the ellipse is a circle. If e were, say, 0.99, the ellipse would be elongated and very flat.

In fig. **1,** I've tried carefully to show an accurate drawing of the actual shape of Phase III-A orbit. If c is the distance from the ellipse center to either focus, $e = c/a$; so perigee will occur at a distance of $a - c = a(1-e)$ and apogee at a distance of $a + c = a(1+e)$ from Earth center.

By C. R. MacCluer, W8MQW, Post Office Box 1858, East Lansing, Michigan 48823

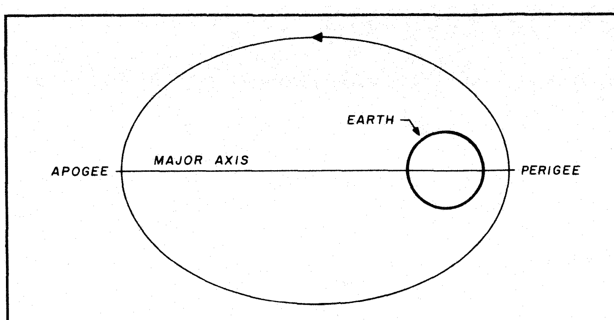


fig. 1. The Earth and the orbit of Phase III-A as drawn by a Tektronix 4051.

The orbital plane is inclined i degrees to the *equatorialplane,* the plane containing the equator of the Earth (**fig. 2**). The *inclination,* i, is the angle between the two upward-pointing normals: one perpendicular to the orbital plane pointing upward and one perpendicular to the equatorial plane (along the axis of rotation) through the north pole.

The period, P, is the time in minutes needed for the satellite to complete one orbit from perigee to perigee.

The last of the orbital characteristics, the *argument of perigee,* is not encountered in circular orbits. This angle, ω, is measured counterclockwise in the orbital plane and is the angle between the *line* of *the ascending node* and perigee (see **fig. 3**). This is only one example of the "Lord Kelvinesque" language that seems to abound in this discipline.

The line of the ascending node is simply the line connecting the Earth's center with the equator crossing, EQX, of the ascending (northward-bound) pass. The northbound EQX is the *ascending node.* For instance, if a satellite had an argument of perigee of 180 degrees, apogee would occur in an EQX of the ascending pass, while perigee would occur at the EQX of the descending pass.

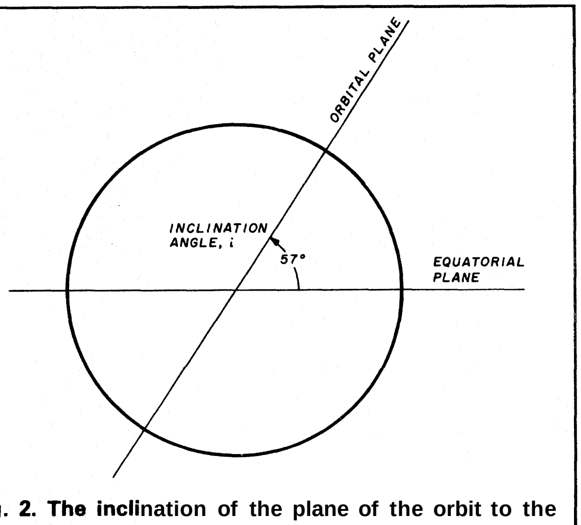


fig. 2. The inclination of the plane of the orbit to the equatorial plane.

argument of perigee — examples

The argument of perigee of Phase III-A satellites will be affected by the oblateness of the Earth. This will cause a *precession* of its orbit; that is, the argument of perigee will increase daily at an estimated 0.07-0.08 degree. Thus the ellipse will slowly rotate counterclockwise in the orbital plane about the center of the Earth. A precession of 0.08 degree per day seems small until compared with Mercury's precession of 574 seconds per century!

After launch, Phase III-A will have apogee only 30 degrees (true anomaly) past EQX, so that a significant portion of some passes will be at low elevation; thus antennas should be able to see to the horizon.

For example, suppose a pass begins with a longitude of perigee of 270 degrees. Then almost 7-3/4 hours of this 9-hour pass will be at elevation angles less than 15 degrees as seen from the Midwest. However, after two years, Phase III-A will have an argument of perigee equal to 270 degrees; thus apogee will occur at the maximum possible latitude of 57 degrees. The satellite will then, for most passes, hang high in the sky and will be accessible by small picnic-table-mounted arrays. At this writing it's estimated that a user will need, at apogee, a transmitting power of 700 watts ERP at 435.1 MHz and, for the average modern 2-meter receiver, an antenna with a gain of 13 dBd at 145.9 MHz with circular polarization. Received signals will, at apogee, be 5-6 dB weaker than those of AMSAT-OSCAR-7B.

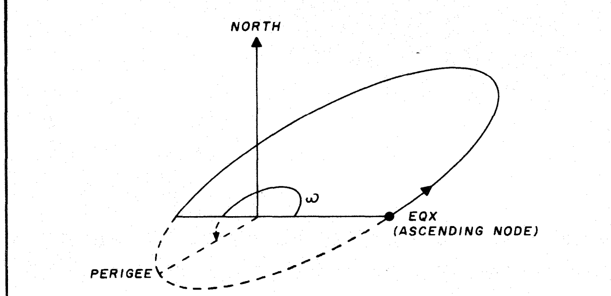


fig. 3. The argument of perigee, ω.

In contrast with the past sun-synchronous satellites, there will be no "typical" passes with Phase III. For instance, if the longitude of perigee is 0 degrees, then at no time during such a pass will the satellite be visible from the midwestern U.S.A. On the other hand, if the longitude of perigee is 180 degrees, the satellite can be worked for 9 hours continuously with 6½ hours of elevations exceeding 45 degrees.

Tracking will never be much of a problem. You'll be able to leave your beam at one setting, often for an hour at a time, without noticeable loss of signal.

ham radio

the ham notebook

using the Radio Shack ASCII keyboard encoder for microprocessor-controlled CW keyboard

I found that the Radio Shack 277-117 ASCII encoder will not interface with the microprocessor-controlled circuit by WB2DFA in *ham radio,* January, 1978, page 80, unless some minor modifications are made.

The problem is that the control key function designed in the Radio Shack encoder is not compatible with the microprocessor circuitry. To overcome this, I've installed an outboard, normally open, single-pole pushbutton switch, which is connected to a 7400 as shown in **fig. 1.** This circuit places the two most-significant outputs of the keyboard in the LOW state when the control button, SW2, or switch SW1 is activated.

I prefer to use my old trusty electronic keyer for contest operating; however the memory in the CW keyboard sure is a work saver for calling CQ. That's why I've installed SW1 in parallel with SW2, so that it's only necessary to press the letter T to send the message stored in the memory. The rest of the sending is done with the electronic keyer in such an operation.

Frequently, when typing at higher speeds, letters would be missed. It turned out that the microprocessor chip did not like the strobe waveform produced by the keyboard. To overcome that problem I installed a 74LS13 Schmitt trigger to change the pulse into a nice, clean square wave, which eliminated the problem nicely.

It's been a good project, and I extend my thanks to Jim, WB2DFA, for his assistance in getting the PROMs programmed and getting on the right track in this project.

Frank Van der Zande, VE7AV

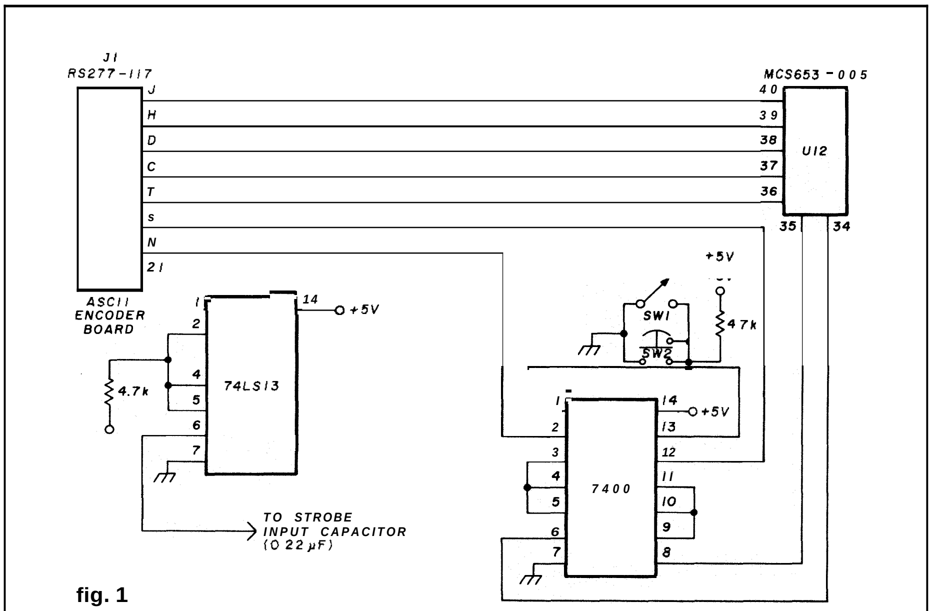


fig. 1

more quad variations

The familiar cubical quad antenna has been twisted into many shapes and variations, to the eminent pleasure of the twisters, and has performed quite well, nevertheless. Here are three more variations, two of which I've tried and found to be worthy of being added to the other modifications.

The first two variations were vertically polarized arrays for two-meter fm operation. First comes the double quad, which consists of two driven elements in the diamond configuration with a common apex **(fig. 2).** The two loops are operated in parallel, thus reducing the load impedance presented to the transmission line, which might be expected to more closely match the line. Note that the parasitic elements have a slightly different configuration, with a crossover of the wires rather than a junction. The reflector was 5 per cent larger than the driven element and spaced a quarter wavelength from it. The two directors were 5 per cent smaller than the driven element and were spaced 0.15 wavelength from the driven element, and from each other.

This antenna was cut to size and assembled without any effort to maximize tuning or dimensions. The VSWR was below 1.5. The antenna gave very good results until it suffered mechanical damage because of violent weather.

This multiple-quad concept can probably be expanded, particularly as a fixed array for use on the lower-frequency bands. The configuration of the driven and parasitic elements would be something like **fig.** 3.

The second variation, which replaced the double quad in an attempt to simplify the construction, was an adaptation of the bi-square beam, where the array measured a half wavelength on a side. Because this array is fed at a voltage point, a tuned stub was needed on the driven element. To bring the feed point nearer to the center of the structure, a half-wavelength open-wire stub was used, with a coaxial balun to obtain a balanced feed. Again, the reflector was 5 per cent larger than the driven element and spaced a quarter wavelength from it.

This was only a two-element beam, so no directors were used, although they should function as well with this configuration as with any other. The dimensions of the final arrangement are shown in **fig. 4.** After a careful adjustment, consisting of trimming the tuning stub and locating the feed points, it was possible to bring the VSWR below 1.05.

This antenna was somewhat smaller and easier to assemble than the other and has given a good account of itself. Not having an antenna testing range, I'm unable to give a measured pattern of it, but the front-to-back ratio appears to be all that can be expected from a two-element affair, and the forward gain is very satisfactory.

The third variation suggests a means of building a balun into the driven element of a monoband quad. I haven't tried this one as yet, but it looks very interesting. As seen in **fig. 5,** the driven element is composed entirely of coaxial line. The feed line is continuous all the way to the top tenter of the driven element, at which point the outer conductor ends. The center conductor connects to the other half of the driven element, which, in the interest of symmetry, is made of the outer conductor of the same-size coaxial line.

Back at top center, currents flowing on the outer conductor of the feed line reach the end of that conductor and flow back on the outside of the outer conductor, thus balancing the currents on the other half of the driven element. The result should be a completely balanced feed, with minimum feedline radiation to complicate the problems of TVI and unwanted rf in the shack. Any takers?

Henry S. Keen, W5TRS

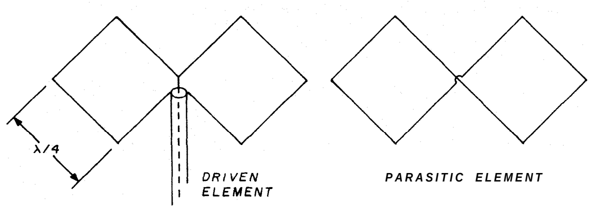


fig. 2

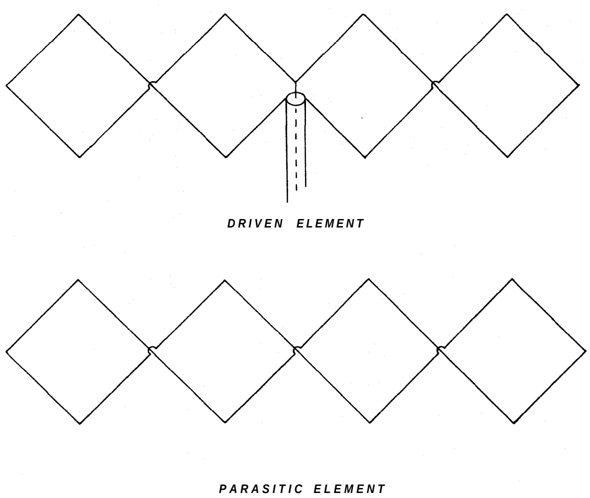


fig. 3

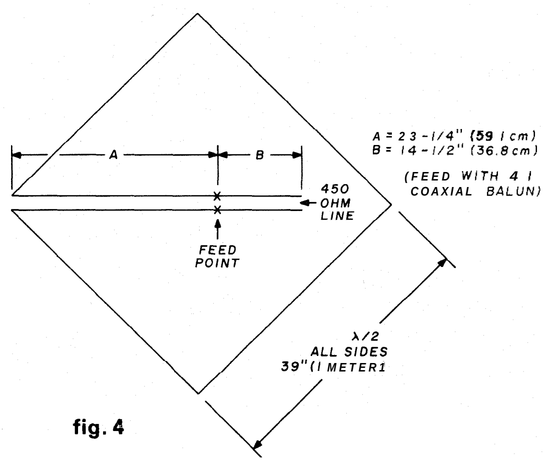


fig. 4

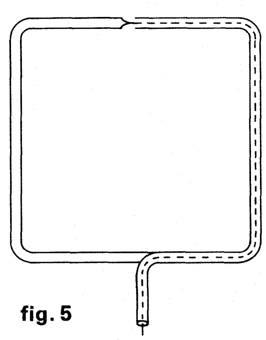


fig. 5

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contents

Observation & Opinion

A new book has come to our attention, a book interesting enough to merit mention in this column. The name of the book is ***From Beverages Thru OSCAR — A Bibliography,*** and the author is Rich Rosen, K2RR, of Littleton, Colorado.

From Beverages Thru OSCAR is not, as the name might imply, a bibliography of reference works dealing with Amateur antennas. It is instead a complete list of every article of interest to the Radio Amateur and professional published over the last 65 years in any of 288 electronics magazines and journals, including ***CQ, ham radio, 73,*** and ***QST.*** The 30,000 articles referenced in this text are divided into 92 subject categories, to make locating any given article largely a matter of determining into which category it should fall. Catchy or cute article titles have been simplified and entered into their proper category. The subject categories include such headings as Preamps, Oscillators, Filters, SSB, Lasers, Alternative Power Sources, Receivers, and Antenna Hardware.

The value of such a reference text is immediately obvious. Having access to this bibliography makes it possible to track down that elusive article on signal enhancement, or rotators, or whatever — that article that you're sure you've read in some magazine or other, but you can't quite remember which magazine it was. Or whether you read the article while in the "Fathers' Suite" of the maternity ward waiting for Junior to be born or on the way to his high-school graduation. Or whether the article was in one of your regular subscription magazines or in one you leafed through at a flea market but decided not to buy. Now, with the help of K2RR's bibliography, that long-lost article can be found with a quick look in the appropriate table.

In addition, the bibliography makes it easy for the researcher or homebrewer to find just the information he needs to get started on the project put off so many times for want of a few tips from someone who's already tried. K2RR's index of articles will not, of course, give you the information you're looking for — but it will tell you where to find it, and that's very nearly as good.

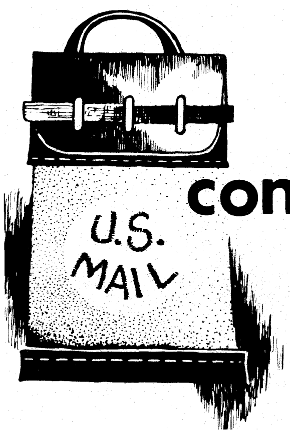
Each subject category consists of a list — some of them quite lengthy — of the articles compiled for that particular subject. The most recent articles come first. There are seven columns of information on each page, the first of which identifies the subject area as denoted by a four-digit number. (All of the subjects with their four-digit identifying numbers are listed at the beginning of the book.) The second column gives an "Abbreviated Title or Topic Synopsis," which briefly describes the article. The third column gives (in coded form) the publication in which the article appeared, and the fourth column the year and month of publication. The fifth column gives the page on which the article begins and the sixth gives the author's name (except for articles appearing in any of the four major ham magazines). The last of the seven columns is reserved for miscellaneous information and notes that might be useful for purposes of identification.

All in all, it's an impressive bit of work, one which the author says took him four years and many thousands of hours to produce. That's easy to believe, looking at (and hefting) this 620-page magnum opus. All the information contained in this book has been stored on floppy diskettes (as an alternative to the original IBM punch cards, of which 180 pounds were needed), and the author expects to be able to provide updates with each passing year. If ever there were an example of the value of computer storage, this must be it.

From Beverages Thru OSCAR — A Bibliography is currently available from Rich Rosen, K2RR, at **6043** W. Maplewood Drive, Littleton, Colorado 80123. Rich says that, in addition to the complete volume, he has also made available individual subject chapters for those who would like the benefits of this index but don't need more than a few subject headings.

In our opinion, this is the sort of reference text that many Amateurs will find useful. Our thanks go out to K2RR for having provided the Amateur fraternity with so valuable a tool.

Martin Hanft, WB1CHQ
administrative editor



comments

battery charging

Dear HR:

I am writing to you regarding the letter to the Comments column by Robert H. Weibrecht, W6NRM. His comments regarding charging batteries at a low rate are only partly correct. The rest of the story is that the nickel-cadmium battery should be discharged to 1 volt per cell for exercise. This will help to remove the memory induced by continuous charging. Gel batteries, on the other hand, need continuous charging.

D.L. Carlson
Burnside, Minnesota

ground systems

Dear HR:

The article "Ground Systems" in your May, 1980, issue was very interesting, It was particularly interesting to know that the inductive reactance of only 9 feet of wire at 4 MHz can be 100 ohms or so, and that the rf resistance of a wire is some seventy times its dc resistance at 14 MHz! The mention of rf resistance immediately brought to mind Litz wire – three or more strands of insulated wire braided together. I immediately replaced my 12 feet of ground wire with three insulated wires braided together, and it certainly decreases BCI interference! I wonder if anyone has formulas (empirical or otherwise) for the rf resistance of ordinary wire and the rf resistance of Litz wire?

incidentally, another thing to try if you have BCI problems is to put Amidon FB-801 ferrite beads on the ac or dc power leads just outside or just inside the case of your transceiver. A bead on the "live" side of the mike lead (again, close to where it enters your transceiver or speech processor) may also do some good. Don't put beads on any ground leads!

Keith Wilkinson, ZL2BJR/JG1YCI
Tokyo, Japan

selfish attitudes

Dear HR:

The Observations and Comments column in the August, 1980, issue of *ham radio* asks, "What can be done about the selfish attitudes of those who interrupt contest operation?"

If this were a perfect world with a perfect society, this condition would not take place. However, on the other side of the subject, why should a contest operator come on a frequency in use by others and call "CQ TEST" until he either gets control of the frequency or drives the others off?

It is my firm opinion neither group is completely free of guilt. Don't you think some better planning of worldwide contests should take place? Almost each weekend there is a contest, sometimes on both CW and SSB at the same time. Would limiting the contest to a band of frequencies be the answer? Why should the operators who like to rag chew or keep skeds each weekend be punished? Should we stand in the way of the contest operator? There is no easy solution to the problem and until each side sees the other side of the coin nothing will change.

On "What about slow-scan TV and interference by SSB operation?" I wouid say this is a very difficult question to answer at the present time. Until the FCC decides to allow General class operators to use SSTV it is not a good idea what a good approach would be. As 20 meters was the band mentioned in the editorial, I have a suggestion. My thoughts at this time would be to ask the SSTV operators to consider moving from 14.230 MHz to ±14.270 MHz as a calling frequency. This would put then near the General class end, but not too close to give or take QRM from each other.

Paul T. Atkins, K2OZ
Park Ridge, New Jersey

Q system

Dear HR:

I think the "Q system" is a good idea but should be in reverse order; that is Q1 would be full copy (first class).

Arthur Masthay, W1IUZ
Avon, Connecticut

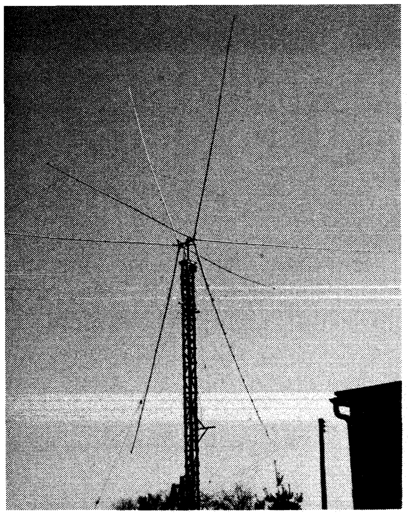
satisfied reader

Dear HR:

Over the years, *ham radio* has had an evolution toward more technical dissertations. Although the math was minimized, I couldn't help feeling that things were too heavy to be enjoyable. On the other hand, I had a fear that I was growing old for the technology at 47.

The August, 1980, *ham radio* seems to return to more readable articles and a few reasonable construction articles. I hope this is a trend and not a maverick edition. For the first time in several years, I read all the articles.

Don Nelson, WB2EGZ
Vorhees, New Jersey



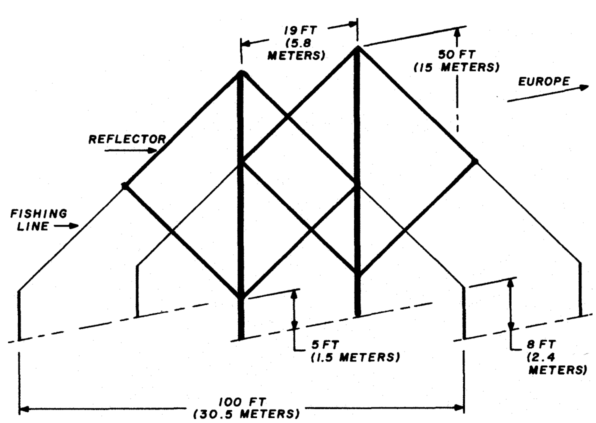
super quad for 7-28 MHz

An ambitious project for the fellow who likes to "roll his own"

My interests in DX and DX contests goes back many years. I often marvel at how the state of the art has progressed. Ordinary dipoles, verticals and long wires on 7 MHz worked satisfactorily in those days, as everybody else was using the same thing. By 1964 many DXers on 7 MHz had "grown" good beams, and it became harder and harder to be a winner in the pileups. To stay competitive, I had to think about drastic changes in the 7-MHz antenna department.

early quad experiments

A quad had always intrigued me, so I started to read books and collect information on this antenna. My quad project began back in 1965 after I acquired two telescoping 50-foot (15.25-meter) TV masts and some No. 18 (1-mm) copper-clad wire. These masts were extended another 5 feet (1.5 meters) using aluminum tubing. They were then erected 19 feet (5.8 meters) apart in my 100-foot (30.5-meter) wide back yard. The antenna pointed directly toward Europe.



A 2-element 40-meter quad in a diamond shape was supported by these masts. The feed point and reflector tuning point were only 5 feet (1.5 meters) from ground: very convenient for tuning and matching the array. The driven element was fed by a 4:1 balun and RG-8/U coax. The quad was adjusted for minimum backward radiation.

A whole new world opened up. I began hearing European signals that were inaudible on a ground-plane antenna. However, I felt frustrated when I wanted to work DX in different directions and resorted to the ground plane antenna, which was always a good performer for long-haul DX.

After using the two-element fixed quad for a number of years and collecting stacks of data, I decided to make the antenna rotatable and also higher.

design criteria

In 1970 I arrived at the fundamental design concepts:

1. All elements to be full size.
2. The longest metallic object in the system to be 13 feet (4 meters) maximum.
3. Incorporate concentric quads for 40-20-15-10 meters.
4. Boomless or very short boom design.
5. Separate feed lines for each quad
6. Nonmetallic tower (see 2 above).
7. Center of quads to be 44 feet (13.4 meters) above ground.
8. One person can raise and lower the array for tuning or repair.
9. Cost to be $250 maximum.
10. Use diamond configuration in the design.
11. Keep the 19-foot (5.8-meter) spacing from driven element to reflector.

It took a year to complete the design and construction of this project with much redesign along the way, and in July 1972 the array design was fixed.

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From necessity many features of the entire system (tower, rotator, winch) are merely touched upon in this article, major emphasis being on design and construction of the antenna.

construction

Fig. 1 shows the construction of the spreader or spider arms. A list of tubing is given below.

Spreaders. To insert the 1-3/8 inch (34.9 mm) tubing into the I-112-inch (37.9-mm) tubing, both parts must be straight, round and burr-free. The insert must be thoroughly lubricated on the outside; the same goes for the inside diameter of the 1-1/2-inch (37.9-mm) tube. I was able to insert all eight pieces with the help of a rawhide mallet by placing one end of the I-112-inch (37.9 mm) tubing against a tree stump. However, it's advisable to slot the 1-318 inch (34.9 mm) tubing lengthwise for 30 inches (76 cm) with a saber saw, then deburr and insert the slotted end first. The 72-inch (183-cm) long tubing was polished on one end, lubricated, then driven into the tubing as shown in **fig. 1.** Make sure the tubing has entered at least 2 inches (5 cm).

Bushings. The insulating bushings are needed to comply with item 2 of the design criteria. The electrical length of the spider arms is 12 feet (3.7 meters) maximum. I turned these bushings on a small bench lathe. Take care to have good concentricity and roundness. For this reason the outside diameter was turned last by pushing the finished inside diameter onto an arbor made of plastic. I used scrap pieces of nylon, Delrin, and PVC. After the bushings were completed they were placed on the respective tubing sections as shown in **fig. 1** and held in place with epoxy.

The spreader arm is now ready to be assembled as in **fig. 1.** Stainless steel hose clamps were used as shown.

List of aluminum tubing. All items listed (Page 14) are 12-feet (3.7-meters) long 6061-T6 drawn round aluminum tubing. All items must be straight and round! No defects accepted.
The total cost of these items in 1971 was $105.52. Three 12-foot lengths of 1-1/8 OD x .058 inch (28.4 x 1.5 mm) wall tubing should be added for the optional reinforcement of parts shown in detail 1, **fig. 1.**

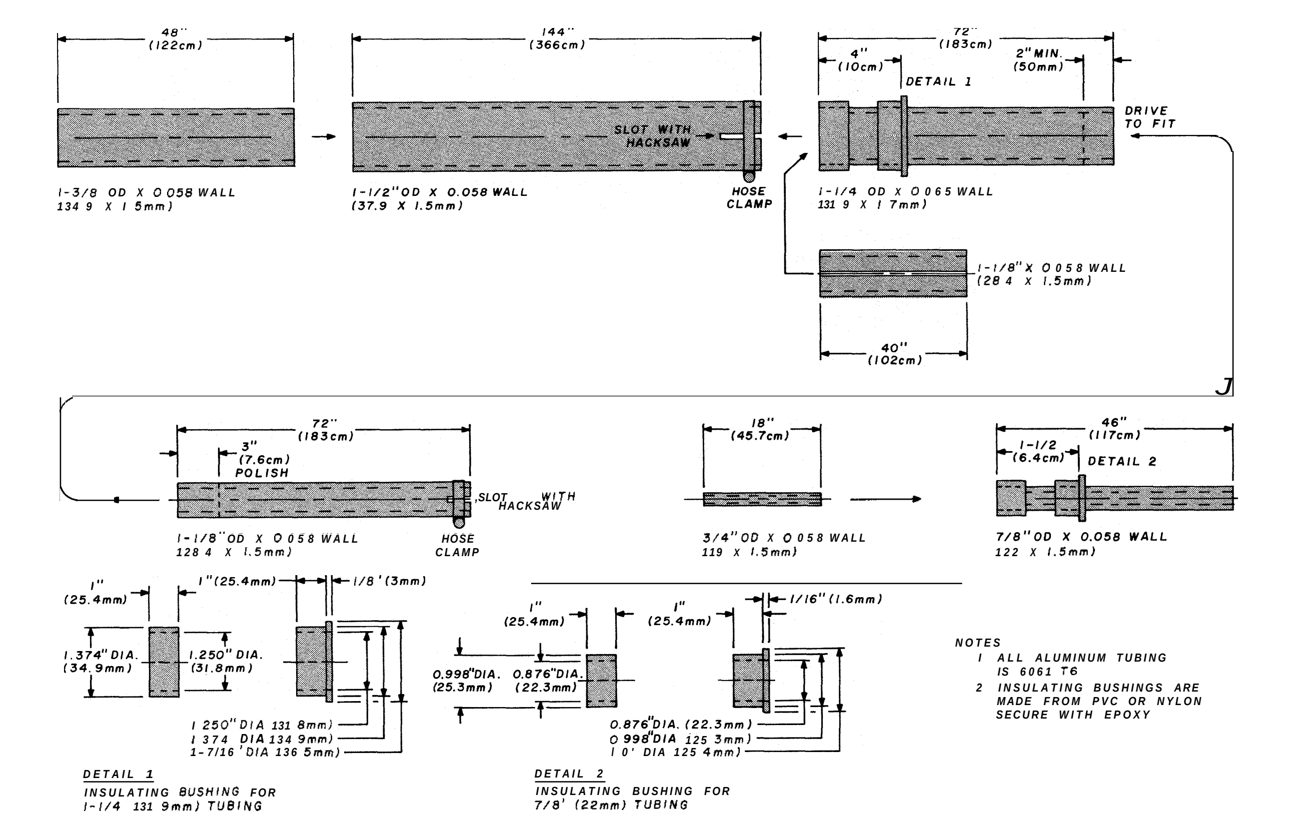


fig. 1. Construction of spreaders for the super quad. Insulating bushings are made from PVC or nylon material. Epoxy cement is used on the bushings. Spreader arms are made from 6061-T6 aluminum tubing.

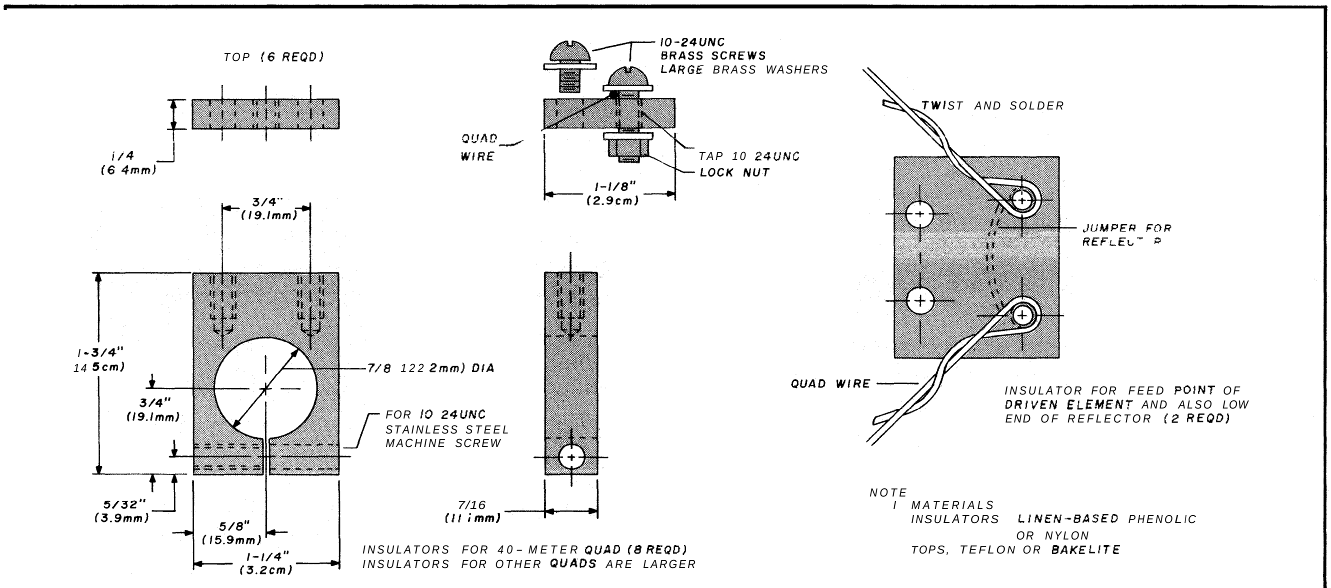


fig. 2. Details of the insulators, which were made from linen-base phenolic material.

quantity	size	weight
8	1½ OD x .058 (38 x 1.5 mm) wall	30 lb. (13.6 kg)
3	1-3/8 OD x .058 (35 x 1.5 mm) wall	11 lb. (5 kg)
4	1¼ OD x .065 (32 x 1.7 mm) wall	13 lb. (6 kg)
4	1-1/8 OD x .058 (28.4 x 1.5 mm) wall	11 lb. (5 kg)
3	7/8 x .058 (22 x 1.5 mm) wall	6 lb. (2.7 kg)
1	3/4 OD x .058 (19 x 1.5 mm) wall	1.77 lb. (1.5 mm)
		72.75 lb. (33 kg)

Insulators and brackets. A suitable insulator had to be designed and made to hold the quad wires to the spreader arms. **Fig. 2** shows the insulator for the 7-MHz quad. The 20-meter quad insulator clamps onto the 1-1/4-inch (31.9 mm) tubing; the 15- and 10-meter insulators clamp to the 1½-inch (37.9 mm) tubing. Variation of the insulator size must be made accordingly. The first insulators I made were without the Teflon part. During the first rain storm, rf voltage arced from the brass screw to the tubing and burned up the insulators on the high-voltage points of the driven element.

The quads are fed at the lower corner of the driven element, so I also made up some supporting brackets for the coax cables that run down the lower front spreader arm. **Fig.** 3 shows details for this support.

These insulators and brackets were made in my work shop using a small lathe and drill press. All the insulating supports were clamped onto their respective spreader arms in the precalculated positions. I've omitted dimensions for these locations. To calculate these points is a good mental exercise, and every high-school student should be able to arrive at the correct numbers. Final adjustments can be made after stringing the quad wires.

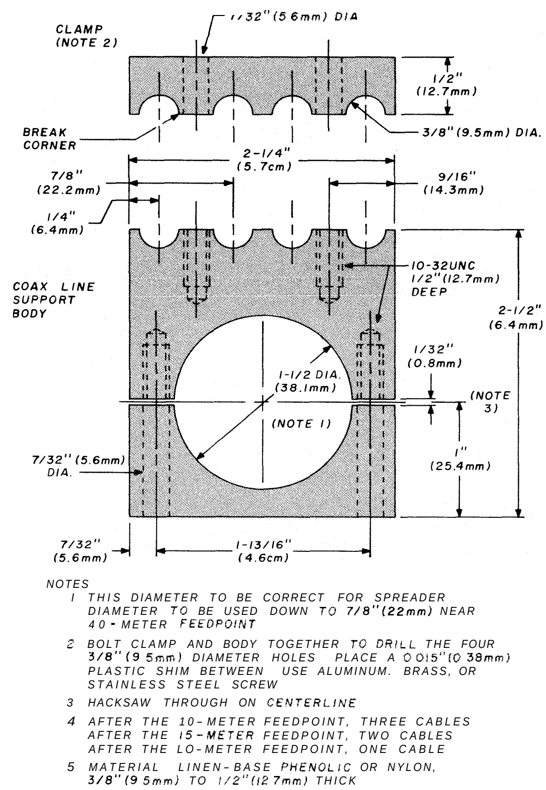


fig. 3. Supports for the coaxial cables that run down the lower front spreader arm.

Boom and associated hardware. The boom is a length of solid PVC measuring 30 x 3 inches (76 x 7.6 cm). Nylon or Delrin could also be used. Great care must be taken when drilling the eight 3/8 inch (9.5

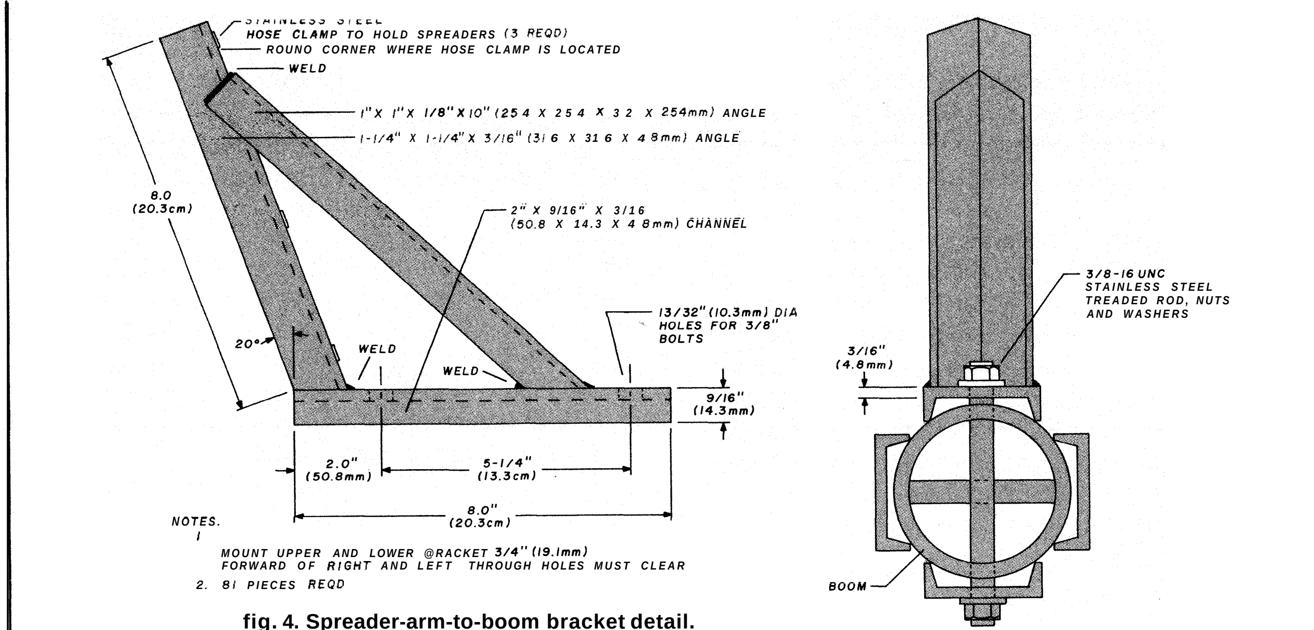


fig. 4. Spreader-arm-to-boom bracket detail.

mm) through-holes to ensure good angular alignment of the spreader arms. Also, the holes for the upper and lower boom-to-spreader-arm brackets must be about 3/4-inch (19-mm) in front of the horizontal brackets so that the 3/8-inch (9.5 mm) through-bolts won't interfere with each other. I drilled these holes on a vertical milling machine in a friend's machine shop.

The boom-to-mast plate is 1/4-inch (6.5-mm) thick steel plate. Four U bolts must be used to clamp the boom to the plate and four U bolts to clamp the plate to mast. At the beginning, I used only two U bolts for each, but the first heavy wind gave me the needed education!

Spacing between driven element and reflector is 19 feet (5.8 meters) for the 40-meter quad (0.136λ). It's not a magic number but it worked well on the fixed quad and it also provided clearance between lower spreader arms and guy wires for the tower.

The theoretical angle from vertical for the spreader arms to point outward is 17°48′. However, some preloading of the tension cords that run from the front to the rear spreader arms near the ends is needed. I chose an angle of 20°, which amounts to 12 inches (30 cm) at the end of each arm.

Fig. 4 shows the spreader-arm-to-boom bracket. I had all parts ready cut, shaped, and drilled. Then, with a template to assure the correct 20° angle, they were taken to a welder.

After welding, I gave the brackets a few coats of zinc chromate (galvanizing would have been better by far).

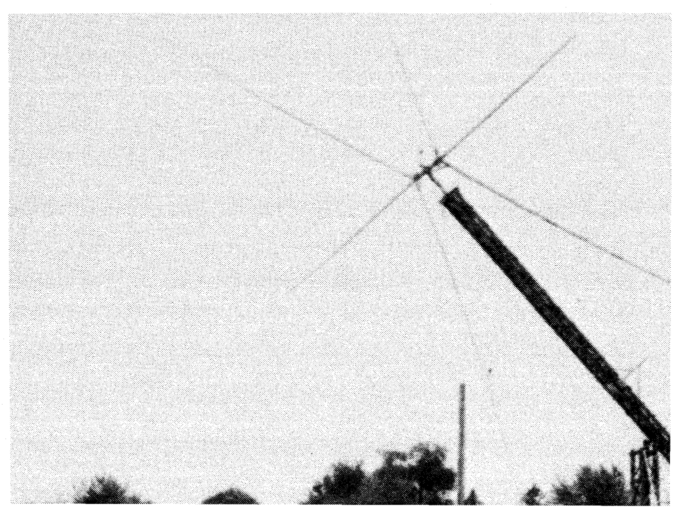
These brackets were mounted to the boom with stainless-steel threaded rods, cut to size, and stainless-steel washers and nuts on each end of the rods.

To conform with items 2 and 6 of the design criteria, I made the tower of wood. It's a foldover tower. When in a vertical position, it has four guy wires (broken with insulators). The uprights for the 40-foot (12-meter) fold-over section are straight pieces of 2 x 4 lumber 20 feet (6 meters) long. The tower is 14 inches (36 cm) square (to the outside of the uprights).

The horizontal braces, which also serve as rungs for climbing the tower, are 1 x 3 lumber; the diagonals are 1 x 2 lumber. No. 10 wood screws 1.5-inch (38 mm) long and waterproof glue were used in the construction. The hinge pin is a l-inch (25.4-mm) diameter stainless-steel rod. Hinge members are aluminum plates 318-inch (9.5-mm) thick, which were bolted to the uprights. The tower hinge point is 16 feet (4.8 meters) from ground. A wooden tower section, 16 feet (4.8 meters) high, is permanently bolted to 4-inch (10-cm) channels, which are embedded in a concrete base. I placed four screw-in anchors equidistant from the center of the foldover section on a 16-foot (4.9-meter) radius. **Fig. 5** shows the tower and the quad.

The winch, which is used to raise and lower the

fig. 5. Photo showing the homebrew wooden tower and quad ready for erection.



tower, is also home built and uses a 50-tooth, 1° pitch, single-thread worm gear for safety. A 112-inch (25.4-mm) diameter reversible 500 rpm electric drill chucked to the tower worm shaft is used to raise and lower the tower.

The rotator also uses worm gears. It's mounted 8 feet (2.4 meters) from ground inside the tower. A 2-inch (51-mm) galvanized water pipe is used for the mast. A thrust bearing is located 8 feet (2.4 meters) from the top. The drive shaft from rotor to mast is also 2-inch (51-mm) galvanized pipe. To conform to my design criteria, it was broken into three sections, which are coupled together with solid PVC couplings as insulators. Again, I want to bring to your attention the enormous stresses that are applied to these parts during strong winds.

assembly

With the tower lowered, resting on a 13-foot (4-meter) high A frame, and guyed to right and left for safety, the antenna assembly begins. An extension ladder was used (see **fig. 6**).

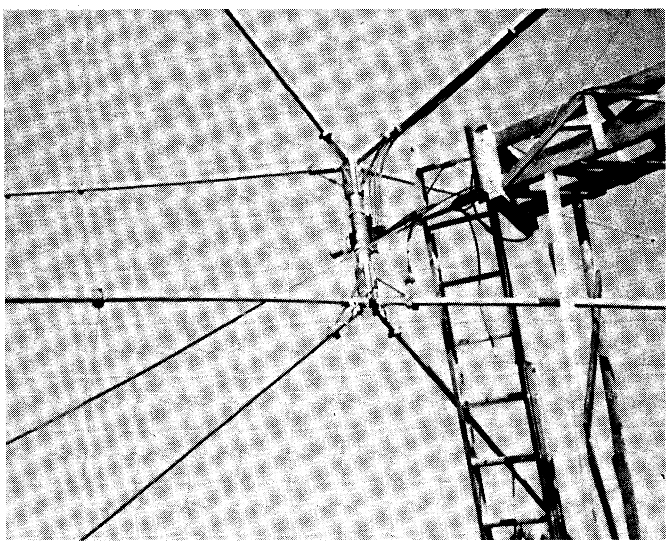


fig. 6. Photo of antenna showing extension ladder, which was used to assemble the elements and feed system.

The boom was mounted to the mast, then the four driven-element spreader arms were mounted to the brackets. Use three stainless-steel hose clamps (or more). (To meet item 2 of the design criteria I insulated the arms from the brackets.) I used strips of vinyl between arms and brackets and also under the hose clamps. I used 19-foot (5.8-meter) long stress cords attached to the upper vertical spreader arm near the end and also to both horizontal arms. I used small weights on the loose ends. At this time the tower was raised to the vertical position. I rotated the antenna 180 degrees, then lowered it. The other four spreader arms were then clamped to the brackets, and the stress cords were fastened to the reflector set of spreader arms.

At this point we're ready to attach the quad loops to the insulators. When I first erected the fixed quad, I used the formulas for the loop length from the ARRL *Antenna* ***Book:***

$$\textit{driven element } (ft) = \frac{1005}{f(MHz)};\qquad \textit{reflector } (ft) = \frac{1030}{f(MHz)} \tag{1}$$

I found that these numbers were wrong in my case. For my rotary quad I used:

$$\textit{driven element } (ft) = \frac{994}{f(MHz)};\qquad \textit{reflector } (ft) = \frac{1019}{f(MHz)} \tag{2}$$

The constants 302 and 309 may be substituted into the numerators of **eq. 2** for calculating lengths in meters. The lengths, from **eq.** 2, are 141.5 feet (43.2 meters) for the driven element and 145 feet (44.2 meters) for the reflector.

The reflector is slightly less than 2.5 per cent longer than the driven element, and the adjustment is critical. In any event, the reflector must be tuned for minimum backward radiation. (I was not concerned with SWR while tuning for maximum front-to-back ratio.) All quad loops were fastened to the insulators, which were placed in the calculated positons on the arms.

feed system

The quad is fed by 52-ohm coax and a 75-ohm quarter-wave matching transformer. One coax line runs to a relay box at the top of the tower where the desired quad is selected by one of four relays. (The inner conductor of the unused lines is *not* grounded.) The five-wire control line to the relays was decoupled 13 feet (4 meters) from the relay box by winding three turns of this control line through a 2-inch (51-mm) diameter toroid.

tune up

For tuning the quad I put a sensitive field-strength meter with a 20-foot (6-meter) long horizontal pickup dipole 6 feet (2 meters) from ground, about 200 feet (60 meters) from the quad. I pointed the quad toward the field-strength meter and fed a small amount of rf at 7050 kHz into the quad to give a full-scale meter reading. I then pointed the back of the quad toward the meter. The reading should drop to about 1/50th of full scale, which is close to zero. At this point I adjusted the meter reading to about half scale then varied the frequency plus and minus but maintained the same output from the transmitter. Wherever the minimum reading on the meter occurred I considered to be the maximum performance (maximum front-to-back ratio) operating frequency. Record it!

I used the same procedure for the 14-MHz antenna. I cranked the tower down, with the reflector facing the ground, then made precalculated adjustments to the reflector only. (In my case the desired frequencies were 7020 kHz and 14020 kHz).

The 40-meter elements turned out to be as mentioned before. The 20-meter driven element is 71 feet (21.7 meters); reflector is 72 feet, 8 inches (22.2 meters), and the spacing is 12 feet, 8 inches (3.9 meters).

standing-wave ratio

The SWR of the 7020-kHz quad is near 1:1. That of the 14020-kHz quad is 1.5:1. The 21-MHz and the 28-MHz quads have not been tuned as described. However, the SWR on 21020 kHz was very high and I substituted a gamma match, which improved the SWR. I think that the gamma match is superior to the matching transformer.

I think the spacing is too great for the 21- and 28-MHz quads. I wasn't able to obtain the excellent results as with the 7- and 14-MHz quads. (Remember the spacing of the higher-frequency quads is not proportional to the 7-MHz quad since the boom is a constant.) But once I had the two lower bands working I never took the time to improve the 21- and 28-MHz antennas. It's too much fun to sit behind the loud-talking 40-meter antenna!

front-to-back ratio test

After the 7-MHz quad was completed and tuned I placed the back toward the field strength meter again. I shorted the insulated drive-pipe section. Nothing happened to the front-to-back ratio. Then I short circuited the guy-wire insulators and short circuited the guy wires to the drive pipe. Nothing happened to the front-to-back ratio. However, when I short-circuited two spreader arms to the mast, a definite deterioration of front-to-back ratio occurred. This test was only made on 7 MHz to satisfy my curiosity.

performance

Many times I switch from the 40-meter quad to my ground-plane antenna and some signals just turn into a faint scratching sound, whereas they were RST 559 on the quad. It's mind boggling when a European station receives me RST 599, and after I turn the antenna backside toward him, he loses me completely.

I also have a tribander TH6DXX at 74 feet (22.6 meters). I can compare the 14-MHz quad with the Yagi at the flip of a switch. [The center of the quad is only 44 feet (13 meters) from ground.] On long-haul DX the high Yagi always out performs the quad.* Signals from Europe and Central America improved up to three S units on the quad under wide-open conditions. When band conditions are low, the Yagi takes over. When the 20-meter band gets wiped out by rain static on the Yagi, I switch to the quad. Lo and behold! no static, only signals.

maintenance

In seven years I replaced the loop wires in the 14- and 7-MHz quads twice. During an ice storm the reflector wire broke on the 7-MHz loop. The ice loading kinked the 1-1/4-inch (31.75-mm) tubing section of the upper spreader arm. The ice on the wire measured 1/2-inch (13 mm) in diameter, and the tubing was 1-3/4 inches 144.5 mm) from the ice buildup. It required only six hours to put the quad back into operation again, and the nicest part was that I could do it all by myself.

cost-reduction tips

To keep within the $250 limit, the use of the junk-box was mandatory. I also visited a few surplus houses where I was able to pick up stainless-steel aircraft cable for $.05 per foot, 18-inch (45.7-mm) turnbuckles for $2.00 each, and worm gears and worms for $5.00 per set. I cultivated the friendship of a machine-shop foreman who saved scrap aluminum plates, old steel shafting, and nylon scraps for me. I bartered a case of beer for the welding of the boom-to-spider brackets. Such is the stuff of which hams are made.

Would I do it over again? The answer is yes! However, since making the different tests I would be brave enough to use a heavy-duty commercial 50-foot (15-meter) foldover tower. Who is adventurous enough?

This has been a fun project. The greatest reward has been getting into a pileup and having those rare ones come back to me.

acknowledgement

Many thanks to N3RD, who was my critic and coordinator. I appreciate the suggestions of N3ANW. Special thanks go to my dear wife, who was always ready to help with holding and signaling chores. She kept the meals warm on many occasions and never blew her Irish fuse!

reference

1. Wayne Overbeck, N6NB, "Quads vs. Yagis Revisited," ham radio, May, 1979, pages *12-21.*

*See Wayne Overbeck's article[1] on this controversial subject. Editor.

ham radio

automatic CQer for RTTY

An alternative method for sending CQ on your RTTY station

If you're an active RTTYer without tape or memory bank capabilities, you know that it's a chore typing out several lines of CQ, especially if you don't get an answer on the first try. Here's a solid-state circuit that will do it automatically for you. It's called the NS-CQ. The board is 4.5 x 4 inches (11.4 x 10.2 cm) and contains only seven ICs. It sends a string of 16 CQs without spacing, then DE, your call three times, and finally CR LF: CQCQCQCQCQCQCQCQCQCQCQCQCQCQ-CQCQCQ DE W4AYV W4AYV W4AYV CR LF. It will continue this until the reset switch is turned to the **HOLD** position.

how it works

Fig. 1 shows the schematic. The oscillator, U1, is a 555 timer IC set to operate at 45.45 Hz, which is the Baud rate for 60 wpm. This frequency is adjusted by R1. Oscillator output is fed to U2, which is a dual 4-bit binary ripple counter. This means that there are two separate sections with separate clock inputs that can count up to, or divide by, 16. Thus a total count of 256 is provided. The outputs are binary-coded four bits, so we can use the various output combinations to drive the binary inputs of the other ICs.

U3 is the CQ IC. This chip is analogous to a 16-position rotary single-pole switch. It is advanced one step at a time by the binary codes from U2. In other words, it selects one input from zero to 15 and sends it to the output, just like turning a rotary switch from 0 to 15. Each of the 16 inputs is wired to either +5V or ground, depending on the Baudot code for C or Q.

As you know, the Baudot code consists of five bits in various combinations to make the machine print a letter or figure or perform a function. In addition there is one start pulse, which is always low and one stop pulse, which is always high. All these pulses are 22 ms long except the stop pulse, which is 31 ms long. To simplify the circuit I've made the stop pulse *two* 22-ms pulses. This way, the machine will operate properly and no discernible difference can be noted. So we now have eight pulses in all: one start pulse, five data pulses, and two stop pulses. With these 16 inputs we can hardwire in the letters C and Q with necessary start and stop pulses using the Baudot code.

We want the inputs U3 (CQ) to be scanned 16 times. This is done by taking off count 256 (16 bits x 16) from the U2 counter and sending it to U6, which is a flip-flop control. This will give 16 CQs without spacing. To insert spaces would require more ICs and circuitry, because for each space another eight bits would be needed. The method used here keeps the unit simple.

When the inputs to U3 (CQ) have been scanned 16 times U3 is cut off and switched to the PROM (programmable read-only memory) IC, U4, which contains DE, call letters three times, then CR LF. These data consume almost all of the 32 characters available; and at the end of the data, the PROM is switched off and goes back to CQ. U7 is a multiplexer, which decides which of the PROM or CQ information goes to the final output.

U4 is programmed permanently with call letters, etc. Since the start and stop pulses are always the same, only the five-bit data are programmed. The PROM outputs are in parallel form; that is, all the five-bit data for the machine appear at the output at the same time. These data must be converted back to serial form for the machine, and this is done with U5.

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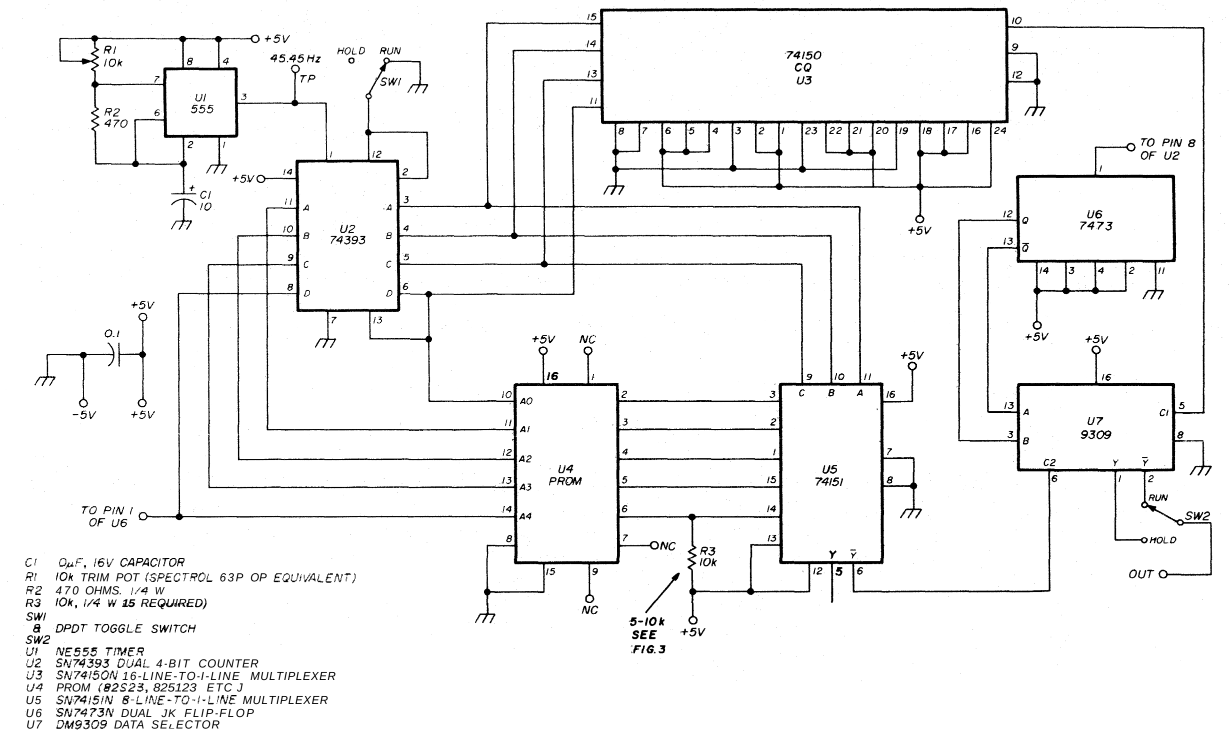


fig. 1. Schematic of the NS-CQ automatic CQer for RTTY. It uses only seven ICs. It sends 16 CQs without spacing; then DE; then your callsign three times; and finally CR LF. Note that five 10-k resistors are required from +5V to each of the five inputs to U5.

This IC is very similar to the 74150 except that it has only eight inputs.

Since the start and stop pulses are always the same, U5 has its first position grounded for the start pulse, and the final seven and eight positions are wired to +5 V for the stop pulses. Now all that's needed from the PROM are the five-bit data. Note there are five 10k resistors from +5 V to each of the five inputs of U5. These are necessary because some of the PROMs used have open-collector outputs.

Because the PROM outputs are in parallel and connected to U5 multiplexer inputs, PROM IC must run eight times slower to give U5 time to scan all eight inputs. This is done by taking off the proper count from U2. The PROM used here is described as a 32 x 8, or organized as 32 words of eight bits each. This can be thought of as a ladder with 32 rungs or "address positions." Each of these address positions has a storage capacity of eight bits. The address positions are advanced one position at a time with the proper binary code from counter U2.

interfacing the CQer

Output from the NS-CQ is taken from either pin 1 or pin 2 of U7. **Fig. 1** shows output from pin **2.** This will give approximately 5 V on mark and 0V on space with the transmitter in LSB mode. The two outputs are opposite. When one is high the other is low. As shown, the output on pin **2** is low when reset pins **2** and 12 of **U2** are above ground. This position resets the counter to zero, which stays there until again grounded. The dpdt switch changes the output of U7 pin **2** to pin 1, which is high, so the machine will hold in a mark condition during standby. If your setup is inverted with connection to pin **2,** just reverse the two leads going to the dpdt switch.

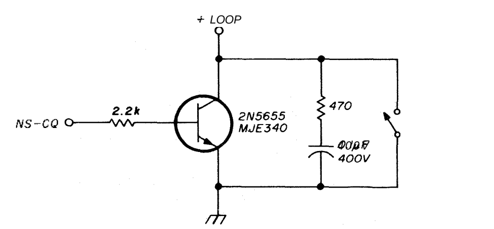


fig. 2. Alternative keying circuit that can be inserted in series with the loop.

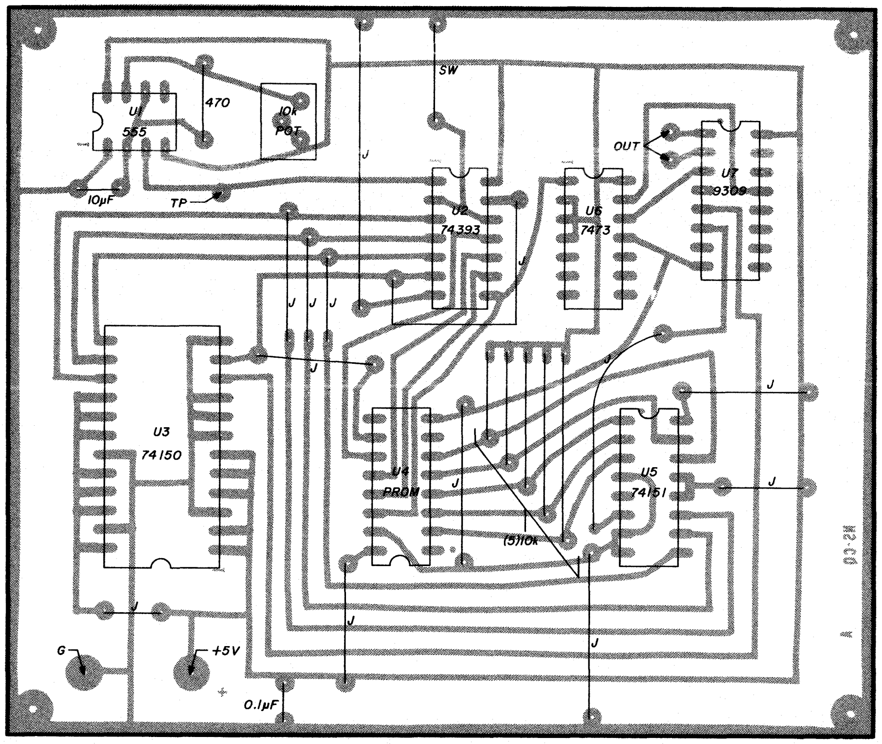


fig. 3. Parts placement on PC board. Note notch on each IC for correct placement on board.

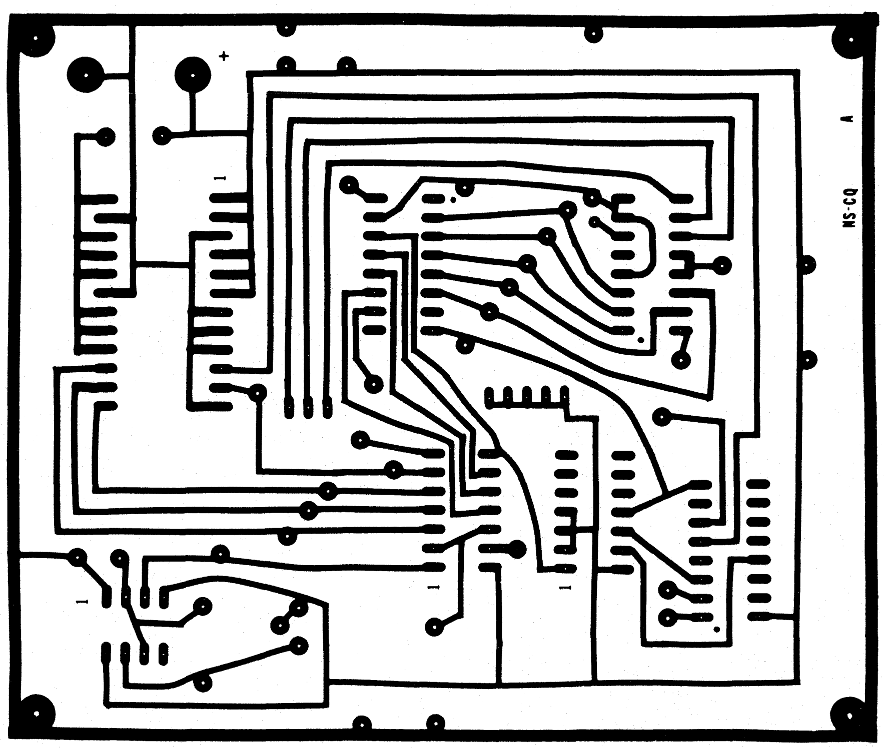


fig. 4. Foil side of the NS-CQ PC board. Boards are available from author (see text).

Most machines have the keyboard in series with the loop supply for local copy, and associated circuitry produces a keying voltage for FSK or AFSK. If your TU has a loop keying transistor, as shown in **fig. 2,** you can lift the base resistor at the far end and connect the NS-CQ output here. Another method would be to insert another keying circuit, as in **fig. 2,** in series with the loop. For AFSK, the NS-CQ output can usually replace the normal keying voltage going to your AFSK tone generator. For example, the NS-CQ will drive the *Mainline* ***AK-2*** directly. All these changes can be made with toggle switches.

construction

Construction is straightforward with a PC board.* Use a low heat soldering iron with small solder, not more than No. 18 (1 mm) in size. Refer to **fig.** 3 for position of components and jumper wires.

It is strongly recommended that sockets or Molex pins be used for the ICs. This simplifies removal of an IC if necessary. If Molex pins are used, first solder the pins then remove the top tab by bending it over once or twice. CAUTION: It is imperative that each IC be inserted in its socket in the proper way, otherwise it may be destroyed. Each IC has a notch or deep circular indentation on one end as viewed from the top. This notch should line up with that on the parts placement diagram **(fig. 3)**.

After soldering all components, check for solder bridges between pins and between circuit traces that are close together. Connect – 5 V and ground to proper terminals. **Fig. 4** shows the foil side of the board.

Set R1 to 45.45 Hz with a frequency counter connected to TP. If no counter is available, you can get proper speed by slowly adjusting R1 while observing printout on the machine. Set R1 so that machine runs and prints smoothly.

the PROM

Several different types of PROMS are available. Some come with all outputs high and some with all outputs low. Using a Baudot table, a letter or function is made by changing one output bit at a time from high to low, or from low to high, as required. Special equipment is needed to do this; but simply stated, it consists of pulsing a certain voltage on an output pin for a length of time in the microsecond-millisecond range. Consult the data sheet of the PROM for proper procedure. I've found the following ICs to be the easiest to program: 82S23, 82S123, 7577, and 7578.

*A few partial kits consisting of board and programmed PROM, and a few wired and tested units are available. Send a SASE to author for prices and information.

ham radio

More Details? CHECK–OFF Page 94

Yagi antenna design: stacking

Data for various stacked Yagi configurations

By James L. Lawson, W2PV, 2532 Troy Road, Schenectady, New York 12309

This article describes the use of multiple Yagi antennas arranged into a coherent antenna system. The number of potential arrangements is unlimited, but certain basic configurations deserve detailed analysis because they have attractive properties. To start, I shall limit the discussion to systems where the individual Yagi antennas are all physically identical and aligned for maximum radiation in the same direction. Moreover, to ensure that each Yagi contributes to the overall main radiated wave front in a coherent manner, I shall limit the configurations to those in which the Yagi positions (say, for example, the reflector end of the boom) lie in a plane perpendicular to boom direction. Usually all of the Yagis are coherently excited by the same driver current (magnitude and phase). Using identical Yagis positioned in such a plane helps maintain a uniform radiated pattern over a desired frequency band. The overall system beam pattern can be pointed in azimuth only by mechanically rotating the entire system.*

The overall system array can be viewed as a large-area aperture illuminated in a quasi-uniform way by the individual Yagi antennas. So long as the individual Yagi antennas are not too far apart (so that illumination is relatively uniform), the system gain should be ***proportional to the total effective aperture area.*** The system beam pattern should also show an angular width inversely proportional to the aperture dimension. Thus, in concept, a horizontal array of Yagi antennas (horizontally polarized) should produce a narrow horizontal system beam pattern; similarly, a vertical array of Yagi antennas (horizontally polarized) should produce a narrow vertical system beam pattern.

We must consider the system array over earth or ground; in this case all of the effects mentioned previously[1] will occur. Recall that ionospheric paths over earth primarily favor low radiation angles (up to say, 20 degrees); moreover, this whole range of antenna radiation angles should be covered to accommodate a continuous earth range as well as different multimode ionospheric paths. We shall see that, by vertically stacking two or more horizontally polarized Yagis over ground, it is possible to improve significantly low-angle performance (over that of a single Yagi antenna over ground) ***without*** reducing the azimuthal coverage. This improved result comes about through a suppression of otherwise useless radiation at the higher angles.

*The radiated beam from a mechanically fixed (system) array of laterally spaced Yagi antennas can, in principle, be steered in azimuth by changing the excitation phase to each Yagi antenna. However, the beam quality generally deteriorates. Such mechanically fixed, electrically steered phased arrays are not considered here.

stacked Yagi antennas

For Amateur Radio communications relatively wide horizontal or azimuthal coverage is generally desirable, not only to make a given contact less sensitive to critical beam heading but to accommodate the many occasions in which the communication path is somewhat skewed due to ionospheric conditions. Wide azimuthal coverage is especially desirable under contest conditions, where it is advantageous to have the beam simultaneously illuminate the largest desired Amateur population. So a horizontal array of Yagi antennas doesn't appear as desirable as a vertical stack;[1] therefore I shall not attempt analyses of such horizontal arrays.

Vertically stacked Yagi arrays are now in reasonably wide use. It is interesting to study the theoretical performance of such systems. Before attempting a formal analysis, I will make two obsevations. First, vertical stacking requires a supporting mast. If the stacking separation is large (which we shall find desirable), the large mast must be entirely rotatable and, of course, very rugged mechanically. Such a mast, including its foundation, is a major undertaking.

Two interesting variations of this system are not as formidable. The first variation is a stacked Yagi antenna array offset from a fixed, or guyed, tower. The offset allows simultaneous rotation of the Yagi antennas over a range in azimuth of about 300 degrees; at either end of this range, the antennas are designed to nest around the mast. I use this construction for a stacked 28-MHz, 6-element Yagi antenna system on a Rohn 45 guyed mast. It works very well and is cost effective.

The second variation is to use a fixed, or guyed, mast with the top Yagi antenna fully rotatable and a second, lower, Yagi antenna fixed in a preferred direction. This is a particularly interesting variation for contest operation, especially on the lower frequencies where the mast must be very high.

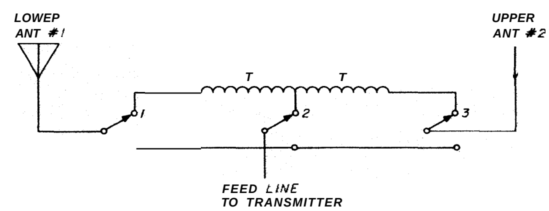
My 7-MHz system is a good example. A full-sized, three-element beam is fully rotatable on top of a 180-foot (55-meter) Rohn 45 guyed mast. A second full-sized, three-element beam is fixed at 90 feet (27 meters), which is aimed at Europe. Thus, in the European direction, full stacking is available; in all other directions the top beam can be used alone. Moreover, it is easy to excite both beams and activate two azimuthal directions simultaneously, or it is also possible to switch instantly from one direction to another without losing the normal time to turn the large Yagi antenna.

[1]For certain point-to-point communications, where the path conditions are marginal, the increased gain from lateral stacking could outweigh the nuisance of the narrower azimuthal angle.

I have found the flexibility of this system to be very helpful in many situations.

antenna excitation

My second observation is that, for all types of stacked arrays, I have found it useful to provide a switching system that allows operation of each Yagi independently or both together. When only high-angle radiation is desired, the lower antenna is usually best. For lower angles of radiation, the combined stack is better. It is easy to arrange such a switch using conventional relays and quarter-wave coaxial transformers; a practical system is shown in **fig. 1** for two stacked Yagi antennas.



CONTACTS FOR RELAYS 1, 2, 3 SHOWN IN NORMAL (UNEXCITED) STATE

RELAY	#1,2 BOTH	#1 LOWER	#2 UPPER
1	OFF	ON	OFF
2	OFF	ON	ON
3	OFF	OFF	ON

RELAY EXCITATION SCHEDULE

fig. 1. Both, lower, or upper (BLU) switching system. T is a λ/4 transformer; 70-ohm coaxial cable, λ/4 electrical length.

The relays may have to be compensated by small shunt capacitors if their series inductance is too large. The relay box should be mounted on the mast about half way between the Yagi antennas. Extension to more than two stacked Yagi antennas is equivalently easy. However, the particular scheme will depend on the way in which power is to be split between all Yagi antennas.

Because of these various excitation techniques, it is desirable to compute not only the properties of a vertically stacked Yagi system, but the properties of the individually excited Yagi antennas.

Two complicating problems arise. First, not only is a single Yagi antenna over ideal ground *not* the same antenna as in free space,[1] but it is further changed by all other Yagi antennas as well as their ground images. This is true even if all other Yagi antennas are not driven. To some extent their elements will be parasitically excited by the single driven Yagi antenna.

This means that the computation for a single Yagi must be carefully made to account fully for all the parasites and images in its local field. Second, if only

table 1. Representative good Yagi beams. All elements are cylindrical with radius p = 0 0005260(λ).

element	3-elements 0.25 (λ) boom length: length (λ)	boom position (λ)	6-elements 0.75 (λ) boom length: length (λ)	boom position (λ)
reflector	0.49801	0.000	0.49528	0.000
driven	0.48963	0.150	0.48028	0.150
D1	0.46900	0.300	0.44811	0.300
D2			0.44811	0.450
D3			0.44811	0.600
D4			0.44811	0.750

the top Yagi is rotatable, the performance of the single lower antenna alone will depend on the relative azimuthal orientation of the two antennas. In this case it is instructive to compute three cases: parallel, orthogonal, and antiparallel orientations.

stacking arrangements

Let us now choose some representative horizontally polarized stacking arrangements over flat, ideal, ground and compute their theoretical performance. I shall present computed H-plane patterns over the range of elevation angles of interest.

The E-plane pattern over ideal ground is, of course, zero everywhere. System forward gain at central design frequency is shown (from zero to 20 dBi) as a function of elevation angle (from zero to 60 degrees). The plots show not only how well the overall system performs at the important low angles, but also what may be sacrificed at the higher angles, which are occasionally useful.

Two basic Yagi designs are used. They are the same three-element beam (boom = 0.25 λ) and the same six-element beam *(boom* = 0.75 λ) shown in table **1** of the previous article.[1] They are reproduced for convenience in table **1**.

I shall start with two stacked, identical beams over ground. In practice, the height of the upper beam will be fixed at the overall mast height. The placement of the iower antenna will be made at some lower position. It is interesting to understand the tradeoffs involved in the height of the lower antenna.

antenna patterns

I shall choose, for illustrative purposes, four different heights, HU, for the upper beam (assumed to be the supporting mast height). For each of these cases, three different heights, *HL*, for the lower beam are chosen. All heights are expressed in wavelengths (λ) at the central design frequency.

Tables **2** and 3 show computed results for all these cases. These tables also refer to figs. **2** and 3, which display detailed H-plane patterns for all cases.

Note that each figure has several graphs: one for the combined stacked performance (labeled 1); one for the lower antenna alone (labeled 2); one for the upper antenna alone (labeled 3); and, where applicable, what the lower antenna only would show if no upper antenna were physically present (labeled 4).

In cases 2 and 3, both antennas are physically present, but only one is driven (all nondriven elements act as parasites). I have assumed, in these calculations, that the unused driven elements are sufficiently detuned so that they play no part in overall performance.

An examination of tables **2** and 3, and especially the H-plane patterns of figs. **2** and 3, reveals a number of interesting and important characteristics of these simple, vertically stacked systems. Table **2** shows the maximum gain and corresponding elevation angle for each case of a stacked pair of 3-element beams. Also shown is the *F/B* ratio, which we now know varies with the exact element complex current(s), which in turn are influenced by the mutual

table 2. Gain in dBi of a 3-element stack, upper height *HU(λ)* **and lower height** *LU(λ)*.

fig. no.	HL	HU	both: max. gain	angle (degrees)	F/B (dB)	lower: max. gain	angle (degrees)	F/B (dB)	upper: max. gain	angle (degrees)	F/B (dB)	lower only: max. gain	angle (degrees)	F/B (dB)
2A	0.30	0.75	14.42	21	24.34	11.48	33	19.72	13.96	19	21.83	11.74	36	19.50
2B	0.375	0.75	14.50	21	23.84	11.78	33	21.70	13.59	18	23.25	12.38	32	21.76
2C	0.45	0.75	14.55	21	21.81	11.83	34	25.81	12.90	18	26.23	12.96	29	23.06
2D	0.60	1.50	15.82	11	32.28	13.69	20	30.21	14.77	10	27.92	13.87	23	31.33
2E	0.75	1.50	16.56	11	21.14	14.51	16	23.97	15.19	10	22.28	14.07	18	21.21
2F	0.90	1.50	16.71	11	21.84	14.53	15	20.84	14.89	10	20.91	14.14	16	19.08
2G	0.90	2.25	15.61	8	17.78	14.32	16	20.00	14.30	6	20.21	14.14	16	19.08
2H	1.125	2.25	16.33	8	19.08	14.32	13	23.06	14.36	6	18.65	14.32	13	25.98
2I	1.35	2.25	16.99	7	17.19	14.62	10	17.49	14.77	6	18.23	14.35	11	19.48
2J	1.00	3.00	15.19	6	20.24	14.19	14	19.28	14.50	5	20.24	14.24	14	20.07
2K	1.50	3.00	16.32	6	18.80	14.38	10	19.84	14.41	5	19.28	14.38	9	19.78
2L	2.00	3.00	17.11	5	18.79	14.56	7	19.04	14.60	5	18.93	14.46	7	19.86

table 3. Gain in dBi of a 6-element stack, upper height $HU(\lambda)$ and lower height $LU(\lambda)$.

			both			lower			upper			lower only		
fig. no.	HL	HU	max. gain	angle (degrees)	F/B (dB)	max. gain	angle (degrees)	F/B (dB)	max. gain	angle (degrees)	F/B (dB)	max. gain	angle (degrees)	F/B (dB)
3A	0.30	0.75	15.22	20	24.23	13.46	44	9.57	15.08	17	23.13	13.40	30	17.14
3B	0.375	0.75	15.61	20	21.00	13.70	43	9.75	14.47	16	19.21	13.97	27	23.57
3C	0.45	0.75	15.63	19	18.85	13.51	43	8.87	13.45	16	14.10	14.52	25	29.80
3D	0.60	1.50	17.47	11	22.71	15.09	21	21.88	16.43	9	30.72	15.18	21	24.19
3E	0.75	1.50	17.28	11	14.07	15.16	20	18.50	15.74	9	18.11	15.59	18	21.52
3F	0.90	1.50	18.09	11	23.48	15.74	16	36.08	16.06	9	24.17	16.16	15	28.57
3G	0.90	2.25	18.00	8	30.06	15.94	14	26.76	16.73	6	44.62	16.16	15	28.57
3H	1.125	2.25	18.41	8	28.56	16.39	12	48.07	16.56	6	30.48	16.31	12	34.93
3I	1.35	2.25	18.60	7	19.87	16.26	11	22.61	16.36	6	21.58	16.38	10	31.95
3J	1.00	3.00	17.33	6	38.94	16.49	14	33.55	16.62	5	40.44	16.36	14	35.77
3K	1.50	3.00	18.70	6	32.04	16.61	9	36.88	16.80	5	35.32	16.58	9	38.61
3L	2.00	3.00	19.23	5	25.19	16.79	7	26.65	16.82	5	25.86	16.67	7	40.94

impedances to all other elements. **Table** 3 shows the equivalent quantities for the stacked pair of 6-element beams.

Note from these tables that the smaller values of overall antenna mast height, HU, do not give as much overall maximum gain as the higher antennas; this gain deficit is more severe for the 6-element beams than for the 3-element beams. This is the same general result previously obtained for single antennas over ground;[1] it results from the same phenomenon; that is, the natural increased free space directivity of the larger Yagi antennas reduces the gain potential at the higher elevation angles required for the lower antennas.

Note also from these tables that the exact placement of the lower Yagi antenna does not markedly influence the stacked maximum gain of the system but usually does significantly affect the angle of the lower antenna radiation. Note also that the excellent free space F/B ratio can be significantly affected by stacking; it is most strongly affected when the stack spacing is small and where the number of (adjacent) parasites is large, for example, especially the first three cases in **table** 3.

To properly assess all of these stacked Yagi antenna systems, it is necesary to look at the H-plane (elevation angle) patterns shown in **figs. 2** and **3.** It is instantly clear that excellent stacked coverage (curve 1) of the crucially important 0-20 degree elevation angles requires a reasonably high system ($HU = 1.0$X) but not too high ($HU = 2.5\lambda$). Above the first main lobe of radiation the patterns are quite varied; it is helpful to understand the basic reasons for these variations. **Fig. 4** shows a simplified sketch of the two Yagi antennas above ground, each one represented on this diagram by a point. The lower antenna is at a height HL (in λ) and the upper one is at a height HU (in λ); also shown are the image antennas below ground at heights of $-HU$ and $-HL$, respectively.

Note that at an elevation angle, 0, the radiation from the lower antenna lags that from the upper antenna by a distance $(HU-HL)\bullet sin\theta$ (also in λ). This phase lag causes the pair of antennas to interfere both constructively and destructively. At certain values of θ, which I shall designate θp, destructive interference will be complete and produce a radiation pattern null. Since the phase lag between the two antennas above ground is identical to that between the two images below ground, the overall radiation will also show these nulls where

$$\theta_p = sin^{-1}\,[(N + 1/2)/(HU-HL)] \qquad \textbf{(1)}$$

where N can take on integer values starting with zero $(0,$ 1, 2, . . .).

Now, from **fig. 4,** note that the radiation from the image pair (which is excited out of phase with the real antenna pair) further lags by a distance (HU + HL) • sin 0. Thus nulls will ***also*** occur in the overall pattern due to ground reflections at values of θ which I shall designate as θ_G where:

$$\theta_G = sin^{-1}\,[M/(HU + HL)] \qquad \textbf{(2)}$$

where M can assume integral values (0, 1, 2, . . .).

As an example, consider **fig. 3L** where $HU = 3.00\,\lambda$ and $HL = 2.00\,\lambda$. **Eqs. 1** and **2** predict that nulls should occur (in the range 0 to 60 degrees shown) as follows:

θ_p = *30 degrees*
θ_G = *11.5* degrees, 23.6 degrees, 36.9 degrees, 53.1 degrees

While **fig. 3L,** which shows gain only above 0 dBi, only suggests these minima, the full calculations show them all quite clearly. Moreover, note from **fig. 3L** that the upper envelope of gain falls off substantially with azimuthal angle; this general result is caused

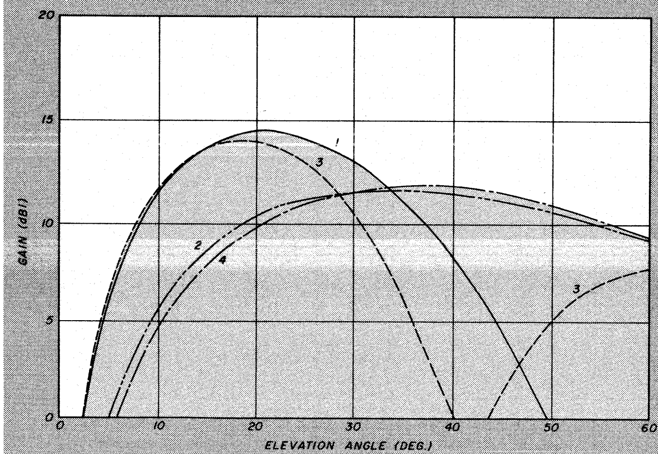


fig. 2A. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

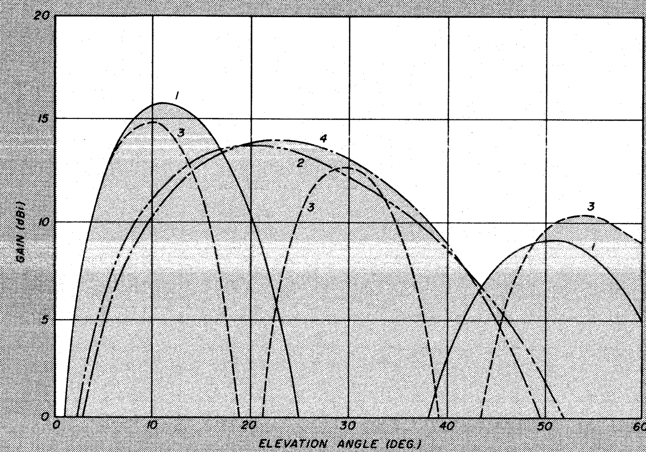


fig. 2D. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

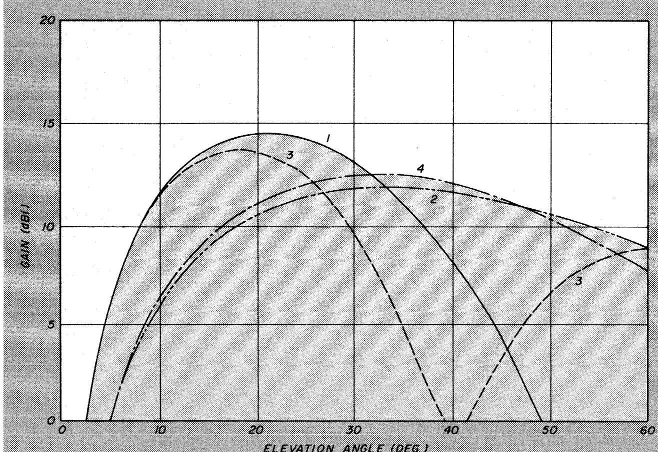


fig. 2B. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physcially absent).

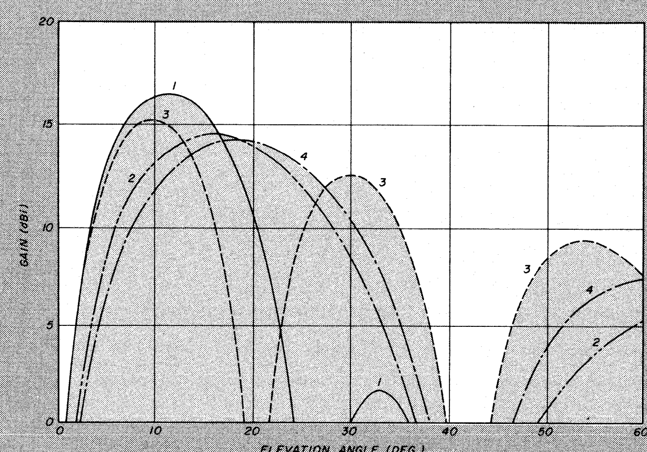


fig. 2E. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

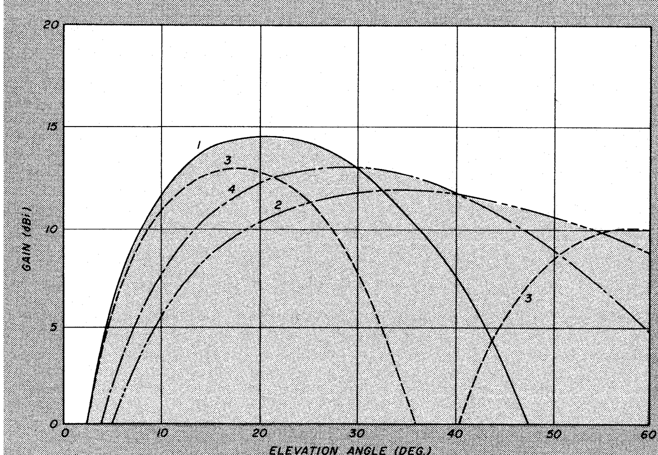


fig. 2C. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

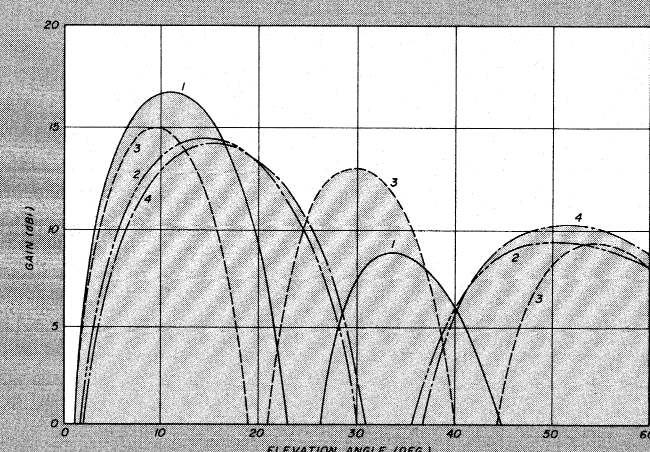


fig. 2F. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

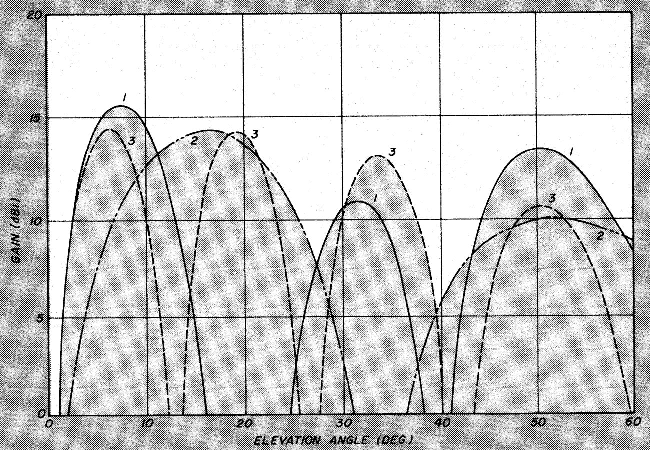


fig. 2G. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper.

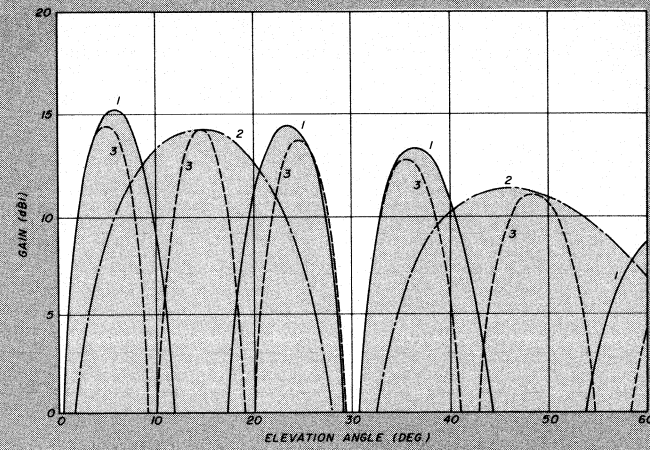


fig. 2J. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper.

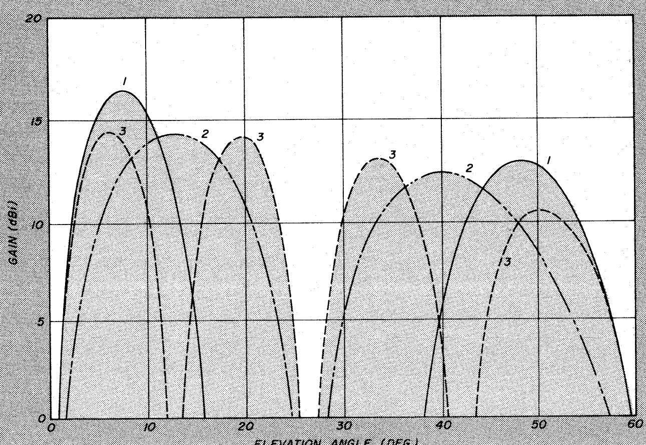


fig. 2H. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

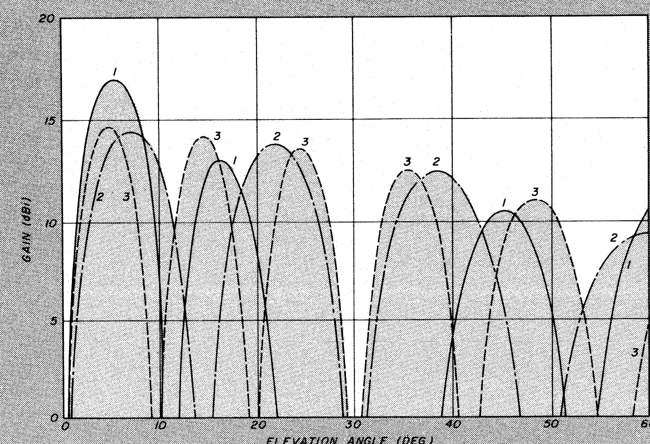


fig. 2K. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

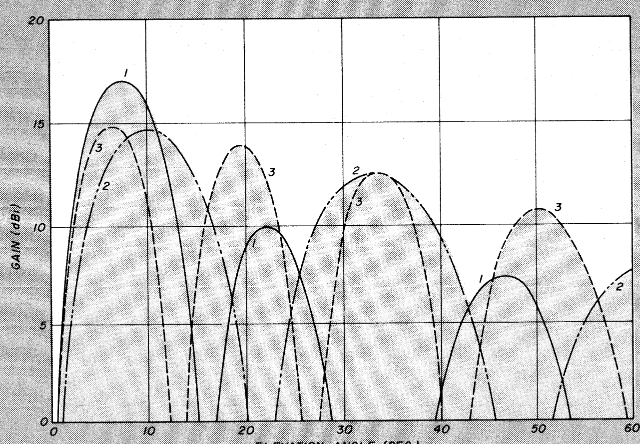


fig. 2I. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

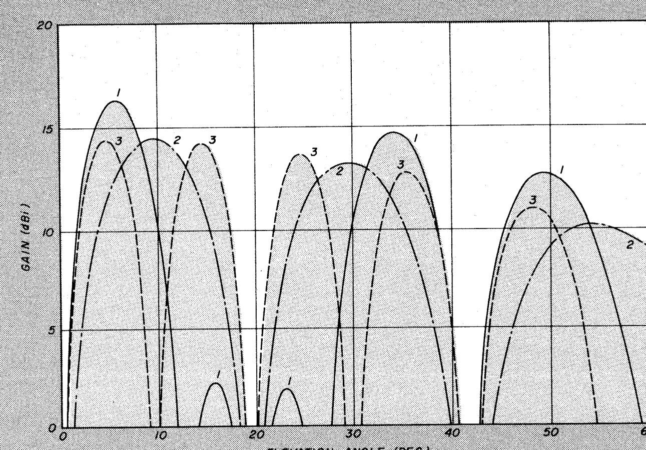


fig. 2L. Gain of a 3-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — upper only (upper physically absent).

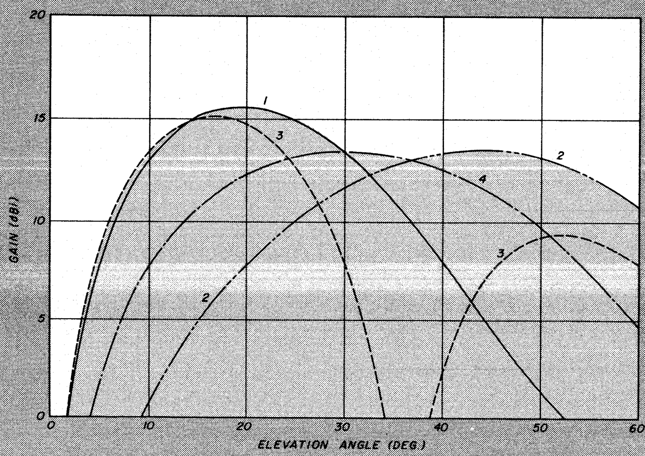


fig. 3A. Gain of a 6-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

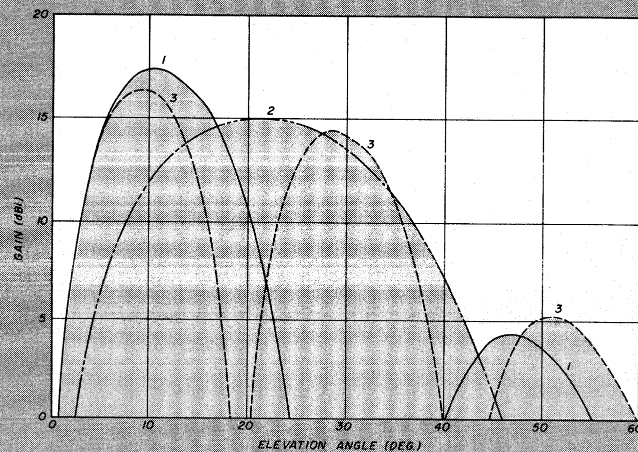


fig. 3D. Gain of a 6-element stack. Curve 1 — both, 2 — lower, 3 — upper.

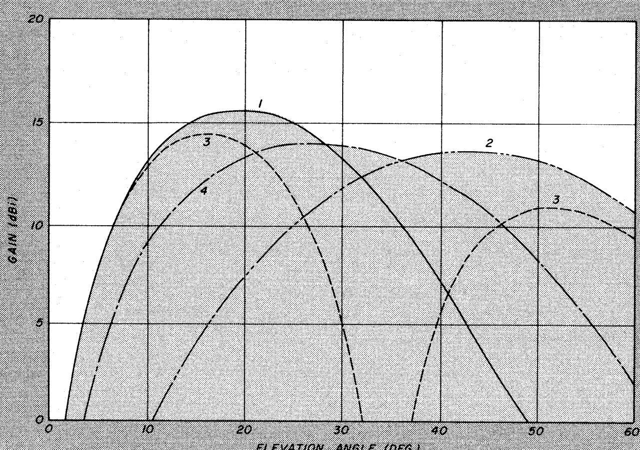


fig. 3B. Gain of a 6-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

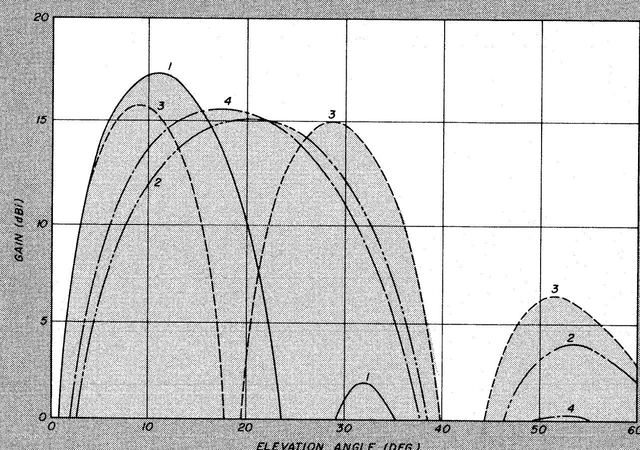


fig. 3E. Gain of a 6-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

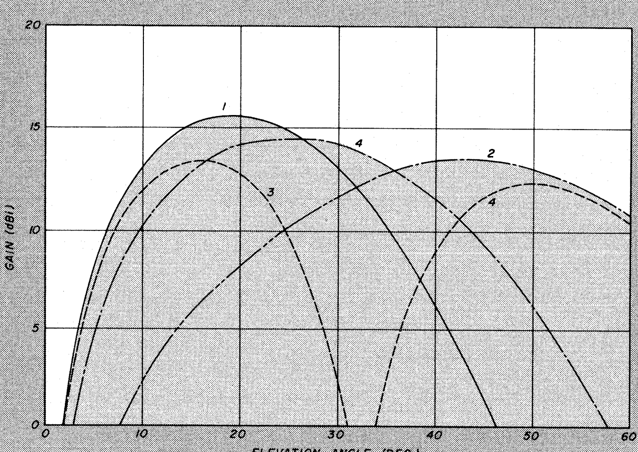


fig. 3C. Gain of a 6-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

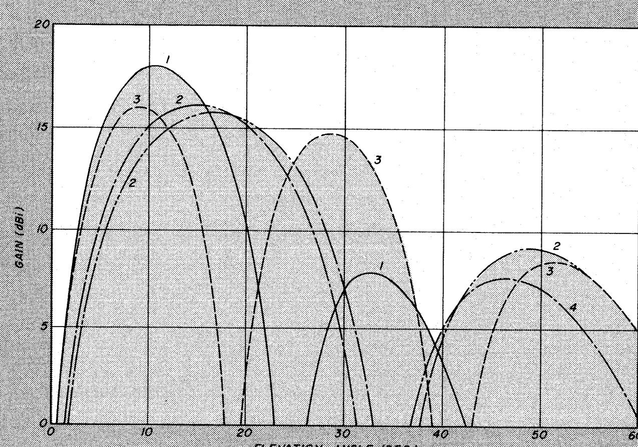


fig. 3F. Gain of a 6-element stack. Curve 1 — both, 2 — lower, 3 — upper, 4 — lower only (upper physically absent).

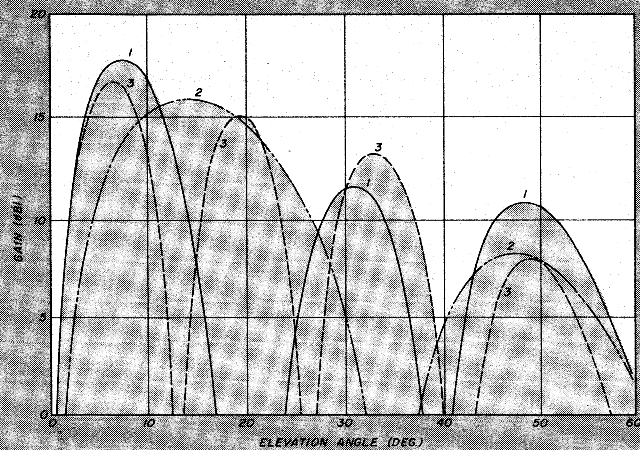


fig. 3G. Gain of a 6-element stack. Curve 1 – both, 2 – lower, 3 – upper, 4 – lower only (upper physically absent).

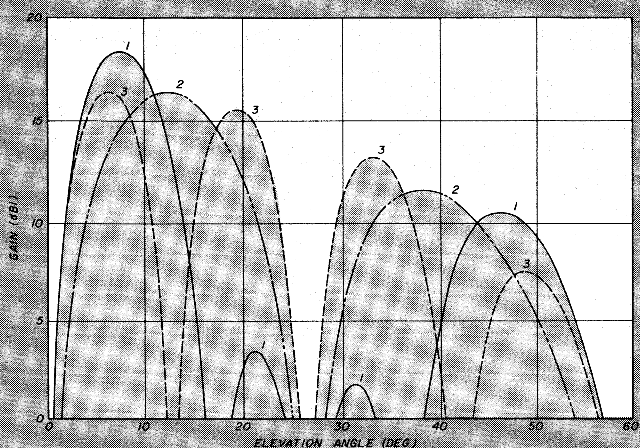


fig. 3H. Gain of a 6-element stack. Curve 1 – both, 2 – lower, 3 – upper, 4 – lower only (upper physically absent).

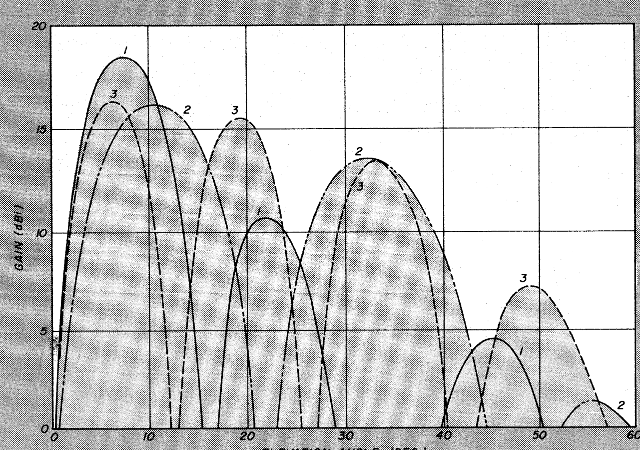


fig. 3I. Gain of a 6-element stack. Curve 1 – both, 2 – lower, 3 – upper, 4 – lower only (upper physically absent).

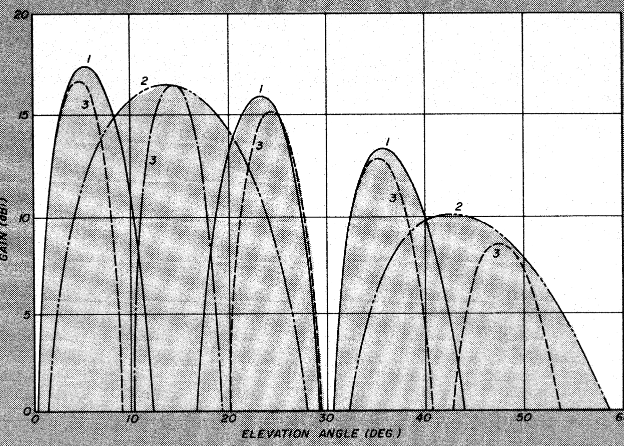


fig. 3J. Gain of a 6-element stack. Curve 1 – both, 2 – lower, 3 – upper, 4 – lower only (upper physically absent).

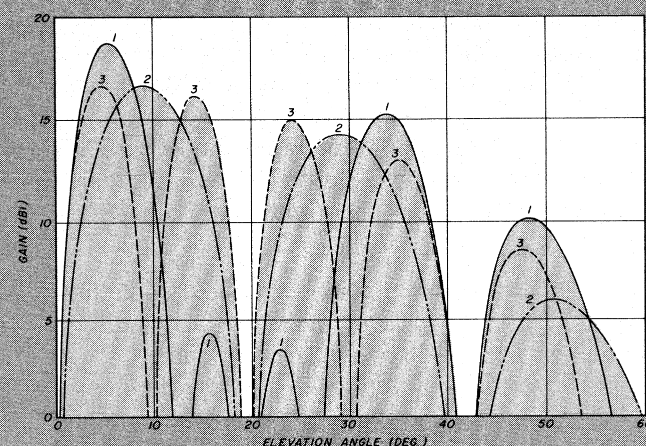


fig. 3K. Gain of a 6-element stack. Curve 1 – both, 2 – lower, 3 – upper, 4 – lower only (upper physically absent).

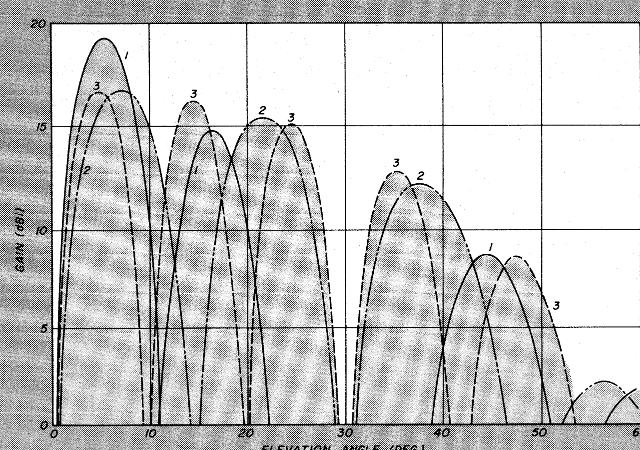


fig. 3L. Gain of a 6-element stack. Curve 1 – both, 2 – lower, 3 – upper, 4 – lower only (upper physically absent).

by the natural free space directivity of the individual Yagi antennas. Note that this effect is much more pronounced for the larger 6-element Yagi antennas (fig. 3L) than for the smaller 3-element equivalent stack (fig. 2L).

Thus, the overall H-plane pattern is the result of three effects: first, the natural free-space directivity of the individual Yagi antennas; second, the interference effect of the two real antennas; and third, the interference effect of the above ground system with its image counterpart. All three effects have different angular dependences; it is therefore not surprising that the overall resultant can be quite varied and complex.

For those readers interested in constructing a vertically stacked Yagi antenna array, a careful scrutiny of tables 2 and 3 and especially all of the relevant figures is quite enlightening. It is apparent that there is no single ideal design; nevertheless, there are a number of salient points that are worth noting.

1. A mast height (upper antenna height) of 0.75 λ is really not high enough to get very much additional gain from stacking, especially with large Yagi antennas.

2. The higher systems provide better low-angle performance than the lower systems but sacrifice (sometimes needed) high-angle performance. They also provide less gain sacrifice due to ground images for big antennas and through increased antenna spacing provide less spoiling of the inherently good individual Yagi free-space characteristics.

3. The important (lowest) first-lobe gain is only weakly dependent on the placement of the lower antenna. The gain alone would favor HL somewhat above $HU/2$ (see for example figs. 3G, 3H, and 3I); nevertheless, a lower placement (wider element spacing) will result in smaller beam interactions.

4. Mutual coupling or interaction between Yagi antennas tends to spoil the otherwise excellent properties of a single Yagi. This spoiling is most pronounced for low systems where spacings are small, not only to ground but between Yagi antennas (see for example figs. 2A, 2B, and 2C). This spoiling can be easily seen in the altered pattern(s) of the lower beam (curve 2) when the upper beam is physically present and (curve 4) when the upper beam is absent. You can also see the effect that stacking has on the F/B ratios (tables 2 and 3) and also (not shown) the effects on the calculated driving point impedances of both upper and lower Yagi antennas.

5. Interactive effects are also more serious when large Yagi antennas are used. This general result is anticipated and is due to the larger number of adjacent parasites; it is illustrated by comparing curves 2 and 4 of figs. 2B and 2C with those of figs. 3B and 3C.

6. Any good (reasonably high) stacked array will benefit by the Both, Lower, Upper or BLU switch arrangement (see figs. 3G and 1) where at high angles a fill in the performance can be made (usually) using the lower antenna only. Best higher angle fill occurs when the placement of the lower antenna is at or preferably below $HU/2$. A good practical height is $HU/3 > HL > HU/2$. Note that a good fill obtained in this way slightly compromises maximum gain; however, this compromise is really not very serious.

7. With the BLU switch available it is interesting to compare performances. In all cases, at the very lowest angles, B and U give essentially identical results, that is, the stack is just as good as the upper antenna alone. However, the stack always accepts a broader range of vertical angles in its first lobe (due to its lower average height) and at its peak has more gain than either upper or lower alone. This gain advantage is one to three dB depending on the particular stack. Although this may not seem very impressive, experience demonstrates that the stack does indeed provide a commanding performance advantage over a single Yagi antenna and, coupled with the broader vertical coverage of the first lobe, will be more consistent.

8. A number of excellent stacked arrays can be chosen from these figures. As a good example note fig. 3D. I have operated a stack very much like this on 14 MHz for several years; experience shows this to be a superb performer even without a BLU switch arrangement. Figs. 3E and 3F also look very attractive, but the closer beam spacing results in increased variations in F/B properties and probably would require a BLU switch for best high-angle fill. For a higher stack note the excellent gain performances of figs. 3G through 3I. However, for any of these cases, a fill seems desirable by the use of a BLU switch; note that for best fill at some higher angles the ***upper*** antenna should be used. For a very high stack fig. 3J provides exceptional stacked gain, and by the additional use of the lower antenna (for fill), it accommodates radiation angles up to nearly 30 degrees. However, at the 30 degree angle the system performance is abysmal, giving essentially zero response for any setting of the BLU switch.

electrically derived fill

I shall now turn briefly to an alternative method of obtaining higher-angle fill, a method that promises to be operationally simple and potentially very effective. Up to this point, I have used identical driver currents table 4. Performance of stack shown in fig. 3D vs. relative phase angle, ϕ, of lower-to-upper drive current. Gain is in dBi, elevation angle in degrees, F/B in dB, R and X in ohms, and ϕ in degrees.

phase	maximum elevation			impedance	
ϕ	gain	angle	F/B	R	X
0	17.47	11	22.71	21.96	−1.22
45	16.95	11	23.53	21.96	−1.09
60	16.47	11	23.80	21.94	−1.05
90	15.00	10	24.45	21.88	−0.99
120	15.81	28	21.46	21.79	−0.98
135	16.43	28	22.41	21.75	−0.99
180	17.26	28	25.33	21.65	−1.07
−45	16.74	11	21.80	21.86	−1.30
−90	14.53	28	37.64	21.73	−1.30
−135	16.69	28	29.43	21.65	−1.20

in both magnitude and phase. Let us now consider what effect is made on (stacked) H-plane patterns if the phase of the drive current in the lower antenna is changed relative to that in the upper antenna. I shall use as a test case the stack of **3D** and will change the relative phase angle, ϕ, of the lower antenna drive current with respect to that of the upper antenna drive current. Computation of system performance under these conditions exhibits some remarkable effects. **Table 4** shows performance as a function of ϕ (in degrees) for several discrete relative phase angles from zero to 180 degrees, and **fig. 5** shows the H-plane patterns corresponding to selected values of ϕ.

The H-plane pattern for any positive value of ϕ is nearly identical to the pattern for the same negative value of ϕ; minor differences (which are also evident in **table 4**) are caused by the detailed way in which all mutual coupling effects take place. It is easy to see from **fig. 5** that reversing the phase (ϕ = *180* degrees) results in excellent system performance at higher angles; basically giving maxima where the original H-plane pattern showed minima. At intermediate values of ϕ an intermediate result is obtained where the resulting H-plane pattern is a combination of both the ϕ = *0* degrees (original pattern) and the ϕ = *180* degrees (out-of-phase pattern). Note that this higher angle fill effectively uses the extra gain potential of both Yagi antennas; it is therefore potentially superior to a single Yagi antenna fill and is also quite easy to implement (by switching in to only one of the antenas a coaxial line whose electrical length is $\lambda/2$).

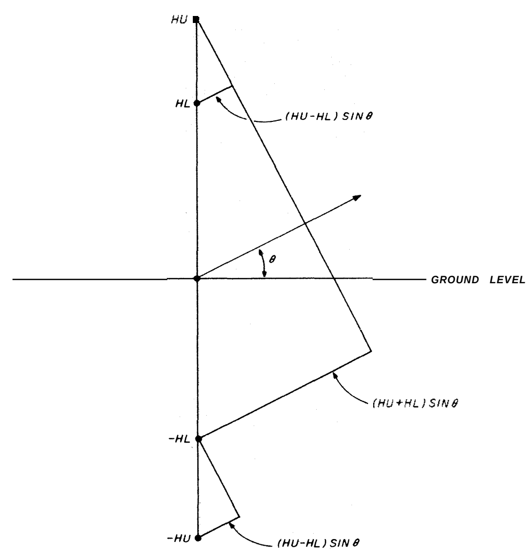


fig. 4. Two stacked Yagi antennas; lower at height HL(λ) and upper at height HU(λ).

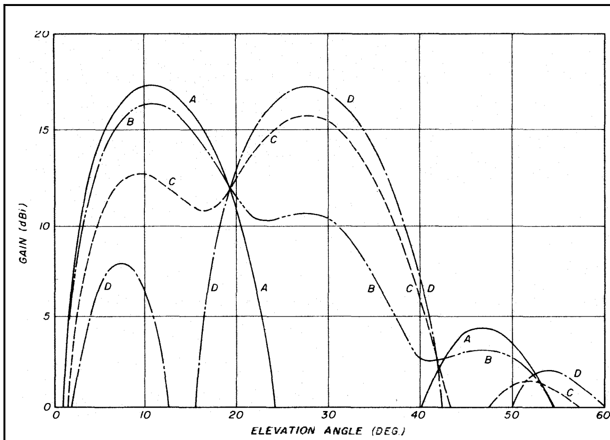


fig. 5. Gain of a 6-element stack, HL = 0.60(λ) and HU = 1.50(λ). Curve A — phase, ϕ, = 0 degrees, B — phase, ϕ, = 60 degrees, C — phase, ϕ, = 120 degrees. D — phase, ϕ, = 180 degrees.

One can also see clearly from **fig. 5** that if ϕ is relatively small little degradation of system performance occurs; this fact potentially allows the stacking of dissimilar Yagi antennas. Nevertheless the use of dissimilar antennas raises questions about how to measure the effective ϕ and certainly increases the complications of controlling ϕ over a reasonable bandwidth of frequencies.

It is important to note that only two values of ϕ are desired. The in-phase case (ϕ = *0* degrees) is best for low-angle performance and the out-of-phase (ϕ = *180* degrees) is best for higher values of elevation angle. All other values of ϕ give inferior results to either one or the other of these cases.

more than two antenna arrays

Let me now consider the possibility of stacking table 5. Gain of multi 6-element stacks, lowest at height $H1(\lambda)$ and next at $H2(\lambda)$, etc. Gain is in dBi.

fig. no.	H1	H2	H3	H4	all max. gain	all angle (degrees)	all F/B (dB)	lowest max. gain	lowest angle (degrees)	lowest F/B (dB)	top max. gain	top angle (degrees)	top F/B (dB)	lowest only max. gain	lowest only angle (degrees)	lowest only F/B (dB)
6	0.75	1.50	2.25		19 14	8	14 81	15 22	20	15 58	16.31	6	22.51	15.59	18	21.52
7	0.75	1.50	2.25	3.00	20.23	6	15.92	15.28	23	16.06	16.22	5	21.04	15.59	18	21.52

more than two Yagi antennas. It is obvious that some additional performance improvement should be possible provided the mast height is sufficiently high. As examples I show computations for two different evenly spaced stacks shown in **table 5.** The first stack shows three of the 6-element Yagis evenly spaced with a top height of **2.25 λ**, and the second shows four 6-element Yagis evenly spaced to a top height of 3.0 λ. **Figs. 6** and **7** show the H-plane patterns for these two systems; they should be compared with **figs. 3G** and **3J**, which are basically equivalent 2-Yagi stacks of comparable mast height. It is at once apparent from **fig. 6** that the addition of the third Yagi antenna gives a main lobe gain over **fig.** 3G of **1.1** dB; all other characteristics are quite comparable. Likewise, **fig. 7** shows that the two additional Yagi antennas give a main lobe gain increase of 1.9 dB over **fig.** 3J; again, all other characteristics are quite similar.

These examples of vertically stacked Yagi antenna arrays using more than two antennas show that a noticeable gain increase is possible over a 2-antenna stack; moreover they open up a wide range of higher-angle fill possibilities. As an example, **fig. 7** shows patterns where all four antennas are excited; for fill at higher angles, the lowest antenna (curve 2) and the highest antenna (curve 3) are shown. Note, however, that additional fill situations are possible if two or even three of the original four antennas are excited coherently. Moreover one can also consider feed line phasing(s) for even better fill. Clearly, a host of possibilities exists, but the practical use of all potentially desirable combinations not only requires a complex switching system but a great deal of trouble in determining experimentally the right combination for the prevailing circuit conditions. Surely the additional complexity and expense of these large vertical stacks reaches a point of practical diminishing returns. Nevertheless, how fortunate we are to be able to predict with reasonable confidence the performance of such large systems, without ever having to build one.

I shall conclude this article on stacking by referring again to the basic two-antenna stack shown in **fig. 3D**. Note that the F/B performance has deteriorated from the excellent free-space performance of the individual Yagi antennas of 48 dB to 22.7 dB. Analogous to the optimization of a single Yagi antenna over ground,[1] it is possible to optimize the basic Yagi design for this stacked system. **Table 6** shows the optimized parameters of the 6-element Yagi antenna first for free space and second for the stack of **fig. 3D** ($HL = 0.6\lambda$, $HU = 1.5\lambda$. **Tables 7** and **8** show the swept-frequency performance of each of these cases close to the design frequency.

table 6. Specifications for 6-element beams optimized by slight shifts in boom positions for D1 and D3. All element radii are $0.0005260(\lambda)$.

element	optimized for free space: length (λ)	optimized for free space: boom position (λ)	optimized for stack fig. 3D: length (λ)	optimized for stack fig. 3D: boom position (λ)
reflector	0.49528	0.000	0.49528	0.000
driven	0.48071	0.150	0.48157	0.150
D1	0.4481 1	0.2991650	0.44811	0.302948
D2	0.44811	0.450	0.44811	0.450
D3	0.44811	0.5999658	0.44811	0.63948
D4	0.44811	0.750	0.44811	0.750

The iterative optimization was carried out by adjustments of the boom positions of D1 and D3 to obtain high F/B (> 90 dB) and by a slight adjustment of driven-element length to minimize reactance at the design frequency. Note again that, because of mutual coupling intereactions, the stacked Yagi antenna is ***not*** the same Yagi antenna as it would be in free space, nor is it the same Yagi antenna as it would be

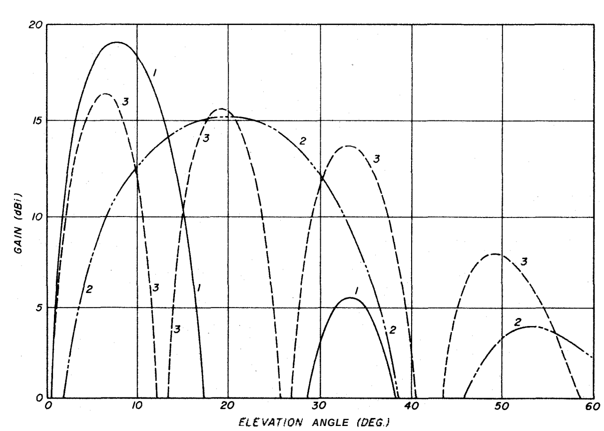


fig. 6. Gain of 6-element stack of 3 Yagi antennas. H1 (lowest) = 0.75(λ), H2 (middle) = 1.50(λ), H3 (highest) = 2.25(λ). Curve 1 — all, 2 — lowest, 3 — top.

table 7. Free-space performance of optimized 6-element Yagi (specificationin table 6).

frequency	gain (dBi)	F/B (dB)	angle (deg.)	driver impedance (ohms) R	X
0.996	10.59	26.59	0	22.42	−6.53
0.998	10.65	32.64	0	22.10	−3.50
1.000	10.70	120.18	0	21.80	−0.42
1.002	10.75	32.69	0	21.51	2.72
1.004	10.79	26.68	0	21.25	5.93

table 8. Performance of (fig. 3D) optimized, stacked 6-element Yagi over ground. (Specification in table 6.)

frequency	gain (dBi)	F/B (dB)	angle (deg.)	driver impedance (ohms) R	X
0.996	17.28	24.90	11	16.76	−6.24
0.998	17.35	30.86	11	16.25	−3.14
1.000	17.42	94.69	11	15.75	0.02
1.002	17.48	30.69	11	15.27	3.23
1.004	17.53	24.57	11	14.82	6.50

singly over ground (compare with **table 5** of reference 1).

orthogonal and antiparallel stacked Yagis

Now that an optimized Yagi design has been found, which provides the superlative performance shown in **table 8** when the two stacked antennas are parallel to each other, we can ask about performance degradation when the two Yagi antennas are orthogonal to each other and also when they are antiparallel. This question is relevant when a stack is used where the lower antenna is fixed in some direction and the upper antenna is rotatable. **Table 9** shows the system performance for the case where both Yagis are supported at the center of the boom.

It is clear that, in principle, optimization can be carried out for only one configuration, and performance will automatically deteriorate somewhat for other geometries. The extent of deterioration will be more severe for stacks with small antenna spacings (both with respect to ground and to each other); that is,

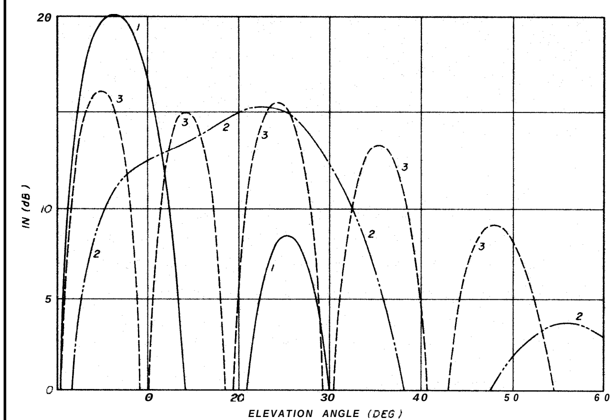


fig. 7. Gain of 6-element stack of 4 Yagi antennas. H1 (lowest) = 0.75(λ), H2 (next) = 1.50(λ), H3 (next) = 2.25(λ), H4 (top) = 3.00(λ). Curve 1 — all, 2 — lowest, 3 — top.

with lower overall mast height. For the stack shown in **table 9, it** is gratifying to see that the performance for all situations is really quite acceptable.

summary

1. Vertical stacking of two Yagi antennas allows both substantial improvement in low-angle system performance and improved flexibility. This flexibility can be used either to obtain fill at some needed higher angles or to illuminate other azimuthal angles (one of two Yagi antennas rotatable).

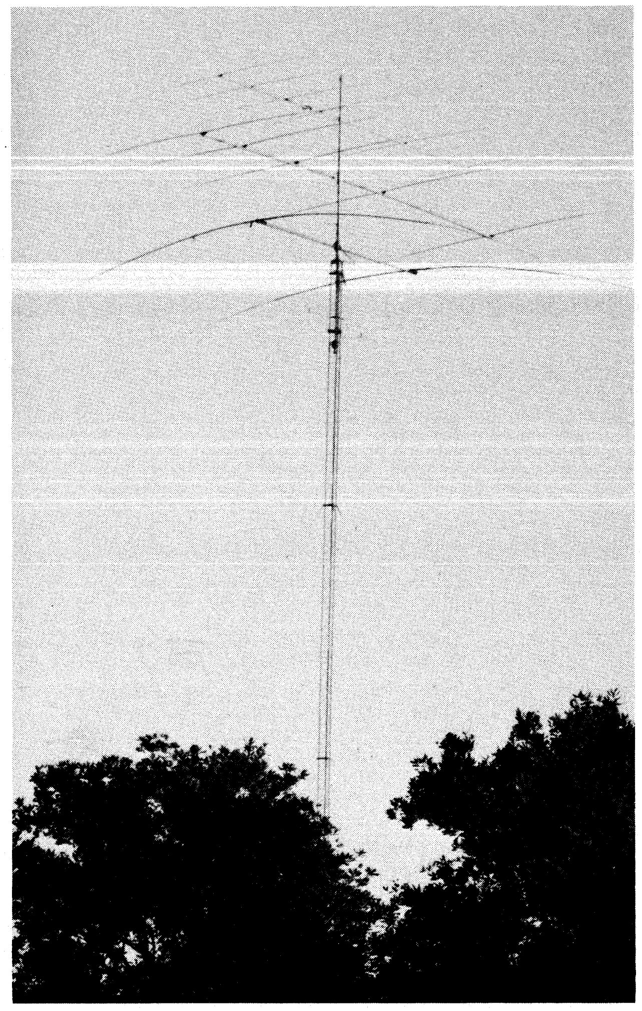
2. Mast heights of between one and perhaps 2.5 λ can provide excellent 2-Yagi stacked systems.

3. Higher masts favor low-angle radiation and also give smaller mutual interaction effects. However, they also treat the (occasionally useful) higher angles unfavorably.

4. For all vertical stacks, improved performance is available if excitation is switchable to both antennas, *B*, the lower antenna, *L*, or the upper antenna *U* (BLU switch). Switching must be done in a way that preserves phase integrity and keeps the total drive impedance matched to the supply coaxial line. For

table 9. Gain in dBi of a parallel-optimized, stacked 6-element Yagi array showing the alternative configurations.

	both				lower				upper			
configuration	gain (dB)	F/B (dB)	R (ohms)	X (ohms)	gain (dB)	F/B (dB)	R (ohms)	X (ohms)	gain (dB)	F/B (dB)	R (ohms)	X (ohms)
parallel	17.42	94.69	15.75	0.02	15.06	30.34	13.66	−0.41	16.40	25.64	15.59	−0.51
orthogonal					15.15	38.09	14.10	−0.80	16.60	20.86	15.87	−0.82
antiparallel					15.03	31.24	13.63	−0.44	16.43	23.88	15.57	−0.54
angle (deg.)	11				21				9			



those antenna stacks where the lower and upper beams remain aligned (rotate together), a highly useful switch is a phase reverser to only one of the beams.

5. Vertical stacks using 3 or **4** Yagi antennas can display even greater performance, but the stacks must be very high and must use for best results a more complex feed-line switching arrangement.

6. Optimization (very high F/B at one frequency) can be obtained for only one physical configuration at a time. Nevertheless, there are practical examples where an optimized antenna design for a 2-Yagi stack will still exhibit excellent properties when only the lower antenna or only the upper antenna is excited; moreover, these excellent properties are retained even if the azimuthal directions of the two individual Yagi antennas are parallel, orthogonal or even anti-parallel.

reference

1. J.L. Lawson, "Yagi Antenna Design: Ground or Earth Effects," *ham radio,* October, 1960, page 29.

ham radio

More Details? CHECK—OFF Page 94

crystal use locator

A computer program for matching crystal frequencies to popular Amateur radios

If you're like I am, you have a box full of crystals. They came from scrapped equipment, 5-for-$1.00 sales that couldn't be passed up, old fm rigs, and abandoned frequencies. I decided that I wanted to find out if the crystals were good for anything in the radios I own. To save a lot of hand calculations I wrote program XLOC. Within XLOC are the crystal formulas for all the radios I own and the frequencies on which these radios can be used. If I type in a crystal frequency, XLOC prints out all the radios in which the crystal is usable and the resulting frequency.

program description

Listing 1 is a sample run of XLOC. After entering the crystal frequency in MHz, XLOC plugs it into the crystal formulas for each radio and checks the result to determine if it's in the specified band for that radio. If one or more of these calculations generates an in-band result, the crystal frequency is printed, followed by pairs of radio descriptions and resulting operating frequencies. If the crystal doesn't produce a valid operating frequency for any of the radios, the message "SORRY!" is printed and XLOC prompts for the next crystal frequency. Entering a 0 in response to the prompt causes XLOC to terminate.

Internally XLOC is straightforward. It's within Technical Systems Consultants' extended BASIC program for the Model 6800 computer.* Looking at **listing** 2, lines 30 through 100 establish the possible operating frequency ranges for the radios.

Example. Line 40 establishes the frequency range for band 1 to be 28 MHz to 30 MHz. Seven bands are set up in XLOC. To add a new band, just add statements after line 100 to set up new values in the *BS* and *BE* arrays. For example, to set up a band from 220-225 MHz as band 8 you'd add the statement

110 BS(8) = 220:BE(8) = 225

Note that *BS* and *BE* are dimensioned at 20. If you want to handle more than 20 bands total, line 30 would have to be changed.

The other information that must be set up is the radio descriptions, the bands on which they operate, and the crystal formulas. Lines 180 through 520 establish the first two. Each string in array *N$* contains a radio description, and the corresponding element in array *B* contains the number of the band on which it is designed to operate. More radio description/band pairs can be added after line 520. *B* and *N$* are dimensioned at 20; therefore, the DIM statement at line 180 would have to be changed to handle more than twenty radios.

Logic. Line 660 reads the crystal frequency into variable X. Then variable FD is set to zero to indicate that no radio requiring this crystal has yet been found. Variable C is set to 1 to indicate that we are starting with the first radio. This is used as an index for the *N$* and *B* arrays. Lines *700* through 1030 consist of pairs of statements. The first statement is the crystal formula for the radio; the second statement is always a GOSUB 1080.

The subroutine at line 1080 checks to see if the operating frequency (*OP*) that resulted from the calculation in the first statement falls within the desired band. If it does, the radio name and operating frequency are printed. In any case, C is incremented so it points to the information about the next radio. If more radios have been added, then the crystal formulas, each followed by a GOSUB 1080, must be added following line 1030. Each formula should be written using the variable OP as the result and X as the crystal frequency. For example, if a transmitter operates on 32 times the frequency of its crystal, the formula would be written:

OP = 32*X

Finally, if no radio is found that could use the crystal (indicated by FD never getting set to 1) the message "SORRY!" is printed and control is transferred back to the prompt message.

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*Southwest Technical Products.

listing 1. **Sample run of XLOC program. Enter the crystal frequency in MHz. The output shows if the crystal is in the specified band for the radio of interest.**

```
RUN

Crystal Use Locator

Enter crystal frequency in MHz (or 0 if done)? 11.67267
SORRY!


Enter crystal frequency in MHz (or 0 if done)? 11.10417

      11.10417

GE 7668326-G1 RCVR
      447.75012  MHz


Enter crrstal frequency in MHz (or 0 if done)? 10.93333

      10.93333

GE 7668326-G1 RCVR
      441.59988  MHz


Enter crrstal frequency in MHz (or 0 if done)? 11,845

      11.845

RCA CMU-15 RCVR
      441.565  MHz


Enter crystal frequency in MHz (or 0 if done)? 12.3194

      12.3194

SWAN FM-2X XMTR
      147.8328  MHz
WILSON 1402-SM XMTR
      147.8328  MHz
RCA CMU-10 XMTR
      443.4984  MHz
GE 7669061-G1 XMTR
      443.4984  MHz


Enter crystal frequency in MHz (or 0 if done)? 37.2

      37.2

GE 2 MTR RCVR
      155.835  MHz


Enter crystal frequency in MHz (or 0 if done)? 0

READY
```

listing 2. **XLOC program listing. Seven bands are set up. To add a new band add statements after line 100 to set up new values in the *BS* and *BE* arrays. Program will handle 20 Amateur bands; if more are desired, line 30 must be changed.**

```
READY

LIST

  10 REM XLOC - Crystal Use Locator SSC 1-80
  20 REH Band start and end values
  30 DIM BS(20),BE(20)
  40 BS(1)=28:BE(1)=30
  50 BS(2)=50:BE(2)=54
  60 BS(3)=28:BE(3)=54
  70 BS(4)=144:BE(4)=148
  80 BS(5)=144:BE(5)=174
  90 BS(6)=438:BE(6)=450
 100 BS(7)=420:BE(7)=470
 160 REM N$ is the radio description
 170 REM B is the desired band
 180 DIM N$(20),B(20)
 190 N$(1)='SWAN FM-2X RCVR'
 200 B(1)=4
 210 N$(2)='SWAN FM-2X XMTR'
 220 B(2)=4
 230 N$(3)='WILSON 1402-SM
 240 B(3)=4
 250 N$(4)='WILSON 1402-SM XMTR'
 260 B(4)=4
 270 N$(5)='RCA CMU-10 RCVR'
 280 B(5)=7
 290 N$(6)='RCA CMU-10 XMTR'
 300 B(6)=6
 310 N$(7)='RCA CMU-15 RCVR'
 320 B(7)=6
 330 N$(8)='RCA CMU-15 XMTR'
 340 B(7)=6
 350 N$(9)='LINK 2240 XMTR'
 360 B(9)=4
 370 N$(10)='LINK 1905 RCVR'
 380 B(10)=5
 390 N$(11)='LINK 2210 RCVR'
 400 B(11)=5
 410 N$(12)='GE 7668326-G1 RCVR'
 420 B(12)=6
 430 N$(13)='GE 7669061-01 XMTR'
 440 B(13)=6
 450 N$(14)='GE 2 MTR RCVR'
 460 B(14)=5
 470 N$(15)='GE 2 MTR XMTR'
 480 B(15)=4
 490 N$(16)='GE ET-6-B XMTR'
 500 B(16)=2
 510 N$(17)='GE ER-6-H RCVR'
 520 B(17)=2
 630 PRINT 'Crystal  Use Locator'
 640 PRINT
 650 FRINT 'Enter  crystal frequency in MHz (or 0 if done)';
 660 INPUT X
 670 IF X=0 THEN 1200
 680 FD=0
 690 C=1
 700 OP=(X*3)+10.7
 710 FOSUH 1080
 720 OP=X*12
 730 GOSUB 1080
 740 OP=(X*9)+10.7
 750 GOSUH 1080
 760 OP=X*12
 770 GOSUB 1080
 780 OP=(X*48)+39.1
 790 GOSUB 1080
 800 OP=X*36
 810 GOSUH 1080
 820 OP=(X*36)+15.145
 830 GOSUB 1080
 840 OP=X*36
 850 GOSUX 1080
 860 OP=X*48
 870 GOSUH 1080
 880 OP=(X*18)-5
 890 GOSUH 1080
 900 OP=(X*4)-4.943
 910 GOSUH 1080
 920 OP=(X*36)+48
 930 GOSUB 1080
 940 OP=X*36
 950 GOSUR 1080
 960 OP=(X*4)+7.035
 970 GOSUR 1080
 980 OP=X*24
 990 GOSUB 1080
1000 OP=X*24
1010 GOSUB 1080
1020 OP=X*6
1030 GOSUB 1080
1040 IF FD=0 THEN PRINT 'SORRY!'
1050 PRINT
1060 PRINT
1070 GOTO 650
1080 REM Check for match
1090 IF OP>BE(B(C)) THEN 1180
1100 IF OP<BS(B(C)) THEN 1180
1110 IF FD=1 THEN 1160
1120 FD=1
1130 PRINT
1140 PRINT TAB(5);X
1150 PRINT
1160 PRINT N$(C)
1170 PRINT TAB(5);OP;'MHz'
1180 C=C+1
1190 RETURN
1200 END


READY
```

transmission-line circuit design

Using distributed resonant circuits for VHF/UHF transmission lines — equations and design relationships

Resonant transmission-line circuits predominate at frequencies above 50 MHz. The reason is that finite practical lengths of transmission line can be readily used to achieve resonance rather than lumped values of capacitance and inductance in the less-than-10 pF and low-nH range respectively. This article shows how various transmission-line configurations can be designed into resonant circuits at these higher frequencies.

The spectrum above 50 MHz becomes ever more important as advances in technology permit more extensive exploitation. Low-noise-figure devices, higher-power devices, greater efficiency, and low cost have yielded hardware today that ten years ago was considered impossible.

Technology advances haven't been limited to devices only. The complex calculations required for rf and other engineering problems have been simplified by programmable scientific calculators such as the Hewlett-Packard HP-67/97 and Texas Instruments

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dedication statement

I *would like to dedicate this article to the memory of Jim Fisk, W1HR, since it was due to his encouragement that this extensive task was undertaken.*

TI-59. These tools permit solution of design problems in hours rather than weeks, with a precision previously considered unrealistic. An interesting by-product of the time saved is that a more thorough trade-off between design parameters can be achieved, yielding a more efficient final design. More than 15 years ago I addressed this same topic (reference 1), necessarily more crudely. This article provides a fresh approach, using more elegant tools, which results in significantly expanded and new data.

The article is divided into three parts. First I address the governing expressions for calculating resonant transmission-line parameters. Included are data on design relationships such as efficiency, coupling, and resonating capacitance. The second part concerns the geometry of twelve different transmission lines in common use and derives the parameters for resonant-circuit design. The third part gives design examples demonstrating the use of the data provided.

This article is the result of my interest in resonant transmission-line phenomena for over 30 years. Comments, corrections, new formulations, or additional data are welcomed.

calculation of resonant circuits

A description of methods to calculate parameters of resonant transmission-line circuits is provided. After deciding on the physical parameters that fit the chosen configuration, efficiency and coupling are discussed. Graphs and tables are provided.

HP-67/97 programs are detailed using the HP-97 printer capability. Therefore, the tables describing the programs are displayed in HP-97 format. To convert the programs for HP-67 use, refer to the ***HP-67 Owners Handbook,*** Appendix *E*, page 324. The programs contained here are usable on the new HP-41C with the same magnetic cards.

Distributed resonant circuits. Distributed resonant circuits are different from lumped-constant circuits: fixed elements of inductance and capacitance are not required for resonance. Instead, the uniform values of distributed inductance and capacitance per unit length of transmission line are used together with a fixed line length and a capacitance, with the physical configuration determining line impedance, Z_0.

Distributed resonant circuits become practical at frequencies of 50 MHz and above because reasonable values of loaded circuit Q can be achieved with realistic geometries. Furthermore, since there are many transmission-line configurations to choose from, space constraints are readily met.

Optimum transmission-line configurations are generally available. Each is chosen to fit within the physical dimensions, materials, and hardware available. To provide maximum flexibility, twelve different configurations are discussed. See fig. **1.**

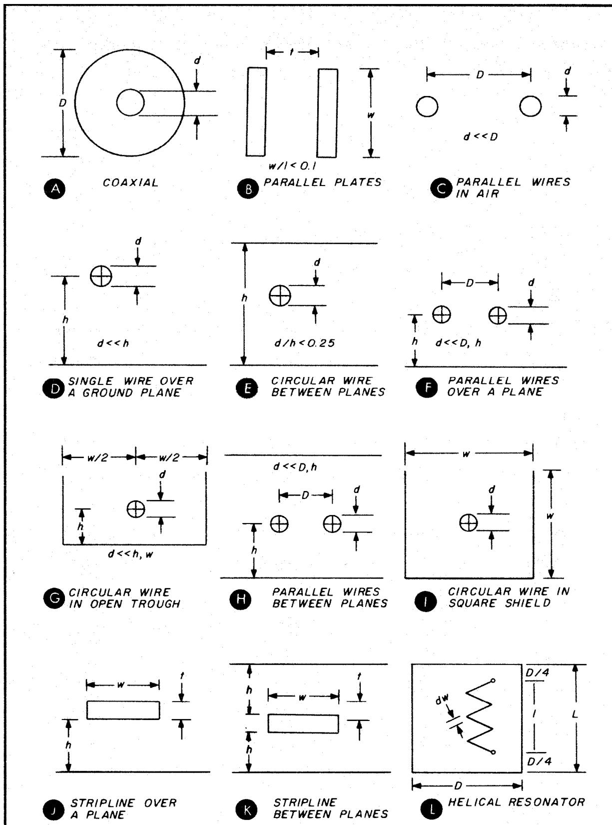

fig. 1. Distributed resonant transmission-line configurations discussed in this article.

To determine the parameters of a resonant quarter-wavelength ($\lambda/4$) segment of transmission line, the following expression is used:

$$X_C = Z_0 \tan \beta \ell \qquad (1)$$

where $X_C = \dfrac{1}{2\pi FC}$, capacitive reactance necessary to resonate line section (2)

F = resonant frequency (hertz)

C = capacitance at the unterminated end of the $\lambda/4$ line (farads)

β = electrical degrees per unit length at the resonant frequency (°/in. or °/cm)

Z_0 = transmission-line impedance (ohms)

ℓ = length of line used (inches or cm)

A coaxial line cross section is shown in fig. **2A,** which illustrates the key parameters considered. The ratio of D/d determines line impedance Z_0. (This parameter is discussed in detail in a following sec-table 1. HP-67/97 program for calculating Z, C, ℓ, or F given any three unknowns.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
			051	RCL7	36 07	103	CHS	– 22
			052	X = 0?	16-43	104	–	– 24
001	*LBLA	21 11	053	RTN	24	105	ST04	35 04
002	STO0	35 00	054	x	– 35	106	RTN	24
003	3	03	055	Pi	16-24	107	*LBL4	21 04
004	0	00	056	2	02	108	2	02
005	0	00	057	x	– 35	109	4	04
006	X⇄Y	– 41	058	x	– 35	110	0	00
007	–	– 24	059	1/X	52	111	STO9	35 09
008		– 62	060	ST06	35 06	112	*LBL6	21 06
009	0	00	061	RTN	24	113	2	02
010	2	02	062	*LBL1	21 01	114	ST-9	35-45 09
011	5	05	063	RCL5	36 05	115	RCL3	36 03
012	4	04	064	RCL2	36 02	116	RCL4	36 04
013	–	– 24	065	x	– 35	117	x	– 35
014	STO1	35 01	066	TAN	43	118	I-X	52
015	PSE	16 51	067	1/X	52	119	1	01
016	3	03	068	RCL6	36 06	120	3	03
017	6	06	069	x	– 35	121		– 62
018	0	00	070	ST03	35 03	122	4	04
019	X⇄Y	– 41	071	PIS	51	123	7	07
020	–	– 24	072	*LBL2	21 02	124	5	05
021	STO2	35 02	073	RCL6	36 06	125	x	– 35
022	PSE	16 51	074	RCL3	36 03	126	RCL9	36 09
023	RCL0	36 00	075	–	– 24	127	x	– 35
024	EEX	– 23	076	TAN^{-1}	16 43	128	STOB	35 12
025	6	06	077	RCL2	36 02	129	RCL5	36 05
026	x	– 35	078	–	– 24	130	3	03
027	ST07	35 07	079	STO5	35 05	131	6	06
028	RTN	24	080	RIS	51	132	0	00
029	*LBLB	21 12	081	*LBL3	21 03	133	x	– 35
030	STO3	35 03	082	RCL2	36 02	134	RCL9	36 09
031	RTN	24	083	RCL5	36 05	135	–	– 24
032	*LBLC	21 13	084	x	– 35	136	TAN	43
033	ST04	35 04	085	TAN	43	137	RCLB	36 12
034	GSB9	23 09	086	RCL3	36 03	138	X⇄Y	– 41
035	RTN	24	087	x	– 35	139	–	– 45
036	*LBLD	21 14	088	ST06	35 06	140		– 62
037	ST05	35 05	089	GSB8	23 08	141	0	00
038	RTN	24	090	RIS	51	142	1	01
039	*LBLa	21 16 11	091	*LBL8	21 08	143	X≤Y?	16-35
040	STOI	35 46	092	RCL7	36 07	144	GTO6	22 06
041	DSP5	– 63 05	093	x	– 35	145	RCL9	36 09
042	GSBi	23 45	094	Pi	16-24	146	1	01
043	*LBL9	21 03	095	2	02	147	1	01
044	RCL4	36 04	096	x	– 35	148	8	08
045	EEX	– 23	097	x	– 35	149	1	01
046	1	01	098	1/X	52	150	1	01
047	2	02	099	STO8	35 08	151	X⇄Y	– 41
048	CHS	– 22	100	EEX	– 23	152	–	– 24
049	x	– 35	101	1	01	153	ST00	35 00
050	ST08	35 08	102	2	02	154	RIS	51

table 2. HP-67/97 program run time for three frequencies.

F	77.70395 MHz	144.03659 MHz	2952.75 MHz
ℓ	17	13.14	0.4
C	35	10	1.2
Z_0	70	70	70
run time	2 min. 10 sec.	3 min. 54 sec.	5 min. 45 sec.

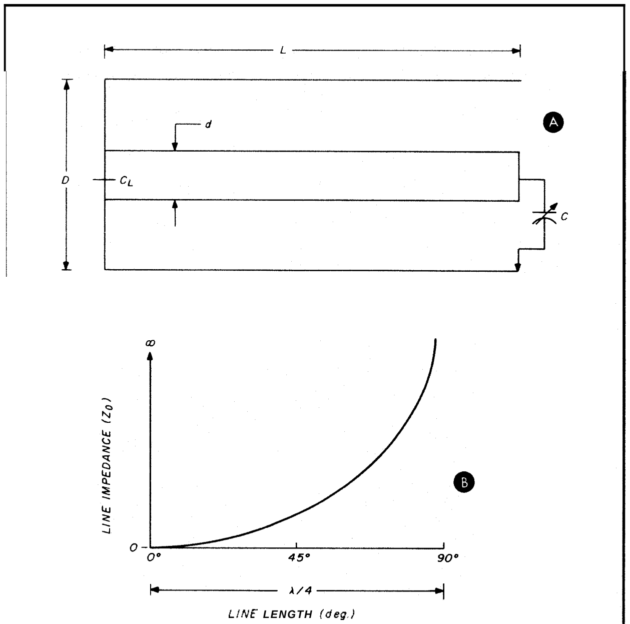

fig. 2. Coax line cross section showing key parameters for resonance, (*A*). Sketch *B* shows line impedance as a function of this line length.

tion.) Parameter β (**eq. 1**) is determined from the following relationship:

$$\beta = \frac{360^\circ}{\lambda} \qquad \textbf{(3)}$$

where λ = wavelength in the same units as line length.

The basic relationship is

$$\lambda = \frac{3 \times 10^8}{F_{Hz}} \qquad \textbf{(4)}$$

for $$\lambda_{in} = \frac{11{,}810}{F_{MHz}} \text{ (air)*} \qquad \textbf{(5)}$$

$$\lambda_{cm} = \frac{29{,}999}{F_{MHz}} \text{ (air) *} \qquad \textbf{(6)}$$

*dielectric constant of air = 1.0006

Also the general relationship of line impedance versus electrical degrees for a quarter-wave section, shorted at one end, is shown in **fig. 2B.** Note that this line impedance is totally different from the capacitive load reactance, X_C, in **eq.** *1*, which resonates the line length.

Computer program. An HP-67/97 program was written to calculate Z_0, C, ℓ, and F, given any three parameters. The program is shown in **table 1.** Since the solution of **eq. 2** for F yields a transcendental function, an approximation convergence solution is used when frequency is requested. The lower limit of this program is 49.21 MHz, where λ = 240 inches, and the program requires a match to 1 per cent to achieve a solution. If lower frequency requirements are necessary, the values in the program may be changed in steps 108, 109, and 110 (**table 1**). The maximum wavelength-to-lowest-frequency ratio is 999 inches or 11.82 MHz (**eq. 5**).

The lower frequency of 49.21 MHz was chosen in this program to minimize the solution time. The higher the frequency away from the beginning trial frequency of 49 MHz, the longer the computing time — minutes are involved. (See **table 2.**)

Table 3 identifies the register contents used in the HP-67/97 program. **Table 4** identifies the formulas used for calculation. **Table 5** shows how the program is controlled for each desired value. **Table 6** shows sample problems entered and the calculated results.

table 3. Register contents for HP-67/97 program when calculating Z_0, C, ℓ, and F.

STO 0	F_{MHz}
STO 1	λ_{inches}
STO 2	$\beta_{°/inch}$
STO 3	Z_0
STO 4	C_{pF}
STO 5	ℓ inches
STO 6	$X_{C\,ohms}$
STO 7	F_{Hz}
STO 8	C_{farads}
{STO 9 / STO B}	transient for loop

table 4. HP-67/97 formulas for calculating Z_0, ℓ, C, and F.

calculates Z_0

$$Z_0 = \frac{\tan \beta \ell}{X_C} \qquad \textbf{(7)}$$

calculates ℓ_{inches}

$$\frac{\tan^{-1} 1/32}{P} = \frac{X_C}{Z_0} \qquad \textbf{(8)}$$

calculates C_{farads} for resonance

$$C = \frac{1}{2 F \pi Z_0 \tan \beta \ell} \qquad \textbf{(9)}$$

calculates F_{MHz}*

$$\frac{13.475 \lambda_{in.}}{C_{pF}} = Z_0 \tan \frac{360^\circ}{'in.} \ell \qquad \textbf{(10)}$$

note: $\tan^{-1}$ = arctan.

"Iteratively solved for a value of λ by trial substitution to an accuracy of 0.01.

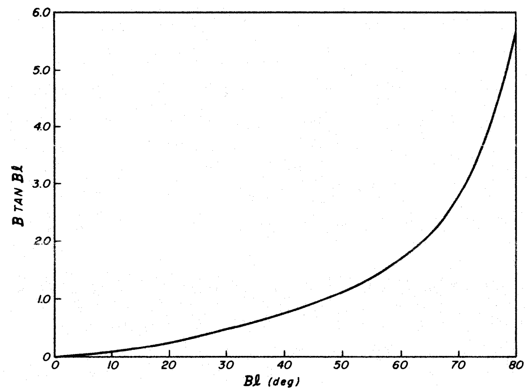

fig. 3. Plot of tangent $\beta \ell$ (degrees) for calculating X_C or Z_0.

Each of four separate calculations is shown with the repeatability for the same values indicated. Note that when F_{MHz} is requested a small error results of about 4 out of 14,440 in frequency (0.03 per cent). This small error results from the convergence solu-

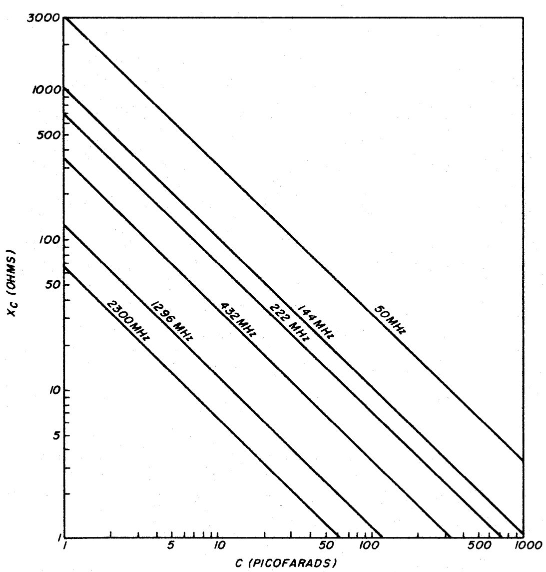

fig. 4. Plot of X_C versus C for the **vhf/uhf** bands.

table 5. HP-67/97 program control for calculating Z_0, C, ℓ, and F.

enter F_{MHz}	press A	(MHz)
enter Z_0	press B	(ohms)
enter C_{pF}	press C	(pF)
enter ℓ	press D	(inches or cm)

Enter any three of the above and calculate the fourth by selecting

1 Z_0
2 ℓ
3 C_{pF}
4 F_{MHz}

and press *fa*

table 6. Sample problems and results.

	1	2	3	4
F_{MHz}	144	144	144	144.03659*†
Z_0	70	70*	70	70
C_{pF}	10	10	10*	10
ℓ inches	13.13522*	13.13522	13.13522	13.13522

* calculated value
† loop run time 3 min. 54 sec.

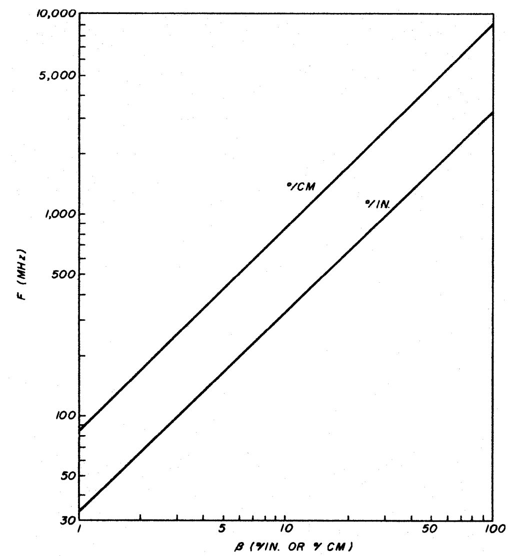

fig. 5. Frequency as a function of β in degrees per inch or degrees per cm.

tion discussed previously but is well within acceptable design limits.

Fig. 3 is a plot of the tangent $\beta \ell$ versus degrees for calculating X_C or Z_0 manually. **Fig.** 4 shows X_C in ohms versus C in picofarads for each of the **vhf/uhf** Amateur bands. **Fig.** 5 shows β in degrees per inch and degrees per cm versus frequency.

Considerations and useful relationships. An important consideration in resonant circuits is efficiency, which is determined from

$$\text{efficiency per cent} = \frac{\text{unloaded } Q - \text{loaded } Q}{\text{unloaded } Q} \times 100 \tag{11}$$

Thus the highest possible unloaded Q is desirable to achieve the best tank-circuit efficiency. Q is generally greater for fully enclosed and shielded transmission-line configurations. Unshielded lines have a low unloaded Q because of radiation losses. Also low values of Z_0 cause less radiation, so a low value of Z_0 should be chosen for unshielded lines. For coaxial geometries the highest values of unloaded Q are realized with Z_0 between 70-85 ohms.

The parameter Q can be estimated by dividing the center resonant frequency of the circuit under consideration by its 3-dB bandwidth (single-section filters or loosely coupled sections). For stripline configurations with dielectric loading, line efficiency is dominated by dielectric constraints (discussed in detail in references 2 and 3).

In most cases it's important to make the transmission lines electrically as long as possible within the $n\ \lambda/4$ constraints. This means that the value of capacitance needed to tune the circuit to resonance should be as small as practicable. This value must include the variability of the active/passive device or devices used. If a desired range of frequency is to be tuned, calculate the required capacitance at the higher frequency then multiply by the square of the frequency ratio; that is:

$$C_{F_{high}} \left(\frac{(F_{high})^2}{(F_{low})^2} \right) = ratio \qquad \textbf{(12)}$$

This equation yields the required capacitance range exclusive of device capacitance. If passive and active devices are used, their minimum capacitances should be subtracted from the value calculated to cover the desired bandwidth with a minimum of capacitance. Nominally values of tuning capacitance for input and output circuits should be designed for ± 25 per cent capacitance tuning range from the center design value. For example, if a single 4CX250B coaxial tank circuit is to tune between 144 MHz-148 MHz with a Z_0 of 77 ohms and a cavity length of 10 inches, what is the capacitance tuning range if the minimum tube-output capacitance is 4.0 pF? From **eq. 1,** the capacitance required is 13.9 pF at 148 MHz. The frequency ratio is:

$$\frac{(148)^2}{(144)^2} = 1.06$$

When multiplied by the resonating capacitance, 13.9 pF yields 14.7 pF. Subtracting the minimum tube capacitance of 4.00 pF yields 10.7 pF as the minimum tuning capacitance required for the design conditions given. The ± 25-per cent rule previously suggested provides a more conservative realization (13.5 pF $\pm$ 25 per cent).

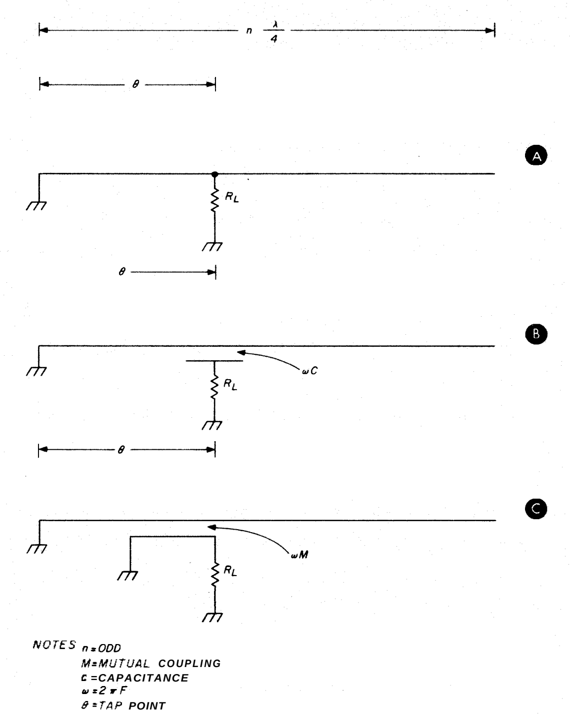

fig. 7. Coupling schemes for a single transmission-line configuration (outer conductors not shown). Direct tap is illustrated in (*A*), probe coupling in (*B*), and loop coupling in (*C*). The tap point may be determined if the other parameters are known (see *eq. 14*).

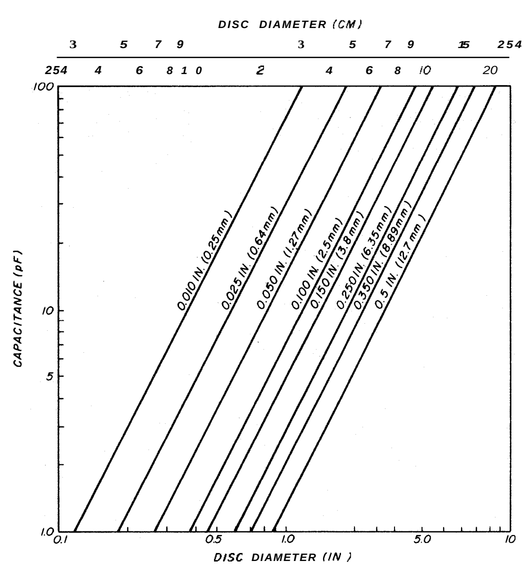

fig. 6. Capacitance as a function of parallel-plate disc diameter with spacing as a parameter.

Tuning capacitor. Construction of the tuning capacitor can be simplified by using parallel plate discs. **Fig. 6** shows capacitance versus diameter for parallel-plate capacitors with various spacings between discs. In some cases a dielectric can be used, so the capacitance value is then multiplied by the dielectric constant. The basic equation is:

$$C_{pF} = 0.225\, \epsilon_r \left[(N-1) \ \frac{A}{T} \right] \qquad \textbf{(13)}$$

where ϵ_r = dielectric constant
N = number of plates
A = area (square inches)
T = spacing (inches)
(For metric dimensions, substitute 0.0885 for 0.225.)

It's important to note that an appropriate spacing between discs must be chosen based on the voltage involved. Very close spacings should be avoided, even under low-voltage conditions, unless excellent parallelism can be achieved. Under high-voltage conditions (greater than 1000 Vdc), no sharp edges are permitted. All edges must be rounded and smooth to prevent build-up of high electrical fields, which can cause flashover.

Coupling into and out of transmission-line circuits can be made using a direct tap, probe coupling, or loop coupling. This idea is shown in fig. **7** for a single transmission line configuration with the outer conductors not shown. Other techniques are possible but are not discussed.

At resonance 8, the tap point, may be determined if the other parameters are known. In fig. **7A,** the equation is:

$$R = \frac{Z_0^2}{R_L} \sin^2 8 \qquad \textbf{(14)}$$

where R = resistance at resonance at 8 on the line (ohms).
Z_0 = transmission-line impedance (ohms).
8 = electrical degrees from ground.
R_L = load resistance (ohms).

If 8 is desired with the other values known,

$$\theta = \sin^{-1}\sqrt{\frac{RR_L}{Z_0^2}} \qquad \textbf{(15)}$$

In fig. **7B** the equation is:

$$R = Z_0^2\omega C^2 R_L \sin^2 8, \qquad \textbf{(16)}$$

which can be solved for C:

$$C_{farads} = \frac{\quad}{Z_0^2\omega^2 R_L \sin^2\theta} \qquad \textbf{(17)}$$

The loop-coupled case (Fig. **7C)** is described by:

$$R = \frac{\omega^2 M^2}{R_L} \cos^2 \$ \qquad \textbf{(18)}$$

where M = **mutual coupling.**

Parameter R is normally equal to R_L when coupling to external transmission lines. A reasonable value for mutual coupling, M, is between 0.5 and 0.8, depending on how tightly the loop is coupled to the line. For the loop, X_L should be tuned out with a variable capacitor in series with the loop to ground if X_L is greater than $R_L/100$. The amount of capacitance required may be calculated from the values of inductance as a function of conductor size given in **table 7.** These values are for straight lengths of conductor but are useful in determining design values for loops. Note that, for equal lengths of conductor with current flowing in opposite directions, the effective inductance is cancelled.

It's clear from **table 7** that the law of diminishing returns prevails when larger-diameter conductors are substituted for smaller sizes to decrease inductance. The inductance is halved from No. 22 to 3116-inch tubing. The only reason for using a larger loop conductor size is for current-handling capability (2 kW PEP output at 144 MHz requires 3116-inch copper tubing).

table 7. Conductor size versus straight pigtail sheet inductance.

conductor	inductance nH/in.	(nH/cm)
114 in. tubing	9.005	(3.55)
3/16 in. tubing	10.446	(4.12)
1/8 in. tubing	12.526	(4.93)
AWG 8 wire	12.374	(4.87)
AWG 10 wire	13.564	(5.34)
AWG 12 wire	14.742	(5.70)
AWG 14 wire	15.92	(6.27)
AWG 16 wire	16.89	(6.65)
AWG 18 wire	18.091	(7.12)
AWG 20 wire	19.23	(7.57)
AWG 22 wire	20.387	(8.03)
AWG 24 wire	21.516	(8.47)
AWG 26 wire	22.691	(8.93)
AWG 28 wire	23.832	(9.38)

A further use of **table** 7 is to calculate lumped resonant circuits at vhfluhf from the length of conductor and a resonating capacitor. It's reasonably accurate and useful if stray capacitances are considered.

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1. H.M. Meyer, Jr., W6GGV, "The Design of VHF Tank Circuits," *73,* November, **1964.**
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3. H.A. Wheeler, "Transmission Line Properties of a Strip on a Dielectric Sheet on a Plane," ***IEEE Transactions on Microwave Theory and Techniques,*** Volume MIT 25, August, **1977,** pages 631-647.

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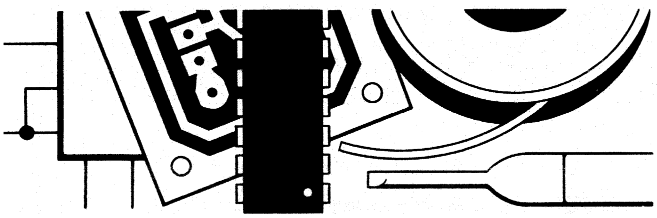
the weekender



simple CW memory

The memory (**fig. 1**) uses standard TTL ICs and a 2102 1k x 1 bit RAM. Information written into the memory can be read out at a faster or slower rate by merely changing the clock rate. A built-in sidetone generator allows monitoring while writing into memory as well as during readout or when keying the transmitter directly. A relay is used to key the transmitter.

Approximately six seconds of recording time occurs at the fastest clock rate, and about 45 seconds occurs at the slowest rate. Some distortion may occur, however, at the slowest setting. Mid setting gives about 30 seconds and should be adequate for most CQs or ID signals.

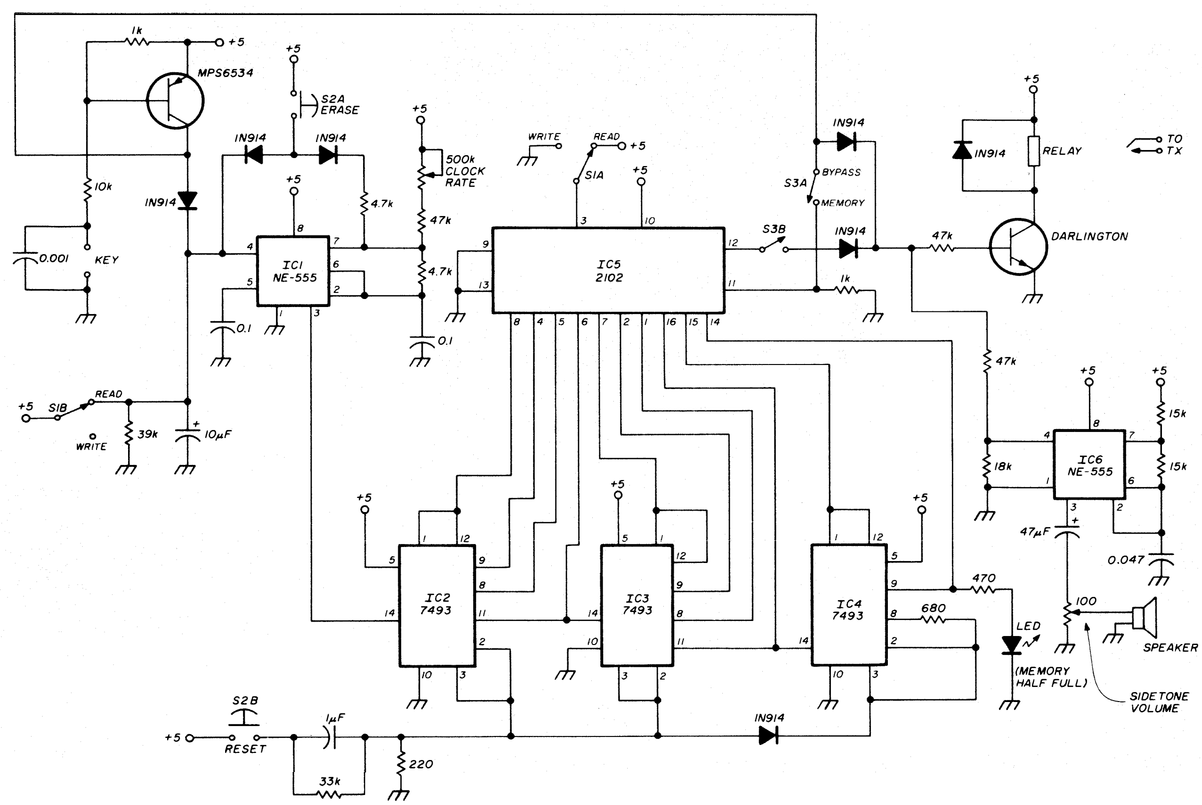


fig. 1. Schematic of the simple CW memory. Circuit may be used with or without a keyer. Although not shown, it's advisable to use bypass caps across supply lines to each IC.

At the risk of overdoing it, I'd like to submit my version of the simple CW memory. I can't lay claim to the design and admit all I did was combine parts that intrigued me the most from several articles.[1,2,3] The circuit can be used with or without a keyer and can be built in a couple of evenings.

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Oscillator. IC1, an **NE-555** timer, is used for the clock oscillator and, in turn, drives the three **7493** 4-bit binary counters. These, in turn, provide the 10-bit address codes to the RAM. IC4 is connected to reset at the count of 3 since the C and D outputs are not needed.

Input power. Power is applied to the clock oscillator only during key-down periods. During this time a charge is accumulated in the 10 μf capacitor connected to IC1 pin **4.** When the key is up, this charge keeps the oscillator running long enough to fill in the normal keying spaces. If you pause or stop momentarily, however, the oscillator shuts down and RAM storage capacity is not wasted.

RAM output. The output data from the RAM drives a transistor that operates the keying relay. The same signal also actuates the sidetone oscillator, another NE-555 timer. An LED indicator shows when the memory is half full. At that point the LED lights and stays lit until the count returns to zero.

Memory. When the unit is first turned on, the memory will store random data and must be cleared. To clear the memory at this time, or when a message is to be rewritten, the **READ/WRITE** switch is set to **WRITE** and the **RESETIERASE** switch flipped to **ERASE.** The switch should be held long enough for **a** complete cycle as shown by the LED. The **ERASE** switch turns on the clock oscillator and also shunts the normal timing resistor with one of much lower value. This action boosts the frequency to a rate that cycles the system rapidly, and the RAM can be cleared in a couple of seconds.

recording a message

Set the **CLOCK RATE** control for the time needed. Switch the **MEMORYIBYPASS** switch to **MEMORY** and **READIWRITE** switch to **WRITE.** Press the **RESET** switch momentarily to return the system to zero. Key in the message. Switch to **READ** and press **RESET.** The message should play back immediately. Message speed may be adjusted by resetting the clock rate. The message will repeat until interrupted. If it's desired to key the transmitter directly, the memory can be bypassed with the **MEMORYIBYPASS** switch, which won't disturb data in storage.

hardware

Laying out a PC board for a one-time project such as this was not considered worth the effort. The parts were assembled and wired on a piece perforated board. The keying relay I used was a reed type with 5-volt coil. Also, the **RESET/ERASE** functions were combined in a spdt momentary toggle switch with center OFF. I used a small sloping panel cabinet to house the device. Although not shown in **fig. 1,** it's advisable to sprinkle some bypass capacitors across supply lines to each IC.

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2. S. Price, G4BWE, "G4BWE CW Memory," Radio Communication, September, **1979.**
3. Eric Unruh, WBØRYN, "Poor Man's CW Memory," 73, June, **1979.**

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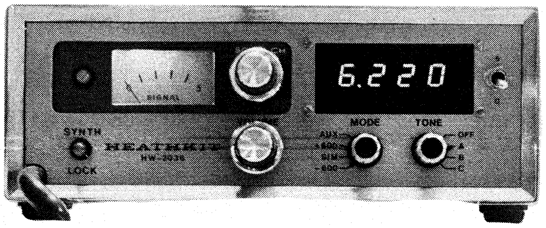


updating the Heathkit HW-2036 for digital readout and scan option

Do thumbwheel switches that can't be seen at night irritate you? Are you frustrated when you know there are more than two repeaters in town and you can't find them? If the answer to these questions is "yes," then hold onto your armchair – you're about to read of a way to upgrade your old HW-2036 to include some of the features of the new Heathkit VF7401 2-meter fm digital scanning transceiver. This project takes the average Amateur two evenings to install and check out, which includes direct replacement of the Micoder™ board and installation of a board to replace the thumbwheel switches. Overall cost of the project shouldn't exceed $55.00 (May, 19801, depending on your junk box.

circuit description

All frequency and Touch-Tone* information is entered through the microphone key pad, therefore I'll start the description with the 2036-MB micboard (fig. 1). Radio Shack National Parts has a 24-conductor microphone cable used on their One-Handler™ CB radio. This cable works just great for passing the required information on to HW-2036 and leaves room for expansion for future projects.

tone generation

Envision the 2036-MB micboard as two separate circuits on one board: one is the tone generator and the other is a BCD frequency generator. Touch-Tone™ generation occurs by depressing a keypad digit through which IC3 produces an audio tone at pins 2 and 15. These tones are coupled to the transmitter audio input. Operating voltage (+5Vdc) for IC3 is available to the chip on transmit only through the PTT switch.

frequency generation

The BDC frequency information is produced by a key depression in receive-mode only, which causes IC2 to generate five outputs. These are BDC information plus one strobe pulse bit. Binary-coded decimal is a means of counting from 0-9, A-F (table 1). Accompanying the BCD information to the 2036-DB (display board) is a strobe pulse, which indicates to the 2036-DB that a key has been depressed on the 2036-MB keypad. Nestled in this 2036-MB circuit is a NE-555, IC1. This chip produces a train of pulses that are proportional to the strobe pulse being present so many microseconds after key depression.

display board

To follow the path of data on the display board (fig. 21, imagine three separate areas: storage (SN74LS298), display (Fairchild 9368), and BCD-to-decimal conversion (IC1,2,3). Keyboard data (BCD A,B,C,D, and strobe) connect to the display board at IC7. The strobe inverts through IC2 then connects to all shift and storage registers (IC4,5,6,7). The storage register produces on its output lines whatever is presented to its input lines at the time of strobe pulse reception. Therefore, by connecting the output of one register to the input of an adjacent register and so on, a digit can be shifted through the registers one at a time with each strobe pulse received.

Surprisingly, the output of these registers is the input to the synthesizer and display drivers. Frequency display is accomplished by feeding the output BCD from IC4,5,6,7 to the 9368 decoder-driver chips. These chips (fig. 3) were used because of ease of adaptation and non-use of dropping resistors on the output lines. Connection is directly to the seven-segment display.

Creation of the 015-kHz signal is as follows. The BCD units digit is fed to IC3, which is a BCD-to-decimal converter. For every BCD digit input, a directly proportional decimal digit is output. I chose to use the digits from zero to four to indicate 0-kHz shift. A logic 1 on any one output of IC3 is inverted through IC2 and fed directly to IC1, a 5-input NOR gate. The output of NOR gate IC1 is connected to the HW-2036 synthesizer board at point X, fig. 2.

Scanning is accomplished by applying a pseudo

*Touch-Tone is a registered trademark of the Bell System.

By Tom French, WA4BZP, Rt. 2, Suburban Shores, Winter Garden, Florida 32787

strobe pulse from IC4 (fig. 4) through the scan-operate switch to display-board, IC2 pin 13. ICs 1, 2, and 3 on the scan board are binary up-down counters and are arranged to facilitate counting up from 000-999. (Q213, located inside the HW-2036 on the receiver board, is tapped for the required signal to stop the scan operation.)

When Q1 (fig. 4) conducts, it forces pins 10 and 7 on IC1 to zero. This action locks the binary counters dead in their tracks. The amount of signal required to lock the scanner is varied by the value of R1, a 90k resistor. The higher its value, the more signal is required to stop the scanner. You might find that, when scanning, your HW-2036 stops 5-10 kHz before the repeater frequency is reached. This problem can be caused by a very strong signal and a low value for R1. My suggestion is to increase the value of R1 in fig. 4 at Q1 base. To re-enable scanning, simply depress the START SCAN button on the microphone.

Remove all knobs and external coverings. Keep the screws in a plastic cup so they won't get lost. The synthesizer lock LED is a tricky item. Be careful when removing the front panel subplate and bezel. Remove and save the thumbwheel switches and mounting screws. You'll need a red or any color plastic lens for the hole covering. Cut it to size with a hacksaw blade and hold the plastic over the window where the display will go. Mark the four new mounting holes and drill them ***slowly*** into the plastic. Trim the edges as closely as possible and install the lens.

I used two 7805 voltage regulators to supply all required power to the display board and microphone board. Two holes must be drilled to accommodate installation of the 7805s. The first hole is beside the speaker just below the synthesizer board and display board (fig. 5).

Now is the tricky part: getting the second hole drilled. Pull the amplifier board a few inches away from the chassis (fig. 6) and drill ***slowly*** into the chassis. At this time it's best to install both 7805 voltage regulators as shown, remembering to use some silicone grease.

Reinstall the amplifier board then remove the microphone cable at both ends. It will be replaced by a 12-to-24-conductor cable. Place some type of insulation over the speaker connections and side magnet, preferably PVC electical tape. This process helps keep the snug-fitting 2036-DB from short circuits. Use 6-32 x 2-1/4 inch (M 3/5 x 57 mm) countersunk mounting screw for the front hole, which will also hold the scan board, and a 6-32 x 1-114 inch (M 3/5 x 31.7 mm) screw for the 2036-DB rear hole.

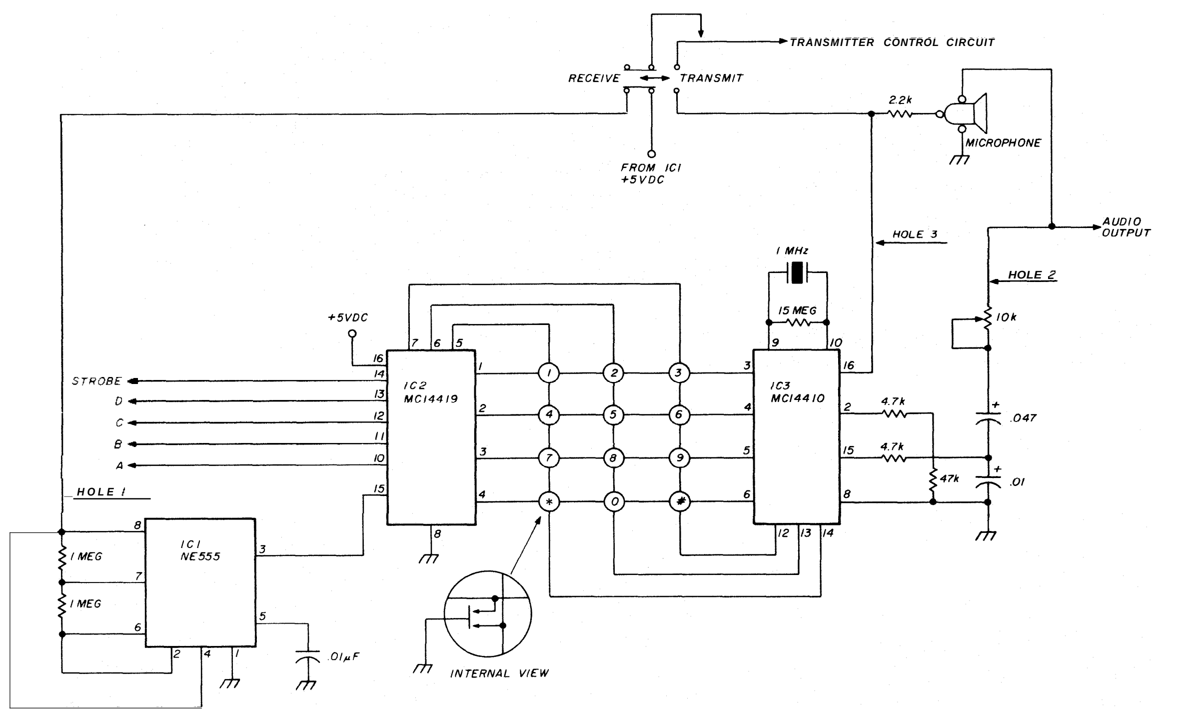


fig. 1. Micboard (2036-MB) schematic. This board replaces the original Heathkit HW-2036 Micoder™ board in your microphone. Circuit consists of a tone generator and a BCD frequency generator.

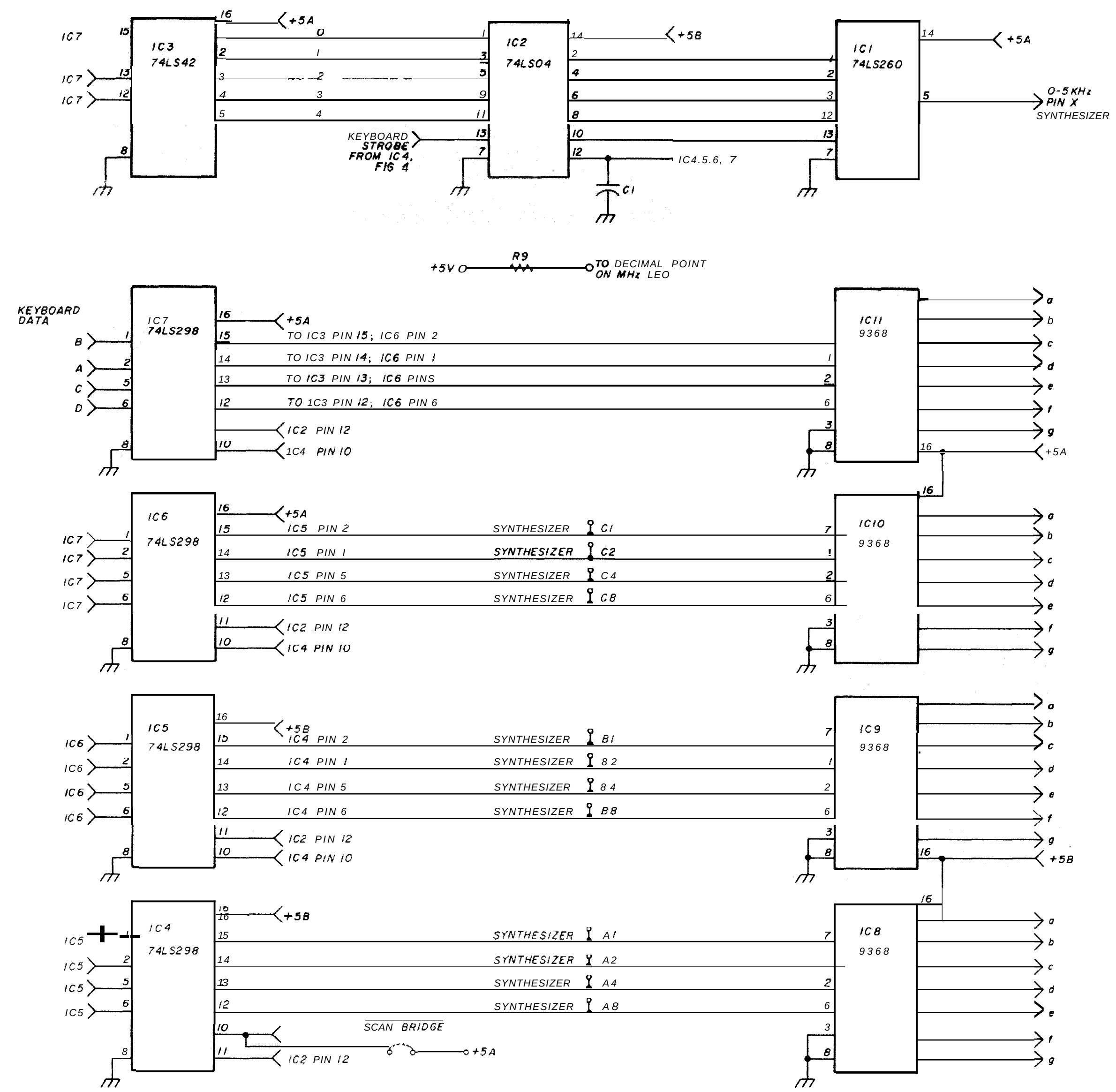


fig. 2. Display board (2036-DB). This board is positioned in the transceiver in the space previously occupied by the thumbwheel switches. The circuit provides storage, display, and BCD-to-decimal conversion.

micboard (2036-MB)

Remove all the parts from your microphone. Leave the microphone element and PTT switch within the plastic housing. Clean the terminal strip of all solder and wire debris. Install the one **2.2k** resistor to one side of the microphone element **(fig. 7)** and the other end to **+5Vdc** on TRANSMIT. Remove all the old keypad pin sockets from your old micboard and reinstall on the 2036-MB.

Install all chips on the component side of the board. Pins **8,8,1** of the **MC14410**, 14419, and NE-555 respectively connect to the ground plane. Note that the .01 μ**F** and .**047** μ**F** caps are electrolytic; observe polarity.

Parts that go on the underside of the board are the 1-MHz crystal and a **.01-**μ**F** disc clock capacitor.

The next step is to cable the microphone, trying to follow some kind of plan on signal-to-color coordination **(fig. 8)**. On the 2036-MB and PTT switch connect a wire to each of the following: +**5** Vdc on

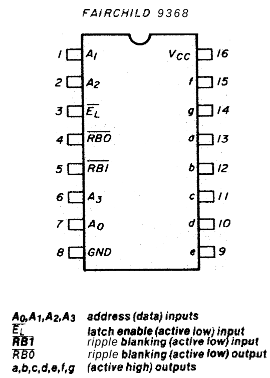


fig. 3. Pin layout of the Fairchild 9368 LED drivers used in the display board.

table 1. Binary-coded decimal format representing the 2036-MB micboard output. Valid outputs are from zero to nine.

	A	B	C	D	
0	0	0	0	0	1 = +5 Vdc
1	1	0	0	0	0 = 0 Vdc
2	0	1	0	0	
3	1	1	0	0	
4	0	0	1	0	
5	1	0	1	0	
6	0	1	1	0	
7	1	1	1	0	
8	0	0	0	1	
9	1	0	0	1	
A	0	1	0	1	
B	1	1	0	1	
C	0	0	1	1	
D	1	0	1	1	
E	0	1	1	1	
F	1	1	1	1	

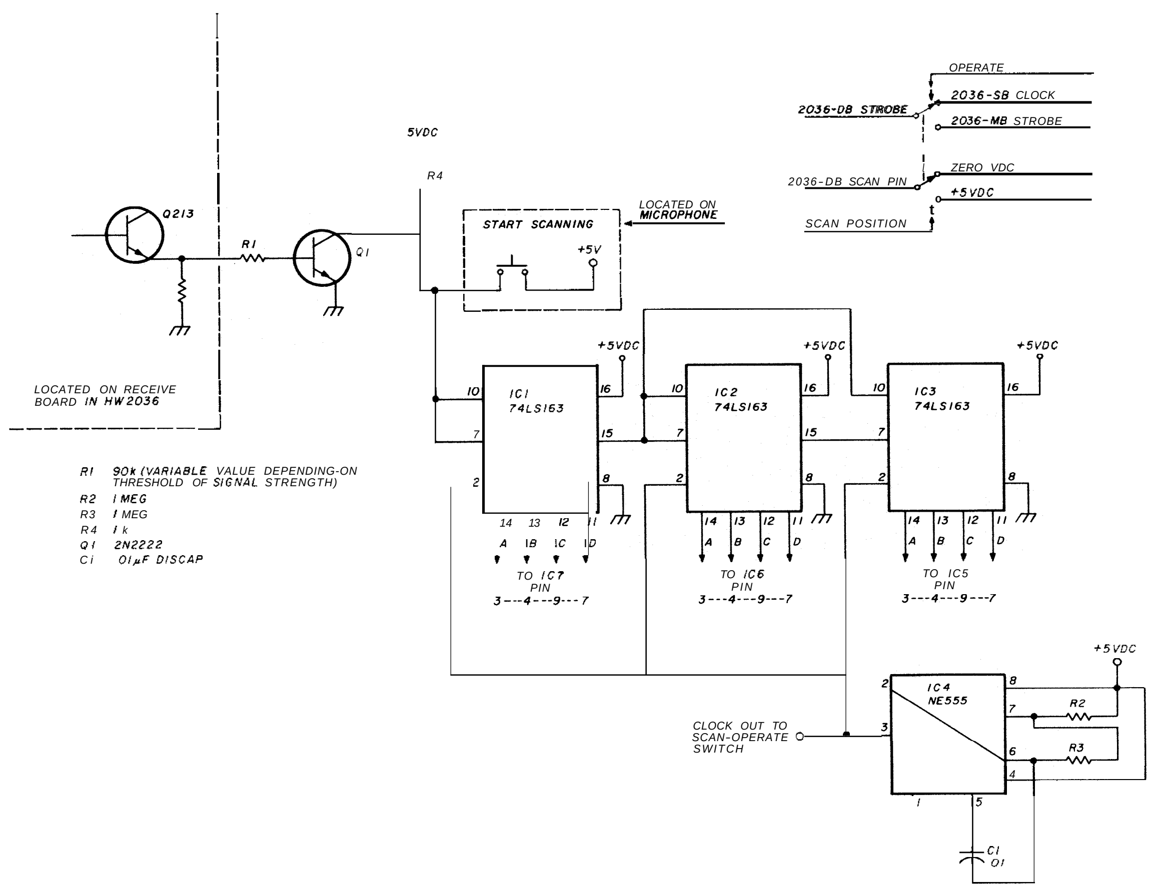


fig. 4. Scan board (2036-SB). This board mounts above the display board for the scanning option (see text).

TRANSMIT, +5 Vdc on RECEIVE, +5 Vdc ground. On the micboard, pin 3 is labeled +5 Vdc on TRANSMIT; pin 1 +5 Vdc on RECEIVE. Pin 2 is the audio output. Power for all circuits in the microphone is obtained from the 7805 (IC1) originally in the rig **(fig. 10).**

display board (2036-DB)

Molex pins or low-profile IC sockets may be used on this board. R9 is a 200-ohm resistor to enable the decimal point on the display. Install it and connect a wire to one end for further connection to the megahertz LED. Capacitor C1 is installed if double or triple digiting happens when the key pad is pressed **(fig. 9)**. Whether or not your SN74LS298 chips match has a direct bearing on the need for C1; its range is between 100 μF-470 pF. To test all of the LEDs after installation, just key in four "eights."

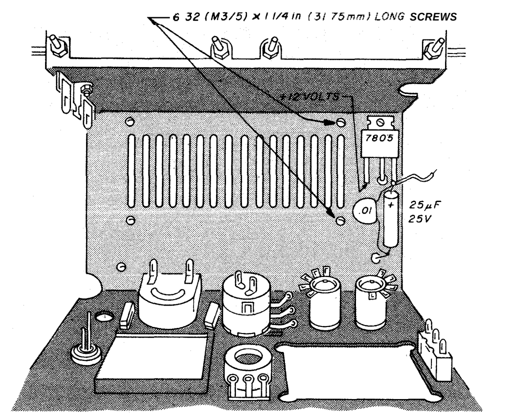


fig. 5. Mounting of components for the voltage regulator. which are installed under the HW-2036 synthesizer board.

HW-2036 synthesizer board

Our next adventure starts on the HW-2036 synthesizer board. Very carefully remove all the pull-up resistors associated with the thumbwheel switches: R401-R409, R411-413. Install a small solder bridge on the 2036-DB, at the scan bridge to pin 10 of IC7 if not using the scan option **(fig. 9)**. A small piece of wire with five leads should be run from the microphone cable to the display board for strobe and BCD data. No shielding is required here. Run the wire beneath

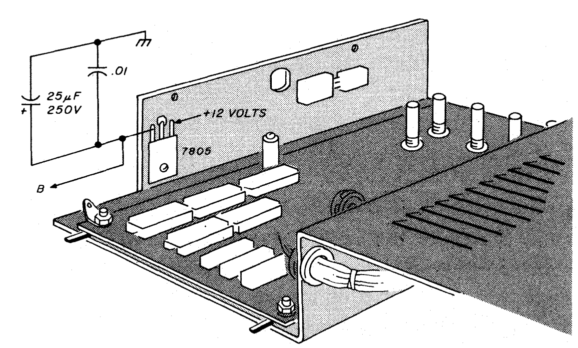


fig. 6. Same as fig. 5, except the regulator is mounted on the back wall of the HW-2036 near the transmitter board.

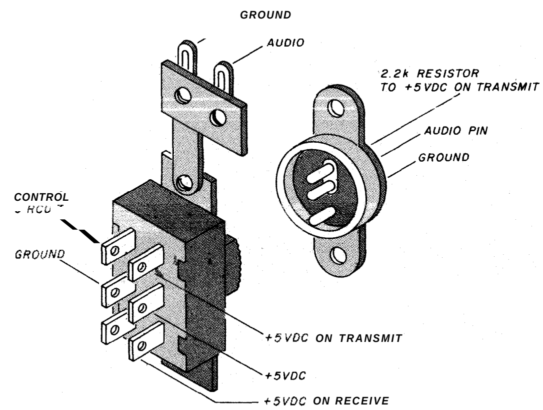


fig. 7. Inside view of the microphone element and PTT switch showing signal connections.

the volume control and mode switches. All common cathode LEDs should be mounted about 1/4 inch (6.5 mm) above the 2036-DB surface. A piece of balsa wood or plastic can be used for this task.

Install the 2036-DB board, and with a pencil mark the window center position. Then super glue the LEDs to the board. Install the 25-μF, 25-Vdc and the

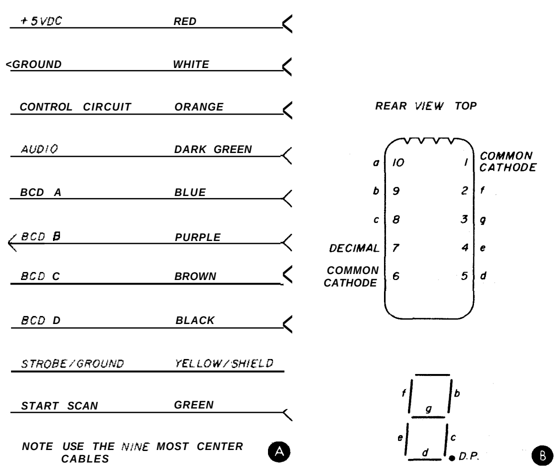
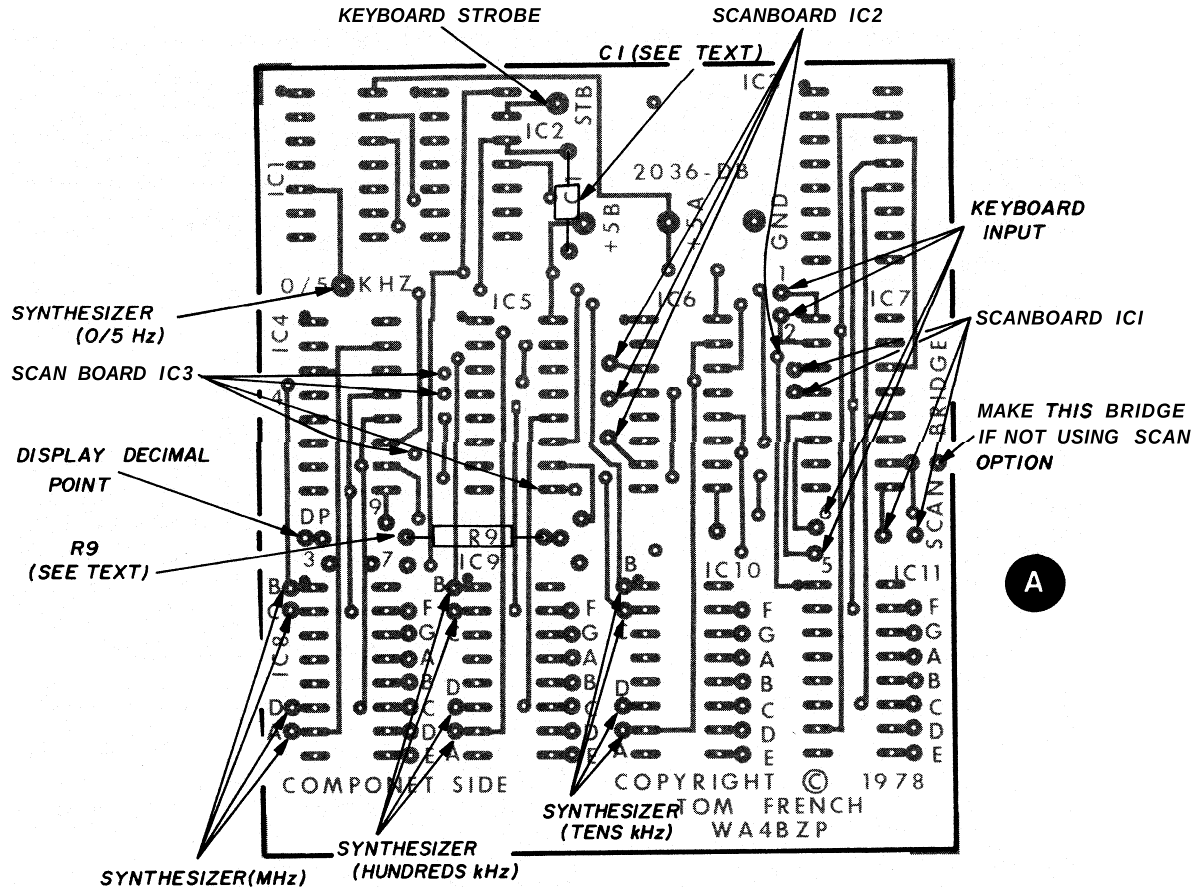


fig. 8. Microphone-cable color code, A, and rear view of the Fairchild 9368 LED driver IC, B.

.01 μF capacitors to both 7805 voltage regulators.

Connect the rear 7805 to the B power pin and the front 7805 to the **A** power pin **(fig. 10)**. All common cathodes of the LEDs can be daisy-chained, then one end can be connected to ground. The wires to the thumbwheel switches should be resoldered to their corresponding drivers at locations A-B-C-D. For IC8, pins 7 and 1 are points A1 and A2 respectively on the



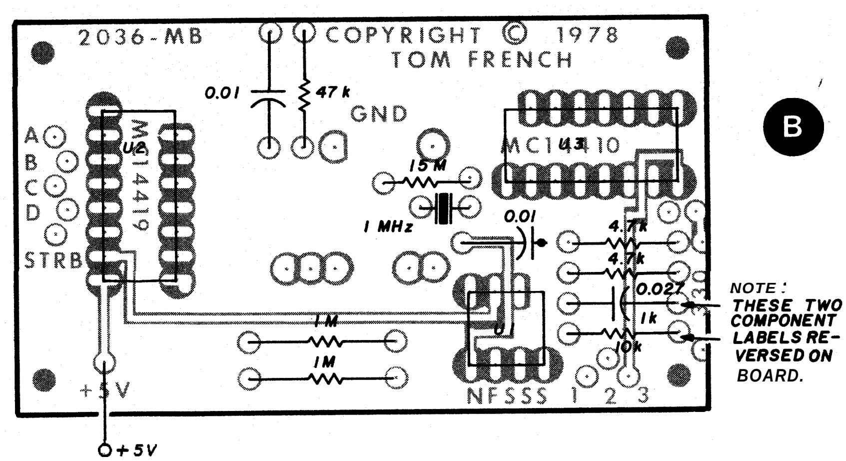


fig. 9. Layout of the display board (2036-DB), A, and the rnicboard (2036-MB) 5, showing corrections. (Labels were reversed on the latter PC board.)

synthesizer board. IC10 displays the tens of kilohertz; IC11 kilohertz (fig. 2).

scan board (2036-SB)

This scan board mounts forward on the 6-32 x 2-1/4-inch (M 3/5 x 57 mm) screw. The scan board is built on a piece of perf board. Install all chips and transistor Q1 with resistor R1. Clock-out from the NE555 goes to the scan-operate switch (0/5 kHz) to provide a strobe pulse to the SN74LS298s. When scanning, if your synthesizer isn't locking onto frequency, the clock frequency of IC4 on the 2036-SB should be slowed down by increasing the values of R1 and R2 to above 1 meg. The leads from IC1-IC3 of the scan board should be connected to the display board using wire wrap.

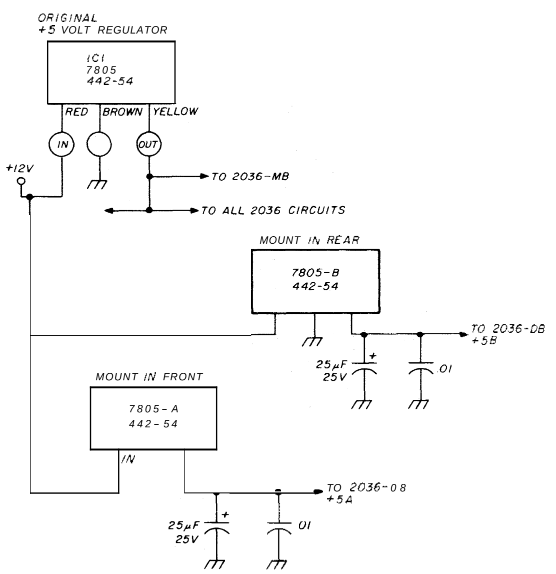


fig. 10. Schematic of the voltage regulators showing 5-volt power distribution.

Wiring the scanloperate switch is next. The 2036-DB strobe is connected to the center pole of one side of this switch. To scan 147.000-147.999 MHz, key in 7-7-7-7, then switch to scan. The switch should be toggled slowly. This scan modification facilitates the location of new repeaters in a new city. By no means is it competitive with professional scanners.

checkout

Complete all connections and reassemble, leaving the skins off. For that matter, you can leave the display board out and to the side of your rig. (No need to hook up the synthesizer wires until the 2036-DB is operational.) Assuming you've done all the above correctly, we're ready to power up. The display will read some random numbers and sometimes even letters.

On the 2036-MB check the NE-555 pin 8 and MC14419 pin 16 for +5 Vdc. If it's present, we'll assume pin 3 of the NE-555 is generating a clock pulse. There are two key strokes considered by the keypad to be invalid in the receive mode. They are # and *. This being true, they will not generate a bit pattern upon key depression, only in receive mode.

Depress some digits. Your display should follow from right to left; if not, let's troubleshoot it. When you depress a valid key, the strobe pulse should appear on the output of MC14419 pin 14. A scope or logic probe is required to view this signal, as it is very short in duration. As long as you hold your finger on the keypad digit, the output loads (A-B-C-D) of MC14419 will represent the desired digit. If not, check all voltages in your micboard and repair. The keypad data and strobe are sent to IC7 on the 2036-DB. At IC7, which is the kHz digit, the BCD data is sent to decimal converter IC3. It will generate a logic one if the BCD input is between zero and four and a logic zero if it is between five and nine. The signal is inverted at IC2 then sent to a five-input NOR gate for the final 0/5 kHz output. A logic zero is equal to zero kHz, and a logic one equals 5-kHz shift up.

On displaying your operating frequency, suppose you desire 147.345 MHz. Simply touch in the digit sequence 7-3-4-5 on the keypad. Upon depressing the push-to-talk switch, the keypad is now a Touch-Tone™ pad and will generate the standard phone tones.

concluding remarks

The mic-Touch-Tone™ kit is available from Data Signal Incorporated labeled EK-2036. It includes a 1-MHz crystal, MC14410 Touch-Tone™ encoder, and various capacitors and components. This kit is required only if your Micoder™ doesn't have the Motorola Touch-Tone™ encoder. Heathkit used the Mostek chip in some versions of the Micoder 2™, so check it carefully

Circuit-board artwork is included for the daring individuals who are familiar with double-sided etching. For those who are not so well off, the author supplies one 2036-MB and one 2036-DB with instructions for $19.95. The 2036-SB (scan board) must be built on perf board, as designing a board was not feasible at the time of this writing.

bibliography

Stephens, Bill, WB8TJL, "Outboard LED Frequency Display for the HW-2036," hem *radio*, July, 1978, page 50.

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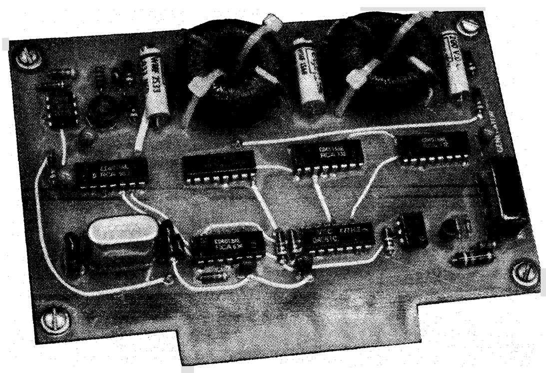

the XK2C AFSK generator

Introducing a CMOS version of the Mainline XK2 — a spinoff design ffering low power consumption

The **AFSK** generator described in this article is a low-power CMOS version of the Mainline XK2.1 The XK2, using TTL devices, required approximately 150 mA at 5 Vdc. The XK2C will operate from a 10-15 Vdc supply; at 12 Vdc it draws only 8 mA, including the additional audio driver stage (fig. 1).

device compatibility

The basic logic of the XK2C is the same as that of its predecessor. However, as Murphy's law would indicate, things are not as simple as direct substitution of CMOS ICs for TTL devices. CMOS operation is slower than that of TTL (about one-quarter as fast at 5 Vdc).

propagation delay

Propagation delay through the programmable divider logic can easily exceed the time for one cycle of the clock, thereby skipping clock pulses and yielding an erratic output frequency. Three steps were taken to get around this problem. First, the clock frequency was reduced to 1606.5 kHz (one-half the TTL clock) and the final divide-by-two stage was eliminated. This provided the same output frequencies while allowing the programmable divider to function at one-half the rate. Second, the number of logic gates for the divider preset was held to a minimum. The preset input polarity for the CD4516 (fig. 2) is opposite to that required for the original 74193s, so an AND gate was required for proper preset. Third, the input voltage was raised to 10 Vdc minimum because CMOS operates faster at higher supply voltages.

power-supply voltages

Operation from a 5-Vdc supply may well be possible, depending on the propagation delays of the chips used and the temperature range over which they will be operated. If 5 Vdc operation is desired, the output frequency should be closely checked with a frequency counter over the desired operating temperature range. The output frequencies should be stable at 2125 Hz, 2295 Hz and 2975 Hz. Propaga-

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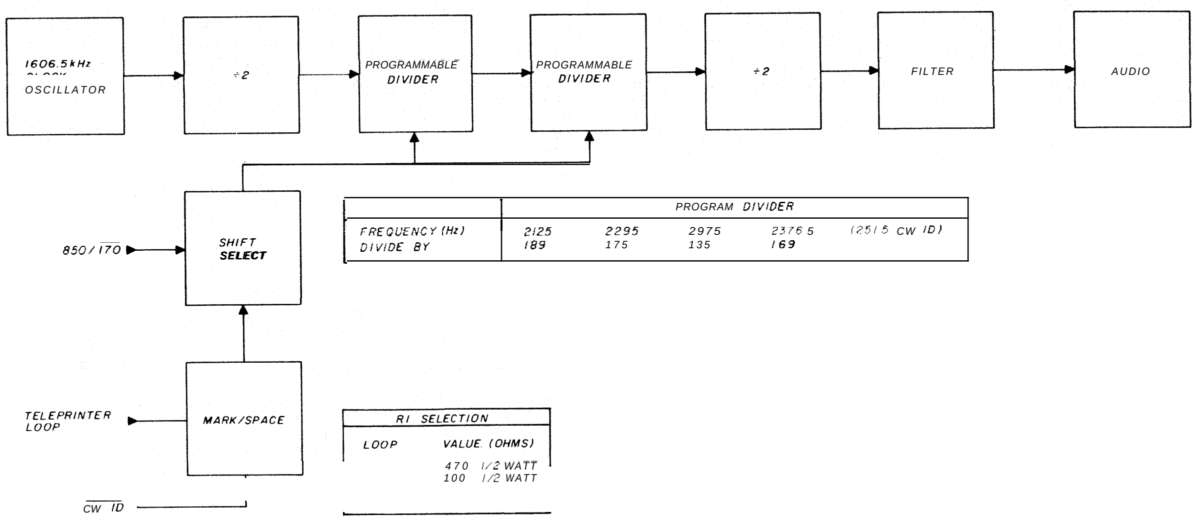
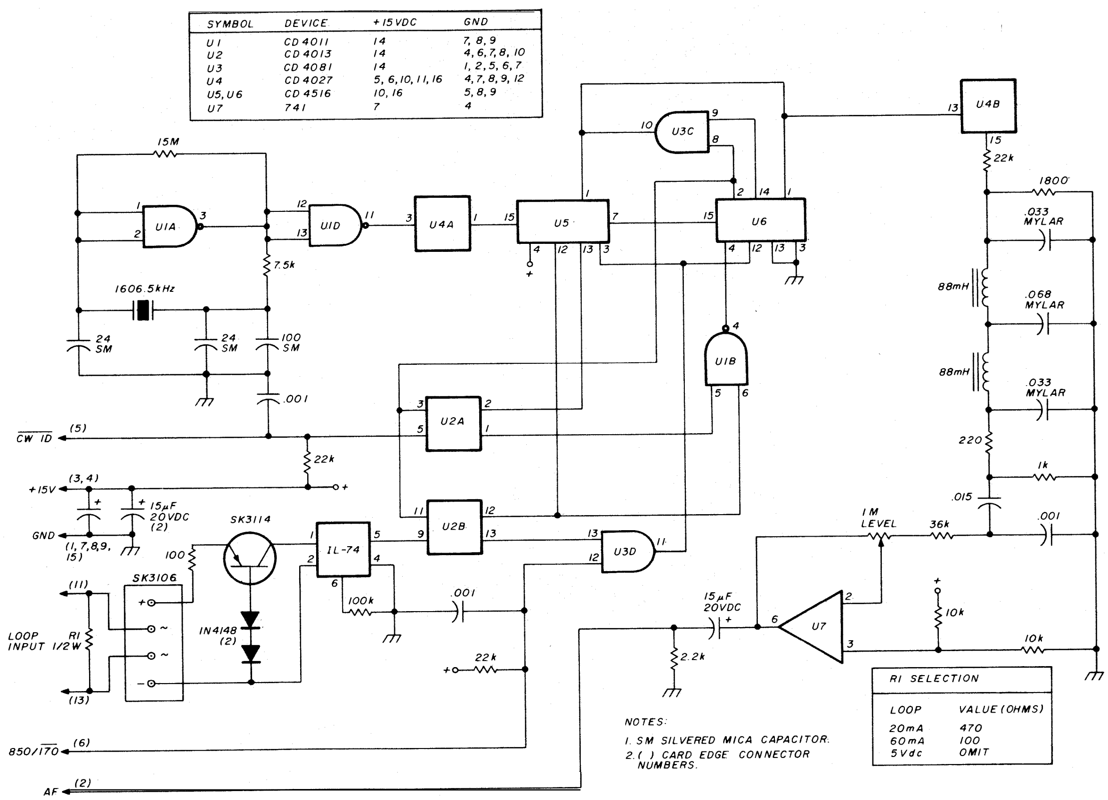


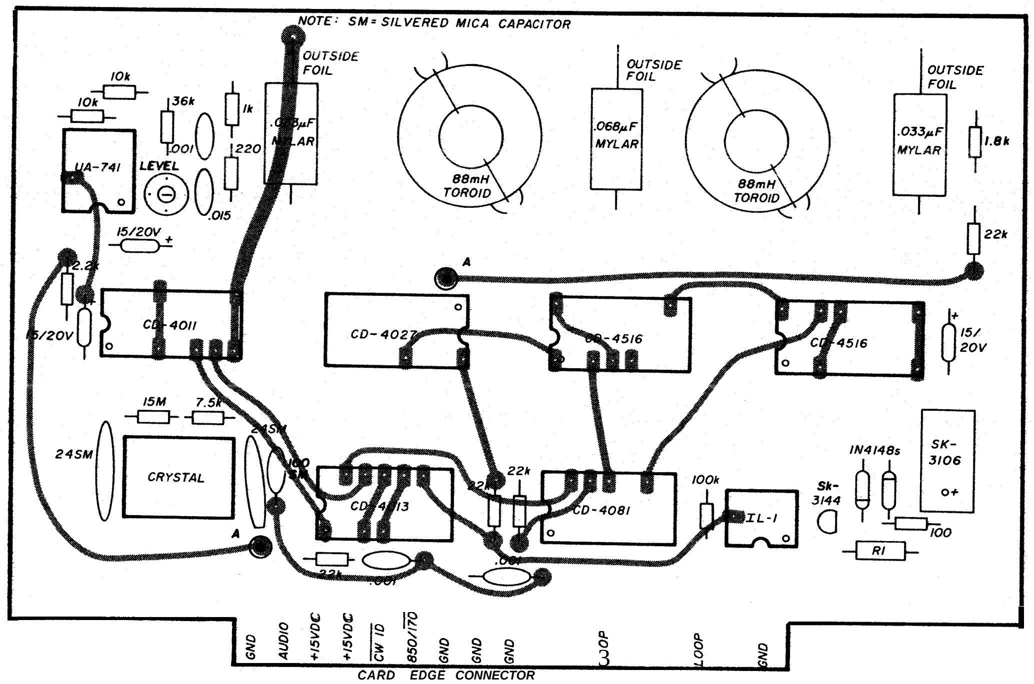
fig. 1. Block diagram of the XK2C crystal Mainline AFSK generator. Power consumption is low; at 12 Vdc circuit draws only 8 mA, even with the added audio driver stage.



1	100 ohm, 1/4 watt resistor
1	220 ohm, lid-waftresirlor
1	1K, 1/4-watt resistor
1	1.8K, 1/4 watt resistor
1	2.2K, 1/4-watt resistor
1	7.5K, 1/4-watt, 5 per cent resistor
2	10K, 1/4-watt, 5 per cent resistor
3	22K, 1/4-watt resistor
1	36K, 1/4-watt, 5 per cent resistor
1	IWK, 1/4-watt resistor
1	15-meg, 1/4-watt, 5 per cent resistor
1	1-meg pot., Spectrol 62-1-1 (12F4219)
1	100-ohm or 470-ohm, 1/2-watt resistor (R1)
2	24pF silvered mica, ARCO DMSCC240J (14F1276)
1	100pF silvered mica, ARCO DM5FC101J (14F1292)
2	.033 μF mylar, Cornell-Dubilier WMF1D33 (15F1453)
1	.068 μF mylar, Cornell-Dubilier WMF1D68 (15F1455)
1	.015 μF ceramic, Cornell-Dubilier DMT1S15 (14F2192)
3	.001 μF ceramic, Cornell-Dubilier DMT1D1 (15F2185)
3	IS μF/20 Vdc tantalum, Sprague 196D156X9020KA1 (47F387)
1	RCA SK 3114 PNP transistor
1	RCA SK-3106 bridge rectifier, 400 PRY, I Amp
2	1N4148 silicon switching diode
1	RCA CD4011, quad 2-input NAND
1	RCA CD4013, dual-D flip-flop
1	RCA CD4081, quad 2-input AND
1	RCA CD4027, dual JK flip-flop
2	RCA CD4516, presettable 4-Bit binary up/down counter
1	UA741 Op Amp
1	Litronix IL 74 or Monsanto MCT271 opto isolator (32F1493)
2	88 mH surplus telephone toroid
1	Crystal, 1606.5 kHz, Tolerance .01 per cent at room temperature. Parallel resonant at 32 pF. International type F-700 Holder. Available from International Crystal, 10 North Lee, Oklahoma City, Oklahoma 73102. Phone: 405-236-3741.

NOTES.
1. All items except crystal and toroids are available from Newark Electronics, 500 N. Pulaski Road, Chicago, Illinois 60624. Numbers in parentheses are Newark stock numbers from Catalog 104
2. All resistors are 10 per cent unless otherwise noted
3. All capacitors are rated at 20 Vdc or more.

fig. 2. Schematic of the XK2C AFSK generator. Operation from a 5-Vdc supply may be possible, depending on several factors (see text).



Note: A = thru-board
Jumper wire soldered to toil on both sides of PC board.
All components must be soldered to foil on both sides of PC board.

fig. 3. Component layout for the Mainline XK2C AFSK generator (component-side view).

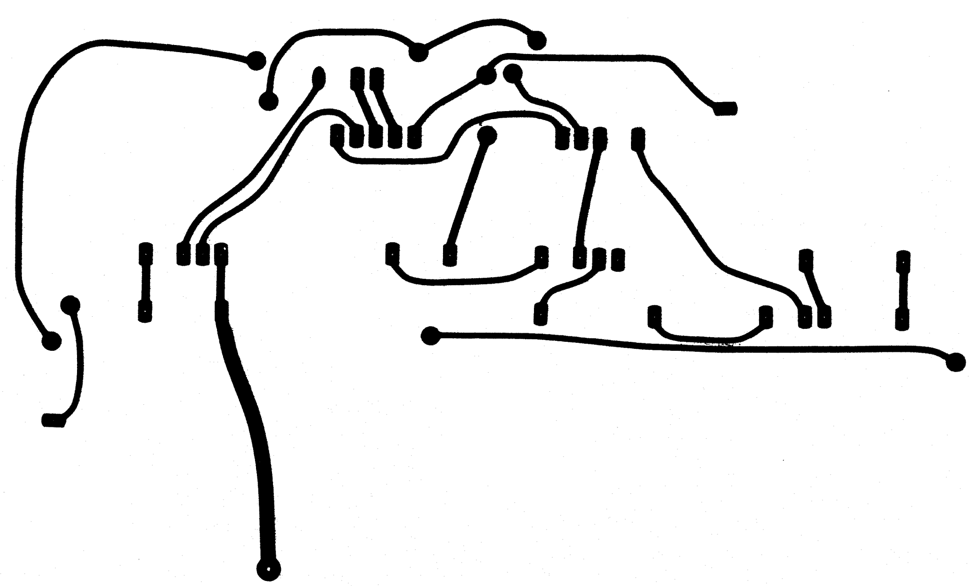


fig. 4. Component side of PC board.

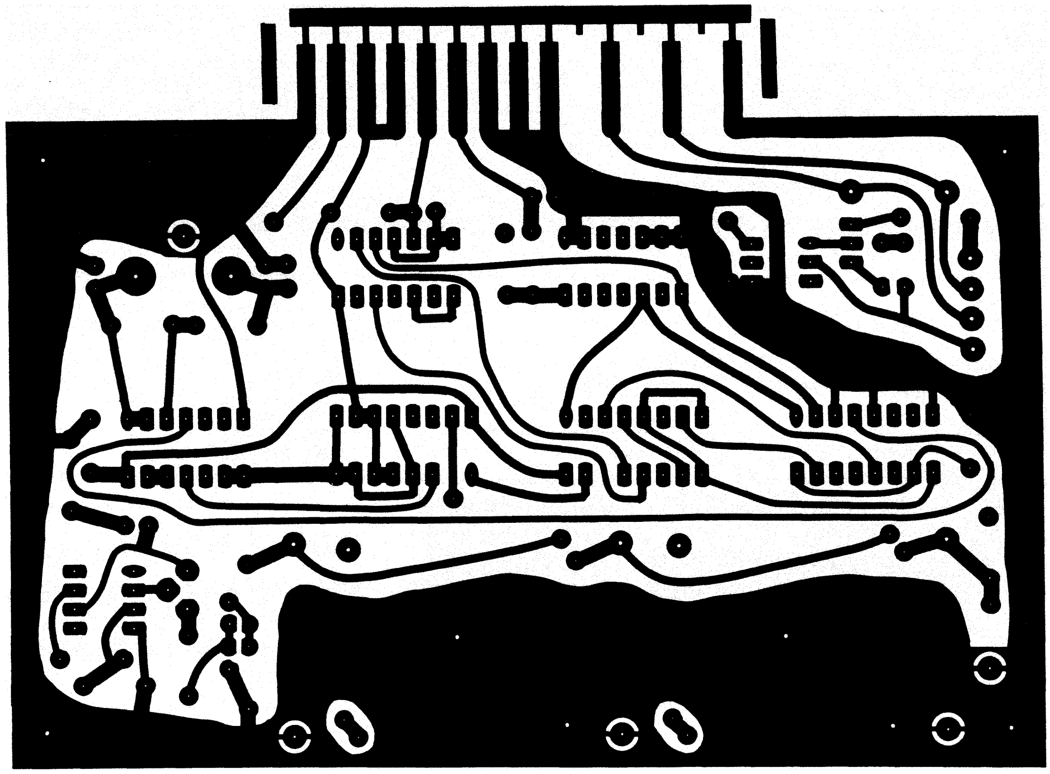


fig. 5. Non-component side of PC board.

tion delays may cause one clock pulse to skip for each preset of the programmable divider, thereby yielding outputs of 2114 Hz, 2282 Hz, and 2953 Hz respectively. At 10-Vdc or more, worst-case propagation delays won't be great enough to cause difficulties.

Elimination of the final divide-by-two stage left one clocked D latch (one-half of a CD-4013) unused. I decided to use this latch to clock in the CW ID data, thereby making it coherent, as with the TTY data. This doesn't really buy anything, but it was available and eliminates the need for one inverter, since both Q and $\overline{Q}$ outputs are available from the latch.

Shift-selection logic was added so that the 170-Hz or 850-Hz selection could be made with a spst switch instead of the spdt switch required by the XK2. I added an audio driver stage (UA741 op amp) to provide plenty of output signal and a good, stiff source capable of driving low- as well as high-impedance transmitter audio inputs. I added a loop-to-logic converter (SK3106, SK3114, and IL74) for direct connection into the teleprinter loop. This input is isolated from ground and will operate with either polarity current.

The value of *R1* was selected from fig. 1 for keying from either a 20-mA or 60-mA loop; or *R1* may be deleted altogether for keying directly across a 5-Vdc supply. High-voltage loops (100-200 Vdc), with selector magnets directly in line, often generate spikes and rf hash, which could get into the XK2C logic. The PC board has been designed so that the loop-to-logic converter can be easily shielded if required. The PC board plugs into a Cinch type 50-15A-20, 15-pin receptacle.

assembly

The XK2C was assembled on a double-sided, plug-in PC board (figs. 4 and 5). Since it was not possible for me to make plated-through holes, it was necessary to solder the components to the foil on both sides of the board. Care must be taken not to overheat the ICs during soldering.

The XK2C has all the features of the earlier XK2 plus a few extra, and it offers a considerable savings in power consumption.

reference

1. Irving Hoff, W6FCC, "The Mainline XK2 Crystal AFSK," *RTTY* Journal, July-August, 1976, page 3.

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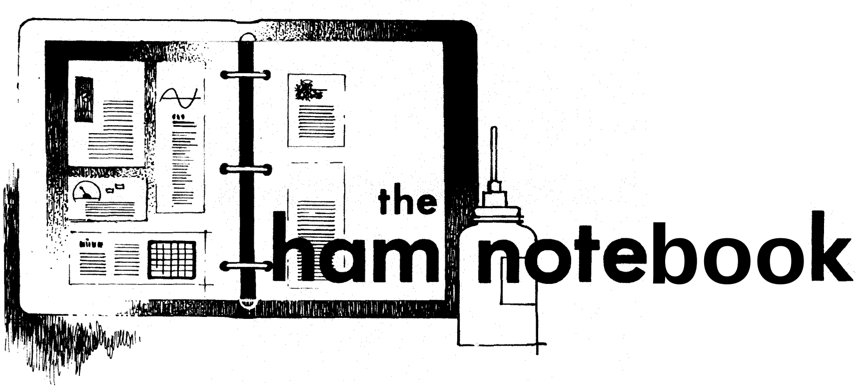

the T coupler

Here's a handy little gadget for your shack or shop that I've found to be as useful as the zip top on a Bud.

If you've ever had a need for a convenient transmitter-to-counter coupler, low power dummy load, matching network for a signal generator, or a little device to help measure repeater desense, this might be just what the doctor ordered. Basically, it's a **50-ohm,** 2-watt dummy load and capacitance coupler made from three standard uhf connectors: a barrel (PL-258), a tee (M-358), and a plug (PL-259).

Construction is simple. Insert a 50-ohm, 2-watt resistor into the back of a PL-259 **(fig. 1)**. Solder and trim the pin end of the connector. Trim the resistor lead at the back end of the connector and fill with solder to prevent any rf leakage. This will serve as your conventional 2-watt dummy load. Not too tricky so far.

Now, modify the T connector as follows. Unscrew the pin from the center section T and replace it with a flat-head screw. Over the head of the screw, place two or three pieces of insulation mica or plastic. Insert another flat-head screw into one end of the barrel connector. Now screw all three connectors together and check continuity to ensure proper insulation.

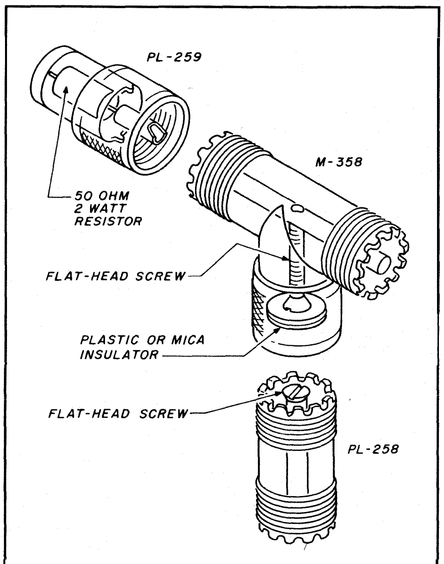

fig. 1. Construction of the T coupler using standard UHF connectors.

When tuning a low-power transmitter, a counter can now be coupled through the barrel. For higher-power transmitters, replace the 2-watt dummy load with a larger one.

With a receiver coupled to the **50-**ohm match, a signal generator can be coupled by means of the barrel connector.

To check repeater desense, connect the duplexer output to a suitable dummy load through the T and connect a calibrated generator through the barrel connector. Measure the signal difference with the transmitter on and at idle.

Now that you've probably thought of at least 27 other uses for this little gem, there's no excuse for not adding it to your test bench.

John LaMartina, K3NXU

improved ground-mounted vertical for the lower bands

When using ground-mounted verticals a good ground system is essential for best results. In the case of a 114-wave or shorter vertical, the largest current in the ground system is near the base. Many radials will result in a small amount of power loss. However, it should be possible to improve on the average ground system by moving the high current portion of the ground into a metal conductor. **A** coaxial vertical antenna is the basis for this idea.

The upper 1/4 wave could be shortened by top loading. The lower 114-wave sleeve also could be cut to a more convenient length, with the feed line passing through the sleeve, as usual. The ground radials would be connected to the bottom of the sleeve at ground level.

Increasing the height above ground of the high-current portion, and allowing current to flow into a low-loss conductor out of the ground, should result in some degree of improvement. Of course, a full-size coaxial vertical would be nice — it wouldn't need any ground radials at all. Quite impressive, too, at 260 feet (79 meters) it would direct passing hams to your location from miles away.

E.R. Larnprecht, W5NPD

modification of Ham-M rotator control box

Early models of the popular Ham-M rotator have one very undesirable characteristic. When power is removed from the rotator motor, power is simultaneously removed from the brake solenoid, causing the brake to slam into the rotator housing. This brings the moving antenna to an abrupt halt, thereby applying severe torsional strain to mast, rotator and tower.

I redesigned the switch in the control unit to change the make-break contacts so the antennas would come to a halt before the brake was applied. I had no way to manufacture a substitute switch, so I sent a drawing of the switch to the manufacturer and suggested the improvement. They said they were not interested! I had no intention of installing a torsion bar (per the manual) on my tower when there surely must be a better way.

Simple wiring changes in the control box of Series-3 units will provide independent brake control with no additional parts or switches and no drilling. My Ham-M is a Series 1, in which I modified the control unit to a Series 3 configuration per the simple instructions in the owner's manual, which came with the unit. Therefore, Series 1 and Series 2 units should be modified to Series 3 before the changes are made.

When the following changes have been made, moving the control lever slightly to right or left will cause the meter to indicate antenna position and will simultaneously release the brake. Moving the lever full right or left will start rotation. When the antenna has reached the desired heading, moving the lever back to first position will allow the antenna to come to a gentle stop. Returning the lever to center position then applies the brake.

I put a piece of masking tape just above the screw terminals on the back of the control box and marked them 1 Blk, 2 Red, 3 Blu, and so on. It is also a good idea to mark out the Series 1 on the control box back and change it to Series 3 for reference, if, indeed, you're modifying one of the earlier models.

One final note: In modifying my unit, I used parts of three schematics to come up with the desired result. I decided to write out the steps required and work from that, rather than pick off each step from a drawing. It worked beautifully for me and I'm sure it will for you.

mod steps

Viewing the control box switch from the top, contact 1 is the first contact on lower left; other contacts progress clockwise. Proceed as follows.

1. Remove eight wires from rear terminal strip. They will be returned to their original position when wiring is completed. Remove four rubber mounting feet. Lift off plastic cabinet. Remove four screws that hold meter assembly to base plate. Move meter assembly outward to provide access to control switch. It may be necessary to remove the power-transformer mounting screws to provide access to the inside of rear terminal strip. In the following wiring changes, when a connection is made, it should be soldered unless another wire is to be connected to that point later, in which case the instructions will say "do not solder."

2. Disconnect wire from SW contact 1. Leave it connected to 5 on rear terminal strip.

3. Remove jumper that is connected between SW contacts 4 and 8.

4. Remove from SW contact 4 the wire that goes to the primary of the *instrument* transformer.

5. Remove wire that connects SW contact 2 to 2 on rear terminal strip.

6. Remove wire from SW contact 3. Leave other end connected to 6 on rear terminal strip.

7. Reroute this wire from terminal 6 and solder to SW contact **8.**

8. Remove the bottom wire from the primary winding of the *power* transformer.

9. Connect the wire just removed from the power transformer to 2 on the rear terminal strip. This now connects SW contact 6 to rear terminal strip 2.

10. Remove the wire from SW contact **4.** (This is one lead of the primary of the *instrument* transformer.)

11. Connect the wire just removed to SW contact 2. Do not solder.

12. Connect a wire from the bottom terminal on the *power* transformer to SW contact 2. Solder two.

13. Install a jumper wire between SW contacts 1 and 3. Do not solder 3.

14. Remove wire that connects SW contact 7 to 3-amp fuse holder on instrument side of fuse.

15. Connect a wire from 3-amp fuse holder on instrument side of fuse to SW contact 3. Solder two.

16. Connect the wire attached to 5 on the rear terminal strip to SW contact 4. In this modification switch contacts 5 and 7 are not used.

This completes the wiring. It might be a good idea to check over the instructions before starting the modification, once the unit is removed from the cabinet. In this way it will become apparent as to just what's happening and why the brake operation will be independent of rotation.

After attaching the eight wires to the rear terminal strip, check out, with 120-Vac connected, should read approximately 30 Vac across terminals 1 and 2 when the switch is operated in either direction. A reading of 31 Vdc across terminals 3 and 7 with the switch operated is normal.

I modified my rotor control about three years ago and it has certainly been a source of pleasure to know that my tower, beams, and rotator are no longer subjected to the severe (and totally unnecessary) torsional forces.

William G. Blankenship, Jr., K4DLA/W1RDR

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contents

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Observation & Opinion

It seems that a West Coast Amateur has decided to make some easy money by publishing material to aid prospective licensees in passing FCC Amateur examinations. His material is crafted so that mere memorization of answers to FCC exam questions practically guarantees a passing grade. His product apparently is derived from FCC exam materials. Such material is gleaned by a well-organized effort to collect questions verbatim from the various exams when they are administered by FCC representatives. Very often this has happened at Radio Amateur conclaves and conventions. We at ham radio magazine deplore such tactics. Amateur Radio has flourished because of its many established traditions. "In today, out tomorrow" publications, such as that referred to above, defeat the entire purpose of the Amateur Radio tradition, which has made our hobby one of the greatest in the nation for over **60** years.

Where do these questions and answers come from? From Radio Amateurs. The publisher in question solicits FCC test questions from those who have recently taken the exam, then publishes these questions along with the proper answers. Pretty neat. All one has to do is memorize the questions and answers, and the exam is a comparative cinch.

The publisher probably is making lots of money publishing the exam questions and answers without apparent legal sanctions (at least to date). But what about the long-range impact on the Amateur Radio Service and U.S. taxpayers at large? We lose.

An interesting sidelight is that the publisher justifies his action in the interest of "socially motivated" hams. His rationale for this rather obtuse reasoning is Part 97.1 (a) of the FCC rules and regulations, Basis and Purpose: "Recognition and enhancement of the value of the amateur service to the public as a voluntary noncommercial communication service, particularly with respect to providing emergency communications." (Italics mine.)

The publisher, however, conveniently overlooks Part 97.1 (b), which states: "Continuation and extension of the amateur's proven ability to contribute to the advancement of the radio art." (Italics mine.)

How can anyone in the Amateur Service comply with regulation 97.1 (b) if a license is obtained by memorizing answers to FCC questions? It is the purpose of this magazine to encourage Amateurs, by publishing articles on current technology, to "contribute to the advancement of the radio art." We believe that, for the most part, Amateurs who obtain their license using only the memorization technique are rarely in a position to contribute to part 97.1 (b) on a technical basis. There are exceptions, of course, but the method of preparing for exams to which we object seems to augur an increasingly less proficient operator in the midst of a rapidly increasing technical operating environment.

What can we Amateurs do to promote the technical integrity of Amateur Radio? Let's learn as much electronic theory as possible before taking the examination. It requires some effort, true, but when we pass the FCC exams based on knowledge rather than memorization we achieve a more significant accomplishment. After all, that's what ham radio is all about. Consider part three of "The Amateur's Code" by Paul Segal: "The Amateur is Progressive . . . He keeps his station abreast of science. It is well-built and efficient. His operating practice is above reproach."

ham radio continues to endorse this philosophy. The Amateur Radio Service cannot survive if licenses are obtained without due regard to technical knowledge: that is, passing FCC exams by learning the questions and answers by rote.

All prospective Amateurs should take a closer look at this problem. We licensed Amateurs who organize training classes and other tutorial endeavors have a special responsibility in this regard. Obtaining an Amateur license requires some effort. It is usually a difficult, time-consuming process. The successful license applicant will find the process rewarding for years to come.

What can the FCC do at this point to promote the technical integrity of Amateur Radio? We have some ideas, but we would like to hear from our readers on this point. Should the FCC look the other way while the abuse of Amateur exams continues? Should the FCC adopt an Amateur exam question series broadly similar to the FAA's several-hundred-question series for the Private Pilot license? More basically, why should newly updated exams be negated by one of us at the expense of us all? Consider this issue carefully, then discuss it among your Amateur Radio associates. Your views on the subject will be welcome at ham radio.

Alf Wilson, W6NIF
Editor

comments

RST feedback

Dear HR:

I read your comments on DL7DO's letter in "Observations and Comments," September, 1980, with some interest and a bit of confusion.

When I was running a '45 with 135 (not 90) volts on the plate, a signal report of S7 would have been somewhat meaningless: it did not gain significance until adoption of the RST system in the late thirties. The proper report prior to that would have been QSA (1-5), R (1-9). At the time of the adoption of the RST system most had converted to non-chirpy crystal control, and a-c on the plate supply brought an immediate citation from the newly formed FCC.

There is a definite need for accurate signal reporting, but if a report on tone is no longer needed (I for one disagree strongly with this reasoning), then let us not go the route of "inventing" a new system when the need is clearly covered in the international Q signals.

My personal feeling is that the RST system is performing admirably, with the exception of some contesters, and a change of the system would not change that. In other words, if it ain't broke, don't fix it!

Rue O'Neill, WØNN
St. Louis, Missouri

Dear HR:

I applaud the idea of junking the RST signal reporting system. But do we really need a new system? Why not simply make use of the existing QSA system which (with "copy" notes added) is as follows:

QSA 1 Scarcely perceptible — no copy
2 Weak — very little copy
3 Fairly good — partial copy
4 Good — almost full copy
5 Very good — full copy

Reports would simply be Q1, 2, 3, 4, or **5.** Where the situation permits, an operator should do the other station the favor of reporting technical signal defects such as distortion, overdriving, VOX clipping, key clicks, poor tone, etc.

The difference between a signal received off the end of a dipole and the same signal received by a properly oriented high-gain beam is tremendous. The signal strength measured in the receiver depends almost entirely upon the character and orientation of the receiving antenna. A signal reported as S5 by a station with a mediocre antenna might easily be reported S9 or more by the station right next door having a superior antenna. So the popular "S" reports are all but meaningless anyhow!

J.W. Kennicott, W4OVO
Lexington, Tennessee

"circuit figure of merit"

Dear HR:

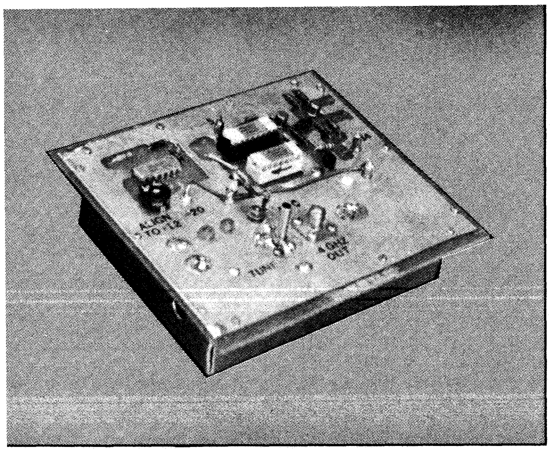
In reference to "Observations and Comments" in the September, 1980, issue of ***ham radio,*** I thought you might be interested in the "Circuit Figure of Merit" used by the State of New York in police two-way fm radio communications in the vhf and uhf ranges.

In writing specifications we usually ask the bidder to guarantee a Circuit Figure of Merit of 3 or better in a defined area of coverage from defined sites and with defined equipment parameters.

Byron H. Kretzman, W2JTP
Huntington, New York

The performance of a two-way radio circuit can be defined by grading the circuit in terms of a "Circuit Figure of Merit" using a scale of 1 to 5 under the following conditions:

circuit figure of merit	grade of circuit performance	voice frequency signal-to-noise ratio	typical receiver quieting
1	Unusable. Presence of speech barely discernible.	Below 8 dB	0 to 6 dB
2	Readable with difficulty. Requires frequent repeats. (Noncommercial)	8 to 16 dB	14 dB
3	Readable with only a few syllables missing. Requires occasional repeats. (Commercial)	14 to 22 dB	20 dB
4	Perfectly readable but with noticeable noise.	20 to 30 dB	25 dB
5	Perfectly readable; negligible noise.	Above 30 dB	Above 25 dB



multipurpose voltage-tuned UHFoscillator

This easy-to-build oscillator features multiple-band application, remote tuning, and phase-lock capability

This uhf oscillator is the result of much experimentation. It has an outstanding record of utility and performance. Despite the opinion of many Amateurs, a good uhf oscillator ***can*** be built without a shop full of machine tools, expensive test equipment, and a high degree of manual dexterity. The PC boards that have been developed for the circuit described here will allow anyone to build a voltage-tuned uhf oscillator.

general description

This oscillator has many applications. It was originally intended for use as the local oscillator in a 1215-1300 MHz TV converter. Later, the board was modified so that the operating-frequency band could be moved up or down to satisfy various other applications. Finally, provisions were made to add either a doubler or tripler circuit to extend the useful output frequency range into the microwave region.

features

The fundamental tuning range of the circuit covers ≈ 1120-1300 MHz. However, by changing the lengths and locations of the frequency-deterrnining circuit elements on the PC board, the operating-frequency range can be adjusted to about 900 MHz and 1400 MHz, giving coverage between 900-4200 MHz with the help of the multiplier circuits.

A varactor provides continuous tuning from a remotely located potentiometer. This feature may be important if you're interested in weak-signal detection, because it allows the entire converter, including the uhf local oscillator, to be located where it belongs – at the antenna.

For television applications, the oscillator may be operated either in the free-running mode or phase locked to a stable reference signal.

The addition of phase-lock capability is easy, because the basic oscillator already includes a tuning varactor. Remote tuning can be used with or without the phase-lock feature. The uhf oscillator is simple. No need for a crystal multiplier chain; therefore no need to struggle with unwanted crystal-oscillator harmonics. Also, if your interest lies in ATV, where crystal control may not be necessary, the design is a natural because of its simplicity.

A divide-by-40 prescaler is mounted on the PC board with the oscillator. The prescaler drives an external frequency counter to monitor the oscillator frequency. Not only is the counter useful as a frequency indicator, it's needed for setting and adjusting the oscillator. The prescaler also provides a signal for the phase detector.

Numerous techniques can be used to phase lock the uhf oscillator to a crystal reference to achieve a high degree of frequency stability; many articles have been written to describe them. In this article, attention is placed on a simple technique that uses a crystal clock as the phase-locked loop (PLL) reference and manual tuning to select the desired lock point. By the proper choice of crystal frequency and divider chains, the uhf oscillator may be locked to any one of a number of desired frequencies. Tuning is done with a ten-turn pot.

applications

Fig. 1 illustrates a typical ATV application that employs the uhf oscillator in the free-running mode as the local oscillator for the mixer. No phase-locked loop is associated with this circuit. A single shielded wire connecting the operating position with the converter serves for tuning, and the converter output is fed over a length of inexpensive transmission line to the receiver. This arrangement avoids the usual degradation in signal-to-noise ratio that generally results from transmitting the rf signal over a long transmission line.

By Norman J. Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126

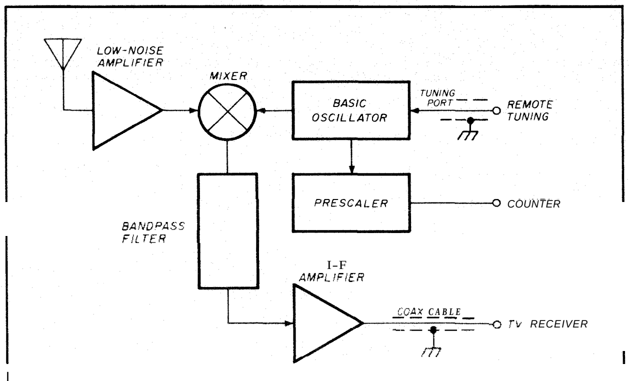


fig. 1. Functional block diagram showing the uhf oscillator in a typical ATV application (free-running mode).

In applications where frequency stability is important, or where a click-stop form of tuning is desired, the basic oscillator can be locked to a stable reference. A block diagram of such a scheme is illustrated in **fig. 2.** The i-f output from the mixer feeds a bandpass filter wide enough to pass the entire band of frequencies of interest, while a wideband fm or television receiver provides the necessary tuning and selectivity. A preselector may be needed between the low-noise preamplifier and the mixer, depending on the application and choice of intermediate frequency.

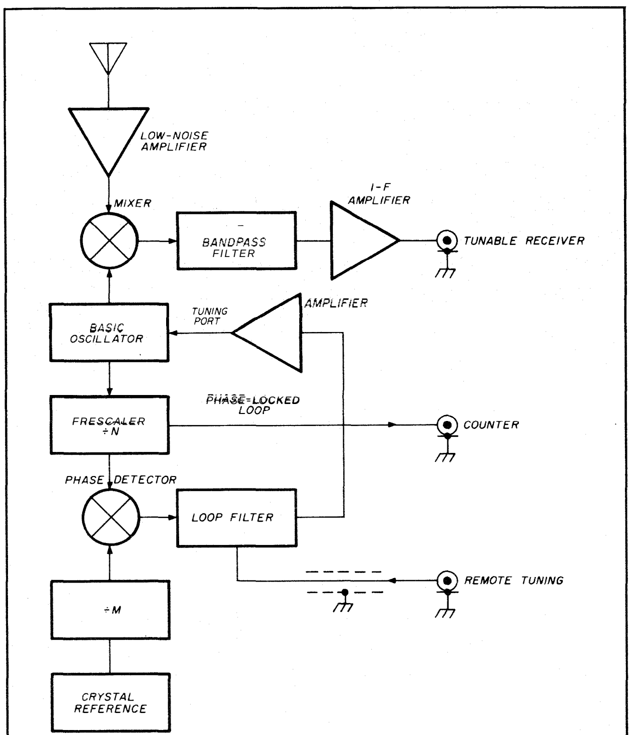


fig. 2. Basic uhf oscillator used in a phase-locked application. Crystal oscillator provides a stable reference for frequency stability.

In both of these arrangements, a frequency scaler drives a frequency counter to permit measurement and continuous monitoring of the uhf oscillator frequency. It's convenient to have this capability, whether the phase-lock feature is used or not. If a programmable counter is available, the readout can display the signal frequency rather than the oscillator frequency.

The advantages to be gained by use of the uhf oscillator described here are now apparent. In some applications the basic oscillator and prescaler alone may do the job, and continuous tuning from a remote location can be used; or a simple PLL may be added for bandswitching, with tuning and selectivity provided by an fm or TV receiver. In either case, a counter can monitor the oscillator (or the equivalent signal) frequency. Other applications can be accommodated using the same PC board with minor modifications, and frequency multiplication can be added for application up into the microwave region.

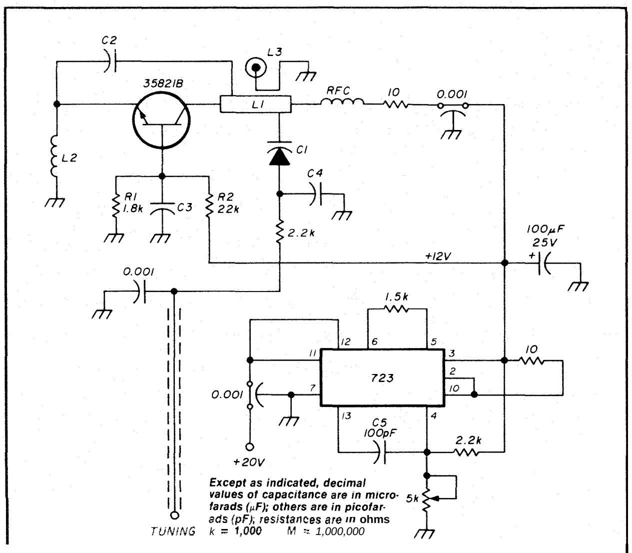


fig. 3. Schematic of the uhf oscillator. Capacitor C1 is the varactor tuning diode (GHZ devices GC-1607 or equivalent — 3.3 pF at – 4.0 volts).

the uhf oscillator

The transistor selected for the uhf oscillator **(fig. 3)** is the HP-35821B. It has an f_t of 4.5 GHz. In the commonbase configuration it's ideally suited for oscillator service. The 35821 has been around for over ten years and is inexpensive. As an oscillator, it can provide 50 mW or more of useful output power with good efficiency.

The base terminals of the 35821 are soldered di-

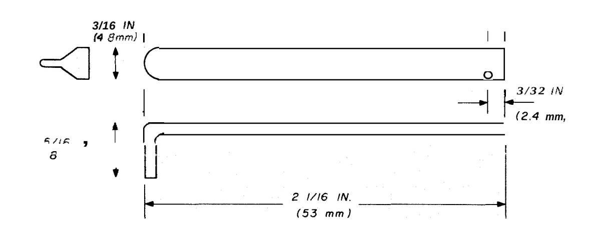


fig 4 Mechanical details of inductor L1 in *fig 3*. Material is 0.032-inch-thick (0.8mm) brass stock.

rectly to the pad provided on the PC board. The board is G10, which is entirely satisfactory for use over the uhf oscillator fundamental tuning range. The board includes all the rf bypass capacitors associated with the oscillator circuit; no chip capacitors are needed.

Fig. 3 is the schematic of the uhf oscillator. There are four special rf circuit elements, L1, L2; C1 and C2. L1 and C1 are the most critical, because they are the principal frequency-determining components. L1 is made of flat brass strip elevated about 0.1 inch (2.5 mm) above the ground plane. The mechanical details of this inductance are illustrated in fig. 4.

Capacitor C1 is a varactor tuning diode connected in series with L1 (fig. 3). It returns to ground through the large pad under L1 but is electrically above ground to accommodate tuning and automatic phase control. The location of C1 sets the effective length of L1. Moving it back and forth adjusts the tuning range up and down in frequency. The distance between the transistor collector and the tuning varactor should be about 1-1/2 inches (38 mm) to tune the range 1120-1320 MHz. The rf ground pad on the PC board was made long intentionally to provide a wide choice of operating range.

Inductor L2 is a four-turn coil wound with No. 18 (1.0 mm) tinned copper busbar with a 1/8 inch (3 mm) inside diameter. The exact inductance of this coil isn't critical.

Capacitor C2 is a feedback capacitor made from 0.010-inch (0.25 mm) shim brass stock 1/2 inch (13 mm) long and 1/8 inch (3 mm) wide. It is soldered to the emitter and extends over the top of the transistor, parallel with the collector inductance, L1. The feedback capacitor is insulated from L1 with 0.001 inch (0.03 mm) Mylar tape. Feedback is controlled by bending the shim to position it closer or further away from L1. Note that the fixed bias divider consisting of R1 and R2 provides very little forward base bias; consequently, the collector current is primarily determined by the amount of feedback from emitter to collector. This is convenient, because it allows a simple means for properly adjusting the feedback. The correct feedback corresponds to the spacing that produces 30-40 mA collector current. Capacitor C3 is a printed-base bypass capacitor. Capacitor C4, which is the rf bypass for the series L1-C1 circuit, is also printed on the osciilator board.

The rf choke is an eight-turn solenoid wound with No. 24 (0.5 mm) enamel copper with a 1/8 inch (3 mm) ID. The junction of the rf choke and the 10-ohm resistor is supported by the terminal of a push-in Teflon standoff insulator.

power output

Overall converter performance can be degraded because of lack of sufficient local oscillator power. Many Amateurs don't have facilities to measure rf power accurately, in which case the adequacy of their local oscillator is unknown. Mixer noise figures less than 5 dB can be realized with 10 milliwatts of LO power. However, as the LO power is reduced below a few milliwatts, noise figure generally increases dramatically. If the mixer in your system needs the help of more than one low-noise preamplifier, chances are that the mixer noise figure is abnormally high. This is most likely the result of inadequate local-oscillator power. It's possible to reduce the mixer's appetite for LO power by various schemes, including applying dc forward bias to the diodes; but for most practical applications, a good design goal for mixer LO power is 10 milliwatts. This point was kept in mind during the design of the uhf oscillator.

The available power from the uhf oscillator described here is, fortunately, quite high, which allows the output to be loosely coupled; in turn this promotes good free-running stability. When the uhf oscillator is used to drive a doubler, power levels well above 10 milliwatts are easily obtained, with the doubler circuit providing the isolation. Power output from a fixed-tuned tripler was measured at +7 dBm minimum when used with an appropriate idler circuit.

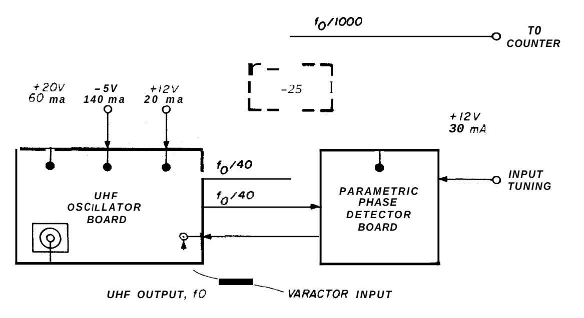


fig. 5. Interface wiring between uhf oscillator board and phase-detector board showing external signal and power requirements.

the phase-locked loop

To provide design flexibility, the oscillator is on one PC board and the phase detector on another. Input signals required by the phase detector are the prescaled signal from the uhf oscillator and the tuning voltage. A single output feeds the VTO (varactor-tuned oscillator) varactor diode for frequency control. **Fig. 5** is a wiring diagram showing a) how these two boards interface, and b) the external signal and power requirements.

The circuit on the phase detector PC board is identical in most respects to the parametric phase detector described in reference 1. This circuit provides considerable design flexibility. In the application here, it operates at about 30 MHz. The circuit **(fig. 6)** also includes provisions for the reference generator, consisting of a quartz crystal and a CD4060B oscillator and divider chain.

Fig. 6 shows the parametric phase detector. This board includes most of the PLL key components, which are the reference generator, spectrum generator, phase detector, and loop filter and dc amplifier. **Fig. 7** shows the phase detector foil and parts layout.

reference signal

The lock points for the uhf VTO are specified in terms of the reference-signal frequency and the prescaling factor. For example, assume the VTO is to be used as the local oscillator in a 23-cm ATV converter and 6-MHz lock-point separation is desired. If a 45-MHz i-f is to be used, the local oscillator frequencies will be 1206, 1212, 1218, and 1224 MHz, corresponding to signal frequencies of 1251, 1257, 1263, and 1269 MHz.

The lock points are 6 MHz apart at the oscillator frequency, but only 150 kHz apart at the phase detector because of the prescaler. The reference needed by the phase detector is therefore 150 kHz. Note that the 202nd harmonic of 150 kHz is 30.3 MHz, which is the spectral line recognized by the phase detector for the 1218-MHz phase lock. Thus, in this type of phase detector, the reference signal must be rich in harmonics. To accomplish this, the phase detector board includes a spectrum generator. On the other hand, if you're interested in a single operating frequency (1257 MHz for example), a crystal-controlled signal at 30.3 MHz is all that's needed. There are, of course, many other schemes that may be used depending on the application.

Tuning and locking to a particular point is easily accomplished by watching the counter. When unlocked, the units and tenths of kilohertz digits will fluctuate due to jitter. When locked, all counter digits will remain steady, and it will be possible to rock the tuning knob back and forth within the hold-in range with no apparent change in the counter status. The final setting should be near the center of the hold-in range.

The pull-in range of the PLL should be less than half the lock point separation; otherwise, if power is momentarily lost, the oscillator may end up locked to the wrong channel. Pull-in range can be controlled

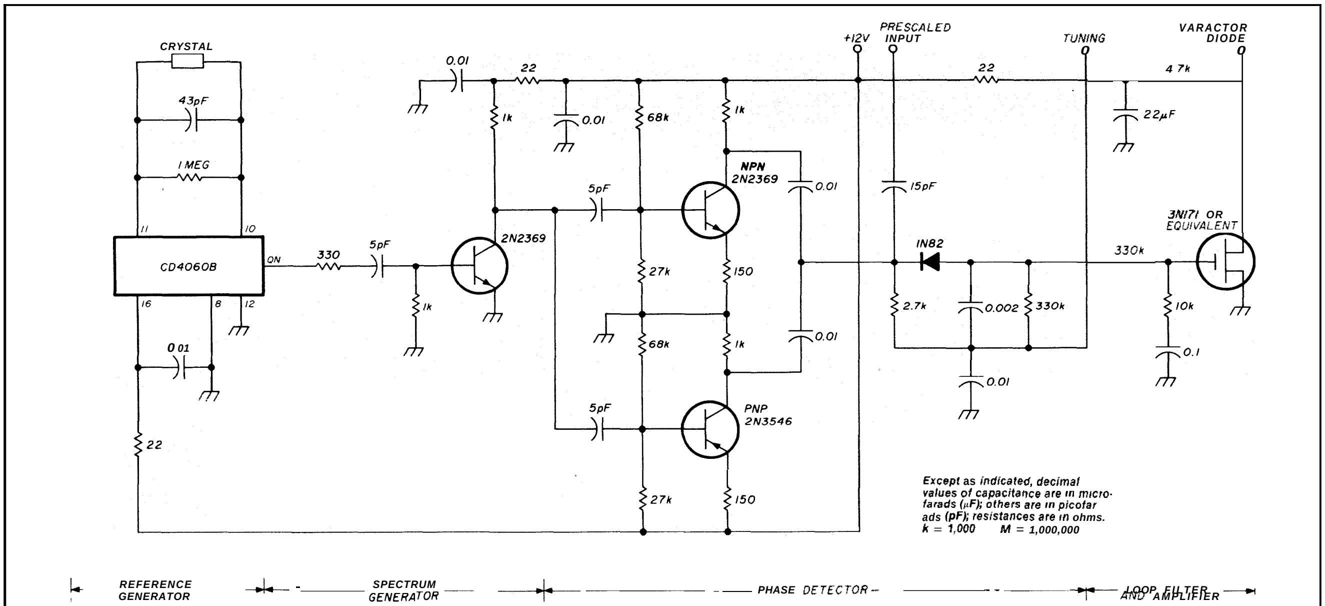


fig. 6. Parametric phase-detector circuit. This circuit includes most of the PLL key components, such as the reference generator, phase detector, and loop filter and dc amplifier.

by adjusting the power level of the prescaled uhf oscillator signal at the input of the phase detector.

prescaler

The Plessey SP-8610 is a 1-GHz divide-by-four prescaler that works well considerably above 1 GHz, even when mounted in a DIP socket. This chip, together with the Plessey 8636 decade divider, provides outputs in the 27-33 MHz frequency range. The circuit is simple and straightforward. One important consideration is that prescalers used at these frequencies require leadless bypass capacitors. Chip capacitors used initially performed satisfactorily from an electrical standpoint, but PC-board flexing caused them to work loose. To solve this problem, leadless capacitors were made by modifying dipped mica capacitors. The insulation was removed with a file, uncovering two metal clamps that hold the stack together. Connections were made directly to the clamps by soldering. This arrangement is entirely satisfactory and considerably less expensive.

The output from the SP-8636 drives a 2N5179 NPN transistor amplifier, which, in turn drives a 2N918 splitter to provide dual low-impedance outputs. One of these is intended to drive the phase detector, while the other can be used to operate the frequency counter. I suggest that an external divide-by-25 circuit be added to increase the overall division factor to 1,000 for the counter. This circuit adds a convenience that relates counter kilohertz to oscillator megahertz. For example, the counrer will dispiay 1200 kHz when the uhf oscillator frequency is 1200 MHz.

A schematic of the prescaler is shown in **fig. 8.** An input signal is coupled to the SP8610 by a small probe bent in an L shape and soldered to pin 4. The bent part of the probe is approximately 1/4 inch (6 mm) long and spaced 3/32 inch (2.4 mm) from L1. The probe should be carefully insulated with Mylar tape to prevent it from coming into contact with +12 volts on L1. Also, to prevent damage, do not overcouple the 8610. The proper procedure is to tune the oscillator to the high end of its range and couple the probe sufficiently for the counter to operate properly.

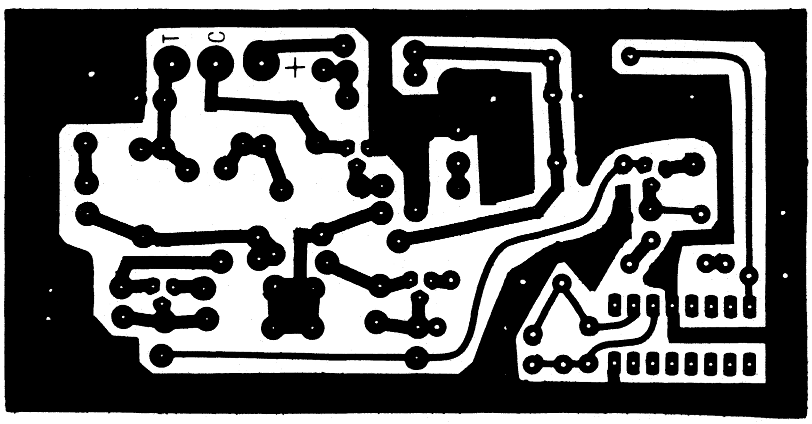
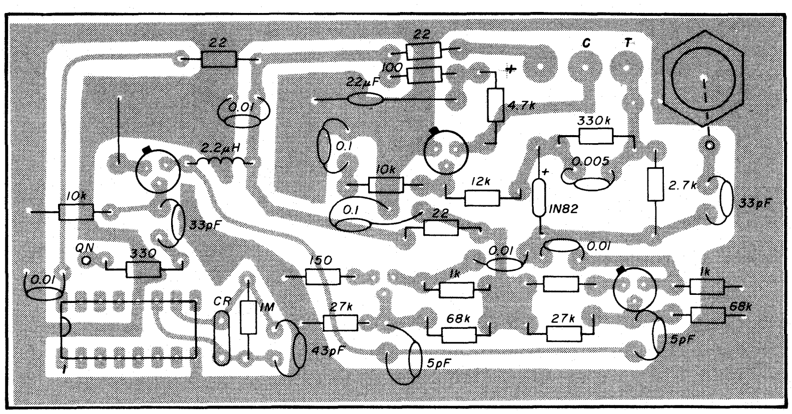
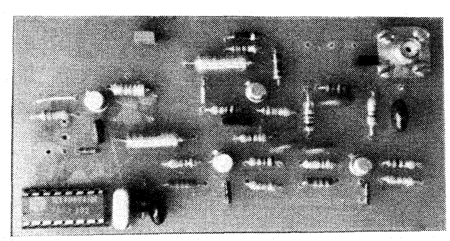


fig. 7. Phase detector board foil side, *top.* and component layout, *bottom.* Assembled board, *lower right.*





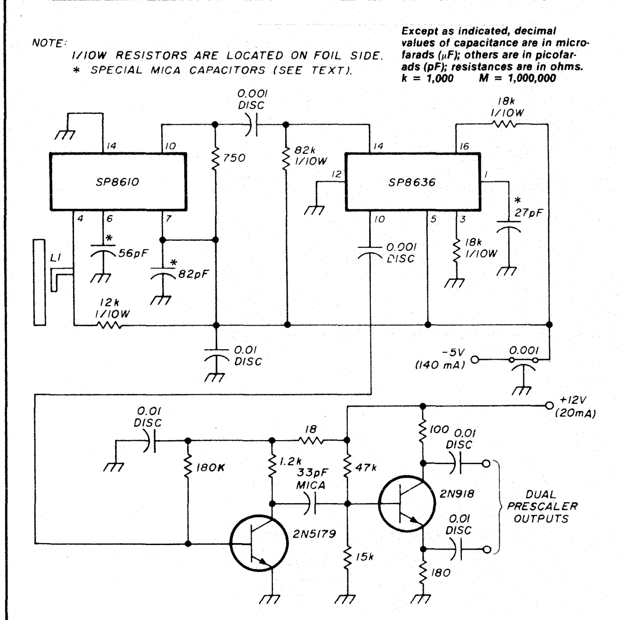


fig. 8. Prescaler schematic. An input signal is coupled to the SP8610 by a small L-shaped probe, which is soldered to pin 4. See text for correct coupling adjustment.

At 1200 MHz, a very small coupling capacitance is sufficient.

construction details

The task of duplicating the performance of the original uhf oscillator is relatively simple when PC boards designed specifically for this project are used. If you don't have the facilities to etch your own boards, they can be obtained from Rock Engineering Supply Company, Inc., 1769 Armitage Ct., Addison, Illinois 60101.

Construction sequence. For the most part, the uhf oscillator assembly is simple except that there is a certain sequence that makes the task easier if followed. I suggest that the feedthrough capacitors be mounted on the board first, followed by the DIP sockets, then all discrete parts not directly associated with the oscillator. Fig. 9 is a drilling template to be used to locate the feedthrough holes, shoulder washers, and Teflon standoff. If the oscillator is to be used at its fundamental frequency, holes should be drilled for the SMA connector. The coupling loop dimensions and assembly are shown in **fig. 10** if an SMA fitting is not available, a BNC type may be substituted.

Connection is made to the rf ground-return pad of the varactor diode by inserting a 2-56 (M2) screw in hole **A,** using a fiber washer to insulate it from the ground plane on the component side of the board. This is the terminal used to bring the tuning and control voltage to the varactor diode.

Varactor diode. The varactor tuning diode should be mounted with special care. Locate it on the rf pad with the cathode side up and solder the anode to the pad. Use a toothpick or pointed object to hold the diode in place during the soldering operation. Apply the soldering iron to the pad, *not* the diode, and only long enough for the solder to flow. Then tin the diode cathode terminal using a fine soldering iron tip. Apply as little solder as possible.

Before proceeding further, cement the two phenolic shoulder washers in the base bypass pad holes with two-part epoxy cement. Use the quick-setting (5-minute) variety to avoid a 12-hour cure cycle.

Collector line. The collector line, L1, should be mounted next. Tin the bottom side of the line where contact will be made with the varactor diode. Insert the pointed end of L1 into the collector shoulder washer hole and solder the line to the varactor diode. Also, to take the stress off the varactor diode, a fiberglass shim should be cemented in place under the line near the rf choke. Trim the shim with a file so that it slides under the line without forcing, then ap-

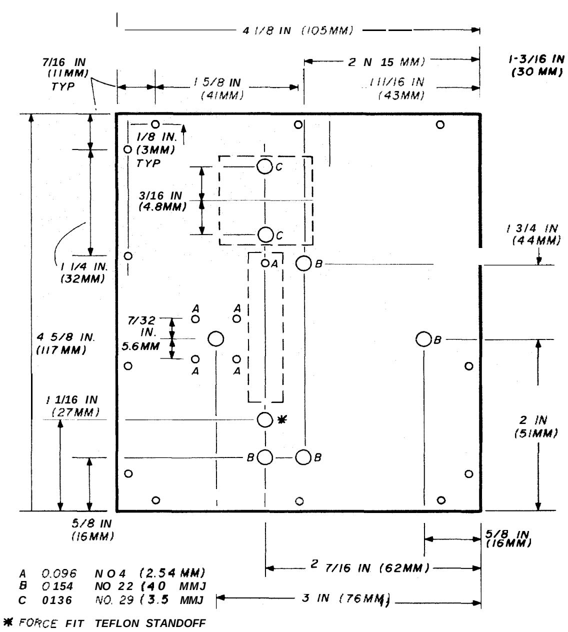
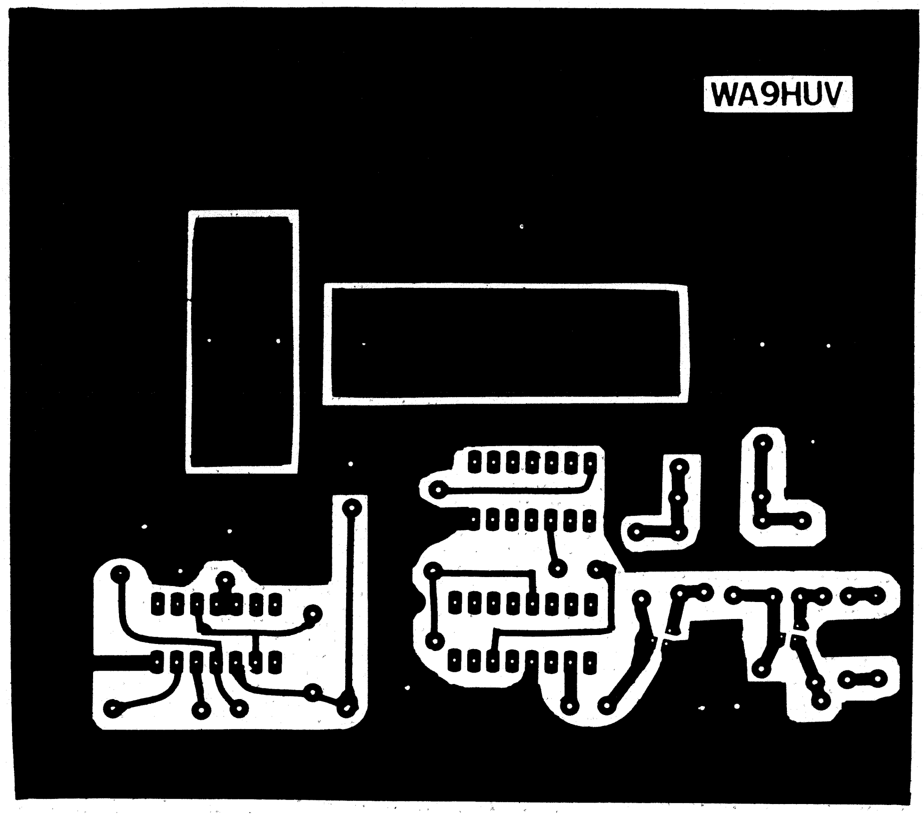
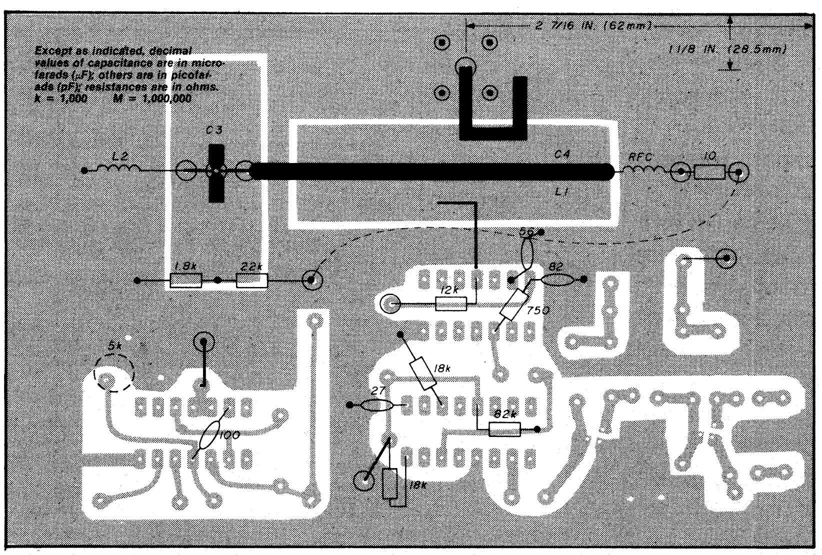


fig. 9. Top view of uhf oscillator board showing mounting-hole locations.





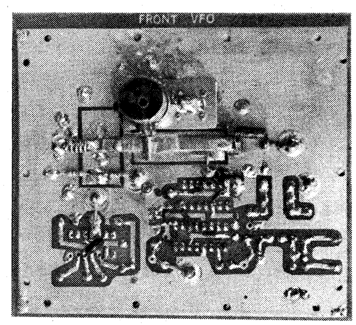
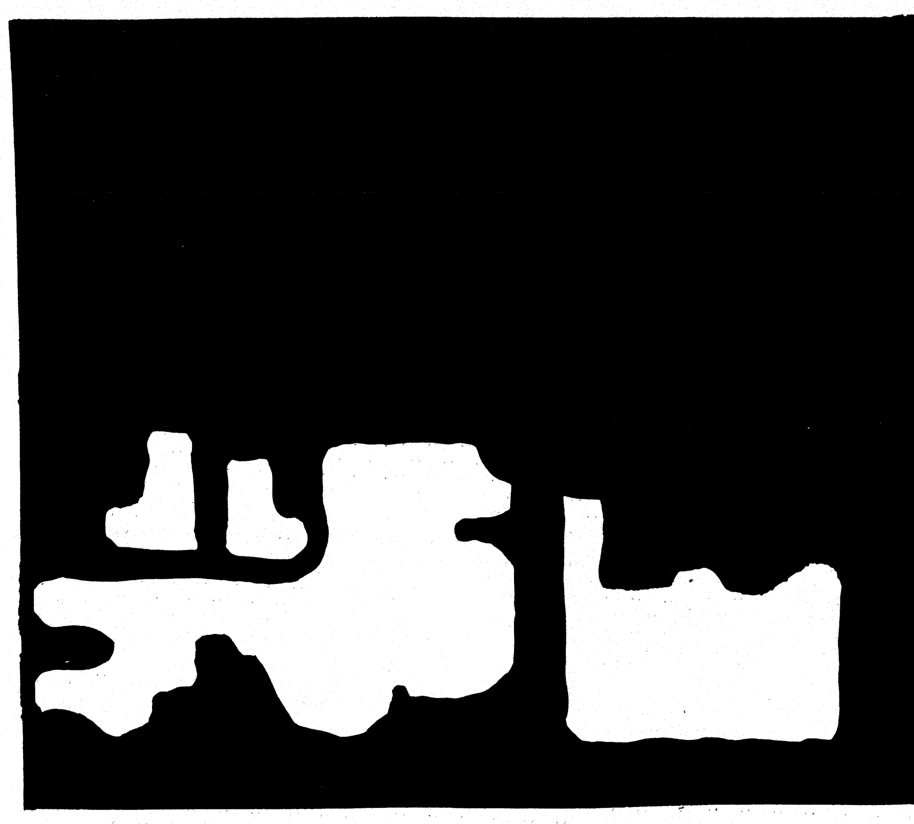
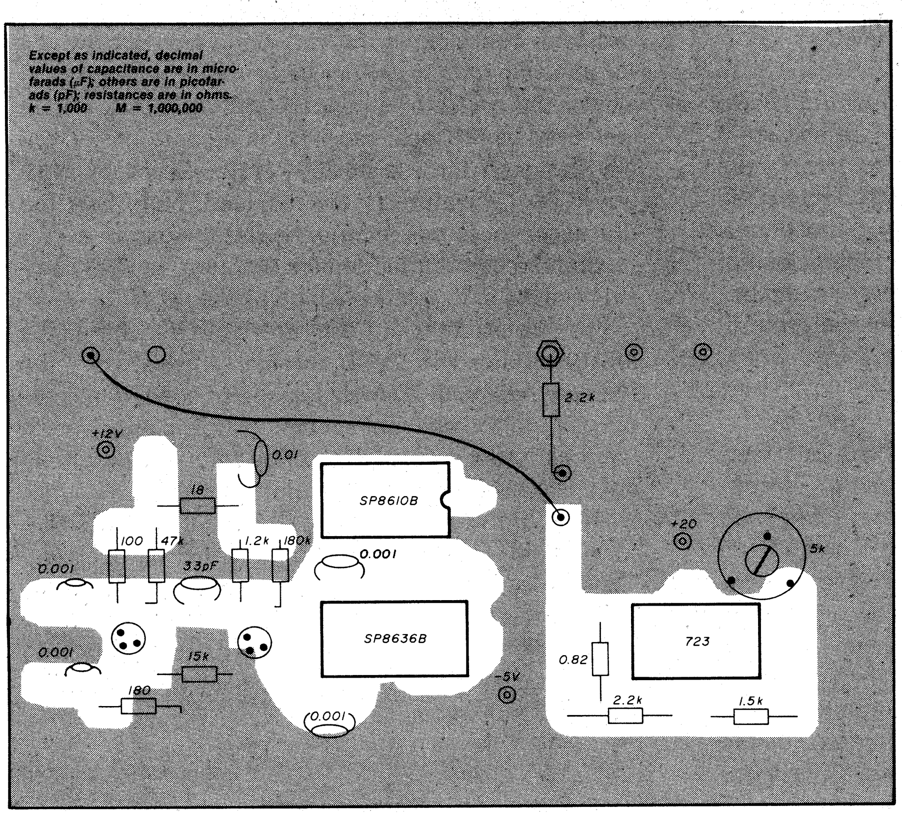


fig. 10. ***Top:*** **Uhf oscillator board, front side.** ***Bottom left:*** **Foil side of oscillator board showing parts placement.** ***Bottom right:*** **Uhf oscillator assembly, top view.**





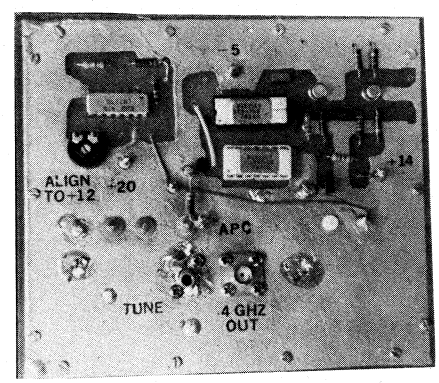


fig. 11. *Top:* Uhf oscillator board, rear side. *Bottom left:* Component-side of oscillator board showing parts placement. Voltage control is a 5k Piheri pot. *Bottom right:* Uhf oscillator assembly, bottom view.

ply a small amount of epoxy cement and secure the assembly in place. Finally, apply a very small amount of epoxy cement into the collector shoulder washer hole to secure L1.

Emitter coil. The emitter coil should be mounted next, and epoxy cement should be applied to the shoulder washer hole to secure it in place. Mount the transistor on the base pad and solder the base leads to the pad. Solder the emitter and collector leads to the emitter coil and L1 respectively, as shown in fig. 10. Solder the feedback shim to the emitter end of L2 (not shown) and insulate the shim with Mylar tape. Space it about 1/8 inch (3 mm) above the collector line.

Before mounting the rf choke and the 10-ohm resistor, check out the 723 regulator and set its output voltage to +12 volts by adjusting the trimpot.

There are five 1/10-watt resistors and three special mica capacitors that are soldered to the *foil* side of the board (see fig. 10). The parts layout on the component side of the uhf oscillator board is shown in fig. 11.

Connect a shielded wire from one of the buffered prescaler outputs to a frequency counter and confirm that the counter displays frequencies between ≈ 27-33 MHz as the tuning control is adjusted.

oscillator enclosure

The mechanical details of the aluminum shield cover that encloses the uhf oscillator are shown in fig. 12. The 2-56 (M) screws used to mount the shield cover on the board also interconnect the groundplane foils on opposite sides of the board. Since initial tests will be made without the enclosure, it will be necessary to insert the screws and temporarily secure them with nuts to simulate the grounding condition.

initial oscillator tests

The uhf oscillator should be checked out first, without the aid of the phase detector board. Temporarily connect a 10k ten-turn potentiometer between +12 volts and ground and connect the arm of the pot to the varactor terminal. Use the regulated voltage from the 723 post regulator. Set the tuning voltage to about 5 volts and monitor the current from the 20-volt source with a milliammeter. When power is applied, the current should be approximately 25 mA. Gradually increase the feedback capacitance until the collector current is approximately 35 mA, but do not exceed 40 mA.

Finally, the phase detector board is integrated into the system as illustrated in fig. 5, and the PLL is then checked out.

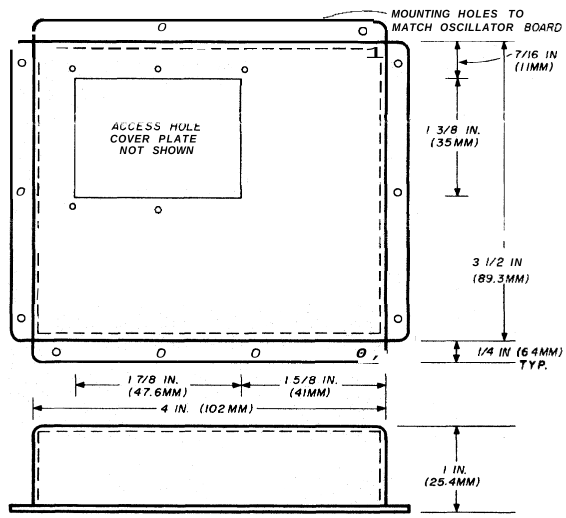


fig. 12. Oscillator enclosure. Material is 0.032-inch (0.8 mm) aluminum.

conclusion

The uhf oscillator described here has many potential applications, depending on your interests. In my case, the performance of an existing 1296 TV converter was considerably improved when the basic uhf oscillator operating in the PLL mode was substituted for the original crystal-oscillator-multiplier chain. A similar uhf oscillator equipped with a doubler circuit was used as the local oscillator in a converter originally designed for use at 2304 MHz. Excellent MDS and ITFS TV pictures were received. Note that the uhf oscillator is not recommended for use in a narrowband receiver intended for CW, am, or SSB service because of its relatively high phase noise.

I've also used the uhf oscillator with a tripler as the local oscillator in a TVRO receiver. In this case, the PLL was built with 20-MHz lock point spacing corresponding to the channel spacing of this class of service. In a future article I'll describe frequency multipliers designed for use with the uhf oscillator.

Some of the parts required to build this uhf oscillator probably won't be found in Amateur parts boxes. These include the prescalers, oscillator transistor, and the tuning varactor. I may be able to suggest sources for some of these parts or help you with other problems. In either case, please send an SASE with your inquiry.

reference

1. Norm Foot, WA9HUV, "High-Frequency Communications Receiver," *ham radio,* October, 1978, page 10.

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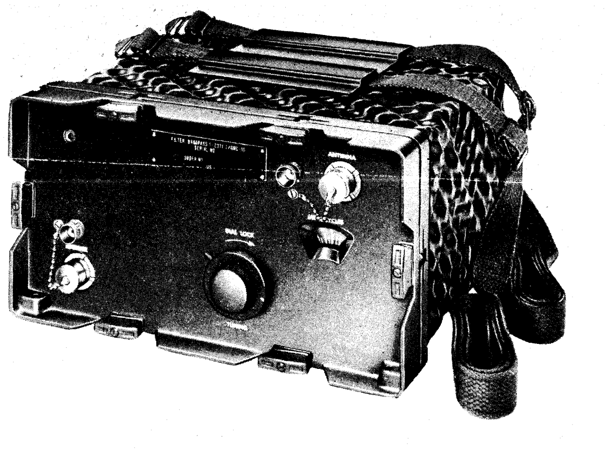

conversion versatility

using the F-237/GRC surplus cavity filter

Good news for VHFIUHF experimenters — this surplus filter can be easily converted for use on 6, 2, and 1-114 meters

In two recent articles,[1,2] **I** described the conversion of several obscure surplus cavity bandpass filters for use in the vhf and uhf Amateur bands. Since then I've found another very interesting surplus cavity bandpass filter* that I've converted for use in the 50-54 MHz, 144-148 MHz, and 220-225 MHz Amateur bands.

The theory and operation of resonant-cavity bandpass filters have been fully covered in the literatures and in my two previous articles. Therefore I'll go right into a description of this surplus "sleeper" and the conversions.

the F-237/GRC-10 bandpass filter

This filter was designed for use with the receiver section of Army radio set AN/GRC-10 and consists of three individual coaxial resonant re-entrant cavities connected in cascade, each tuned with its own variable capacitor ganged for single-dial control.

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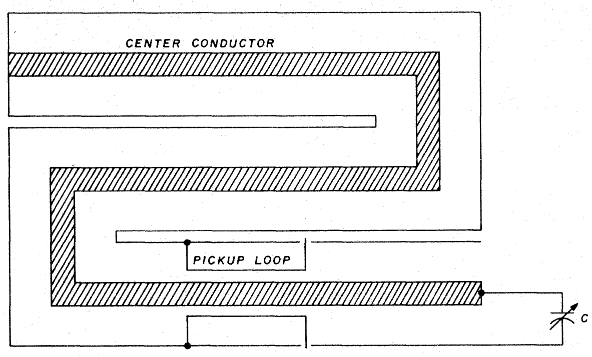

fig. 1. The F-237/GRC-10 bandpass filter showing the zig-zag configuration to compress 22 inches (55.9 cm) of coax cavity into a compact package. Note that the pickup loops are close to the open end.

Each cavity is about 20 inches (51 cm) long but compressed into a compact package by using a snake-like configuration as shown in **fig. 1.** The cavities are of sturdy copper, and the center conductor is silver plated for high conductivity.

Normally, rf pickup loops are located near the shorted high-current end of coaxial type re-entrant resonant cavities where the electromagnetic field is at a maximum. Note that in this cavity, however, the pickup loops are located closer to the open end, evidently to provide looser coupling. This will provide greater selectivity at the expense of a higher insertion loss, which becomes a little over 2 dB per cavity.

The three cavities are similar electrically and physically except that the input and output pickup loops L1 and L6, **(fig. 2)** are a little larger than the others. Also the coaxial cable connection to each cavity varies slightly.

Receiver and antenna jacks on the front panel are made to accommodate a type-C UG-573 connector, which is a jumbo type BNC that's not in general use. If you wish, an N type or uhf type socket can be used in its place by removing the existing socket. Some filing of the socket flange may be necessary to fit into the recessed opening on the front panel.

The F-237 has an input and output impedance of 50 ohms and covers 54-70.9 MHz with continuous tuning. The bandwidth at the 3-dB points is 250 kHz. The attenuation is 40 dB at 4.5 MHz. Insertion loss is 7 dB at resonance. The complete assembly in its cabinet weighs about 16 pounds (7.3 kg) and is approximately 6 × 11 x 11 inches (15 × 28 × 28 cm).

simple conversion to the 6-meter band

Fortunately, the three air-dielectric trimmers C1002-3-4, which are mounted directly on the three-gang variable capacitor C1001 A-B-C, **fig.** 3, have sufficient spare capacitance so they can be adjusted to cover the 50-54 MHz band. After adjustment, the range is 49.5-60 MHz.

Because of the high selectivity, the following procedure is suggested. Set the tuning dial at the lowest frequency position, 54 MHz, and feed a 53-MHz signal into the antenna terminal from any convenient source, such as a grid-dip meter or signal generator. Adjust the three trimmers for maximum output as measured at the receiver terminal using an rf meter or receiver S-meter. A simple rf meter can be made using a germanium diode such as the 1N34 in series with a microammeter.

Repeat the above procedure in small steps until 49.5 MHz is reached; the trimmers should now be at almost maximum capacitance with some to spare for final adjustment. If this filter is to be used with a receiver only, it can be inserted into the transmission line and, with a weak signal around 52 MHz, the filter tuning dial can be tuned for maximum output. The trimmers can then be repeaked for maximum output.

If the filter is to be used with a transmitter or transceiver, an SWR indicator should be used between transmitter and filter. The trimmers should be adjusted for minimum SWR at 52 MHz. The tuning dial can then be calibrated in any manner you choose.

lowering the insertion loss

For general Amateur use, 7 dB is quite a large bite to take out of the received or transmitted signal. The F-237 filter assembly can be modified to provide less insertion loss at the expense of a little selectivity by using only one or two of the original cavities instead of all three. Even with a single cavity, selectivity is adequate for most Amateur applications.

To lift out the cavity assembly and its ganged capacitors in one piece, remove all the screws from

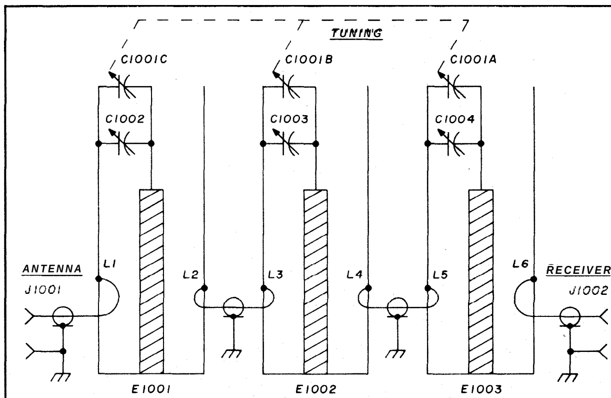

fig. 2. Schematic of the F-237/GRC-10 cavity. The antenna and receiver pickup loops are larger than those connecting the assembly.

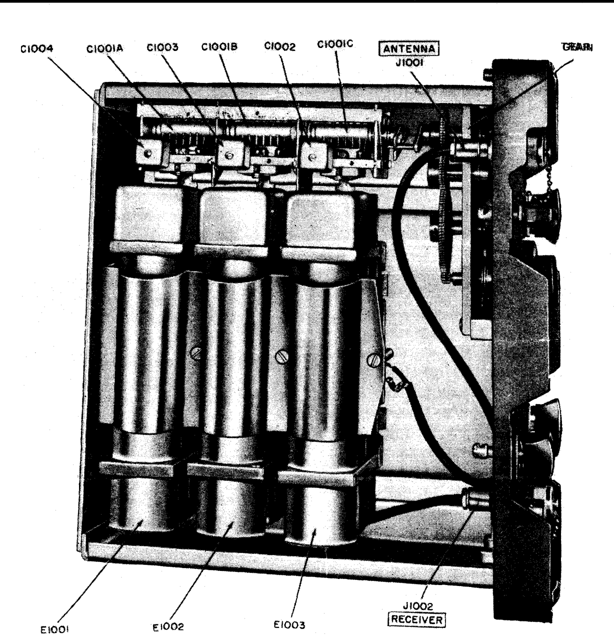

fig. 3. Interior of the F-237/GRC-10 cavity. The three air-dielectric trimmers have enough capacitance to allow coverage of 50-54 MHz.

the underside and unsolder the two coaxial cable leads leading to the front panel. To eliminate a cavity section, remove the Phillips-head screw and unsolder the ground strap. Unsolder the cavity center conductor from the variable-capacitor stator plates and the cavity will unpiug from its adjacent cavity (**fig. 4**).

If only one section is to be used, any of the cavities will do. If two sections are to be used, then eliminate the center cavity and interconnect the remaining two with a short length of RG-58/U coaxial cable. This arrangement is necessary to ensure proper tracking. Adjustment follows the original procedure.

even less insertion loss

The insertion loss can be reduced to under 1 dB per cavity section by rearranging the cavity so that the pickup loops are placed in the high-current end of the cavity. This can be done by reversing the cavity sections as shown in **fig. 5.**

Unsolder the closed end plate at **A** and resolder it to the other end, B. Make certain that very good electrical contact is made between the center conductor and the housing at this high current end, B. Unsolder the ground strap and relocate as shown. Cut a short length of copper or brass rod and insert it into the center conductor at **A** so that it will reach the tuning-capacitor stator. Finally, unsolder the mounting bracket and replace it at the other end as shown.

Reassemble the cavities to the ganged capacitors and you now have a bandpass filter with an insertion loss of less than l-dB per cavity section. The selectivity is still adequate even if you use only one cavity to do the job. The adjustment and tuning is as previously described.

for use with higher power

The F-237 bandpass filter is tuned to resonance by a three-gang variable capacitor of excellent quality with 0.06-inch (1.5-mm) spacing between plates. It should withstand power levels in the order of several hundred watts. The weak point in the filter is the very small air dielectric trimmers, which will probably arc over with rf power in excess of 30-40 watts. To overcome this limitation, the trimmers can be removed and replaced with the APC type of trimmer, 20 pF or more, and with a plate spacing of at least 0.03 inch (0.76 mm). The larger trimmer will also extend the low range a few MHz below 49.5 MHz.

conversion to the 2-meter band

This conversion can be made from either left-over cavities from the 50-54 MHz conversion or from another F-237. A length of 22 inches (56 cm) of coaxial re-entrant cavity is too long for 144-148 MHz and must be shortened to allow for variable capacitance loading.

Fig. 6 shows a convenient method of obtaining a workable length, while at the same time placing the pickup loops very close to the shorted high-current end of the cavity. In addition, the open end is terminated in a handy housing for the variable capacitor.

As shown in **fig. 7,** carefuily eiiminate the shaded portion with a sharp hacksaw; this will leave about 11 inches (28 cm) of cavity for the 2-meter band. File all rouugh edges to a flat and smooth finish and tin thoroughly at both ends for soldering. Unsolder the

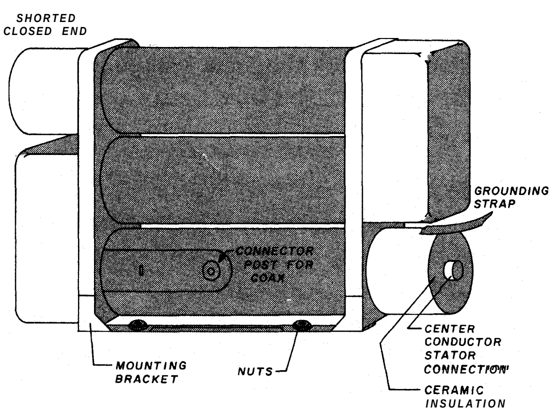

fig. 4. Unmodified cavity section detached from the F-237/GRC-10 assembly.

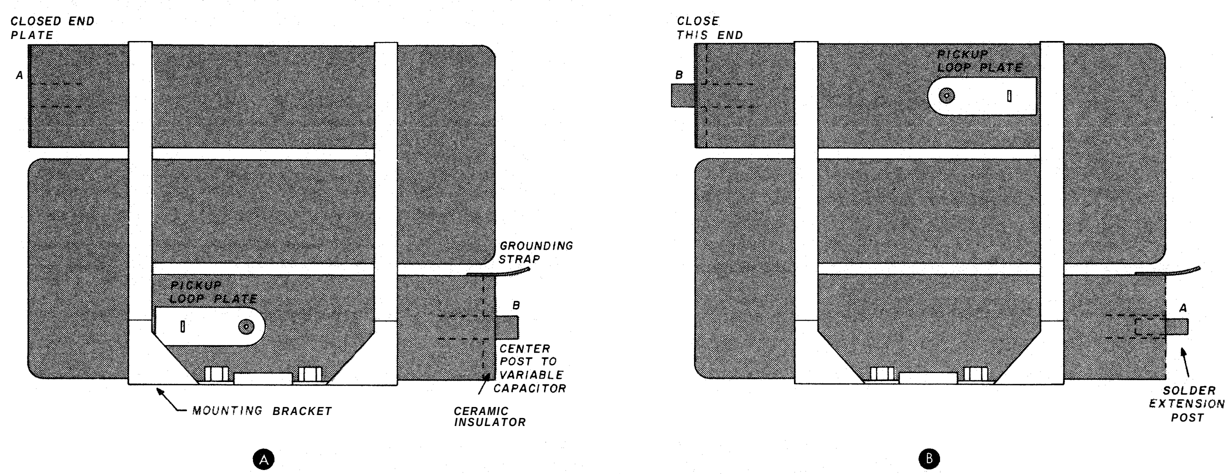

fig. 5. Reversal of cavity to place pickup loops in the high-current area for lower insertion loss. Sketches (A) and (B) show before and after mods.

right-angle portion of the inner conductor as shown.

The two pickup loops will now be visible and accessible from the short open end. Using a screwdriver, bend the center of each loop toward the housing away from the center conductor as shown by the dotted line in **fig. 8.** Try to make the loops as symmetrical as possible.

To close up the end near the pickup loops, unsolder the end plate on the cut-off portion or cut a piece of flashing copper to 1-112 inch (3.8 cm) diameter with a 114-inch (0.6-cm) opening in the center. Solder either one securely to ensure good electrical contact at this high-current area.

Select an APC air dielectric trimmer capacitor and install in the cubical housing as shown in **fig. 9.** A capacitance of about 25 pF with an air gap spacing of at least 0.03-inch (0.76-mm) should fit into the available space and provide adequate tuning range. Solder the stator plates to a heavy lead and attach to the center conductor. The rotor wiper arm should be soldered directly to the housing wall. Try to obtain an APC trimmer with a standard 114-inch (0.6-cm) shaft so

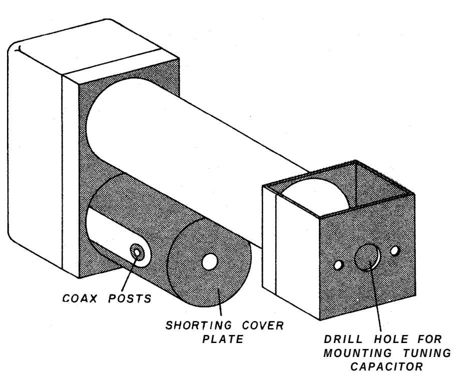

fig. 6. Modified cavity section for the 2-meter band. A workable length is obtained, and pickup loops are close to the shorted high-current end of the cavity.

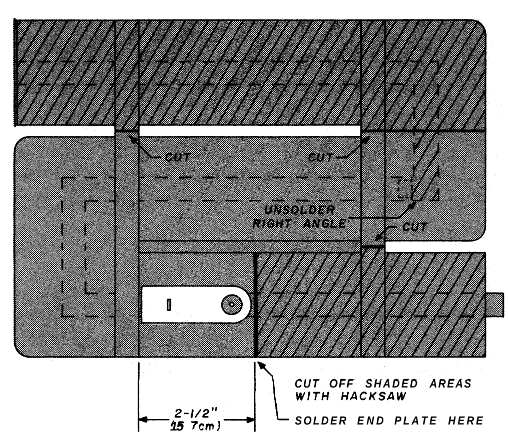

fig. 7. Shaded portions are removed with a hacksaw to leave about 11 inches (28 cm) of cavity for the 2-meter band.

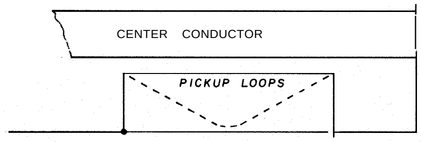

fig. 8. Pickup ioops are bent as shown for the 2-meter conversion.

that a knob can be used instead of the inconvenient screwdriver adjustment.

To test the unit for frequency coverage, attach a 314-inch (1.9-cm) loop to either coaxial terminal and couple a grid-dip meter to it. A sharp dip will indicate resonance, which should occur about midrange with plenty of spare capacitance on either side of resonance. The open end of the cavity can then be closed with flashing copper or left open as you wish.

conversion to 220-225 MHz

This modification is identical to the 144-148-MHz conversion except for the tuning capacitor. At this frequency, even the minimum capacitance of the APC trimmer is too high; therefore, a simple very low capacitance trimmer can be built using two copper pennies. Solder one penny to the inner conductor and the other to a brass machine screw as shown in **fig. 10.** Solder a brass hex nut to the outside of the housing and use a second hex nut to lock in the frequency adjustment. A grid-dip meter can be used to check the frequency range, which should be between approximately 180-240 MHz.

an experimenter's delight

The several conversions discussed in this article are just a small sampling of what can be done with the F-237. One assembly will supply three cavities; one for each band, or all three for one band.

For those who wish to experiment, a length of cavity somewhat shorter than the 11 inches (28 cm) used for the 144-MHz band can be used with a 50-pF air trimmer to provide coverage of both the 144- and 220-MHz bands with one cavity. Also, by using a shorter length of about 3-5 inches (7.6-12.7 cm), this cavity section can be made to resonate in the 440-MHz band.

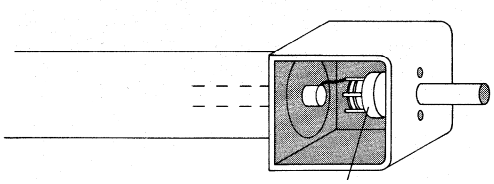

fig. 9. installation of the APC trimmer capacitor for the 2-meter conversion.

The size of the pickup loops, which serve an important role in impedance matching and determining cavity selectivity, can be changed by unsoldering the elongated mounting strip for easy access. Also, for convenient cable connection, small sockets such as the BNC, F, or RCA type can be used as they are small enough to be mounted into the strip.

Another suggestion: You can attach three modified cavities, each for a different band, to the stators of the three-gang tuning capacitor. Separate sets of coaxial cables can be run to sets of separate terminals on the front panel, or a three-position switch can be used to select the cavity to be used. Depending on the length of each cavity, the individual capacitor sections can be used to tune the desired band. If the capacitance is too high, rotor plates can be easily removed to lower capacitance to fit the application'. The main tuning dial can be calibrated with three separate scales, as required.

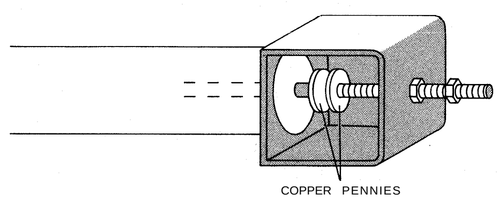

fig. 10. Air trimmer installation for the 22-MHz conversion using two copper pennies to replace the APC trimmer capacitor.

summary

With 66 inches (167.6 cm) of good-quality coaxial cavity available, a three-gang variable capacitor, three shielded miniature air dielectric trimmers, a precision tuning assembly, and a sturdy metal cabinet, vhf and uhf experimenters can really have a field day with the F-237/GRC-10.

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1. William Tucker, W4FXE, "How to Modify Surplus Cavity Filters for Operation on 144 MHz," ham radio, February, 1980, page 42.
2. William Tucker, W4FXE, "More Conversions of Surplus Cavity Bandpass Filters," ham radio, March, 1980, page 46.
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ham radio

Yagi antennas:

practical designs

Last in the Yagi design series, with emphasis on scaling and element taper

In all the previous articles of this series the specifications for a Yagi antenna have been stated only in terms of strictly cylindrical elements. Each element is characterized by an x coordinate or position along the boom, a physical length, *LE*, and a radius RO; each of these three quantities is expressed in terms of wavelengths, λ, at a central design frequency. Such specifications have led to a number of rather good antenna designs, and I shall shortly list a brief selection of such designs. However, when a real Yagi antenna is constructed it will rarely ever be convenient to adhere rigorously to the given cylindrical element design. To start, the element diameter is usually adjusted to fit a mechanical requirement (wind loading, etc.); moreover, the element itself is usually not a cylinder, but a series of telescoping tubes starting with a large-diameter section at the boom and tapering to a small-diameter section at the outer end of the element. In addition, the element is fastened to the boom with a clamping arrangement that may be a plate or angle bracket U-bolted to both boom and element. Some mechanical designs even put the element directly through the boom. Thus, the path from the cylindrical design to a practical antenna will involve three tasks: scaling the original design to an equivalent new design using a different (average) element radius, computing the potentially significant change in element length as a result of the chosen (telescoping) taper schedule, and making (usually minor) corrections to allow for the boom clamping system. Methods for carrying out each of these three tasks will be given following the next section on preferred antenna designs.

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preferred antenna designs

In this section I shall discuss one preferred design for a two-, three-, four-, five- or six-element Yagi antenna. Recall that simplistic Yagis[4] (element spacing uniform and all directors having a common length) are as good as any other design up to a boom length of one wavelength. It was shown that a good two-element beam would have a boom length of about 0.15λ; the exact length is not critical and is a compromise between better gain and lower efficiency and bandwidth. Best parasite element length is a compromise between better forward gain and lower FIB ratio. For a three-element beam it was shown that a boom length of about one-quarter wavelength produces a naturally high FIB and similarly for four-, five-, and six-element beams a boom length of about 314 wavelength gives a naturally good FIB ratio.

Table 1 shows the characteristics of these good Yagi designs. These particular antenna designs are not unique; for example, the boom length can be varied somewhat. Longer booms, in general, give larger forward gain, but the frequency for highest FIB ratio drops somewhat below the center of the band, where gain remains high.

A procedure has also been described that allows fine tuning or optimization to improve the FIB ratio;[5] this optimization procedure can be done for Yagi antennas having four or more elements, Optimization must be done for a specific end use. **Table 2** shows optimized six-element beams first for free-space use, next for operation at 1.0λ over ground, and finally for operation in a two-Yagi stack at heights of 0.60λ and 1.5λ. These parameters are mathematically correct. But note that approximations used in the model really do not justify complete confidence in the precise values in **table 2.** Nevertheless, I suspect that practical antennas constructed from this table (for use over ground) will exhibit superior properties to the (free-space) 6-element case shown in **table 1.**

scaling

Any of the Yagi antenna designs, such as those in **table 1,** can be scaled either to other center frequencies or to elements of different diameter at the same center frequency. Because all design parameters include dimensions expressed in wavelengths at a central design frequency, the design itself is independent of frequency scaling; therefore, the behavior of the antenna will not be affected by the choice of central design frequency. However, this is true only if the design is truly unchanged; that is, *all* physical dimensions (including element radii) are adjusted proportional to the desired wavelength.

table 1. Preferred Yagi antenna designs. All elements with radius, RO, of 0.0005260 (λ_0), length, *LE*, in (λ_0) and boom position. *X*, in (λ_0).

element	*X*	*LE*	*X*	*LE*	*X*	*LE*	*X*	*LE*	*X*	*LE*
R	0.000	0.49366	0.000	0.49801	0.000	0.49185	0.0000	0.49994	0.000	0.49528
DR	0.150	0.47050	0.150	0.48963	0.250	0.47900	0.1875	0.48040	0.150	0.48028
D1			0.300	0.46900	0.500	0.46319	0.3750	0.45232	0.300	0.44811
D2					0.750	0.46319	0.5625	0.45232	0.450	0.44811
D3							0.7500	0.45232	0.600	0.44811
D4									0.750	0.44811
number elements	2		3		4		5		6	
gain (dBi)	6.88		7.86		10.62		10.45		10.70	
FIB (dB)	7.94		23.60		41.62		32.27		52.71	

table 2. Optimized 6-element Yagi antenna, RO is 0 0005260 (λ_0), *LE* in (λ_0), and X in (λ_0).

	A		B		C	
element	*X*	*LE*	*X*	*LE*	*X*	*LE*
R	0.0000	0.49528	0.0000	0.49528	0.0000	0.49528
DR	0.1500	0.48071	0.1500	0.48028	0.1500	0.48157
DI	0.2992	0.44811	0.3039	0.44811	0.3029	0.44811
D2	0.4500	0.44811	0.4500	0.44811	0.4500	0.44811
D3	0.6000	0.44811	0.5959	0.44811	0.6395	0.44811
D4	0.7500	0.44811	0.7500	0.44811	0.7500	0.44811

Note:
A. Optimized in free space.
B. Optimized at 1 0λ_0 over ground.
C. Optimized in a stack/ground at 0 6λ_0 and 1 5λ_0

Experience has shown that ***desired*** element radii expressed in wavelengths is not constant; at low frequencies (long wavelengths) relatively thin elements are used, while at high frequencies relatively fat elements are normal. How, then, can a given design be altered to an equivalent design where element radii are changed? The clue is to make the impedance of the changed, or scaled, element exactly the same as the impedance of the original unscaled element at the central design frequency; in this way exactly the same element currents will flow, resulting in the same detailed antenna performance. Because the (radiation) resistance of the element is essentially unchanged, we need only to make the reactance invariant to scaling-element radius.

Recall[2] that element reactance, X, near resonance can be expressed as:

$$X = RQ(F/FR - FR/F) \quad (1)$$

where R = the (radiation) resistance
Q = the effective Q
F = the frequency referred to central design frequency
FR = the element resonant frequency, also referred to central design frequency.

Recall also that RQ can be (rather accurately) empirically expressed as:

$$RQ = (215.15 \log K - 160) \quad (2)$$

where $K \equiv 1/RO$
RO = the radius of the element expressed in wavelengths at F = 1, the central design frequency.

From eqs. 1 and 2:

$$X = (215.15 \log K - 160)\ (F/FR - FR/F) \quad (3)$$

and at the central design frequency ($F = 1$):

$$X_{(F=1)} = (215.15 \log K - 160)\ (1/FR - FR) \quad (4)$$

Thus, if we wish to scale the element radius from an original value to a new value, we must ensure that $X_{(F=1)}$ is unchanged. Note that $X_{(F=1)}$ contains two variables, (K and FR), which are a function of element radius RO. Recall[2] FR is calculated from the physical length of element LE and physical resonant length LER; both of these lengths are measured in wavelengths, λ_0, at F = 1:

$$FR = LER/LE \quad (5)$$

Empirically,[2]

$$LER = [1 - (10.7575 \log K - 8)^{-1}]/2 \quad (6)$$

Thus, from eqs. 5 and 6:

$$FR = [1 - (10.7575 \log K - 8)^{-1}]/(2LE) \quad (7)$$

We now have the tools to convert a given antenna, such as one in table 1, to a new (scaled) antenna where the element radii are changed; the new scaled antenna will perform exactly in the same way as the original antenna at the central design frequency ($F = 1$). However, the frequency-swept behavior of the (scaled) antenna, while qualitatively similar to the original, will show a broader or narrower bandwidth, depending on the change in element Q (see eq. 2).

The procedure is simple. For any given original element (subscript 1) we are given LE_1 and RO_1. The new (scaled) (subscript 2) radius is designated as RO_2. Compute the new (scaled) element length, LE_2:

$$K_1 = 1/RO_1 \quad ; \quad K_2 = 1/RO_2 \quad (8)$$

$$FR_1 = [1 - (10.7575 \log K_1 - 8)^{-1}]/(2LE_1) \quad (9)$$

$$X_1 = (215.15 \log K_1 - 160)\ (1/FR_1 - FR_1) \quad (10)$$

Having calculated reactance (at $F = 1$), compute the value of FR_2 that will give the same value of X with the new element radius, RO_2:

$$X_2 = X_1$$

$$(1/FR_2 - FR_2) = X_1/(215.15 \log K_2 - 160) \equiv A \quad (11)$$

$$FR_2 = [-A + (A^2 + 4)^{1/2}]/2 \quad (12)$$

$$LE_2 = [1 - (10.7575 \log K_2 - 8)^{-1}]/(2FR_2) \quad (13)$$

It is simple and convenient to set up the entire procedure (eqs. 8-13) on a small programmable calculator.

An example illustrates the results. Consider the antenna design for the six-element antenna in table 1;

table 3. Six-element Yagi; element length, LE (λ_0).

	reflector			driver			director		
	1	2	3	1	2	3	1	2	3
LE (λ_0)	0.49528	0.49489	0.49465	0.48028	0.47876	0.47785	0.44811	0.44431	0.44204
FR	0.97252	0.97042	0.96917	1.00289	1.00311	1.00325	1.07489	1.08090	1.08451
X (ohms)	30.40800	30.40800	30.40800	−3.14700	−3.14700	−3.14700	−78.58200	−78.85200	−78.85200

Note:
Column 1 $R_0 = 0.0005260\ (\lambda_0)$, from table 1
Column 2 $R_0 = 0.0008\ (\lambda_0)$
Column 3 $R = 0.0010\ (\lambda_0)$

this would be a reasonable design for a 14.2-MHz antenna where $\lambda_0 = 69.3$ feet (21.13 meters) and where an RO of 0.0005260 (λ_0) would correspond to an element physical diameter of 0.875 inch 12.22 cm). This would be a reasonable dimension for a mechanically adequate element. Now, suppose that we would like an equivalent antenna for 28 MHz, where RO probably should be increased. The results of **eqs. 8-13** are shown in **table** 3. Note that the (scaled) changed values for LE are not wholly intuitive, because two things happen simultaneously. As RO increases the Q decreases, requiring a greater spread in resonant frequencies of reflector and director; however, at the same time, the resonant ***physical*** length, LER, also changes. Note that, if one scales the actual physical dimensions of boom length up by a factor, S (from, say, a smaller high-frequency antenna model), and the element radius dimension is not also scaled up equivalently, it is wrong, conceptually, to scale element length by the same factor S. Moreover, it is also wrong, in this case, to scale down element resonant frequency by the same factor, S. The only correct way to scale an antenna element is to design it (length and radius) to give the same electrical reactance.

element taper corrections

To this point, antenna designs and all antenna calculations have been made for strictly cylindrical elements, and the results will apply directly to most high-frequency (small) Yagi antennas where the general practice is to use cylindrical elements. However, for frequencies less than about 30 MHz, mechanical considerations usually require that the elements consist of one or more telescoping sections of tubing. At the lower frequencies (say ≤ 7 MHz), the Yagi antenna becomes gigantic, and it is no small mechanical engineering task to construct even a good element. Small diameters favor smaller wind forces, but these diameters are insufficiently rugged for long elements. It is, therefore, a practice to make these large elements of several telescoping sections. The largest-diameter section is clamped to the boom, and succeeding monotonically smaller-diameter sections make up the outer portions of the element. The resulting element taper can introduce a significant change in the required element length.

It's important to understand how to relate the actual detailed taper schedule of an element (diameters and lengths of all sections) to the equivalent length of a cylindrical (untapered) element having the same average or mean diameter. Equivalence is intended to mean that the resonant frequency and the Q are the same for the actual tapered element as for the equivalent cylinder.

To start, I shall introduce the concepts of element pipe inductance and pipe capacitance. Consider a cylindrical element of length s and radius RO as shown in **fig. 1.** A length coordinate, x, is defined with the origin at the center of the element and a related (angle) coordinate, 8, where $8 = \pi x/s$. Note that electrical excitation of this element in the neighborhood of the resonant frequency, **f,** will produce a current and voltage distribution:

$$I_\theta = I_0 \sin (2\pi ft) \cos (8) \qquad \textbf{(14)}$$

and

$$V_\theta = V_0 \cos (2\pi ft) \cos (8) \qquad \textbf{(15)}$$

The electrical driving-point impedance of the element consists of a resistance (which is directly related to far-field energy radiation) and, of course, a reactance.

All reactance effects, including resonant frequency and electrical Q are caused by near-field (non-radiating) energy storage. Energy storage occurs in two ways: the magnetic flux surrounding the current distribution in **eq. 14** and the electrical field produced by the voltage distribution in **eq. 15.** Note that at certain instantaneous times $(t = n/2f)$, the current everywhere is zero, and all stored energy resides in the electrical field. Similarly, at certain other times $(t = n/2f + 1/4f)$ the electric field vanishes, and all stored energy resides in magnetic flux.

As time progresses the (constant) total stored energy transfers back and forth between magnetic and electrostatic fields. This transfer or exchange frequency is, of course, the element resonant frequency. As a result of this complete nonradiative energy transfer, the peak or maximum magnetic stored energy must exactly equal the peak electrostatic stored energy. Note also that the resonant or natural exchange frequency must decrease as the total stored energy is increased.

Now, consider the effect of inserting an infinitesimal length of pipe (of the same radius, RO) into the element of **fig. 1** at the center $(x = 0.)$ The original

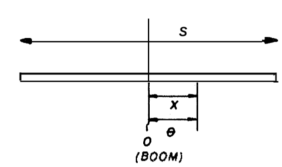

fig. 1. Coordinates of a single Yagi element. S = overall length, X is coordinate of length with zero at center (boom), and θ is corresponding angle ($\theta = \pi \cdot S/X$).

(subscript 1) element driving-point reactance,* X, was shown to be:

$$X = (430.30 \log K - 320)\ (F/FR_1 - 1) \qquad \textbf{(16)}$$

where $K = \lambda/RO$

At the (original) resonant frequency, FR_1, the reactance vanishes; inserting an additional infinitesimal length of pipe, As, at X = 0 will change the resonant frequency to FR_2. At this new frequency the total reactance again vanishes. The added reactance due to the inserted pipe must be balanced by the original pipe reactance at the new frequency:

$$0 = (430.30 \log K - 320)(FR_2/FR_1 - 1) + 2\pi f \Delta L \qquad \textbf{(17)}$$

where f = actual (resonant) frequency
AL = increased inductance due to Δs.

The inserted pipe at x = 0 can produce only inductive effects (stored magnetic flux) since the electrical potential is strictly zero. Now, FR_2 is clearly related to FR_1 by the overall length(s) of the element:

$$FR_2/F_1 = s/(s + \Delta s) \qquad \textbf{(18)}$$

from which

$$\Delta L/\Delta s = (430.30 \log K - 320)/(S2\pi f)$$

and

$$\Delta L/\Delta s = (430.30 \log K - 320)/(\pi c) \qquad \textbf{(19)}$$

where c is the velocity of light.

Thus, the addition of the small infinitesimal pipe section causes the element to behave just as though a pure series inductance were added. The effective inductance per unit length, which I designate by IND, is given by eq. 19 and is easily expressed in conventional units as:

$$IND = (43.03 \log K - 32)(1.061 \times 10^{-8}) \text{ henries/meter} \qquad \textbf{(20)}$$

From the simple model of a resonant circuit it is easy to relate the magnitude of voltage on the reactive components to magnitude of input current by:

$$|V_0| = |I_0(RQ)| \qquad \textbf{(21)}$$

with:

$$RQ = (215.15 \log K - 160) \qquad \textbf{(22)}$$

Now, consider extending the element in **fig.** 1 by length Δs (of the same radius, RO) at its outer end ($x = s/2$). Here the current is zero so the small pipe increases only the electrostatic energy (capacitive effect). Since in this case eq. **13** is still valid, the total increase in stored energy should be just the same as it was for insertion at x = 0. Therefore:

$$\Delta L(I^2)/2 = \Delta C(V^2)/2 \qquad \textbf{(23)}$$

where AC = the capacitance increase due to As at the element end.
AL = the increase in inductance due to As at the element center.

From eqs. 21 and 23:

$$AC = \Delta L/(RQ)^2 \qquad \textbf{(24)}$$

Using eqs. **22**, **24**, and **19**:

$$\Delta s/\Delta C = (43.03 \log K - 32)(25\pi c/10) \qquad \textbf{(25)}$$

or in conventional units

$$\Delta s/\Delta C = 1/CAP \qquad \textbf{(26)}$$

$$= (43.03 \log K - 32)(2.356 \times 109) \text{ meters/farad}$$

where CAP = the capacitance per unit length.

Note that 1/CAP is directly related to IND, differing only in a constant multiplier.

Thus, we now can think of a cylindrical section of element pipe as contributing to element inductance (**eq. 20**) and element capacitance (**eq. 26**). Each contribution is a function of $K(\lambda/RO)$, and therefore RO, and each will depend on the current or voltage on the pipe section.

Let us now see what happens if a small section of pipe of length $\Delta B/2$ is first removed at a position x (or corresponding θ) and for symmetry also at $-$x or $-\theta$ from the element shown in **fig. 1.** Now replace these removed sections with equal length sections ($\Delta B/2$) of larger radius RO. The overall length of the element remains s, but cylindrical "bumps" occur at X and $-$X. As a result of these bumps the stored energy of the system is changed and therefore the resonant frequency is changed. Designate the value of K for the original pipe as K_1 and for the short bumps as K_2. The contribution of the bump(s) to stored energy, W_2, will be

$$2W_2 = \Delta B\ [IND_2\ (I^2\cos^2\theta) + CAP_2\ (V^2\sin^2\theta)] \qquad \textbf{(27)}$$

The relationship of V at the end of the element to I at x = 0 is essentially unchanged from the original element, that is, $CAP_1V^2 = IND_1I^2$ (see eq. **23**). Note also that (**eqs.** 19 and **25**):

$$CAP_2/CAP_1 = IND_1/IND_2 \qquad \textbf{(28)}$$

so that eq. 27 can be rewritten as

$$2W_2 = AB\ [IND_2(I^2\cos^2\theta) + (IND_1^2/IND_2)(I^2\sin^2\theta)] \qquad \textbf{(29)}$$

Let us now find an equivalent length, $\Delta A/2$, of the original pipe which, when placed at the same positions as each of the bumps, contributes an equal stored energy.

$$2W_1 = AA[IND_1I^2(\cos^2\theta + \sin^2\theta)] = \Delta A IND_1I^2 = 2W_2 \qquad \textbf{(30)}$$

so that

$$\Delta A/AB = (IND_2/IND_1)\cos^2\theta$$
$$+ (IND_1/IND_2)\sin^2 8 \qquad \textbf{(31)}$$

Now, for a longer section (longer bump) going from θ_1 to θ_2, the equivalent length of the original pipe can be easily calculated. Designate $IND_2/IND_1 \equiv m$, the length of the long bump as S_B, and the length of the original pipe, which gives equivalent stored energy, as S_A.

$$S_A/S_B = m\ \overline{\cos^2\theta} + (1/m)\ \overline{\sin 28} \qquad \textbf{(32)}$$

The angular functions are to be averaged over the complete bump section. **Eq. 32** is easily integrated and averaged; the result is

$$S_A/S_B = (m + \frac{1}{m})/2 + (m - \frac{1}{m})F(\theta)/2 \qquad \textbf{(33)}$$

where

$$F(\theta) = (\sin 2\theta_2 - \sin 2\theta_1)/(2\theta_2 - 2\theta_1) \qquad \textbf{(34)}$$

with θ measured in radians.

We can now compute from a given element taper schedule (involving several sections with different pipe diameters) the equivalent lengths of sections of "standard" cylindrical pipe. The procedure is to first choose the "standard" cylinder that is expected to provide equivalent Q. This is, of course, the pipe size at the center of each half element; that is, the average or mean pipe size. Next, for each section of the tapered element, compute the starting θ_1 and ending θ_2. For each section compute m; it is easily derived from **eq. 20,** or

$$m = (43.03 \log K_2 - 32)/(43.03 \log K_1 - 32) \qquad \textbf{(35)}$$

From **eqs. 35** and **33** compute S_A/S_B, which, multiplied by the (tapered) section physical length, gives the equivalent section length of the standard pipe. Adding the lengths of all equivalent sections gives the overall length of the standard cylindrical element that should perform essentially the same as the chosen taper schedule.

Perhaps an example will illustrate the procedure. **Fig. 2** shows schematically a half element with five different sections whose physical diameters range from 1.250 inches (3.25 cm) at the boom ($x = 0$) to 0.500 inch (1.3 cm) at the outer end. Readers will recognize this taper schedule as one in common use (by Wilson) for a 14-MHz Yagi reflector antenna element. The middle pipe section, **7/8** inch (2.2 cm) in diameter, will represent the "standard" pipe. At a frequency of 14.2 MHz, $\lambda_0 = 831.76$ *inches (21.13 meters),* $RO = 0.0005260$, and $K_1 = 1901.17$. **Table** *4* illustrates how to calculate the equivalent cylinder section lengths. For each section column 2 shows the actual physical length, S_B, column 3 shows pipe diameter, column 4 the K value, column 5 the value of m computed from **eq. 35,** column **6** values of θ_1, column 7 values of θ_2, column **8** values of $F(\theta)$ computed from **eq.** 33, and column 9 equivalent section lengths, S_A, also computed by **eq.** 33. Note that the overall actual length of the tapered half element is 215 inches (5.46 meters), whereas the overall length of the equivalent cylindrical standard **7/8** inch (2.2 cm) pipe is only 206.54 inches (5.25 meters). In other words, just due to the taper schedule alone the total (full length) tapered element must be made 16.9 inches (42.9 cm) longer than an equivalent cylinder! This taper correction is surprisingly large; it shows clearly that element length alone is a totally inadequate specification.

The physical reason why the tapered element must be longer than an equivalent cylinder is that the inner (larger) sections have smaller inductance than a standard cylinder and therefore must be made longer; similarly, the outer (smaller) sections have smaller capacitance than the standard cylinder and must also be made longer. The taper correction will be quite

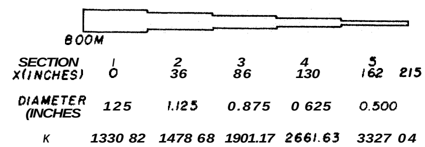

fig. 2. Diagram of a tapered half-element example.

table 4. Equivalent length computations for element in fig. 2.

section	S_B (inches)	d (inches)	K	m	θ_1 degrees	θ_2 degrees	$F(\theta)$	S_A (inches)
1	36	1.250	1330.82	0.93890	0.000	15.070	0.95452	33.904
2	50	1.125	1478.68	0.95695	15.070	36.000	0.61449	48.696
3	44	0.875	1901.17	1.00000	36.000	54.419	−0.00718	44.000
4	32	0.625	2661.63	1.05764	54.419	67.814	−0.52851	31.102
5	53	0.500	3327.04	1.09586	67.814	90.000	−0.90300	48.835
	215							206.537

small if the taper is small, but quite significant if the taper is large.

In the derivation of taper correction calculations, I have assumed that radial "bumps" are treated as small perturbations on the strictly cylindrical case and that the current and voltage distributions are sinusoidal. Note that K values for the heavily tapered element of fig. 2 differ from unity by only a few per cent; thus tthe calculation, even though made by a perturbation method, should be reasonably good. Moreover, the current distribution should still be reasonably sinusoidal over the tapered element. Nevertheless there may be some small inaccuracies in the overall calculation. It is important to note, however, that we are after a length *correction* of only a few per cent due to taper, and therefore some inaccuracy in the computation of the (small)correction is tolerable.

One further point merits elaboration. The procedure just outlined allows only a computation of cylinder equivalents from a given taper schedule; how may we compute a suitable taper schedule starting from a given cylinder? I have found that the simplest procedure is to initially specify all of the taper schedules from mechanical considerations, leaving as a variable only the length of the outermost section. Choose a guessed or estimated length for this section and compute the overall equivalent cylinder. It will generally miss the desired length by a differential length, A. One can now readjust the length of the outermost section by $-Am$ and recalculate. One or two such iterations will bring the tapered element equivalent cylinder length into adequate agreement with the desired figure.

boom clamping correction

I now come to the subject of the boom-to-element mechanical clamping system and its effect on the element reactance and, hence, resonance. It is clear that a wide range of clamping systems are in common use; it is virtually impossible to make valid calculations for all varieties. Nevertheless there are two major kinds and it is helpful to understand them.

The first clamping system is simply to put the element directly through the (round) boom. In this construction a length of element equal to the complete boom diameter is replaced with the boom itself. Since this replacement occurs at a voltage node, we must determine the effective inductance of the replacement; once this is done it can be considered the first section of a tapered element from which an equivalent cylinder length can be calculated. I have not attempted a rigorous calculation of (boom) inductance; instead, I refer to the measurements of Viesbicke[9] in which his **fig. 10** shows that element length due to the presence of a (round) boom should be increased by about 0.7 the diameter of the boom. This is tantamount to saying that the inductance of the boom section of the element is very low compared with normal element inductance; physically this is an expected result. The low inductance, of course, is due to the blockage of magnetic flux by the boom.

The second clamping system is much more widely used since it permits easier element maintenance and replacement. In this sytem either a flat, metal, rectangular plate or an angle bracket is interposed between element and boom; two U-bolts fasten the boom to plate or bracket and two more U-bolts fasten the element to plate or bracket. The U-bolts may also use saddles or cradles, which are mechanically better and which further tend to separate boom and element. For this clamping system we wish to know the inductive effect of the boom itself and more importantly the inductive effect of the plate or bracket. i have found experimentally that for this clamping system the boom itself has remarkably little effect. Even though the (round) boom and (round) element are in physical contact, the element length should be increased by only **6** per cent of the boom diameter; this small correction rapidly disappears as the element is spaced away from the boom (even by a small amount). The reason this result is so different from the through-the-boom result is the relative ease with which the magnetic flux (which results from element current flow) can squeeze between boom and element, especially if there is any gap between them.

The correction in length due to the mounting plate or bracket is readily calculable. The method is to first calculate the equivalent radius of the element plus bracket (which produces the same inductance) and second to use this equivalent radius as the first (short) section of a taper design. The theory for equivalent radii of single and multiple parallel conductors is given by Mushiake and Uda.[10] In their notation the equivalent radius, ζ_ϵ, of a flat thin plate of total width, a, is simply:

$$\zeta_\epsilon = a/4 \tag{36}$$

and that for a right-angled bracket of width a and b is given by a rather complicated expression, which depends only slowly on the ratio b/a. For ratios between 0.3 and 1.0 a good approximation (error <5 per cent) is:

$$\zeta_\epsilon \cong 0.2(a+b) \tag{37}$$

Mushiake and Uda show that for two parallel conductors, it is possible to calculate the equivalent radius of the combination. If a_1 and a_2 are the lengths of the peripheries of the cross sections, ζ_1 and ζ_2 the equivalent radii of the two conductors, d_m the mean distance between them, and ζ_ϵ the equivalent radius of the combination of both conductors, then

$$\log \zeta_\epsilon = (a_1^2 \log \zeta_1 + a_2^2 \log \zeta_2 + 2a_1a_2 \log d_m)/(a_1+a_2)^2 \qquad (38)$$

Eqs. 36 and **38** permit a calculation of the equivalent radius of an element which is proximate to a plate; similarly, **eqs.** 37 and **38** provide a way of calculating the equivalent radius of an element proximate to an angle bracket. To check this method of calculation, I have determined the experimental detuning effect of a plate just touching a 1 inch (2.54 cm) diameter element resonant at **46** MHz. **Table 5** shows both theoretical and experimental results for two different plates. These experiments were not particularly accurate because the resonant frequency is difficult to measure accurately; nevertheless the agreement of theory and experiment within estimated experimental accuracy is gratifying.

Note that element length corrections due to a proximate mounting plate or bracket can easily be as much as 10 percent of plate length. These corrections are not especially large in practice, but should be made wherever there is a relatively large boom-to-element clamping system.

scaling and taper example

It may be helpful to show how to specify a good three-element beam starting with the cylindrical design in **table 1.** I shall go through necessary scaling, then taper schedule calculations for element length(s), and finally apply reasonable boom clamping and boom corrections; this procedure is used to specify a 14.2-MHz beam, a 21.3-MHz beam and a 28.5-MHz beam.

First I choose an average cylinder size that is sufficiently strong. I shall assume that the final element is made of aluminum tubing such as 6061-T6 with seamless 0.059 inch (1.5 mm) wall thickness. For all three bands I choose a cylinder size of 0.875 inch (2.2 cm) OD, although for 28.5 MHz a slightly smaller size is probably permissible. Second, I choose a convenient taper schedule which is easily made from standard 12-foot (3.7-meter) lengths, leaving the length of the outermost section to be adjusted for correct overall length. The sections of seamless tubing (except the last section) are slit back about 3 inches (7.6 cm) at the outer ends (I use one slit only), and a common stainless steel hose clamp fastens sections together. Tubing overlap of about 8 inches (20 cm) gives good joint strength. For 14.2 MHz, the second section is a full 12-foot (3.7-meter) section, over which is slid the shorter first section; this procedure gives added (central) strength and improves the ease of clamping with U-bolts and saddles. For 21.3 and 28.5 MHz this extra inner section is unnecessary.

Table 6 shows the specifications for these tapered half elements where x_1 and x_2 represent the start (inner) and end (outer) positions (in inches) of each section. Note that the tubing requirements for all three elements are shown in 12-foot (3.7-meter) lengths.

table 5. Increase in resonant frequency due to a proximate plate. Element radius is 0.50 inch (1.3 cm) and length produces resonance at 46 MHz.

plate	plate dimensions: length (inches)	plate dimensions: width (inches)	change in resonant frequency: per cent theory	change in resonant frequency: per cent expected
1	4.5	3.625	0.304	0.325 ± 0.1
2	6.0	4.000	0.530	0.521 ± 0.1

table 6. Taper schedules (half elements), for the 3-element Yagi.

section	d (inches)	x_1 (inches)	x_2 (inches)	3-elements tubing lengths (12 feet)
14.2 MHz				
1	1.125	0	24	1
2	1.000	24	72	3
3	0.875	72	136	3
4	0.750	136	176	2
5	0.625	176	< 215	2
21.3 MHz				
1	0.875	0	72	3
2	0.750	72	112	2
3	0.625	112	< 145	2
38.5 MHz				
1	0.875	0	72	3
2	0.750	72	< 105	2

Third, for these three cases it is necessary to scale the original design of **table 1** to use the desired average cylinder size. **Table 7** shows the scaled cylinder lengths (in λ_0 for all three beams using scaling techniques discussed previously.

We are now ready to compute the effect of taper schedule. For the 14.2-MHz element(s), **table 8** shows the flow of calculations; x_1 and x_2 (inches) show the start and finish of each section. First, a trial guess at the overall reflector length is made; I guessed 212 inches (5.38 meters) in this case. For each section K, m, $F(\theta)$ and S_A (in inches) are calculated by the previously described technique. Note that the sum of all cylinder equivalents S_A is 207.63 inches (5.27 meters); what was desired was 207.11 inches

(5.26 meters). This was a lucky guess; however, a small correction should be made to section 5. This correction is m times the needed cylinder correction. Next in **table 8** is shown a second reflector calculation after the correction is made; note that the new cylinder equivalent is exactly what was desired. Thus the overall length of the half element (last x_2) is 211.45 inches (5.37 meters).

By using the correction procedure, the next calculation derives the overall length of the driven element and a small iteration sets it (last x_2) at 208.0 inches (5.28 meters). The same procedure is used for the director, whose overall length (last x_2) is 198.83 inches (5.05 meters). **Table 9** shows exactly the same calculation procedure for the 21.3-MHz beam elements, and **table 10** shows the results for the 28.5-MHz beam elements.

We are now ready for the final small boom and boom clamp corrections. For this purpose I assumed the elements are U-bolted with saddles to flat plates, which in turn are U-bolted with saddles to the boom. Boom diameters are assumed to be 3 inches (7.6 cm) OD (14 MHz) and 2 inches (5.1 cm) OD (21 and 28 MHz). Full plate dimensions are assumed to be 6 inches (15.2 cm) wide and 8 inches (20.3 cm) long (14 MHz); 5 inches (12.7 cm) wide and 6 inches 115.2 cm) long (21 MHz); and 4 inches (10.2 cm) wide 4 inches (10.2 cm) long (28 MHz). These plates reduce central pipe inductance and thus cause an electrical shortening of the half element. This shortening is easy to calculate by techniques previously described. It amounts to about 0.66 inch (1.7 cm) (14 MHz); 0.44 inch (1.1 cm) (21 MHz); and 0.24 inch (0.6 meters) (28 MHz).

table 7. Scaling computations (3-element beam of table 1).

Freq. (MHz)	λ_0 (inches)	d (inches)	K	RO (λ_0)	R	DR	D
14.2	831.76	0.875	190.17	0.0005260	0.49801	0.48963	0.46900
21.3	555.81	0.875	1270.42	0.0007871	0.49790	0.48916	0.46765
28.56	414.42	0.875	947.25	0.001056	0.49769	0.48819	0.46490

table 8. Taper calculations at 14.2 MHz.

	SEC.	d (inches)	x_1 (inches)	x_2 (inches)	K	M	$F(\theta)$	S_A (inches)
R_{TRIAL}	1	1.125	0.	24.	1478.68	0.95695	0.97905	22.990
λ_0 = 831.76 inches	2	1.000	24	72.	1663.51	0.97713	0.74164	47.189
CYLINDER	3	0.875	72.	136.	1901.17	1.00000	0.02854	64.000
RO = 0.0005260 (λ_0)	4	0.750	136.	176.	2218.03	1.02641	−0.66514	39.320
LE = 0.49801 (λ_0)	5	0.625	176.	212.	2661.63	1.05764	−0.95324	34.133
HALF LENGTH 207.11 inches								207.632
R	1	1.125	0.	24.	1478.68	0.95695	0.97894	22.989
	2	1.000	24.	72.	1663.52	0.97713	0.74038	47.190
	3	0.875	72.	136.	1901.17	1.00000	0.02467	64.000
	4	0.750	136.	176.	2218.03	1.02641	−0.66945	39.316
LE = 0.49801 (λ_0)	5	0.625	176.	211.45	2661.63	1.05764	−0.95540	33.609
HALF LENGTH 207.11 inches								207.104
			x_2 $LAST$ = 211.45 inches					
DR	1	1.125	0.	24.	1478.68	0.95695	0.97820	22.990
	2	1.000	24.	72.	1663.52	0.97713	0.73174	47.200
	3	0.875	72.	136.	1901.17	1.00000	−0.00145	64.000
	4	0.750	136.	176.	2218.03	1.02641	−0.69796	39.286
LE = 0.48963 (λ_0)	5	0.625	176.	207.8	2661.63	1.05764	−0.96192	30.135
HALF LENGTH 203.63 inches								203.611
			x_2 $LAST$ = 208.0 inches					
D	1	1.125	0.	24.	1478.68	0.95695	0.97619	22.992
	2	1.000	24.	72.	1663.52	0.97713	0.70843	47.226
	3	0.875	72.	136.	1901.17	1.00000	−0.06997	64.000
	4	0.750	136	176.	2218.03	1.02641	−0.76731	39.214
LE = 0.46900 (λ_0)	5	0.625	176.	198.75	2661.63	1.05764	−0.97859	21.538
HALF LENGTH 195.00 inches								194.97
			x_2 $LAST$ = 198.83 inches					

table 9. Taper calculations at 21.3 MHz.

	SEC.	*d* (inches)	x_1 (inches)	x_2 (inches)	*K*	*M*	$F(\theta)$	S_A (inches)
R_{TRIAL}	1	0.875	0.	72.	1270.42	1.00000	0.61831	72.000
λ_0 = 555.81 inches	2	0.750	72.	112.	1482.16	1.02836	– 0.45812	39.503
CYLINDER	3	0.625	112.	140.	1778.59	1.06191	– 0.93549	26.476
RO = 0.0007871 (A)								137.979
LE = 0.49783 (λ_0)								
HALF LENGTH 138.35 inches								
R	1	0.875	0.	72.	1270.41	1.00000	0.62020	72.000
	2	0.750	72.	112.	1482.16	1.02836	– 0.45319	39.509
LE = 0.49783 (λ_0)	3	0 625	112.	140.4	1778.59	1.06191	–0.93404	26.857
HALF LENGTH 138.35 inches								138.366
			x_2 *LAST* = 140.2 inches					
DR	1	0.875	0.	72.	1270.42	1.00000	0.60720	72.000
	2	0.750	72.	112.	1482.16	1.02836	– 0.48664	39.471
LE = 0.48884 (λ_0)	3	0.625	112.	137.7	1778.50	1.06191	– 0.94368	24.289
HALF LENGTH 135.65 inches								135.760
			x_2 *LAST* = 137.6 inches					
D	1	0.875	0.	72.	1270.42	1.00000	0.57571	72.000
	2	0.750	72.	112.	1482.16	1.02836	–0.56284	39.386
LE = 0.46675 (λ_0)	3	0.625	112.	131.65	1778.59	1.06191	–0.96375	18.547
HALF LENGTH 129.71 inches								129.933
			x_2 *LAST* = 131.4 inches					

table 10. Taper calculations at 28.5 MHz.

	SEC.	d (inches)	x_1 (inches)	x, (inches)	K	M	$F(\theta)$	S_A (inches)
R λ_0 = 414.42 inches	1	0.875	0	72	947.25	1.00000	0.37839	72.000
CYLINDER	2	0.750	72	104	1105.12	1.02988	–0.85138	31.209
RO = 0.001056 (λ_0)								103.209
LE = 0.49769 (λ_0)								
HALF LENGTH 103.13 inches								
DR	1	0.875	0	72	947.25	1.00000	0.35882	72.000
	2	0.750	72	101.89	1105.12	1.02998	–0.86433	29.140
LE = 0.48819 (λ_0)								101.140
HALF LENGTH 101.16 inches								
			x, *LAST* = 101.90 inches					
D	1	0.875	0	72	947.25	1.00000	0.31133	72.000
	2	0.750	72	97.08	1105.12	1.02998	–0.89377	24.429
LE = 0.46490 (λ_0)								96.429
HALF LENGTH 96.33 inches								
			x_2 *LAST* = 96.97 inches					

Thus, to compensate for the boom clamp, each half element should be lengthened by an equivalent amount; it should be further lengthened by the empirical 1/16 boom radius described previously.

With all these corrections, the overall physical length of each half element is shown in table **11.** None of these taper schedules is severe; therefore, the actual element lengths are not a great deal longer than the cylinder lengths shown in tables **8, 9,** and **10;** nevertheless the differences are there. Also shown in table **11** is the boom position x_B for each element expressed both in λ_0 and in inches. Although I have not tested any of these particular three-element beams experimentally, I am confident that their performance will be excellent and, moreover, they all should be easy to construct.

summary

Let me now summarize briefly the results of the entire Yagi design series.

1. A computational methodology was developed and validated[2,3] that allows the important Yagi antenna table 11. Overall element half lengths (in inches) and boom positions (in λ_0 and inches); 3 element beams with taper schedules of tables 8, 9 and 10.

(MHz) freq	element	initial taper (inches)	clamp (inches)	boom (inches)	final length (inches)	boom location, x_B (λ_0)	(inches)
14.2	R	211.45	0.66	0.09	212.20	0	0
	DR	208.00	0.66	0.09	208.75	0.15	124.7
	D	198.83	0.66	0.09	199.58	0.30	249.5
21.3	R	140.2	0.44	0.06	140.70	0	0
	DR	137.6	0.44	0.06	138.10	0.15	83.4
	D	131.4	0.44	0.06	131.90	0.30	166.75
28.5	R	103.91	0.24	0.06	104.21	0	0
	DR	101.90	0.24	0.06	102.30	0.15	62.2
	D	96.97	0.24	0.06	97.27	0.30	124.3

properties to be computed. Such computations produce results which are judged accurate to a few per cent; such an accuracy probably exceeds the accuracy of state-of-the-art experimental techniques.

2. Computations have been made throughout the series which have led to many new insights to Yagi antenna behavior. Among them are:

a. Simplistic designs (all elements spaced equally along the boom, and all directors of equal length) are as good as any other design for the same boom length as long as the boom is shorter than one wavelength.4

b. Yagi forward gain basically depends only on boom length (in λ); it is essentially independent of number of elements as long as element spacing along the boom is not too large.[4] Conceptually, the boom can be considered an aperture illuminated in a quasi-uniform way by the discrete elements. The illumination produces a diffraction pattern (the radiated antenna pattern) whose details are controlled by the precise illumination schedule.

c. Yagi *F/B* ratio is (naturally) best when the diffraction pattern has a null in the back direction. This occurs approximately when the boom length is an odd multiple of $\lambda/4$.

d. A procedure exits whereby a Yagi antenna having four or more elements and roughly favorable boom length can be fine tuned by slight changes in element positions on the boom to give an indefinitely high *F/B* ratio; this astronomical *F/B* (that is, $>$120 dB) exists only at a single frequency. It occurs due to vectorial cancellation of individual element contributions and is equivalent in concept to a notch frequency filter which is carefully adjusted to give an exceptionally deep notch.5

e. Yagis, quads and quagis all behave alike qualitatively. Conceptually a quad can (if properly adjusted) have a somewhat higher gain (a fraction of one dB) than a single Yagi; for horizontal polarization the increased gain comes about from slightly increased vertical directivity. This conceptual advantage may be eroded in practice by the difficulty of experimental quad adjustment compared with the accurate construction of a Yagi to a valid computed design.6

f. The gain and impedance of any equilateral quad loop is *strictly independent* of the position of the feed point.

g. Ground effects are extremely important and lead directly to preferred antenna heights (*1* to 2λ) with corresponding preferred radiation elevation angles.[7]

h. Stacking (horizontally polarized) Yagis vertically over ground is very effective if the top Yagi is sufficiently high (*1* to 3λ). Stacking does result in significant mutual coupling effects, which can degrade normally expected performance, especially *F/B* ratio.[8]

i. A new method is suggested for raising the radiation acceptance angle for stacked beams. This method uses phase reversal for one of two antennas in a stack; the apparent advantage is the retention of stack gain at the higher angles.[8]

j. Fine tuning, or beam optimization, for high *F/B* ratio depends on the ultimate end use. Designs are different for free-space conditions, a single Yagi antenna over ground, and Yagi antennas to be used in a stack.8

3. Practical computation procedures are provided in this article for *scaling* a given design to use elements of different radii, for *length corrections* due to element taper schedule, and for length corrections due to mechanical boom-to-element clamps.

4. The entire series provides a way for anyone to make a Yagi antenna system having high computed performance, starting from his own computed designs, or starting from designs which have been suggested in this series. Moreover, it is also shown in this article how to make a Yagi antenna which will accurately emulate the performance of any existing Yagi design; the performance will be just as good (or just as bad) as the emulated design.

final comments

In the development and exposition of this series of related articles, which I found both technically challenging and requiring considerably more effort than originally anticipated, I have attempted to proceed from basic electromagentic theory to a model of a Yagi antenna system which could ultimately be used in a practical way. All of the required steps and tools have been described. However, along the way I have noticed a number of areas in which further work by interested people could be very helpful. Among these are the foliowing:

1. Valid theoretical treatment of mutual impedance where element length is not $\lambda/2$, and where the current distribution is not sinusoidal but consistent with the element function and environment. A particularly difficult question exists with regard to the imaginary part of this impedance at small distances.

2. Valid theoretical treatment of the screening effect of closely adjacent dipoles on the electric field normally present at a given dipole.

3. Valid theoretical treatment of the mutual coupling between quad loops, especially including the imaginary component of coupling at all loop distances.

4. Valid theoretical treatment of the reactance of a full quad loop as a function of its length (perimeter) in the neighborhood of λ.

None of these tasks is easy. All require good physics followed by tractable mathematics. Moreover, even if "solutions" are claimed, they must be viewed with some suspicion until ***accurate*** experimental results confirm their validity.

In addition to these theoretical tasks, it would be extremely helpful if good experiments could be made in one or more of the following areas:

1. Experiments on model Yagi antennas, similar to those reported by NBS[9], but carried out with improved instrumentation and especially improved control of the physical environment. Such experiments could be exceedingly useful in attempting to validate not only the models I have used, but improved models which I am sure will occur in the future.

2. Find a way to better characterize real (rough, contoured, or both) ground sites. Such characterization should also include the electromagnetic properties of ground. The objective of such work is to provide valid models for a wide spectrum of real-world sites; the use of these models should lead to better understanding of ground effects and perhaps methods for minimizing ground problems.

3. From (flat) ground sites at several magnetic latitudes measure the (statistical) arrival angles of incoming signals. Such measurements should be made at a number of widely separated useful frequencies; at each frequency the results should be correlated with the measured state of the ionosphere. These measurements should be made, not only over a yearly cycle, but over at least one complete solar cycle. Only in this way will a real understanding of the relevant behavior be reached. The end result of this understanding is, of course, to allow specifications for needed incoming arrival angles and hence specifications for optimum antenna height(s) and stacking arrangements.

It is clear that all of these suggestions require an uncommon competence and dedication, as well as the development of sophisticated experimental instrumentation. They also require a great deal of effort.

In the meantime I am convinced that the tools now available will not only permit the design of improved antenna systems, but in many aspects also permit a practical design that is unlikely, even in principle, to be significantly improved.

It is my wish that many readers will construct these superior Yagi antenna systems, make meaningful measurements of their properties, and report results accurately in the literature.

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ham radio

mobile kilowatt for DX

One way to put out a big signal from your car

After several years of **DXing** with a six-element quad, I thought it would be a real challenge to put out a big signal from a mobile rig and see what could be done. It turned out that working DX from a moving automobile is enjoyable and well worth the effort of building the equipment to provide a full kilowatt input.

In my mobile a **TS-120S** drives a modified HA-14 amplifier. The **TS-120S** is powered from the standard 55-ampere automotive system. To power the HA-14 linear, I use a three-phase alternator-powered supply.

high-voltage mobile supply

A three-phase Leece-Neville alternator is used as a primary source, which I bought for $10. It has a rating of 7 volts at 60 amperes. The alternator circuit is shown in fig. 1.

I mounted the alternator on the car and used a belt drive from the crankshaft pulley on the engine. (It takes a tight belt to prevent slippage under maximum load.)

The high-voltage supply (fig. 2) is a three-phase delta configuration with voltage from each phase applied to a full-wave voltage doubler.

The outputs of each voltage doubler are connected in series to obtain 2400-2600 Vdc. I used three surplus transformers with 12-volt primaries and 170-volt secondaries. Other transformers can be used, and if the turns ratio is correct, voltage doublers aren't necessary. Regular 12-volt, **60-Hz** filament transformers with a 220-volt winding can be used.

Applying 25-30 volts to a transformer rated at 12 volts can be alarming but because of the alternator output frequency, the impedance is acceptable, and the transformers will work well without any heating.

Although the alternator is rated at 7 volts output, the output voltage is dependent on the regulator. The regulator (fig. 3) sets the alternator output at 25-30 volts. The alternator works quite efficiently at the elevated output voltage. I've had no problems while running it this way. The field current must be taken into account, however, and the 5.6-ohm resistor (fig. 1) limits it to a safe value. I've test-loaded this power system at approximately 2400-2600 watts with no problems.

regulator

The regulator is a modified version of a circuit published several years ago. No battery is needed in this power system. Several regulator designs were tried and worked well; this is the one I like best. The alternator will usually self-excite when turned on, but if not a momentary push button switch will do it (S2, fig. 1).

This power system has been trouble-free and very dependable. Since the high power drain doesn't affect the automotive power system or its battery, rundown battery problems don't exist. I can run full power with this mobile setup for hours on end with no overheating or other problems. The limitation, of course, is that the engine must run at idle rpm or more to operate the linear.

installation

I mounted the power supply and linear amplifier in the car trunk and the regulator under the hood away from engine heat. If the antenna is bumper mounted, it ***must*** be well grounded. While transmitting with this high power, allow no one to touch the antenna — severe burns will result. Even the outside of the car can give rf burns.

While pulling into my drive one night I was surprised to see what I thought was lightning on a clear night.

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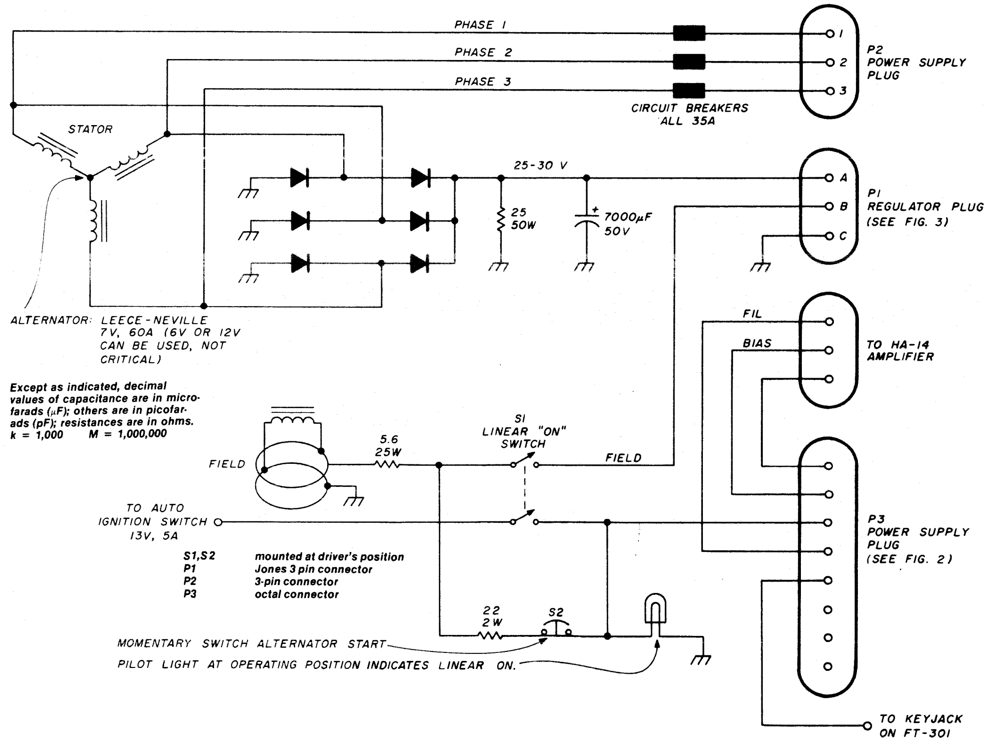

fig. 1. Primary power source for the mobile linear amplifier uses a three-phase Leece-Neville alternator. Voltage from each phase is applied to a **fullwave** voltage doubler.

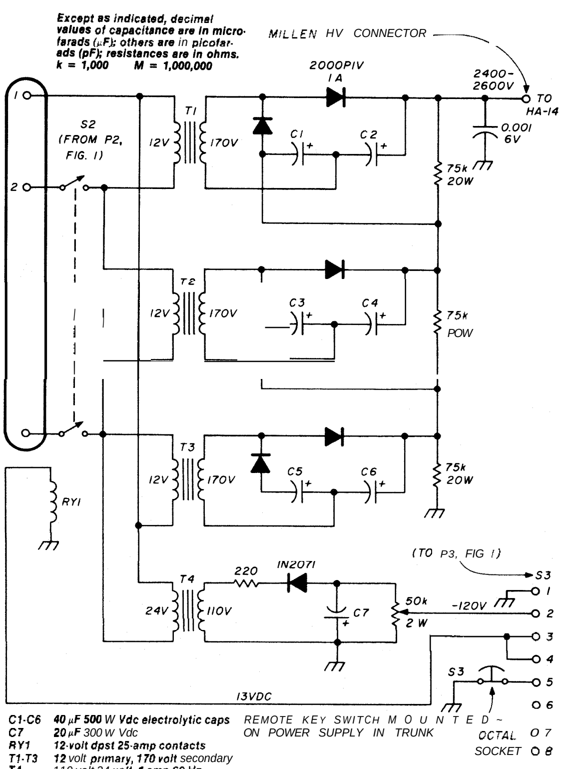

fig. **2.** High-voltage supply. Outputs of each voltage doubler are connected in series to provide **2400-2600** Vdc for the mobile final amplifier. The system has been test loaded at **2400-2600** watts.

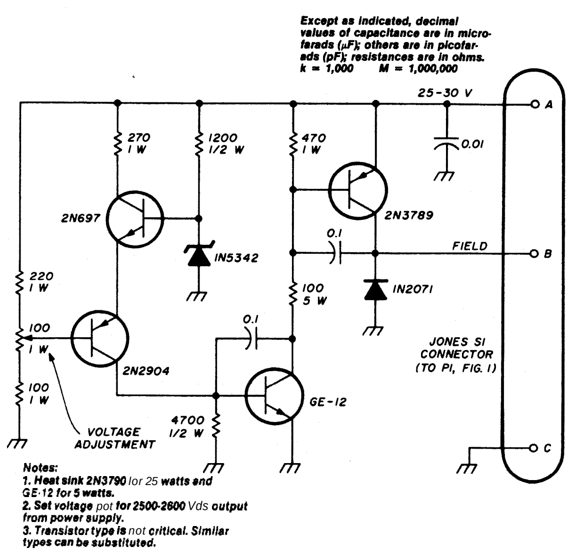

fig. 3. Power-supply regulator, which sets alternator output at **25-30** volts. The 5.6-ohm resistor in series with the alternator field (fig. **1**) limits field current to a safe value.

The top of the antenna was touching low tree limbs as I transmitted, and the damp limbs drew arcs from the antenna, with one of the limbs smoldering and on ***fire.*** I've since learned to shut down when under trees with low limbs.

results

After everything is in place and working, what kind of results can be expected from a kW in the car?

DX stations such as D4, ZS, EL2, 6W8, XT2, H44, VR8, and TR8 have been worked with 5-9 or better reports from the 20-meter mobile. I've enjoyed many contacts with DX friends such as ZS6DN, F3EG, and VR3AR while driving to and from work. In my case, that's a 35-minute trip on the interstate usually with light traffic. Just right for a little mobile DXing.

During peak band conditons, reports are routinely received from both coasts of 30-40 dB over S9 and occasionally "pegging the S meter." Numerous comments such as, "You're too strong to be a mobile," have occurred. I usually honk the horn to convince the doubters.*

Other bands are worked also, and, what with the excellent conditions during the fall of 1979, the 10- and 15-meter band propagation was so good that the mobile was just as good as a fixed station. Many DX stations were worked on first call on these bands in pile-ups during this time. During the winter months, 75 meter DX is worked routinely into most areas of the world. I use CW from the mobile also. A memory keyer is a great help.

The biggest limitation to DX work from a mobile is the ability to receive. On today's crowded bands, with the nondirectional vertical, interference is a problem, as is noise while operating mobile in populated areas. Noise blankers help a great deal. The most common problem with the mobile occurs when a CQ is called. The average ham expects a mobile not to be too strong, and when he hears one calling CQ and answers him, he finds it hard to believe that the mobile can't copy his signal on a simple antenna.

I've enjoyed this mobile for about 1½ years and can recommend mobile DXing as another means of enjoying ham radio. For a mobile station to be able to jump into a huge pileup on a rare station on 20 meters and come up with a contact is something that apparently never ceases to amaze the Big Guns at their multikilowatt stations with huge antennas scraping the clouds.

I'll be glad to help in planning your super mobile DX station on the receipt of a large, self-addressed stamped envelope.

ham radio

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*Using Morse code, of course. Editor.

More Details? CHECK—OFF Page 126

amplitude compandored sideband

Narrowband techniques for vhf mobile communications

It is obvious to most observers in larger metropolitan areas (New York, Los Angeles, Chicago, San Francisco) that saturation is beginning to occur on the 2-meter band. Even with the extra megahertz provided by the added repeater sub-band, with a total possible repeater population of 60 or so machines above 146 MHz, and 20 or so in the 144-145 MHz region, there are times when a ham population of 10,000 or more in such regions taxes these systems to their limit. Timers of 60, 40, or even 30 seconds are not really the answer.

If hams have been experiencing a problem, consider the plight of commercial users of vhfluhf. It has been impossible for some time to obtain vhf licenses in many areas, and uhf channels are in short supply as well. Common carrier multiplexing schemes and/or 900-MHz channels have been proposed, but individual vhfluhf or semi-shared channels have many advantages to the ultimate user, not the least of which is long-term cost.

Sideband use on vhf has long been used by Radio Amateurs (and the military). With the recent introduction of multi-mode 2-meter rigs, a surge in interest and activity has been sparked using this mode. Below fm threshold, SSB provides distinct advantages in sensitivity and range capabilities. Unfortunately, many of the convenience featues of fm operation do not work with our current sideband transceivers, and the signal-to-noise ratio on stronger signals, as well as the audio bandwidth and quality, do not match the better fm rigs.

amplitude compandored sideband

Recent developments promise to change the situation. In his report to the FCC after an extensive two-year research program into narrowband techniques for vhf land mobile,[1] Dr. Bruce Lusignan of Stanford University's Satellite Planning Center has come to some very interesting conclusions. By modulating a standard single-sideband transceiver with specially processed audio and processing the recovered audio through a similar system on the receive end, equal or even better performance can be obtained than when using NBFM. Because less than one-fifth the spectrum is required for equivalent channel-to-channel protection, five times as many stations can occupy the same spectrum space.

ACSB, or amplitude compandored sideband, combines several common techniques especially tailored for SSB. The system, developed by Dr. Lusignan in conjunction with Dr. Fred Cleveland of the University of the Pacific and VBC, Incorporated, features 4:1 amplitude compandoring, a pilot subcarrier system, and 12-dB/octave pre-emphasislde-emphasis. The resultant ACSB system provides:

1. 50-70 dB adjacent channel protection using 5-kHz channels (as opposed to 20-25 kHz spacing for fm).

2. 10-dB power advantage due to both processing and bandwidth.

3. Automatic frequency locking and carrier identification.

4. Very rapid AGC (20 Hz) to greatly reduce mobile flutter.

5. A degree of quieting performance that, combined with its greater sensitivity, equals or exceeds normal fm.

6. Extended, reliable range by a factor of two, up to

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about 25 miles (40 km), limited to a factor of 1.5 times only by earth curvature beyond this distance.

Furthermore, noise during fading is much less distracting (and less tiring as a result). This feature is the result of compandor characteristics, which reduce both noise and signal at poor signal-to-noise ratios rather than producing the noise bursts common to fm. Unlike normal sideband, ACSB provides a 5-dB capture effect that is several dB better than fm's normal 6-8 dB capability.

description of a typical system

The microphone audio is first passed through a preamplifier to bring it up to the proper level for the compressor circuitry. It is then passed through the first of two 2:1 compressors so that the normally desired 40-dB dynamic range of speech is compressed into 20 dB.

This compressed audio is then mixed with a 2850-Hz pilot tone set – 7 dB below peak audio output. Both signals are then passed through a second 2:1 compressor, which compresses the 20-dB dynamic range of the first compressor into a 10-dB dynamic range. As one might expect, the pilot tone will be reduced during voice peaks by the amount of gain reduction produced by the audio peaks. This works out to about 10-dB reduction of the pilot on voice peaks, or 17 dB below peak reference level. Obviously, if we were to monitor this signal at this point, very compressed audio with a high pitched tone would be heard.

The final processing technique is to pre-emphasize the speech at a rate of 12 dB per octave. This is done to equalize the inherent differences in power levels in human speech, which tends to be concentrated in the low frequency areas.

The processed signal is transmitted on an otherwise standard single-sideband transmitter. It is also received on a standard single-sideband receiver using its normal AGC techniques (perhaps modified slightly to complement ASCB characteristics).

The received signal is passed through an AGC-controlled audio stage, which is controlled by a detector tuned to the pilot tone frequency. Its time constant is set very fast so that up to a 20 Hz-per-second change in input signal will be kept nearly constant in output level. Additionally, as a reduction in pilot level will cause an ***increase*** in output level, the suppression of the pilot in the second transmit amplitude compandor will be translated by the pilot AGC system into expansion at the same rate. Thus, a strong signal with no modulation will be quieted by the presence of the pilot signal. As the pilot is at its peak when there is no modulation, maximum quieting will occur.

After processing by the pilot-derived AGC, the leveled, expanded signal is again passed through a 2:1 expander. The pilot-derived expansion restores the 20-dB dynamic range from the transmitted 10-dB dynamic range signal. The second expander restores the original 40 dB dynamic range from the 20-dB pilot-derived expansion. The resulting audio is then processed through a 12-dB per octave de-emphasis filter to restore the original frequency response.

ACSB, then compresses 40 dB of speech information into a dynamic range of 10-dB, transmits it, then restores the 40-dB dynamic range at the receiving end of the system. This means that a signal-to-noise ratio of just over 10-dB is all that is required for an effective restored dynamic range (signal dynamics ***and*** signal-to-noise) of 40 dB. Additionally, a 2:1 quieting curve is established due to the constant presence of the pilot tone in the AGC and pilot expander receiver circuits.

A close look at the dynamics of ACSB will show that a carrier-to-noise ratio of only 5 dB will provide the equivalent of 20 dB quieting (to use fm terminology). Indeed, a 10-dB carrier-to-noise will give almost the full 40 dB dynamics and signal-to-noise we started with, except for the addition of a few dB of noise due to proximity to the noise floor. This certainly explains the weak signal superiority of this mode.

Dr. Lusignan estimates that ACSB has about a 15-dB advantage over normal SSB (he assumes 20-22 dB signal-to-noise is required for "high intelligibility" . . . all consonants audible). It also has a bandwidth advantage over fm, giving less interference from impulse noise (ignoring fm limiting) and higher signal-to-noise for a given power level at the receiver. This combination provides the measured 10-dB advantage of ACSB over 5-kHz deviation fm.

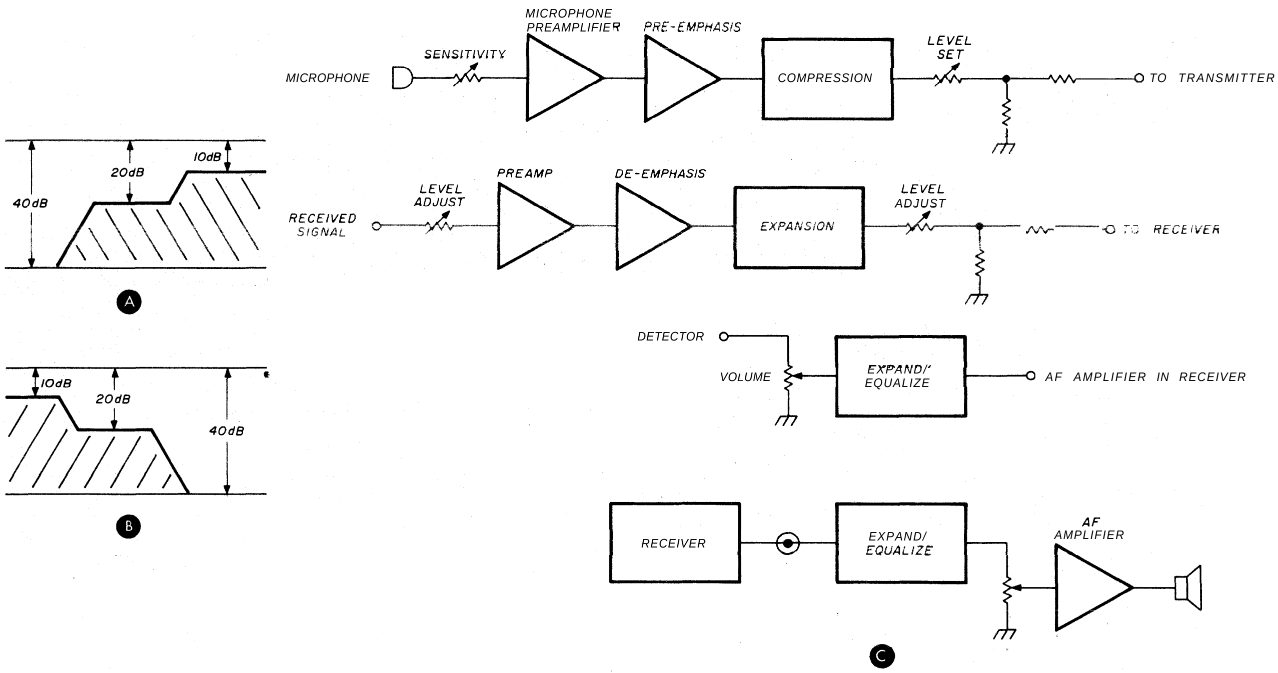
ACSB and NBFM comparison

Dr. Lusignan's report to the FCC[1] compares ACSB with NBFM as follows:

Signal to noise. ACSB shows a 10-dB advantage over fm at equal peak power levels.

Power required. ACSB requires 1/10th the power of fm for equal signal to noise. Additionally, ACSB requires 1/3 to 1/2 the average power of fm when the transmitters have equal peak output power, since the unmodulated output of ACSB is 7 dB less than its peak output.

Range. ACSB provides a reliable range equal to twice the fm range at distances up to 25 miles (40 km). Beyond 25 miles (40 km) this is reduced to 1.5 times due to the earth's curvature, which then becomes the limiting factor.



Characteristics of the Amplitude Compandored Sideband (ACSB) **system proposed by Dr. Bruce Lusignan of Stanford University's Satellite Planning Center. A and *B* show respectively the 4:1 and 1:4compression and expansion levels of the system. The block diagram, C, is a suggested starting point for experimenters for an** ACSB **Level 1 system.**

During 8-watt PEP tests, simultaneously transmitting ACSB and fm combined on a common transmitting antenna and receiving on an ACSB and fm receiver fed from a common receiving antenna, the fm signal was lost in the south San Jose, California, area, while the ACSB signal was lost near Gilroy, California – some 16 miles (26 km) and 35 miles (56 km) from the Stanford transmitting site respectively.

Fading/multipath noise bursts **(kerchunking).** Field tests and bench tests show ACSB burst noise is 10 dB less than fm burst noise. Additionally, ACSB should be less prone to multipath distortions due to its narrower bandwidth and lack of sensitivity to phase relationships.

Message completion. ACSB is 3-5 times more reliable at a 9-mile (15-km) range than fm at equal power levels. ACSB at this range gives an 85 per cent completion rate compared with fm's 20 per cent rate.

Co-channel protection. On-channel rejection is 2-3 dB better with ACSB than with fm. Capture ratio for ACSB is about 5 dB compared to 7-8 dB for fm.

Adjacent-channel rejection. At 5-kHz spacings, ACSB provides 50-70 dB rejection of adjacent channel interference (depending on linearity and frequency stability). Fm at 25-kHz channel spacing yields 65-75 dB; at 20-kHz spacing it yields 55-65 dB.

According to the report,' the protection of 50 dB is sufficient, because other factors (intermodulation, co-channel interference) become equally problematical beyond this point.

"In typical applications the probability of loss from adjacent channel transmissions compared with 50 dB isolation is negligible compared with . . . shadowing or co-channel transmissions. Increasing . . . from 50-70 dB would not result in a noticeable change in the probability of successful transmissions."

Stability requirements. The ACSB system developed by VBC, Incorporated, for this study will automatically lock signals that are ± 800 Hz from the center of the channel. At 160 MHz this is not outside normal stability for current fm equipment.

Digital transmissions. ACSB can handle up to 4 Kb/second in the main 2-kHz audio channel as well as about 20 b/second superimposed on the pilot carrier.*

Doppler shift in mobile service. The AFC circuit will control Doppler shifts normally encountered at all frequencies through 900 MHz (± 800 Hz).

Fm/ACSB shared channels. It is possible to use ACSB and fm from a common repeater site providing the two channels are separated by 12.5 kHz. That is,

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'Experiments with wider audio bandwidths (up to 3 kHz) are in progress. This should increase digital rates as well as improve audio fidelity.

an fm repeater could also provide two ACSB channels each 3 kHz wide centered 15 kHz away without interference from the ACSB channels to the main channel. (This might be a solution to the 15-kHz split situation on 2 meters between 146-148 MHz, for example.)

hardware

Commercially available LSI chips that perform all ACSB functions should be available in one to two years, depending on FCC action, market acceptance, and other normal factors relating to volume and production. in the meantime, experiments with ACSB Level 1 is within easy reach of the experimentally inclined ham. The Signetics NE 5701571 Compandor IC is available from Jameco Electronics, 1021 Howard Ave., San Carlos, California 94070. Their price is $4.95 (1980 catalog), but they also have a $10.00 minimum.

The NE570, an LM324 op amp, and an rf-tight box will allow everything necessary for 2:1 compandoring with pre-emphasis/de-emphasis. My own experimentation shows a marked improvement on all but the weakest signals (signals under 4-5 dB signal-to-noise ratio show no apparent improvement, even though background noise with no signal *will* be improved). The block diagram on the preceding page is recommended as a starting point.

conclusion

Out here in the west we like to talk about the wide open spaces. Well, you can still drive to those wide open spaces without too much effort. In the crowded city, however (and we ***do*** have some crowded cities), one soon learns that it is best to give one's neighbor plenty of elbow room whenever possible. On vhf, ACSB promises a good way to do just that.

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ham radio

More Details? CHECK–OFF Page 126

first building blocks for microwave systems

Simple and stable 1152-MHz multiplier chain for Amateur microwave bands

There is an apparent abundance of commercially built high-frequency and vhf equipment available, little of which is adaptable for use above 1 GHz. Purchased equipment may be used to provide a 1296-MHz station (generally a varactor tripler driven by a 432-MHz transmitter, and a relatively high-noise-figure receiving converter with no rf preamplification). It's virtually impossible to purchase any station equipment specifically designed for weak-signal communications above 1296 MHz. Thus far only the most intrepid experimenters have ventured above 1296 MHz, generally hand-in-hand with a master machinist and expensive power tools (lathes and the like).

All is not lost, however. Because of two interesting factors, building a microwave station is now possible for most experimenters willing to spend a few evenings etching PC boards and soldering components. That's right — no more machinists, at least not for 1296-MHz and 2304-MHz equipment.

frequency relationships

The first factor to help resolve the microwave dilemma lies in the arithmetic of our microwave bands. Within all our bands above 1300 MHz is at least one frequency that is a multiple of that "magic number" — 1152 MHz. Even 1296 MHz is related to 1152 MHz. The former frequency was originally selected for weak-signal work because it is the third harmonic of 432 MHz and therefore can be obtained by tripling. A difference frequency of 144 MHz exists between 1296 and 1152 MHz, which becomes the receiving i-f. Note also that 1152 MHz is the eighth harmonic of 144 MHz. The relationships between the 1152-MHz magic number and weak-signal frequencies in our uhf and microwave bands are listed in table 1 and graphically illustrated in **fig. 1.**

low-order frequency multiplication

Another interesting mathematical feature is that the frequency of 1152 MHz can itself be generated by a chain of low-order (and therefore relatively good efficiency) multipliers. This chain, made up only of frequency doublers and/or frequency triplers, allows filtering to reduce undesired (spurious) signals at the multiplier-chain output. Many writers have insisted that starting frequencies be in the range of about 50-100 MHz to avoid producing undesired harmonics in the 144-MHz and/or 432-MHz bands. This requires an overtone crystal. As the frequency of such crystals is notoriously difficult to pull, a variable-crystal-frequency source was developed that allows use of

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crystals operating in the fundamental mode, below about 20 MHz.

One optimum chain (shown by the heavy-bordered boxes in **fig. 2**) thus starts at 16 MHz, triples to 48 MHz, doubles to 96 MHz, doubles a second time to 192 MHz, doubles a third time to **384** MHz, then triples to 1152 MHz. The use of this chain requires that the unwanted third harmonic of 48 MHz be very greatly attenuated. If present, the third harmonic will fall into the low end of the 2-meter band (at the weak-signal EME portion around 144.000 MHz). Radiation of any significant amount of energy at that frequency will tend to irritate neighboring 2-meter CW operators. In a vhf-contest environment, the third or ninth harmonics may very well QRM your own 2-meter or 70-cm station. These two undesired harmonics, however, appear to be the only problem harmonics. The ability to suppress undesired harmonics is enhanced by proper partitioning of the multiplier chain. The basic-frequency (for example 16-MHz) oscillator and only a few of the total number of multipli-

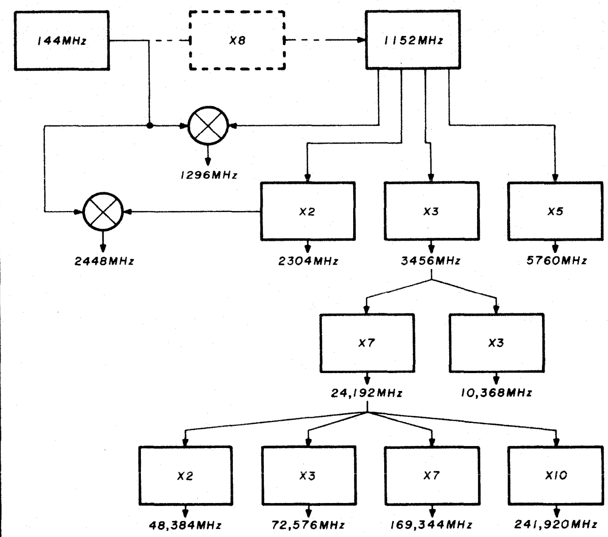


fig. 1. Relationship between 1152 MHz and desirable frequencies for highly stable, weak signals in the Amateur bands above 1 GHz.

table 1. Relationship between "magic number" 1152 MHz and weak-signal frequencies in the Amateur uhf and microwave bands.

band (MHz)	desirable frequency (MHz)	by mixer	by multiplier
1240-1300	1296	1152 + 144	432 x 3 or 108 x 2 x 2 x 3
2300-2450	2304 2448	(1152 x 2) + 144	1152 x 2 or 102 x 2 x 2 x 3 x 2
3300-3500	3456		1152 x 3
5650-5925	5760		1152 × 5
10,000-10,500	10,368		1152 × 9 = 1152 × 3 × 3 = 3456 × 3
24,000-24,250	24,192		1152 × 21 = 1152 × 3 × 7 = 3456 × 7
48,000-50,000	48,384		1152 × 42 = 1152 × 3 × 7 × 2 = 3456 × 7 × 2 = 24,192 × 2
71,000-76,000	72,576		1152 × 63 = 1152 × 3 × 7 × 3 = 3456 × 21 = 3456 × 7 × 3 = 10,368 × 7 = 24,192 × 3
165,000-170,000	169,344		1152 × 147 = 1152 × 3 × 7 × 7 = 3456 × 49 = 24,192 × 7
240,000-250,000	241,920		1152 x 210 = 3456 x 70 = 48,384 × 5

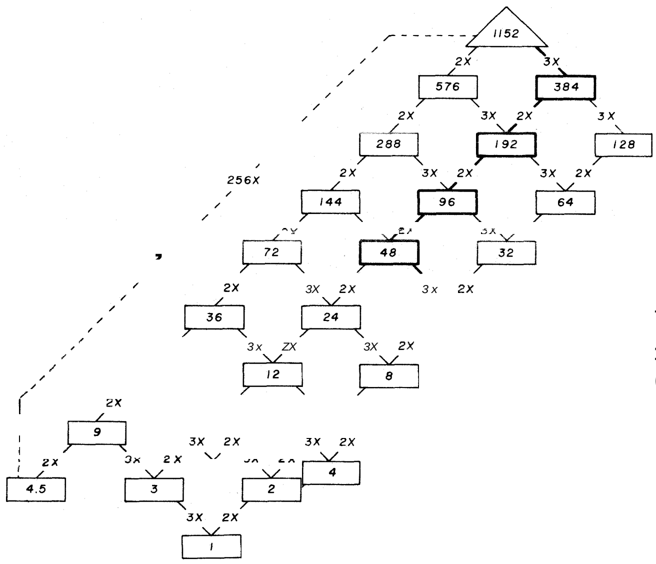


fig. 2. Frequency-multiplication tree for generating a 1152-MHz signal using only low-order multiplication (doublers and triplers) from a low-frequency source.

ers are packaged in a low-frequency building block. The remainder of the multipliers are packaged in a separate, second building block. The low-frequency block output may then be made to have very low levels of signals at undesired frequencies.

The second important factor is the present-day ability to generate the desired 1152 MHz signal in a practical manner from a lower frequency driving signal. In this regard, great thanks should be given to Paul Shuch, N6TX, for his design of a PC board 96-1152 MHz multiplier unit.[1] This microstrip unit, for which a printed circuit board and set of tuning capacitors are available from N6TX, was apparently designed to replace a multiplier chain2 using a packaged oscillator, at 96 MHz, driving a pair of 2N5179 transistor frequency doublers to 384 MHz; a pair of 2N3866 power amplifiers, providing several hundred milliwatts at 384 MHz;[3] and a step-recovery-diode tripler to provide about 5 milliwatts at 1152 MHz.[4] Having built three such frequency-multiplier chains, I must concur with the general undesirability of vhf multipliers using step-recovery diodes.

The replacement of the entire 96-1152 MHz chain with three stages of transistor multipliers (using the Motorola MRF 901) results in a great saving of time, labor, and parts cost. I've built several of the 1152-MHz sources (described later in this article) as well as a 1296-MHz solid-state transmitter, based on the microstrip multiplier of reference 1, and have selected that basic design for the 96-1152 MHz portion of this common microwave system. While some may desire to be purists and design ***all*** their equipment themselves, i believe that judicious use of the contributions of others often makes for the best (and the most rapid) attainment of the end goal: to get as many stations on the microwave bands as quickly and inexpensively as possible.

96-MHz VXS

As mentioned, the N6TX unit was designed for use with a fifth-overtone oscillator, which is replaced with the variable-crystal-frequency source (VXS) shown in the block diagram of **fig.** 3. I've arbitrarily chosen a tuning range, at 2304 MHz, of 2303.928-2304.086

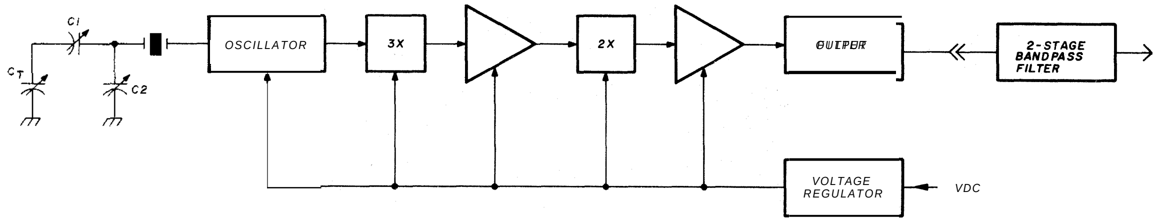
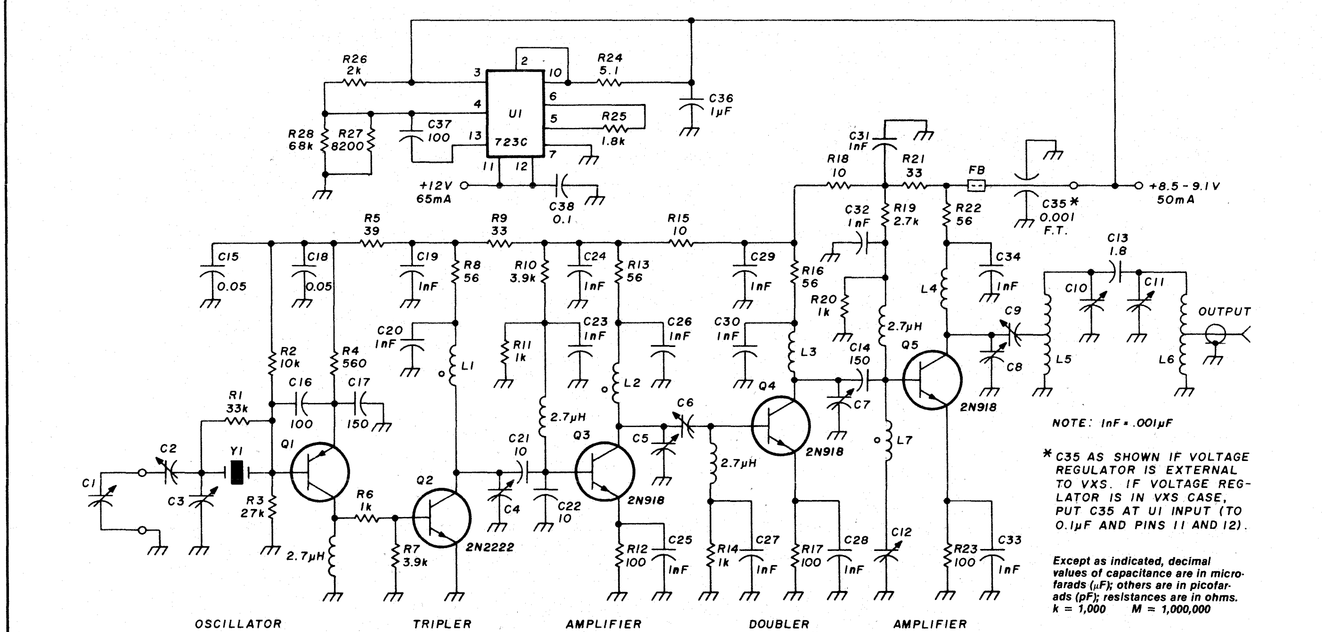


fig. 3. Block diagram of the VXS/BPF system. Output is at 96 MHz using a 16-MHz overtone crystal oscillator.



C1	1-10 pF tuning	L5,L6	8 turns No. 20 (0.8 mm) 1/4" (6.4 mm) 5/8" (16 mm) long
C2	10-60 pF trimmer	L7	8 turns No. 28 (0.3 mm) on T37-10 core
C3	5-50 pF trimmer	Q1	2N2907, 2N3906 PNP
C4,5,6,7, 8,9,10,	5.25 pF trimmer	Q2	2N222, 2N918, 2N2857 or 2N5179 NPN
		Q3,Q4,Q5	2N918, 2N2857 or 2N5179
		U1	LM723 (DIP)
L1,L2	6 turns No. 28 (0.3 mm) on T37-10 core	Y	16,001 kHz Crystal (HC-6U holder)
L3,L4	4 turns No. 2210.6 mm) on T37-12 core		

fig. 4. Schematic of the 96-MHz variable-crystal-frequency (VXS) source. Circuit is built on a single-sided PC board.

MHz, corresponding to an oscillator frequency range of 15.9995-16.0006 MHz (therefore, an 1100-Hz range at 16 MHz gives a 158.4-kHz range, when multiplied 144 times, to frequencies around 2304 MHz). This requires that the crystal frequency be pulled about 0.0066 per cent, which certainly can be achieved with almost any fundamental crystal.

The crystal frequency was chosen as 16,001 kHz with 20 pF parallel capacitance, and thus is slightly higher than the nominal 16,000-kHz frequency. By paralleling the crystal with a bit more capacitance, provided by the main tuning capacitor C1 and its series and shunt band-setting capacitors C2 and C3, the desired frequency range can be realized. The schematic of the 96-MHz variable-crystal-frequency source is shown in fig. 4, the PC-board layout is shown in fig. 6, and the parts placement in fig. 7.

PNP transistor Q1 is the crystal-controlled oscillator, driving a frequency tripler, Q2. Transistor Q3 is a 48-MHz buffer, Doubler Q4 and a tuned buffer, Q5, at 96 MHz, follow. The 96-MHz output filter is a double-tuned bandpass configuration. An additional double-tuned bandpass filter (fig. 5) may be used at the high-frequency multiplier block or placed in a separate shielded box outboard of the source and multiplier blocks to provide an additional 20 dB suppression of the undesired signals provided by the VXS. The tuning range, with the components listed, is sufficient to allow the VXS to be used with crystals between 15-18 MHz. In the first case (15-MHz crystal) the final multiplier output is 1080 MHz, which is used for doubling to 2160 MHz. This frequency is used for local-oscillator output in 2304-MHz receiver converters with a 144-MHz i-f. The 18-MHz crystal produces a final multiplier output of 1296 MHz for use in exciters in the 23-cm band.

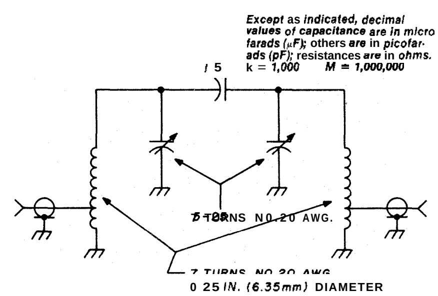


fig. 5. An auxiliary outboard bandpass filter that may be used at the high-frequency multiplier block to provide an additional 20 dB suppression of unwanted signals from the VXS.

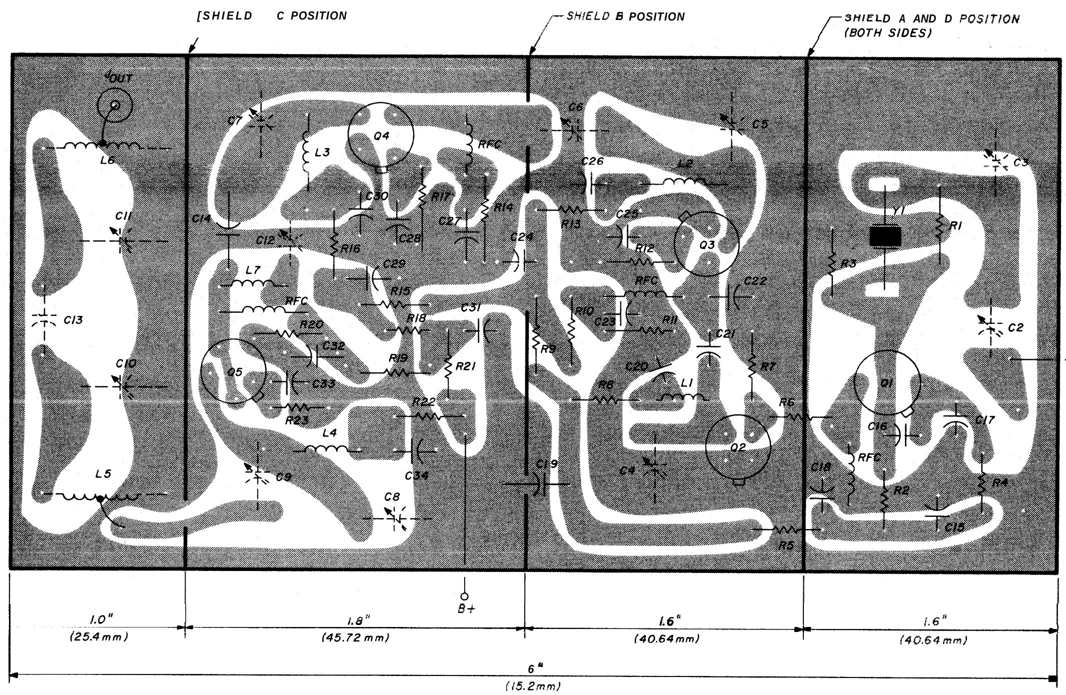


fig. 6. PC-board layout for the VXS-96 microwave signal source.

fig. 7. Component side of the VXS-96 board (copper side). Mount all variable caps on this side; all other components are mounted on the reverse side.

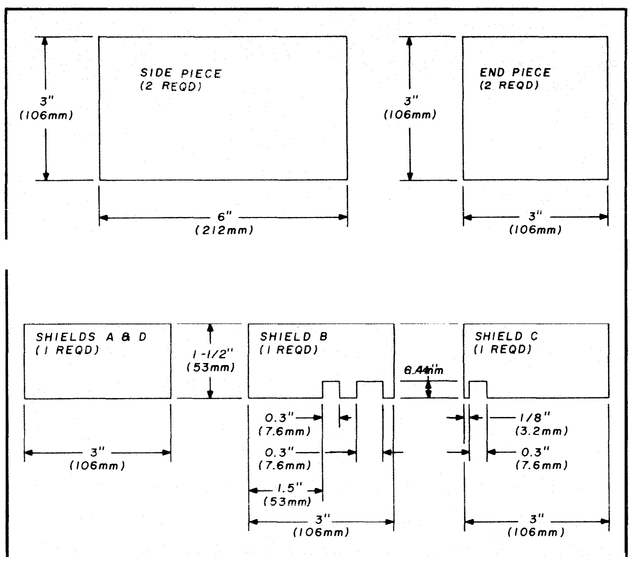


fig. 8. Shield, side, and end pieces for the VXS constructed from double-clad PC-board stock. Covers (top and bottom) are 3¼ × 6¼ inch (8.25 × 15.9 cm) PC-board pieces.

some other uses

The output of the VXS can be:

1. Set to 116 MHz (by using a 19.334-MHz crystal) for use as a 2-meter local oscillator.

2. Used with a frequency doubler to generate a 192-MHz signal for use as a 220-MHz local oscillator.

3. Set by a 16.834-MHz crystal to provide a 101-MHz signal for input to a cascaded pair of frequency doublers to generate a 404-MHz local-oscillator signal for use in 70-cm equipment. (See **fig. 9.**)

In the VXS schematic of **fig. 4,** both crystal leads are above ground in the circuit. This might be a problem if crystal switching is desired. For higher stability the crystal will be placed in a thermally isolated environment (such as a crystal oven positioned above the PC board or in a block of styrofoam).

shielding considerations

Note, in **fig. 8,** that pieces of double-clad PC board form three shield partitions, A, B, and C, directly soldered to the copper-clad side of the PC board. A similar partition, D, is soldered to a PC-board case built around the entire board above shield A (between the oscillator and the multiplier stages) for added attenuation of oscillator harmonics. The oscillator is enclosed in a shielded compartment separated from the tripler-buffer area, which is separated from the doubler-buffer area. The output filter is in its own compartment, shielded from all oscillator, frequency multiplier, and buffer stages.

The VXS circuit also includes a high degree of power-supply decoupling. An IC voltage regulator, U1, provides a constant voltage to the circuit; this is necessary not only to prevent oscillator frequency changes with varied input voltage (in my case, from the battery in my automobile during mobile operation from any convenient mountaintop), but also to keep all transistors operating at fixed biased points, which causes the transistor input and output impedances to be stabilized. This stabilization of device impedances prevents changes in tuning with changing input voltage and contributes to the overall spectral purity of the VXS output signal. Note the use of a BNC connector for the rf output of the source, and the use of a feedthrough capacitor to bring the voltage into the VXS enclosure. Both components are used to maintain the shielding integrity and provide minimum amplitude of undesired signals.

Also note that the voltage regulator IC, U1, and the associated resistors, R24-R28, and capacitors C36-C38 are mounted on a wire-wrap 18-pin IC socket, with the end pins on either side extending full length and soldered to the inside of the case. The remaining 14 pins are bent at right angles, close to the bottom of the socket; the regulator-circuit resistors and capacitors are soldered between the bent pins. See **fig. 10.**

spectrum analysis

The VXS is aligned by using any of the well-known

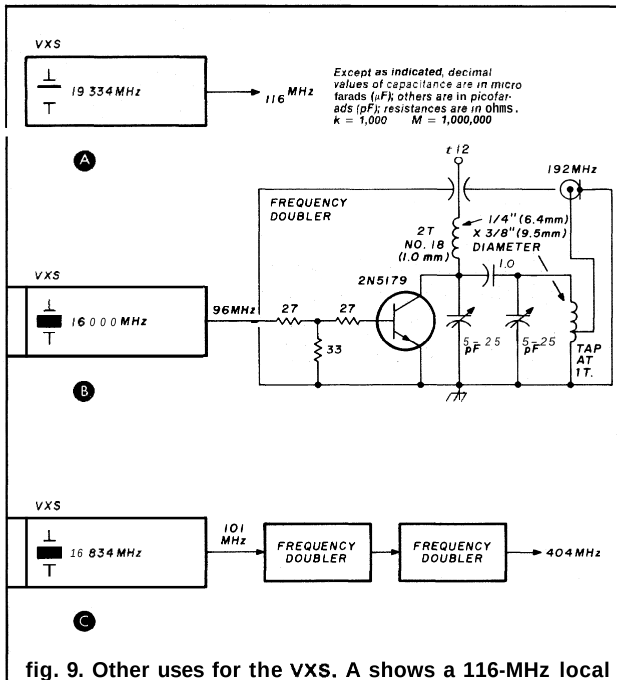


fig. 9. Other uses for the VXS. A shows a 116-MHz local oscillator for a 2-meter converter with a 28-MHz i-f. A frequency doubler for generating a 192-MHz local-oscillator signal for a 220-MHz converter, using a 28-MHz i-f, is shown in B. A pair of cascaded frequency doublers, C, generate a 404-MHz local-oscillator signal for a 432-MHz converter with a 28-MHz i-f.

tuning procedures including: a) monitoring the emitter or collector current of the stage following the stage you're tuning for an increase in current, and b) using a test receiver, grid-dip meter and so forth. If a spectrum analyzer is available (and its use is highly desirable aithough not mandatory) an output signai spectrum similar to that shown in fig. 11 may be obtained. In fig. 11, spectrum (A) is for the basic circuit, built on the circuit board, but without the output filter (L5, L6, C10, C11, and C13). Note that the second harmonic is at a level of only – 14 dBc (dB below the desired carrier, at 96 MHz). Adding the output filter, but without shielding, typically provides the (B) spectrum, wherein the greatest-amplitude undesired signal is still the second harmonic, now suppressed to a level of – 40 dBc. Adding the shields and a shielded box (fig. 8) results in the (C) spectrum (shown in solid lines in fig. 11). With the shields and shield box, the greatest-amplitude undesired signals are those spaced above and below the desired signal by the fundamental frequency; for example, at 80 and 112 MHz.

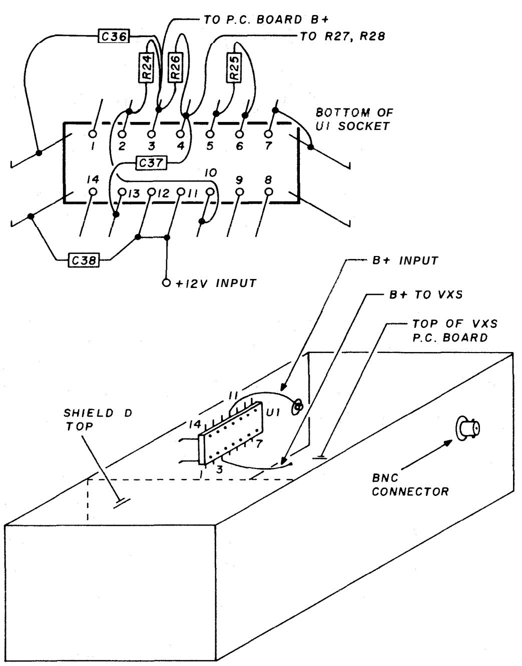


fig. 10. Method of mounting the U1 (18-pin) socket on the VXS case wall with four end pins used for support. The remaining 14 pins are bent outward and hold R24-R26; C36-C38.

With the use of the outboard additional filter (labeled BPF-96) the only signals found, up to 1500 MHz, are as shown in spectrum (D):

frequency (MHz)	16-MHz oscillator harmonic	96-MHz output harmonic	attenuation (dBc)
80	5		– 77
112	7		– 79
192	12	2	– 70
288	18	3	– 70
480	30	5	– 74

Minor signals occur at the 65th, 66th, and 67th harmonics of the crystal frequency (16.001 MHz), with respective amplitudes of – 73, – 75, and – 76 dBc.

Even with the additional two-section BPF-96 filter, the desired 96-MHz output has a level of 16 dBm (40 milliwatts). Because a significantly lower level, on the order of 0 dBm (1 milliwatt), is required for driving the first doubler in the high-frequency multiplier circuit, additional bandpass filters, or a lowpass filter having a cutoff frequency on the order of 150 MHz, could be easily used. Note that the presence of the second and third harmonic of the desired output signal is not particularly troublesome, since these frequencies will be generated in subsequent multiplier circuitry anyway

To achieve the required Q, the on-board double-tuned bandpass filters use air-wound rather than

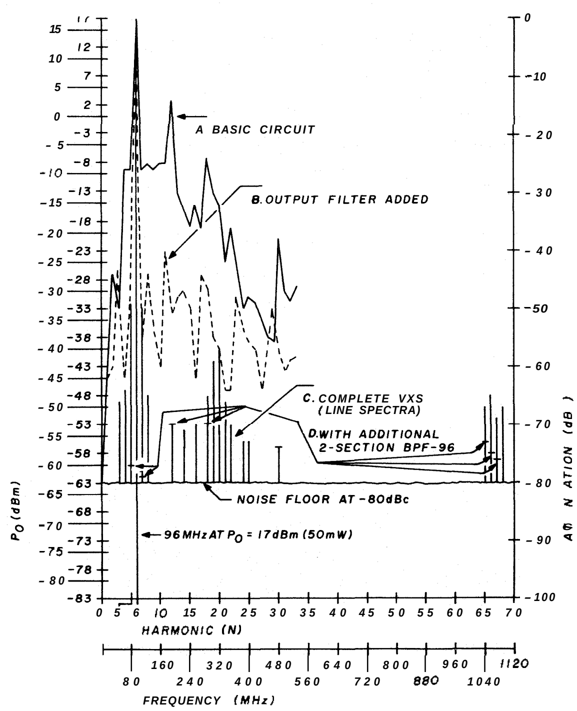


fig. 11. Output of the VXS as seen on a spectrum analyzer. Attenuation of undesirable signals Is shown as a function of frequency for three different configurations. Note the effect provided by adding the bandpass filter.

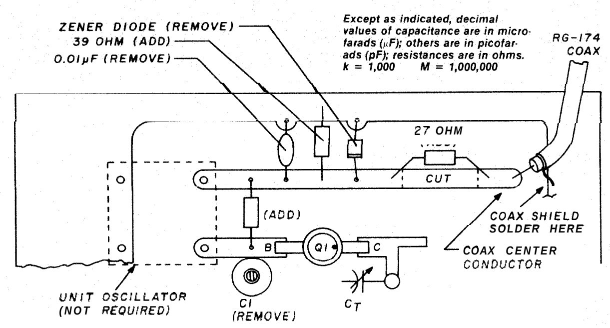


fig. 12. Modification of the N6TX multiplier board for use with an external signal source.

lower-Q toroidal inductors. It is probable, because of the relatively high insertion loss of the bandpass filter sections, that the filters are not completely optimized. However, the ability to provide easily tuned filters using low-cost components was deemed more important than squeezing out an additional few dB of harmonic rejection. Whether or not additional filtering is used, at least 10 dB of attenuation (a T-pad with 22-ohm series arms and a 33-ohm shunt arm) is used at the N6TX high-frequency multiplier board input to ensure that a relatively constant output terminating impedance appears, as well as to reduce the drive level. (I've burned out several MRF 901s but haven't harmed any 2N5179s, in the first doubler stage, with only 6 dB attenuation.)

I prefer the 2N5179 in this stage with an increase in the tuning capacitance of the 192-MHz circuit; this is especially advantageous because the 2N5179 is not only less expensive but is also more readily available than the MRF 901. Of course, any change in terminating impedance can detune the filter or filters and reduce the ultimate suppression of undesired harmonics. Similarly, the ultimate suppression of harmonics of the 1152-MHz signal is a function of the suppression provided by the N6TX circuit and any additional filtering applied thereafter. See **fig. 13.**

construction

After building the basic PC board of **fig. 6** and drilling all component mounting holes, mount the crystal socket on the non-copper side of the board with 4-40 (M3) by 0.37 (9.5 mm) screw, lockwasher and nut. If a crystal oven is to be used, don't mount the crystal socket but wire the oven crystal leads to the appropriate PC-board pads after assembling the board and mounting the crystal oven on it. Before mounting the components, solder the two box sides, cut as shown in **fig. 8,** to the long edges of the PC board. About 1-1/2 inches (38 mm) of the sides extend above and below the plane of the circuit board. A hole for the feedthrough capacitor and for the BNC connector can be drilled in the appropriate side, either before or after soldering.

Solder shield A between the two sides and also to the copper-clad PC-board side. The counterpart of shield A (shield D in big, 8) is positioned against the non-copper side of the PC board and soldered to the pair of opposed box sides. At this side, all components should be mounted to the PC board. Variable capacitors C2-C12 are soldered to the copper pattern on the bottom of the board, while all remaining components are mounted from the top (non-copper bearing) surface of the board.

After installing all components, carefully mount shield B then shield C before soldering the end pieces between the two sides and to the ends of the PC board. A hole may be drilled in the end piece, at the oscillator end of the board, for tuning capacitor C1. However, if the VXS is to be used as a fixed-frequency source, in which capacitor C1 is merely adjusted to set the output to a particular frequency and not to be continuously tuned (as in setting a local-oscillator frequency in a receiver), then capacitors C1 and C2 are dispensed with; frequency is adjusted with C3. Note that output filter inductors L5 and L6 and the 48-MHz trap inductance L7 are also mounted beneath the PC board. The voltage regulator IC socket, with its components, can now be mounted by soldering to one copper side piece, as shown in **fig. 10.**

tune up

Tack solder the top cover to all four sides, but don't completely solder. Install the crystal in its socket and apply at least +12 but less than +20 volts to the B+ feedthrough. Note the voltage at regulator pin 3 (which will be pin 4 of the socket, since pin 1 is attached to ground). The regulator output voltage should be between +8.5 and +9.1 volts dc. Total

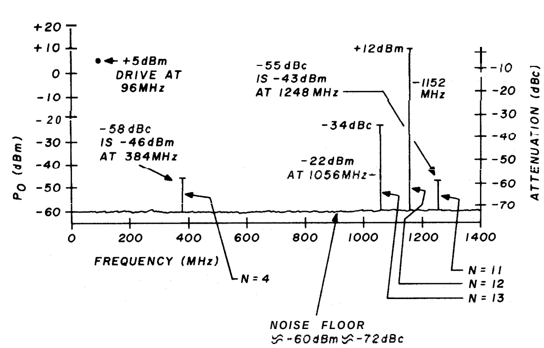


fig. 13. Output spectra of the N6TX multiplier block using the VXS/BPF circuits.

current into the box will be no more than about 75 milliamperes and will probably be considerably less at this time. The base lead of Q2 can be monitored for a 16-MHz signal, indicating that the oscillator is working. Monitor the base lead of Q3 with a 48-MHz rf indicator and tune C4 for maximum rf voltage. Shift the rf indicator to the base lead of Q4 and tune C5 and C6 for maximum voltage at 48 MHz. Retune the indicator to 96 MHz and monitor the base of Q5; tune C7, then C6 and C5, for maximum voltage.

Move the monitor to the tap of filter coil L5 and tune C8 and C9 for maximum voltage. Now connect the monitor to the output connector and tune C10, C11 for maximum output. Then retune C9, C8 for maximum 96 MHz signal. Note that a commercial fm receiver, with carrier-strength meter, may be used for the 96 MHz monitor indicator.

After tuning the bandpass filter for maximum 96-MHz signal, reset the tuning monitor to 48 MHz and adjust C12 for minimum 48-MHz signal. The outboard filter can now be tuned, if used, for maximum 96-MHz signal. As indicated previously, if you can beg or borrow a spectrum analyzer, set the analyzer to display the spectrum from at least 15 MHz to at least 150 MHz (and preferably to at least 500 MHz). Finely adjust C4-Cl1 several times in sequence for best suppression of undesired harmonics while maintaining the desired 96-MHz signal at a reasonable maximum.

Capacitors C6 and C9, especially, are used to adjust the symmetry of the amplitudes of the undesired fifth and seventh harmonics of the crysta! oscillator next to the desired sixth-harmonic signal at 96 MHz. Capacitor C12 has some effect on the tuning of C7. Furthermore, if you use a spectrum analyzer, the 68-k resistor in the voltage regulator circuit may be replaced with a 25-k pot in series with a 56-k fixed resistor, and the pot will vary the circuit voltage. Varying the regulated voltage will often allow you to find a specific voltage at which maximum harmonic suppression is achieved, although power output will change [but, as previously mentioned, it isn't particularly important so long as at least 20 milliwatts (+13 dBm) are available at the attenuator input to be added to the N6TX multiplier].

multiplier modifications

The N6TX multiplier board (fig. 12) is modified by removing the 9.1-volt zener, the 0.01-μF capacitor in parallel with the zener, and the 180-ohm resistor to the zener (not shown). A 27-ohm, 118-watt resistor is soldered from the base lead of the first multiplier transistor to the circuit trace that was the unit oscillator B+ line. A 39-ohm resistor is soldered from the B+ trace to ground, and one end of another 27-ohm resistor is also soldered to the B+ trace. The other end of the second 27-ohm resistor is soldered to the outer conductor of a piece of RG-174 coaxial cable, whose shield is soldered to multiplier ground.

A coaxial cable is connected from the input of the outboard bandpass filter, if used, to the BNC connector on the VXS. If transistor Q1 of the multiplier is a 2N5179 transistor, tuning capacitor CT, on the collector side, should be increased from 1 to 5 pF. The original C1 capacitor (at the first doubler input and unit oscillator output) is no longer needed.

The multiplier should be tuned in the same manner as specified by N6TX in his article. I've found that the tripler input and three output filter capacitors should be the suggested Triko 202-08M, although the pair of 384 MHz tuning capacitors may have to be increased to 2-10 pF, to adequately tune the modified multiplier board. Fig. 13 illustrates the output spectra of the modified multiplier block when driven with the VXS and 96-MHz outboard bandpass filter.

Some uses of the VXS and multiplier blocks are shown in figs. 14 through 17. In fig. 14, one possible way that high transmitting power may be eventually economically realized within the next several years in the 2300-2450 MHz band will probably be by use of microwave oven magnetrons (a magnetron being es-

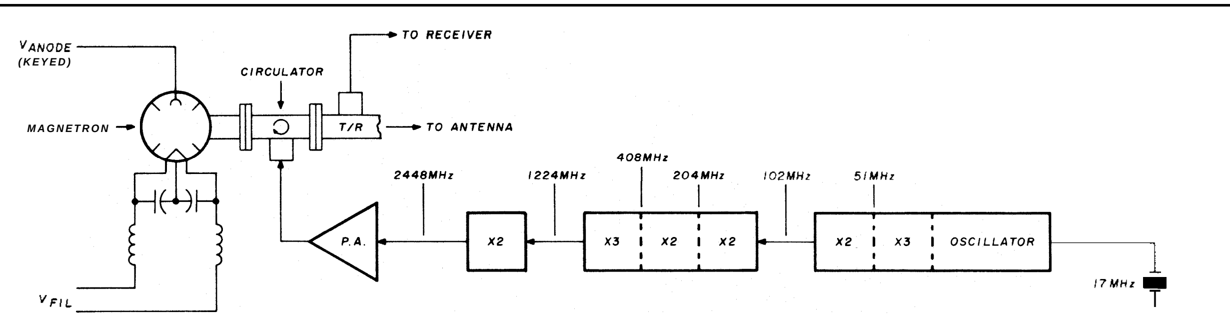


fig. 14. Possible transmitter for crystal-controlled injection locking of a microwave magnetron producing 600 watts output at 2448 MHz.

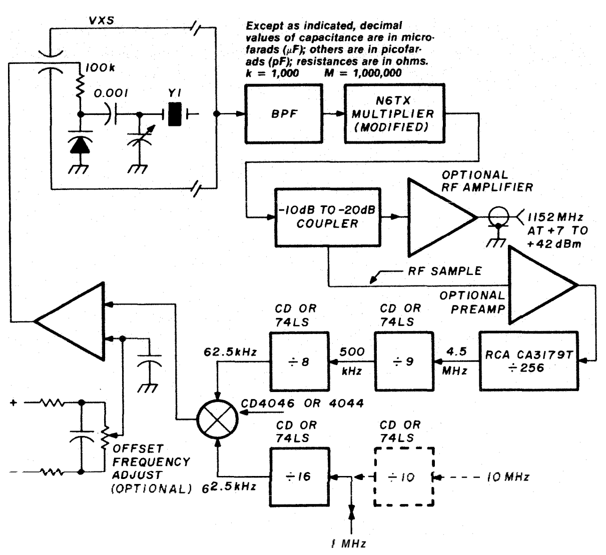


fig. 15. Block diagram of a phase-locked 1152-MHz subsystem that will provide a highly stable signal for multiplication into the 13-cm, 9-cm, or 5-cm bands.

sentially a diode tube in which oscillations occur at microwave frequencies because of the finite time required for electrons to travel or drift between the tube elements). Available magnetrons, which cost about as much as a vhf power tube of the 4CX250 type, provide up to 600 watts of output power but are normally pretuned at the factory for oscillation at about 2450 MHz.

The tuning adjustment is not normally accessible (apparently being inside the vacuum envelope of the tube), but some tuning can apparently be accomplished by varying the tube anode current. Many operators interested in magnetron use have concluded, although none (to my knowledge) have yet proved, that it should be possible to reduce the magnetron frequency to be just within the upper edge of the 2300-2450 MHz band. Advantageously, another multiple of 144 MHz is present at 2448 MHz, which is also a 144-MHz i-f above 2304 MHz, itself a second harmonic of 1152 MHz. It may well be possible, using equipment as shown in **fig. 14,** to injection-lock a 600-watt output magnetron with less that 10 watts of power from a very-high-frequency-stability source, whereby the magnetron assumes the same stability as its locking source. The 10-watt power level is obtainable, now, with fully-transistorized amplifiers.

Probably the greatest obstacle in achieving high power in the 13-cm band is the requirement for a 600-watt circulator. I know of no such unit commercially available, although the technology appears to exist. I am confident, however, that some experimenter will eventually design, or design around, a circulator for this frequency and power level, allowing an injection-locked, high-power source to be realized. **Fig. 15** is a phase-locked 1152-MHz subsystem that will provide a highly stable signal for multiplication into any of the 13-cm, 9-cm, or 5-cm bands.

Fig. 16A is a keyed 1152-MHz source having about

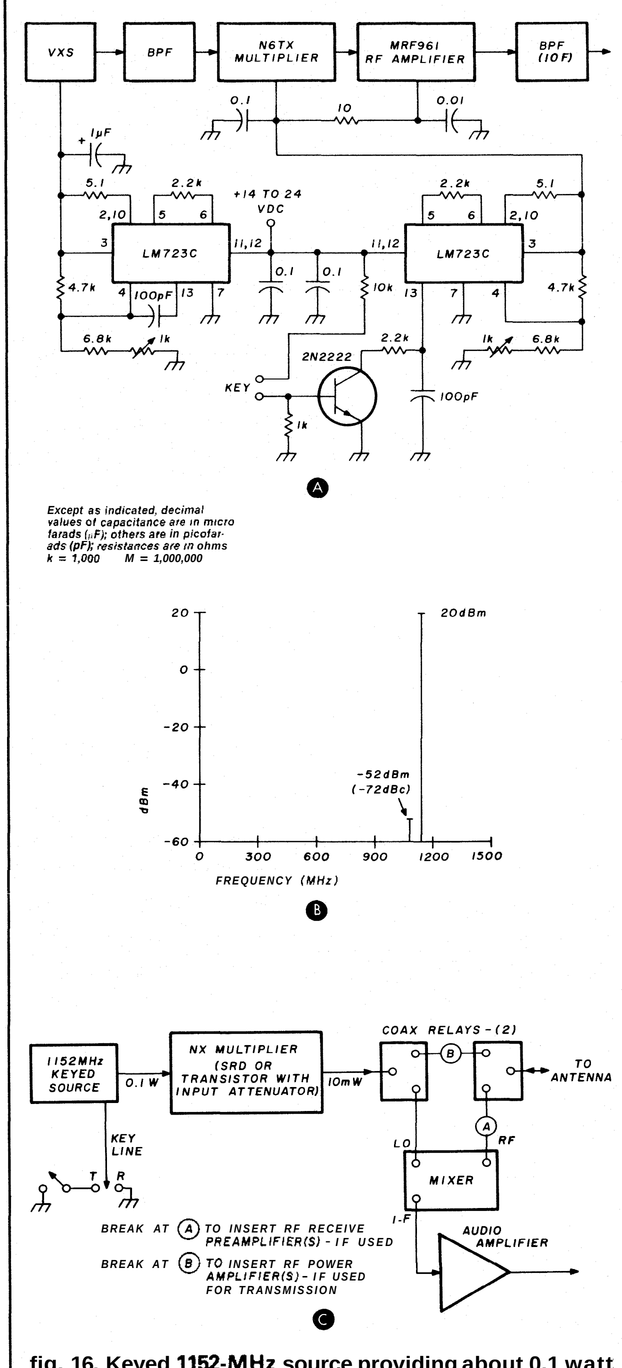


fig. 16. Keyed 1152-MHz source providing about 0.1 watt output, A. Output spectrum, B, shows only one spur at slightly more than −70 dBc. A direct-conversion transceiver using the 1152-MHz source is shown in C.

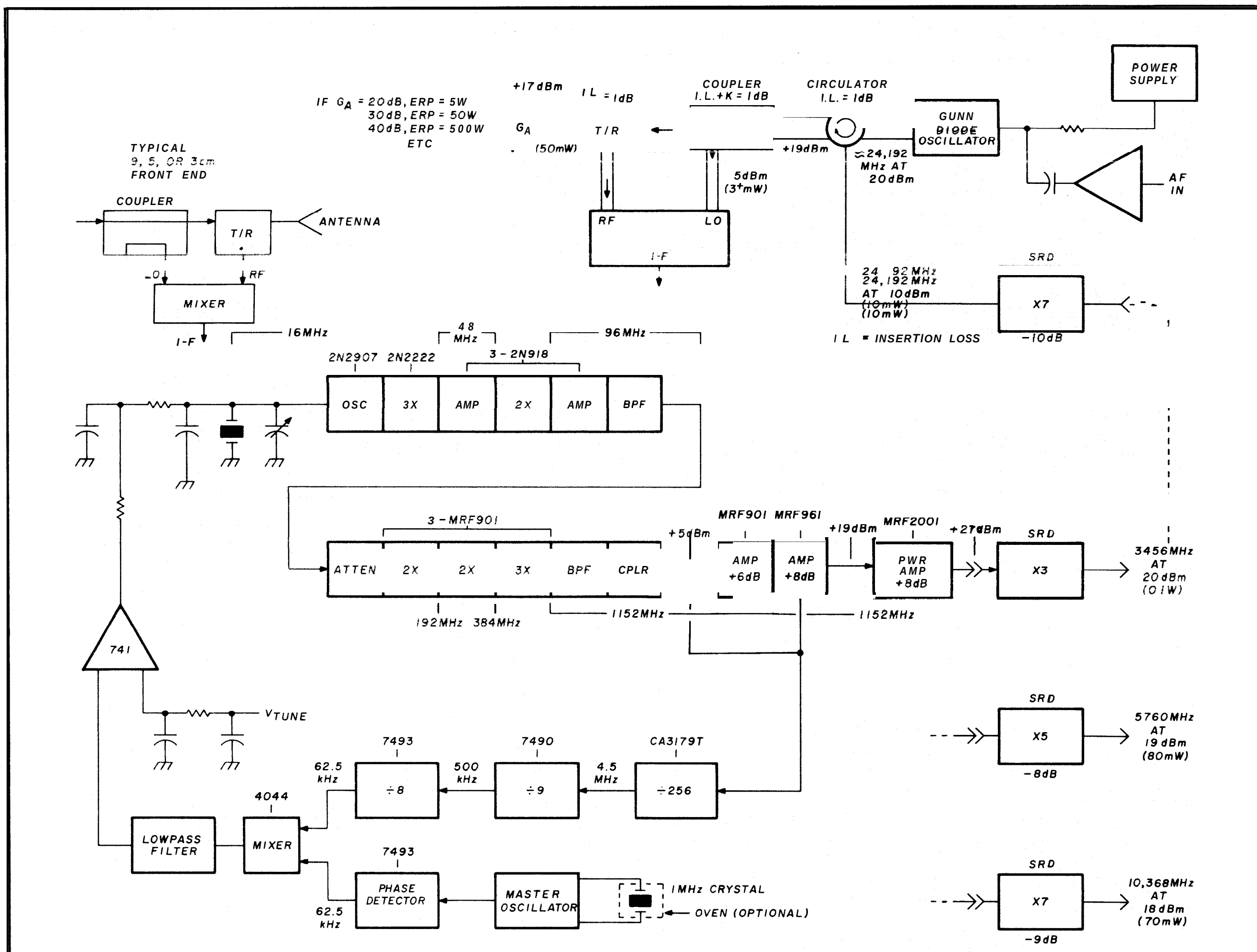


fig. 17. Some other microwave configurations based on the multiplication scheme described in this article.

1/10th watt output, while **fig. 16B** shows its output spectrum (only a single spurious output at slightly more than 70 dB below the carrier). **Fig. 16C** shows a direct-conversion transceiver using the source of **fig. 16A. Fig. 17** shows other microwave source configurations, all based upon multiplication of the 1152-MHz signal.

summary

All of our microwave bands have one frequency that's related to 1152 MHz. By building a power source at 1152 MHz, multiplication to the microwave bands becomes possible. A relatively simple, yet stable, 1152-MHz chain is necessary; one such chain is described. The power amplifier, producing 100 milliwatts at 1152 MHz, is an adaptation of a circuit designed by Dick Frey, WA2AAU. Simple frequency doublers and receiving mixers for 2304 MHz have been described in many articles (check your ***ham radio*** and QST indexes). Thus it's possible to find easily built components for 2304 MHz right now. Higher-frequency blocks and subsystems are being worked on, and further results, from this writer or others, should be forthcoming.

acknowledgments

I would like to thank Dick Frey, the other half of the present Mt. Greylock microwave gang, for his help and encouragement; all the local microwave people for their interest; and my four-year old son, Jeremy, and nine-year-old daughter, Alyssa, for helping to mount parts onto PC boards and for tuning and measuring.

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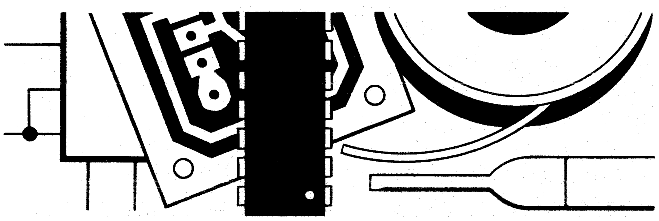
ham radio

the **weekender**



Inrush current protection for the SB-220 linear

Do you have adequate surge protection for your SB-220? If you own this fine piece of gear or similar equipment without the benefit of built-in surge protection, this article should be placed at the top of your project list. For about $10 in parts and six hours of bench work, you can breathe easy when you push the power switch. I call it the $10 insurance policy.

The subject of surge protection has been addressed by many in the past few years. In my opinion, one of the better articles was written by K. M. Gleszer, W1KAY, entitled "Upgrading Your SB-220 Linear Amplifier," which appeared in QST, February, 1979. Specific solutions were offered for operation with 117-Vac for filament inrush current, diode-transient and voltage-equalization protection, plus other items. But conspicuous by its absence was a scheme for diode inrush current protection. This protection is easily obtained with the simple circuit described here.

One other area where I'd suggest a change is the time-delay relay. The time-delay function is automatic with a standard relay coil and a current-limiting resistor. Therefore the high cost, plus purchase time and final alteration, of a time-delay relay can be avoided.

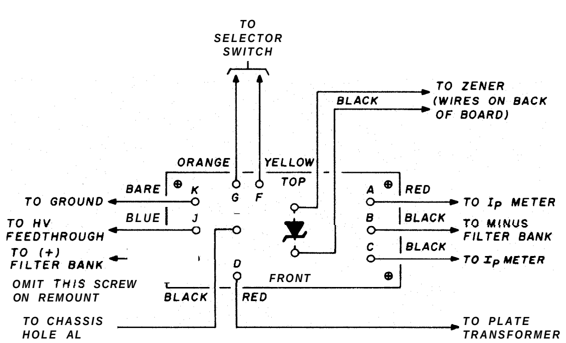


fig. 1. SB-220 rectifier board.

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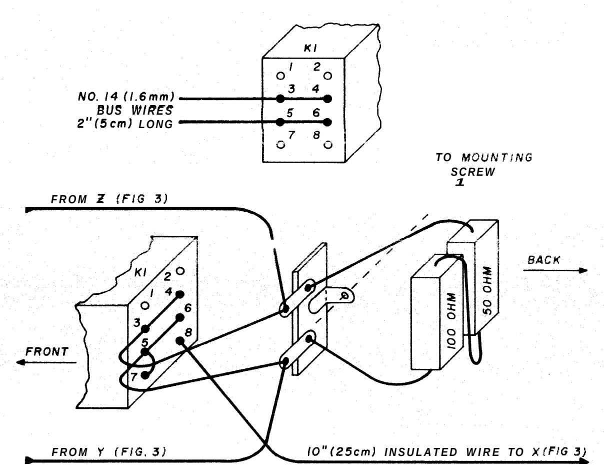


fig. 2. Relay connections for the surge-protection circuit.

The mods I've installed are not unfamiliar, as they've appeared in several 1970-series of the ***Radio Amateur's Handbook.*** However, I've described the procedures in a detailed order using short, sometimes elementary, phrases for clarification. I'm a stickler for the smallest detail, so you needn't bother with assumptions.

With the mods instailed, the foilowing benefits wili be added to your SB-220:

1. Rectifier transient surge protection.

2. Rectifier reverse voltage equalization.

3. Rectifier inrush current protection.

4. Inrush current protection for the 3-500Z filaments.

This procedure is divided into two parts: rectifier protection and surge protection. You can elect to cancel one, but because the amplifier must be uncaged for installation of either, it seems wise to include both.

The fourteen original diodes in the SB-220 were not replaced with higher PIV units. This action is not necessary unless you break some during disassembly. These diodes are rated for 1 ampere average forward current at a PIV of 600 volts. The ratings are adequate for this application, and, combined with the modification, they will have a long life.

The nominal delay was selected as 5 seconds. This time can be altered by varying the total limiting resistance. A resistance of 200 ohms caused a long delay, and the resistors dissipated much power. At the opposite extreme, 100 ohms provided insufficient delay. Therefore, a satisfactory value of 150 ohms was selected. Note that the time delay and resistance values were selected using a line voltage of 220 Vac. I intended to operate this linear only on the higher line voltage for increased efficiency.

rectifier protection

1. Remove amplifier case, top shield cover, and right-side shield.

2. Remove the four rectifier board hold-down screws.

3. Make a wiring map of all twelve wires connected to the rectifier board and identify by color designator **(fig. 1)**.

4. Unsolder all twelve wires at the board end, then remove diodes.

5. Wick twelve wire pads and all diode holes. Remove flux.

6. Drill out all *diode* holes using a No. 47 (2 mm) drill bit from the pad side of the board (assuming all boards are the same).

7. Using a No. 15 (4.5 mm) drill bit, deburr the new holes from the component side. Do not deburr the pad side.

8. Install resistors (470 k ½ w) from the pad side, then install diodes and capacitors (0.01 at 1 kV) from the component side. Next:

- **a.** Solder each pad with its three wires.
- **b.** Clip component pigtails as you go.
- **c.** Clean board to remove flux.
- **d.** Ohmmeter check — note highs will be 470 k.

9. Connect board to SB-220 using the following sequence:

- **a.** Solder red wire to hole D.
- **b.** Solder blue wires at holes H and J.
- **c.** Mount board using three screws — omit lower LH.
- **d.** Solder bare wire at hole K.
- **e.** Solder black wire at hole E.
- **f.** Solder black wires to holes and pads for the zener. Observe proper polarity.
- **g.** Solder orange wire to hole G.
- **h.** Solder yellow wire to hole F.
- **i.** Solder red small wire to hole A.
- **j.** Solder black wire (minus filter bank) to hole B.
- **k.** Solder black wire (I_p meter) to hole C.

This completes the rectifier-board wiring. Dress all wires at right angles away from the board, then

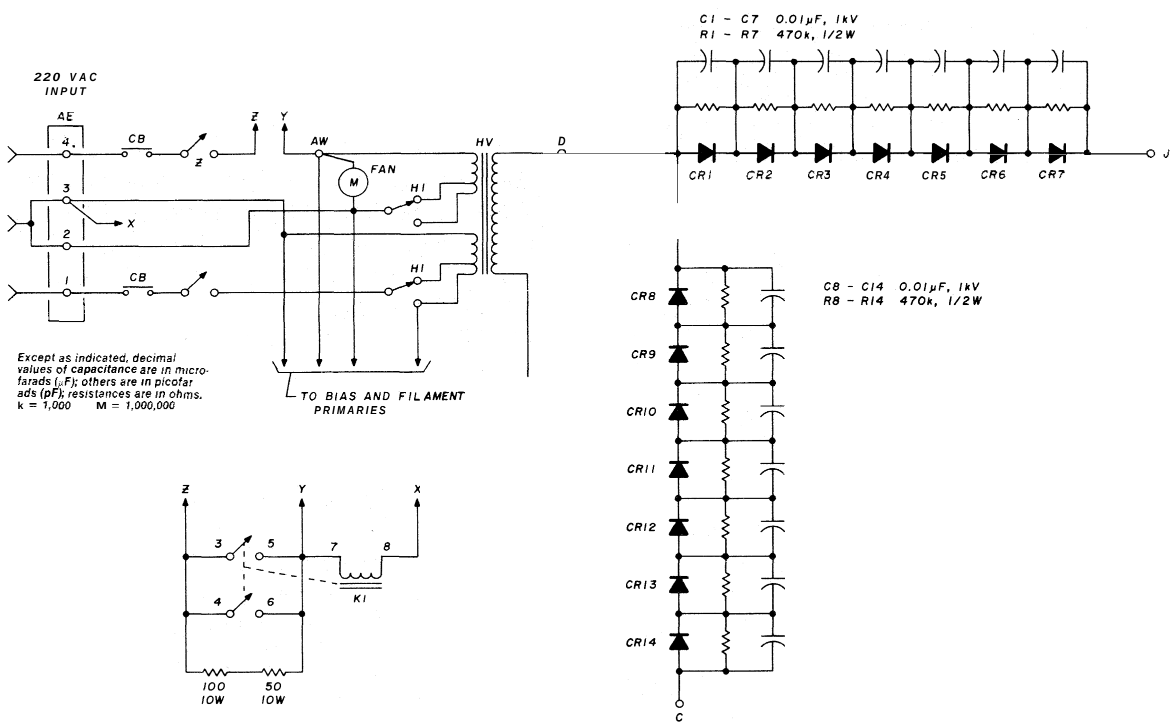


fig. 3. SB-220 surge modifications.

10. Reinstall right-side shield.

11. Oil felt pads on fan motor while top cover is off.

12. Install top shield cover.

13. Test the amplifier using a dummy load.

14. If OK, proceed to the next section.

surge protection

1. Solder No. 14 (1.6 mm) bus wire 2 inches (5 cm) long to pins 3 and 4 of relay K1 (fig. 2).

2. Solder No. 14 (1.6 mm) bus wire 2 inches (5 cm) long to pins 5 and 6 of relay K1.

3. Bend the two wires and solder to a two-lug tie strip.

4. Connect pin 5 to 7 using No. 20 (0.8 mm) bare wire.

5. Connect a black insulated wire (rated for 220 Vac, 10 amperes) about 10 inches (25 cm) long to K1 pin 8.

6. Stack the two current-limiting resistors (100 and 50 ohms) and connect in series. Solder this pair to the lower holes in the tie strip.

7. Mount the completed surge-protection into the SB-220 using the center ground lug on the tie strip and the existing chassis screw located about 2 inches (51 mm) forward of terminal strip AE. The relay case should rest against the chassis, being supported by the bus wires.

8. Connect the 10-inch (25-cm) black insulated wire (trim as required) from relay K1 pin 8 to terminal 2/3 on terminal strip AE of the linear.

9. Remove existing black jumper wire between power switch Z and front standoff AW.

10. Connect Z to pins 3 and 4 of K1 using the tie strip. Use insulated wire with (220 Vac, 10-ampere rating).

11. Connect Y from standoff AW to pins 5 and 6 using the tie strip. Use insulated wire with 220-Vac, 10-amp rating.

12. This completes the surge relay installation.

From the Heathkit manual, these codes are used:
AE 1101220 Vac input terminal strip.
AW front-mounted standoff tie point.
AL front corner hole.
Z power switch.

operation

Checkout of the surge protection circuit can be monitored each time the linear is fired up, assuming the filter capacitors have discharged to a low level. Place the selector switch in the HV position, while the mode switch can be in either the CW/TUNE or SSB position. After the power switch is pushed, there will be a time period of a few seconds of dead silence. This delay time is controlled by the value of the limiting resistors. During this period the plate voltage meter can be observed to slowly increase from zero to about 1500 Vdc. Additionally, the meter illumination lamps will ***slowly*** energize to about half brilliance. Since the 3-500Z filaments are in parallel with these lamps, they will be responding in the same way. If in doubt, turn off your room lights while energizing the linear and peer down through the case top.

The cooling fan will be turning very slowly while gradually building up speed. Therefore there will be no noise from this source during the initial few seconds.

After the five-second surge-delay period, adequate voltage will be available for surge relay K1 to pull in. During a brief interval K1 contacts will close and hold, thus shorting the limiting resistors and applying full line voltage to the transformers. Instantly the plate voltage will increase from 1500 Vdc to its normal maximum value. The 3-5002 filaments will glow with their normal brilliance, and the cooling fan will attain maximum speed. Don't be alarmed when you hear a brief buzzing sound as the relay closes. This sound is caused by K1 contacts bouncing (as all mechanical relays do) combined with slight inductive arcing.

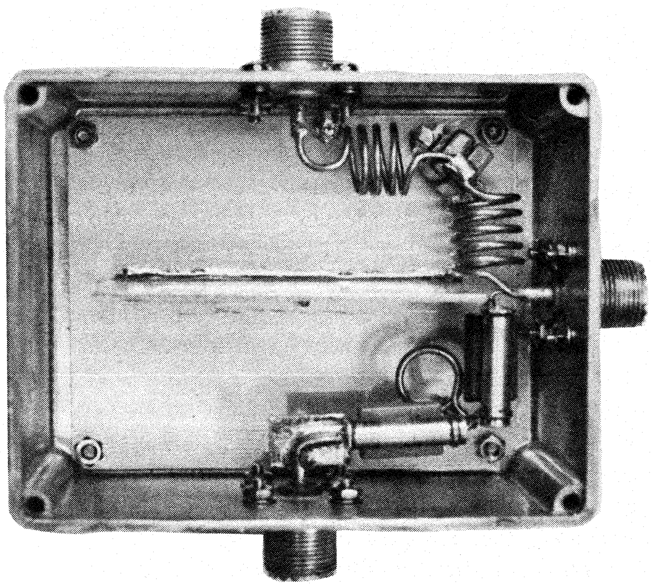
Although this article is written specifically for the SB-220, other similar equipment could be surge protected using these mods.

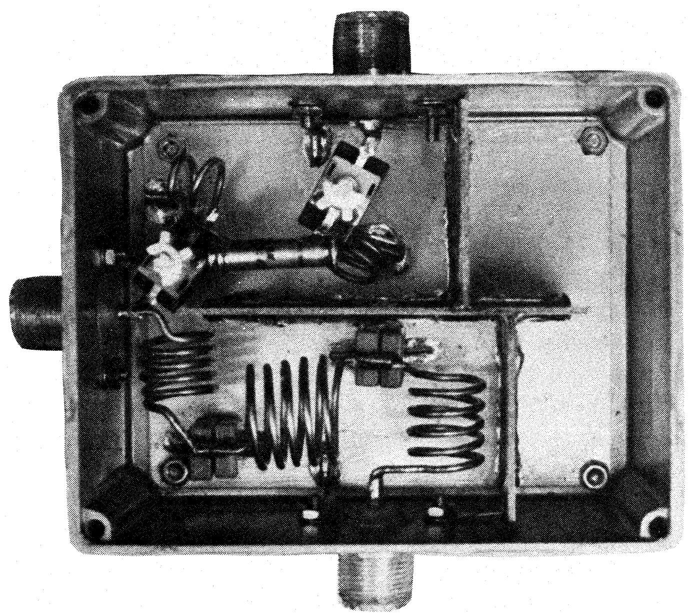
For additional information on rectifier diode protection I suggest the April, 1980, edition of ***Worldradio,*** which has a fine article written by Joe Carr, K4IPV.

Once you've installed the mods as shown in fig. 3, you can place the problem of surge protection on the shelf for a well-deserved rest. I've used these circuits on two other homebrew linear amplifiers with total success. In addition I've used them on power supplies for several transmitters using the lower line voltage. The only difference is the selection of the limiting resistance for a satisfactory delay period.

Note: KI is a dpdt relay, 5000-ohm coil, 120 Vac. Contacts are rated at 10A, 125 Vac. Dimensions: 1-5/8 x 1 x 3/4 inches (41 x 25.4 x 19 mm).

ham radio





Diplexer consists of a matched pair of highpass and lowpass filters, which allow a vhf and high-frequency antenna to share a common feedline. The filter at the antenna end has three sections *(top)*; that at the station end has five sections *(bottom).*

transceiver diplexer:

an alternative to relays

Frequency-selective filters allow vhf and hf antennas to share a common feedline

In many cases it's desirable to reduce the number of feedlines between the ham station and the antennas. One of the more important reasons is the price of high-quality coax cable. It's easy to spend as much money on transmission lines as on a small commercially manufactured 2-meter Yagi antenna. A second reason may be the need to tidy up your installation to please neighbors. If antenna restrictions exist in your area, and you're trying to avoid detection, the presence of several coax cables can be too much to hide.

One of the more popular ways of making the best use of feedlines is to use switching relays at the

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antennas to select the desired antenna. Several systems to accomplish this are available commercially, and homebrewing such an arrangement is not technically difficult.

There are disadvantages to such schemes. What happens when you're chasing a rare station and still want to listen to the local DX repeater on 2 meters? If you have only one feedline, this can be inconvenient. Care must be taken to avoid transmitting on the wrong-frequency antenna to prevent possible damage to both transmitter and antenna.

enter the diplexer

An alternative to relays is frequency selective networks to select the proper antenna automatically. The networks can also allow simultaneous combination of more than one transceiver on the same coax cable.

These networks are called *diplexers*, since they allow two transmitters (or receivers) to use the same feedline at the same time. They differ from *duplexers,* as used in repeaters. Duplexers permit simultaneous operation of one transmitter and one receiver on a common antenna.

Although possible, it would be difficult to construct networks that would permit several different high-frequency antennas to share the same feedline. Relays are probably best used for this purpose. Because 144 MHz and 220 MHz are commonly used for local communications, I designed a simple network to permit either of these vhf bands to coexist with high-frequency signals on one coax cable. I did not include the 50-MHz band because this design would have required more complex networks. (The 420-MHz band will pass through the filters, but the impedance match is marginal.)

I used two networks. The one at the station end (**fig. 1**) allows both the high-frequency and vhf rig to access the coax simultaneously. The network at the antenna end (fig. **2**) does the same for the high-frequency and vhf antennas. Each network consists of a mated pair of highpass and lowpass filters to accomplish the separation and combination of the two different frequencies.

Some disadvantages occur in the use of filters to perform these functions. A small amount of loss is added to the system. This is minimal, however. Also the impedance presented to the transceivers is modified. By proper filter design this mismatch can be kept to a minimum.

One added benefit of the filters should be noted: Lowpass filters are in the circuit to the high-frequency antenna, so some reduction in harmonic radiation is evident, which may reduce TVI to the point that an additional filter isn't needed.

designing the filters

The highpass and lowpass filters are simple Chebychev units that can be designed from tables of normalized filter prototypes or by calculating normalized inductor and capacitor values. I found it easier, however, to use the network design programs available on the engineering computer at my place of employment.

Reflection coefficient. To minimize the amount of mismatch introduced by the filters, I designed them to have a maximum reflection coefficient of 0.065. Since two filters are in tandem, the worst-case reflection coefficient with a 50-ohm load could be twice this amount, or 0.13, which corresponds to a maximum SWR of 1.3. The worst-case situation at the transmitter for a load with a 2-to-1 SWR would be SWR of 2.7. Because of the designs I used, the frequencies of worst match don't coincide, and such a degradation is unlikely. The match may also be better at some frequencies because of the small variations in the impedance transformation through the filters.

Cutoff frequencies. I set the filter cutoff frequencies about 7 per cent above and below the required maximum and minimum frequencies to avoid the loss appearing near the filter corners caused by the finite *Q's* of the inductors. The resulting cutoff frequencies were 32 MHz for the lowpass filters and 135 MHz for the highpass filters.

Isolation. The number of filter sections is governed by the isolation required between high-frequency and vhf equipment. Isolation at the transceivers must be much greater than at the antennas. For protection against receiver overload, at least 50 dB isolation was desired between the high-frequency transmitter and the vhf receiver. Such isolation reduces 1000 watts to 10 milliwatts at the receiver front end. Because of the wide frequency separation, no undesirable intermodulation occurs in the vhf receiver.

The isolation required between vhf transmitter and high-frequency receiver is usually not as great, because most stations use much lower power on vhf than on hf. Even so, I designed the filters to be symmetrical, which should give the same isolation in both directions.

At the antennas, I set the isolation at 30 dB. This isolation should prevent high-frequency-antenna radiation from causing any significant reduction in the front-to-back ratio of a directional vhf antenna.

To obtain the desired isolation, I made the networks at the station end with five sections each and those at the antennas with only three sections each.

After I designed the filters I increased the reac-

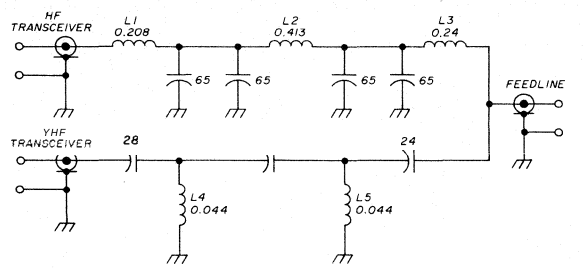

fig. 1. The five-section filters used at the transceivers. Inductor data is given in table 1.

tances of the components at the common port by the same ratio to compensate for the shunting effect of the other filter. I did this with an interactive network analysis program. To make the impedance match as good as for the highpass or lowpass filter alone, I increased the end inductor of the five-section lowpass filter by 15 per cent and decreased capacitor of the five-section highpass filter by the same amount. For the three-section filter, the change of the end components was 30 per cent.

I made allowances for the parasitic capacitances of the inductors to ground in the lowpass sections, which add in parallel with the shunt capacitors. I made allowance of 3 or 4 pF in the capacitors I used. I added small metal tabs about 0.4 inch (1 cm) square to the highpass filters. This restored symmetry to the highpass sections and improved the match at 220 MHz. The final design of the filters appears in **Figs. 1** and **2.**

construction

The filters were built in cast aluminum boxes and a piece of unetched copper-clad PC board was attached to the inside of the box with machine screws. The shunt components were then soldered directly to the copper board with the shortest possible leads. The series components were supported by the shunt components (this arrangement can be seen in the photos). This construction provides a rigid mounting for the parts with minimal stray inductance and capacitance.

Shields were placed between the highpass and lowpass filters in each box to reduce mutual coupling. If you don't include the shields, isolation between the vhf transmitter and high-frequency receiver will be seriously impaired.

For the five-section networks, additional shields were required. The shields were made of double-sided copper board. They were soldered all along the seams together with the groundplane copper boards and the other shields, then fastened to solder lugs on the connectors where possible.

All coils were placed at right angles to each other in the same shielded area to avoid mutual coupling, which can cause filter performance to depart drastically from the theoretical predictions.

components

The fixed caps were micas with a 1000-volt rating. This rating is adequate for power levels up to the legal limit. Because of the high currents flowing in the shunt capacitors in the lowpass filters, the required capacitance was obtained by using two capacitors in parallel, which reduces any possible heating in the capacitors. Currents are highest when operating near the filter cutoff frequency and can easily reach 5 amperes with 1000 watts of input power.

Air variable capacitors could be used throughout, in place of the micas, provided the voltage rating is adequate. In the highpass sections, the micas were paralleled with air variables and glass piston capacitors to allow tuning. After the filters were tuned, it appeared that fixed units of the calculated values would have worked just as well, as judged from the positions of the variables.

All the inductors were wound of No. 12 (2.1-mm) tinned copper wire. Winding data were obtained from charts in the ARRL ***Handbook.*** Information on the dimensions of the coils appears in **table 1.**

tuning the filters

By far the best way to tune Chebychev filters is with a swept reflectometer. These filters were tuned this way, adjusting the coils by stretching and squeezing and by tuning the capacitors until the impedance match across the passband of each filter was within the desired limits. Not everyone has the facilities to adjust the networks in this manner. As an alternative, the filters should be adjusted one at a time into a dummy load with an SWR meter or a noise bridge set to 50 ohms. The frequencies to use are given in **table 2.** It's important not to vary the components too far from the calculated values; do-

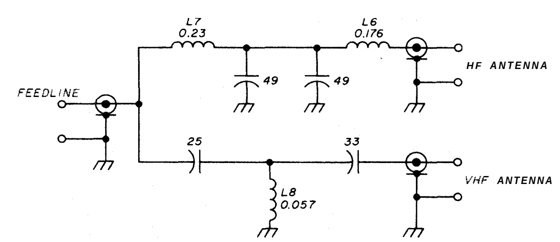

fig. 2. The three-section filters used at the antennas,

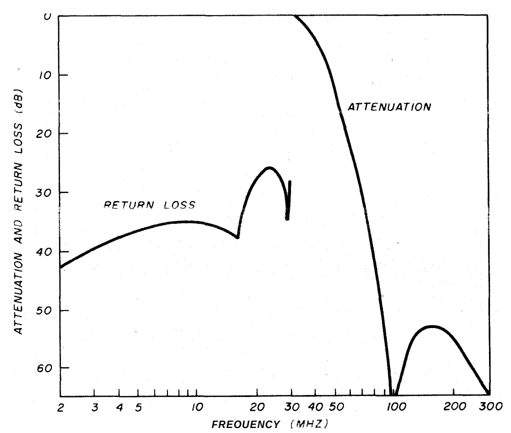

fig. 3. Measured frequency response and return loss for the five-section **lowpass** filter.

ing so may cause the isolation to be upset.

After tuning for best match at the frequencies indicated, check the match at other frequencies within

table 1. The inductors should be wound according to this data. The wire used is solid No. 12 (2.1 **mm**) with spacing between the turns equal to the wire diameter.

inductor	nominal inductance μ**h**	inside diameter inches	**(mm)**	no. turns
L1	0.208	0.5	(12.7)	4.5
L2	0.413	0.75	(19.0)	5.0
L3	0.24	0.5	(12.7)	5.25
L4	0.044	0.5	(12.7)	1.25
L5	0.044	0.5	(12.7)	1.25
L6	0.176	0.5	(12.7)	4.0
L7	0.23	0.5	(12.7)	5.0
L8	0.057	0.375	(9.5)	2.0

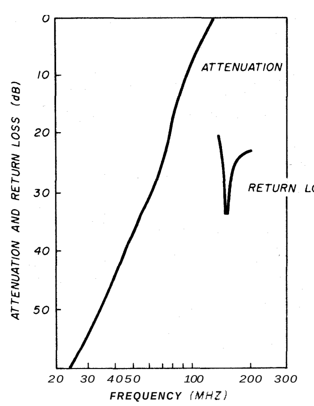

fig. 4. Measured frequency response and return loss for the five-section **highpass** filter.

table 2. Adjust the inductors (and variable capacitors, if used) for best match into a 50-ohm load at these frequencies.

filter	adjustment frequency **(MHz)**
5-section lowpass	28 3
5-section highpass	148.0
3-section lowpass	28.0
3-section highpass	147.0

the filter passbands. It may be necessary to retune somewhat if the impedance match is poor. Remember that the match should not necessarily be perfect at all frequencies, but the SWR should not be worse than 1.2 anywhere in the passband of either filter.

diplexer performance

The two networks were measured with 50-ohm terminations on the unused ports. The results of the

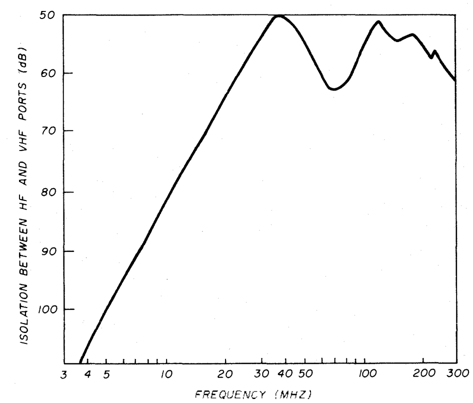

fig. 5. Measured isolation between the high-frequency and vhf ports of the five-section filters. The common port was terminated with 50 ohms.

measurements are given in **figs.** 3 through **8.** The impedance match is plotted as ***return loss.*** This quantity is 20 times the logarithm of the magnitude of the reflection coefficient. It was measured directly by the test equipment used. The reflection coefficient for which the filters were designed, 0.065, represents a

table 3. These are actual measured losses in a **50-ohm** circuit with the unused ports terminated. Resistive and mismatch losses are included.

filter	maximum loss **(dB)**	frequency **(MHz)**
5-section lowpass	0.1	21.0
5-section highpass	0.22	220.0
3-section lowpass	0.07	28.0
3-section highpass	0.05	225.0

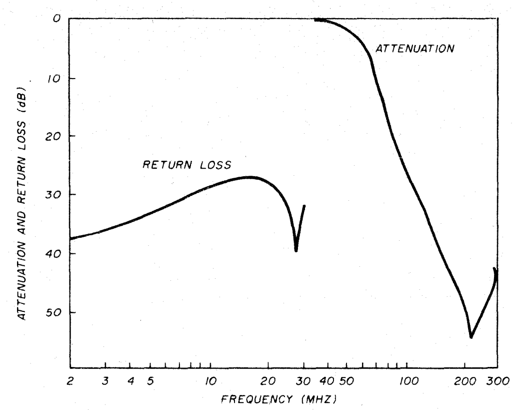

fig. 6. Measured frequency response and return loss of the three-section lowpass filter.

return loss of 23.7 dB and an SWR of 1.14. The diplexer insertion loss was surprisingly low. **Table** 3 summarizes the measured losses through the filters.

Use of the filters shows that the isolation between the high-frequency and vhf equipment is more than adequate. The equipment was a Yaesu FT-301 with an FL-2100B and an Icom IC-22S. The only problem areas were at harmonics of the high-frequency transmitter that fell on frequencies in the 2-meter band. However, this was also a problem when operating with separate feedlines. Significant fifth-harmonic energy was picked up by the 2-meter transceiver even when it and the high-frequency transmitter were connected to dummy loads.

possible improvements

The layout of the filters would be much better if the boxes were long and narrow, with the common connection near the center of the assembly. Then the high-frequency and vhf ports would be separated by the greatest distance. Another layout improvement would be to shield separately each inductor in its own small compartment. This would greatly reduce mutual coupling between the coils.

The other possible improvement is to reduce the effective stray inductance of the shunt capacitors in the lowpass filters by paralleling more than two capacitors to obtain the required value. The self-resonant frequency of smaller capacitors would be moved higher in frequency, and the stopband attenuation and isolation would be greater.

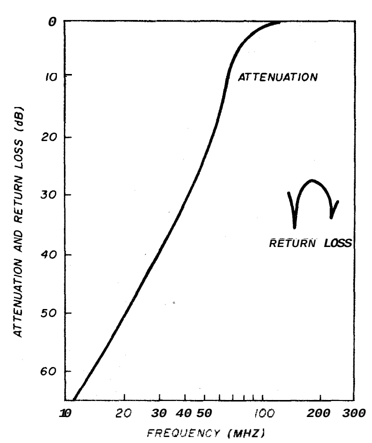

fig. 7. Measured frequency response and return toss of the three-section highpass filter.

using the diplexers

If antenna tuners or TVI filters are in use at your station, they must be placed between the transceiver and the diplexer, which can be a problem if the antenna tuner is used to compensate for fairly high standing-wave ratios. Possible voltage and current stresses on the components in the filters could easily damage them. It would be wise to restrict operation at maximum legal power to standing-wave ratios no higher than 2.5 on the main feedline.

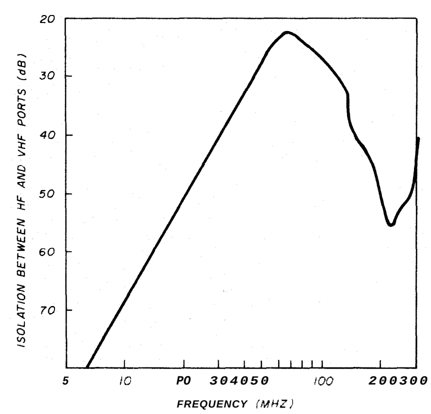

fig. 8. Measured isolation between the high-frequency and vhf ports of the three-section filters. The common port was terminated with 50 ohms.

For normal exciter power levels (under 300 watts input), there should be no problem with standing-wave ratios up to 5 under normal use, especially below the 20-meter band.

If your SWR meter is capable of operation on both hf and vhf, it may be placed in the common feedline and measurements can be made in either frequency range.

ham radio

the ham notebook

spring mounted beam saves rotor gears

What ham has not, at some time or another, had the gears torn out of his rotor drive motor when the beam has been whipped suddenly by a strong gust of wind or by a bad storm?

After experiencing this disaster several times in my sixty years in ham radio, I finally decided to do something about it. This time, when I put up my Mosley TA-33, I made sure the gears would stay in no matter what the wind velocity.

the cure

It's a simple measure and easy to accomplish **(fig. 1)**. The mast from my rotor is a 2-inch piece of pipe. I slid a 6-foot (1.8 meter) piece of 1-1/2 inch pipe (this could be any other length of course) down inside the 2-inch pipe about 2 feet (0.6 meter) (this could vary). I slipped a heavy automobile shock absorber coil spring over both pipes so that the center of the spring came to the top of the 2-inch pipe. Then I welded the coil to the pipe: the top end of the coil to the I-112-inch pipe; the bottom end to the 2-inch pipe. I made three weld spots around each pipe. The spring I used fit snugly around the 2-inch pipe, so welding directly to the pipe was easy.

At the top, I shimmed the spring with three pieces of 314-inch (2 cm) strap iron cut to about I-inch (2.5 cm) long. This made the weld spots fit snugly to the I-112-inch pipe. This precaution probably wouldn't be necessary, but it didn't take much more time and it made a neater looking weld.

Because of the lightweight construction of the TA-33 antenna, I didn't bother with an end thrust bearing at the bottom of the 1-1/2-inch pipe. The spring was heavy enough to take up the beam weight. However, with heavier and more complex beam antennas, it might be wise to do something along these lines. One simple method would be to slide a 2- or 3-inch (5 or 8 cm) cut of the 1-1/2-inch pipe inside the 2-inch pipe at the place you want the bottom end of the 1-1/2-inch pipe to rest, then drill through both pipe walls and secure the pipes with a bolt to hold the piece inside the 2-inch pipe. To avoid as much friction as possible, of course, the bottom of the 1-112-inch pipe and the top of the small inserted piece should be ground as flat as possible and packed with heavy machine grease.

springs

The heavier the spring the better. I came across a spring about 10 inches (25 cm) long made from 318 inch (9.5 mm) spring steel and 2-inches (51-cm) inside diameter. Many such springs are available in auto-part shops, usually from discarded shock absorbers. But I was lucky. I was driving past a shop one day and noticed a sign that said, HEAVY DUTY SPRINGS OF ALL KINDS. It turned out to be a spring manufacturer who

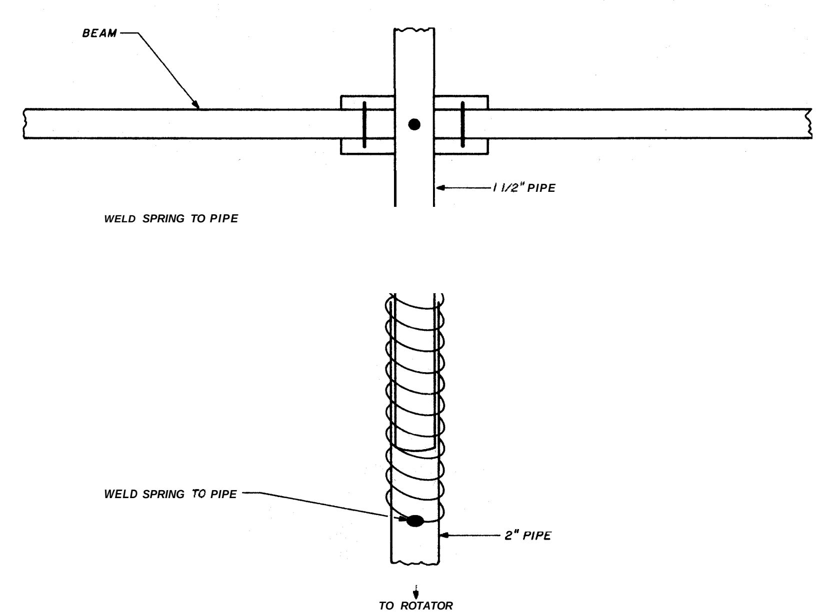

Fig. 1

made springs for the shock absorber people. I explained what I was looking for, and the shop foreman produced just what I wanted. When I asked, "How much?" he said, "Take it. It isn't worth the paperwork." Still some nice people around yet.

My beam has been up for six years. We have had all kinds of high winds, near-tornadoes, and gusts that shook the house. But the beam and the rotor gears are still intact. The beam bounces around a bit in high winds, but there is very little shock to the rotor gears. If I had it to do over, I'd try to find a heavier spring; but of course the nearer you get to a rigid connection, the less effective the arrangement becomes.

Russ Rennaker, W9CRC

calculator care

Many of the less-expensive small calculators aren't too well sealed against moisture and dirt. After living with the results of dirty contacts on the calculator keyboard of my unit, I decided to do something about it.

I opened the machine and squirted some aerosol switch-contact cleaner onto the bottom of the keyboard. I then cut and shaped a sandwich bag to fit around the calculator and taped the ends of the bag with Scotch™ tape. I poked a hole in the bag with a toothpick to accept the charger plug.

Now the calculator is protected from cigarette smoke, dirt, and grime. No more problems with contact bounce resulting in wrong entries when working long problems. The cost: about 0.5 cent.

Alf Wilson, W6NIF

varactor tuning tips

In tuning power varactor doublers, triplers, etc., there is often a sharp or sudden discontinuity in the tuning of one or more of the tuned circuits; a condition known as hysteresis.

While hysteresis is caused by some nonlinearities in the diode function, it seems that it may also be a result of the circuit *Q* aggravating diode nonlinearities. I figured that it might be possible to lessen the effect by a reduction in circuit *Q*. Accordingly, I reduced the bias resistor in my 144-to-432 MHz tripler from 92 to about 12. I was pleased to note that circuit performance was actually improved — tune up was easier, and there was no appreciable loss of power output.

Richard N. Coan, N3GN

power dissipation

Described here is a power-absorbing device commonly known as a dummy load. The circuit contains an active element so I have changed the name from dummy to active load.

an active load

The need for this circuit developed when I was trying to repair a 5-volt, 3-ampere power supply. No hot-dog-sized, 1.66-ohm resistors were available for load testing, so the circuit of **fig. 2A** was constructed and tested on the supply. Load current is controlled in both circuits **(figs. 2A** and **2B)** by R1. R2 limits the maximum base current to a safe value for the transistor used. One-hundred ohms is a nominal value. If the active load is to be used for more than a few seconds, adequate heatsinking must be provided for the transistor.

A provision for metering the current being consumed is included. I used the Simpson 260 volt ohmmeter on the 10-ampere scale.

other applications

This active load, when coupled to a properly designed heatsink, could be used in place of the Hot Mugger X1.[1] While these phenomena have not been fully investigated, an aluminum plate would probably exhibit an SWR of less than 3:1 over the operating range of the "coffee cup." Unfortunately, exact specifications for such a Hot Plate Matcher are beyond the scope of this article.

acknowledgments

I must acknowledge the contributions of David M. Newell, ex-K1KRG, who first introduced me to this circuit idea, and Donald S. Patterson, PS7ZAC, who developed the PNP version shown in **fig. 2A**.

reference

1. Burton, "The Hot Mugger X1," 73, February, 1979, page 163.

Wm. Denison Y. Rich, PS7ZAD

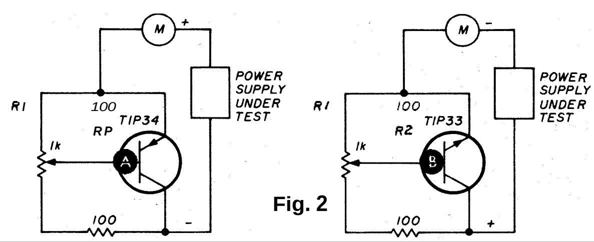

Fig. 2

ham radio cumulative index

1971-1980

a note on this index

To make the index easier to use only the years 1971-1980 are included, because most of the earlier material is now of limited interest. Refer to any December issue between 1970 and 1977 for a cumulative index covering 1968-1970. Copies of ham *radio* for December, 1977, may be purchased from Ham Radio's Bookstore for $2.50 postpaid.

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