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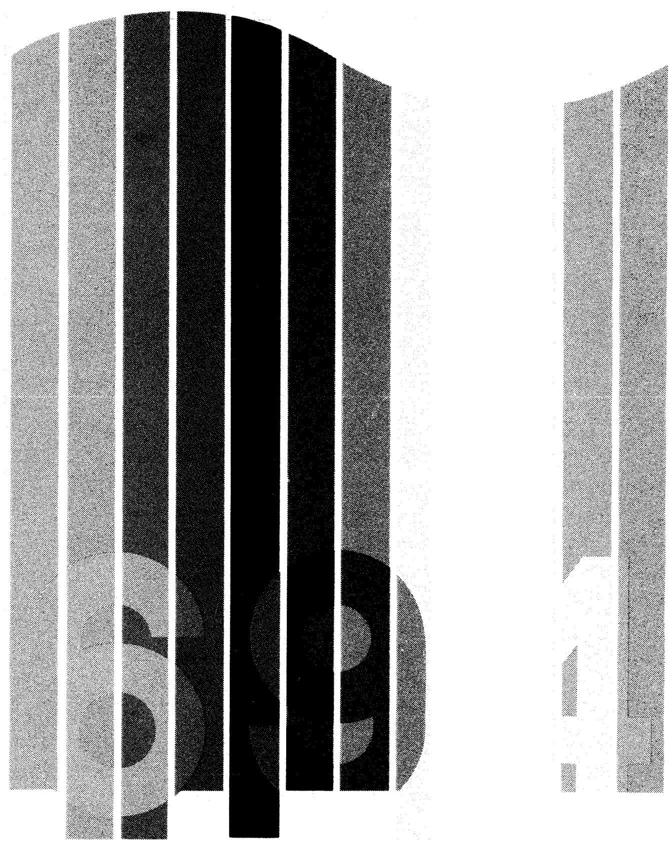

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JANUARY 1979

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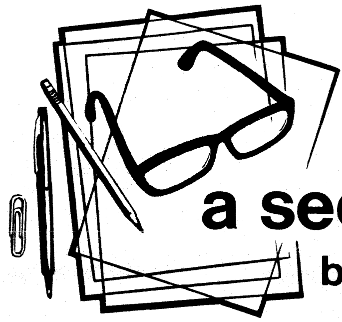


a second look

by Jim Fisk

As the editor I would like to think that our articles have more effect on you, the reader, than anything else; if not the most immediate effect, certainly the longest lasting. To be completely realistic, however, the one department which has the greatest impact on readers is circulation. If your issue is mangled by the Postal Service, or is late in arriving, or doesn't come at all, little time is wasted in letting our Circulation Manager know about it! I would hope that our response is just as immediate.

In the magazine business the word "fulfillment" is used to describe the internal business procedures which ensure that you get your mailed copy each and every month of your paid subscription. All magazines use a computer for this task, and we're no different. In the past all subscription orders were keypunched here in Greenville, and the punched cards were sent on to a computer house in Boston which filed the information on magnetic tape in zip code order. That two-step procedure has worked well for a number of years, but the growth of *ham radio* and the introduction of our sister publication, *Ham Radio Horizons,* has begun to strain the system. To both reduce errors and improve service to our subscribers, we recently contracted with a professional magazine circulation fulfillment service to do the entire task. That means that the subscription information must be transferred from one computer to another.

If this were a perfect world the changeover would go without a hitch, but Murphy's Law being what it is, there almost certainly will be some mistakes and garbled digits. We have instituted every safeguard we have available, but when you are faced with the humongous task of transferring nearly 50,000 names, callsigns, addresses, and subscription expiration dates, a few errors are inevitable. As the old saying goes, "Computers are not perfect – they're only as smart as the data given to them!"

We have been laying the groundwork for this changeover for several months, so we don't foresee any *major* problems. However, if your address label is garbled in the data transfer, please write to Ham Radio, Subscription Fulfillment Service, Post Office Box 711, Whitinsville, Massachusetts 01588. A correction will be made just as quickly as possible.

Although all subscription renewals, changes of address, and the like are to be mailed directly to our fulfillment service in Whitinsville, all correspondence to our editors or advertising department must be sent to our offices in Greenville. In the past, when readers have written to us about a subscription matter, they have often taken that opportunity to pose a question to our staff, or to comment on one of our previous articles. Such comments and questions are immensely useful as we plan future material for the magazine, but in the future such questions and comments should be separated from subscription matter and mailed directly to Greenville. Otherwise our staff won't have the benefit of your suggestions.

If you have an occasion to write to our fulfillment service in Whitinsville, please be patient (for fastest service, be sure to include the mailing label). Just remember that the computer does its work several weeks before the magazine goes into the mail, so there is considerable lead time involved (up to six weeks). This presents a problem for us, too, because we won't know a mistake has been made until you tell us about it, and we won't be certain the problem has been corrected until the computer prints the address labels for the next issue. However, with patience and understanding from you, our readers, the task will go much more smoothly. Thank you for your help.

Jim Fisk, W1HR
editor-in-chief

comments

bandspreading techniques

Dear HR:

The suggestions by Robert Heider, WØEJO, on a different form of standard capacitor ("Bandspreading Techniques," February 1977) are deserving of some comment.

The technique of using a guard for accurate calibration is a good one, but it must be used with care. The following points apply:

1. Careful examination of the referenced article's **fig. 6** will show the variable insulated from the case and thus guarded if the case ground is used.

2. Guard circuits must be used with bridges designed for that purpose (see, for example, *Electronic Measurement* by Terman and Pettit, McGraw-Hill, 1953, pages 102,103).

3. Guard circuits used with "ordinary" bridges, Q Meters, etc., will simply include the coaxial cable capacitance. In this case it would be better to use ***two*** coax cables, one on each capacitor terminal, in order to halve the stray capacitance. Cables should remain in place for resonating with small values of C, since they contribute around 10 to 20 pF shunt.

Any sort of coax at high frequencies (about 1 MHz or above) will begin to show transmission line effects. A couple of feet of coax at 30 MHz will change calibration at the low end of the C range, compared with measurements made at 3 MHz. The series inductance addition was implied when the article suggested "heavy wire and short lengths."

It's surprising that even pros and old timers forget inductance of connection. Small inductance of leads makes the original noise bridge wideband. The largest stray inductance in the Hart bridge is the potentiometer arm connection, physically variable due to mechanical rotation. The arm is effectively balanced to ground on both sides of the bridge into the vhf region.

Leonard H. Anderson
Sun Valley, California

Dear HR:

After reading the September issue of *ham radio,* I feel the urge to write. That one issue paid for the whole year's subscription! The article by W7VK on CATV cable fittings was a godsend to me – I've had twelve 112-inch-cable to SO-239 fittings on order for five months, at $7.50 each. I called them up and said send my money back.

The article by K1XX of your staff gave me the "big answer" for matching 75-ohm hardline. Thanks guys!

By the way, when preparing hardline for use with fittings that terminate in a type N female, you must file the end of the center conductor into a round-nosed bullet shape, because the square end will spread and break the female pin on the cable side of the fitting. When buying male type-N connectors, be advised that thk Amphenol #82-61 is a 50-ohm connector, and the center pin will break the female pin in a 70-ohm hardline fitting. For the 70-ohm connector the Amphenol part number is 82-84, and, apparently they are not a stock item, at least around here. The foregoing tips cost me $22.50 to learn. Pass the word, and save others some money.

Thanks again for a great magazine and if you would like some hardline look me up. I've got a bunch!

Don Ryan, WB4NND
Virginia Beach, Virginia

high resolution hf synthesizer

Dear HR:

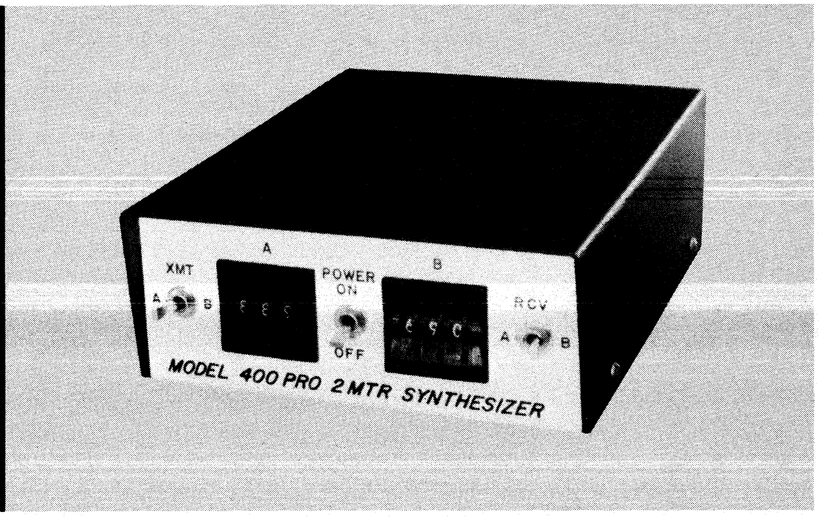
The article in the August issue, "High-Frequency Resolution for an HF Synthesizer," describes a principle which is used in the Collins 651S1 receiver for which Collins Radio has a patent. While this system is apparently very attractive at first glance, it has the following disadvantages:

1. Because of the frequency selection, a large number of birdies are present if filtering is inadequate.

2. The theoretical high-speed locking is degraded because of the algebraic logic. This slows down the synthesizer so much that it can't be used efficiently for search operation. In the 651S1 receiver Collins engineers used an out-of-lock detector which mutes the receiver for virtually all tuning.

Finally, I would like to point out that my company holds the patent for the combination up/down counter with optical shaft encoder which was suggested for frequency synthesizer control (patent 97780, issued July 13, 1962).

Ulrich L. Rohde, DJ2LR
Rohde & Schwarz Sales Company
Fairfield, New Jersey



800-channel 2-meter synthesizer

A 400-channel 2-meter synthesizer featuring single-board construction, 15-kHz splits, and multiple output frequencies

Back in 1975, after operating a converted Motorola 80D for several years, I decided to move up to a more versatile rig that I could synthesize and also use mobile. I looked at the available commercial rigs and synthesizers, every construction article I could locate, and talked to fellow hams who had gone this route. I found several rigs I liked, but no synthesizers; so, I decided to buy a rig (an HW202) and build a synthesizer. My first impulse was to build one that had appeared in a magazine construction article. I studied the circuit and started trying to locate parts, becoming quickly discouraged. After talking with a local ham who built a synthesizer from the same article, I was further discouraged.

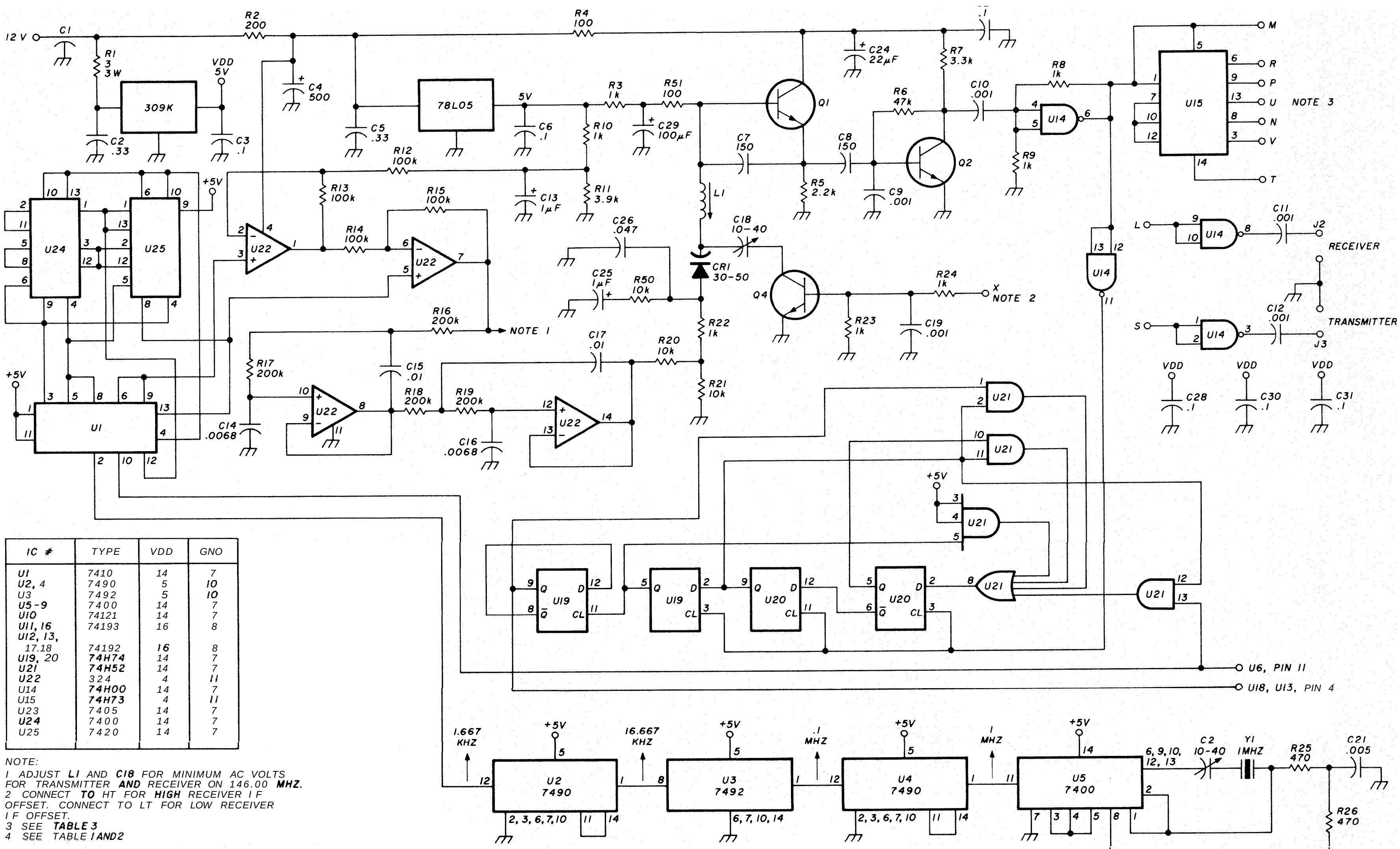
At this point, I decided to design and build from scratch. Having found several rigs that I liked, I felt the synthesizer should be universal enough that I could simply reprogram the i-f offset and output frequency for another rig. In addition I wanted only one crystal oscillator, since that would reduce the stability problems and eliminate spurious outputs associated with mixers. I also wanted the entire synthesizer on one board; parts had to be inexpensive and easy to get. The end product was called the "400 PRO" (400 channels receive, 400 channels transmit with Programmable Receive Offset).

circuit description

The crystal oscillator, as shown in **fig. 1,** determines the overall frequency stability of the synthesizer. The crystal is a 1-MHz, parallel resonant cut for 32-pF load capacitance. For temperature stability, the crystal tolerance should be no more than .003 per cent from −23.5 to 66 degrees C (−10 to 150° F). U5 is a 7400 TTL NAND gate used as the oscillator.

U2, U3, and U4 divide the 1-MHz frequency by 600, producing a 1.666-kHz reference for the phase detector. The phase detector is made up of U1, U24, and U25. These three ICs were chosen over the more popular MC4044 phase detector strictly for a cost savings of about $1.50.

By Bob Fanning, K4VB and Gary Grantland, WA4GJT. Mr. Fannings's address is 1332 Four Mile Post Road, Huntsville, Alabama 35802. Mr. Grantland's residence is RFD 2, Somerville, Alabama 35670.



IC #	TYPE	VDD	GNO
U1	7410	14	7
U2, 4	7490	5	10
U3	7492	5	10
U5-9	7400	14	7
U10	74121	14	7
U11, 16	74193	16	8
U12, 13, 17.18	74192	16	8
U19, 20	74H74	14	7
U21	74H52	14	7
U22	324	4	11
U14	74H00	14	7
U15	74H73	4	11
U23	7405	14	7
U24	7400	14	7
U25	7420	14	7

NOTE:
1 ADJUST L1 AND C18 FOR MINIMUM AC VOLTS FOR TRANSMITTER AND RECEIVER ON 146.00 MHZ.
2 CONNECT TO HT FOR HIGH RECEIVER I F OFFSET. CONNECT TO LT FOR LOW RECEIVER I F OFFSET.
3 SEE TABLE 3
4 SEE TABLE 1 AND 2

fig. 1. Schematic diagram of the **VCO**, phase detector, active filter, and crystal divider portions of the two-meter synthesizer. Starting with the basic 1-MHz crystal frequency, the counters divide this down to 1.667 kHz for one input to the phase detector. The second phase detector input comes from pin 11 of U6 in the counter section (see fig. 2). To keep down costs, the phase detector in this synthesizer is composed of three individual **ICs (U1**, U24, and **U25)** instead of a single dedicated **IC**. The active filter, **U22**, attenuates the 1.667 kHz sidebands on the error signal from the phase detector before it is applied to the **VCO**. As discussed in the text, **L1** and **C18** should be adjusted for **minimum** ac signal on pin 7 of **U22**. The input to Q4 should be connected to HT for receivers with a high i-f offset, or to LT for a low i-f offset. The output from U15 can be strapped to provide the appropriate frequency for the transmitter to be synthesized.

The next portion of the circuit description may be a little more difficult to understand. There are two counter chains and a two-modulus prescaler which make up the dividers necessary to divide the VCO output frequency down to 1.666 kHz (see **fig. 2**). U19, U20, and U21 make up the two-modulus prescaler. This circuit is arranged to divide the VCO frequency by 10 or 11, depending on the dc level at pin 11 of U6. The output of the two-modulus prescaler is fed to both counter chains.

The channel select divider, U11, U12, and U13, is programmed to divide by 400 plus the thumbwheel switch setting. If a frequency of 146.94 MHz is selected on the thumbwheel switches (6.94 is selected since all channels are in the 140-MHz band), the channel select divider divides by 400 plus 694, which equals 1094. This method is used so that it is impossible to select a frequency out of the 144 to 148 MHz range. When in the transmit mode, the i-f program divider, U16, U17, and U18, is always programmed to divide by 1360. When in the receive mode, the i-f frequency of the receiver being used is subtracted or added to 1360. For example, if the receiver has an i-f frequency of 10.7 MHz, high side injection, then 107 would be added to the 1360 when in the receive mode. This is done by programming the i-f program counter to the HT and LT buss on the printed circuit board. The HT buss is high, or a logic 1, when in transmit; it is low, or a logic 0, when in the receive mode. The LT buss is just the opposite of the HT buss. A ground buss and a +5 volt buss are also provided for the counter inputs, which remain the same in transmit and receive. The i-f program counter can be programmed to divide by 1200 to 1599, which corresponds to minus 16.0 to plus 23.9 MHz in 100-kHz steps. See **table 1** for some standard i-f programming information and **table 2** for developing any i-f program.

table 1. Connections for the i-f offset dividers for standard receiver offsets.

Receiver Offset	A	B	C	D	E	F	G	H	J	K	X
+5.5 MHz	HT	LT	LT	HT	0	HT	LT	LT	0	0	HT
−5.5 MHz	1	0	0	HT	0	HT	LT	LT	0	0	LT
t8.0 MHz	HT	LT	0	1	0	HT	0	0	0	0	HT
−8.0 MHz	HT	0	0	HT	LT	HT	0	0	0	0	LT
+10.7 MHz	HT	LT	0	1	0	1	LT	LT	0	LT	HT
−10.7 MHz	HT	0	LT	1	0	HT	LT	0	0	LT	LT
+12 MHz	HT	LT	0	HT	LT	HT	0	0	0	0	HT
−12 MHz	HT	0	0	1	0	HT	0	0	0	0	LT
+13.1 MHz	HT	LT	LT	HT	LT	HT	LT	0	0	0	HT
−13.1 MHz	HT	0	0	HT	0	1	LT	0	LT	0	LT
−11.7 MHz	HT	0	0	1	0	HT	LT	0	0	LT	LT

The output of U16, a negative-going 20 ns pulse, is lengthened to about 50 nS by U10. The subsequent output of U10 is inverted and buffered by U6 and used to load the counter chains with the jam inputs when the i-f program counter counts down to zero. The remainder of U6, which is connected as an RS latch, has a high output when the channel select divider counts down to zero, and a low output when the i-f program divider counts down to zero. When the output, pin 11, is low, the two-modulus prescaler divides by eleven; when pin 11 is high, the prescaler divides by ten.

Assume that the receive frequency is 146.94 MHz and the receiver has an i-f frequency of 10.7 MHz, with low-side injection. The i-f program divider would be set at $1360 - 107 = 1253$. The channel select divider would be set at $400 + 694 = 1094$. The outputs of U6 load both counter chains and set the prescaler to divide by eleven. After 1094 counts, the prescaler is set to divide by ten; the prescaler is reset to divide by eleven after $1253 - 1094 = 159$ counts. Therefore, the prescaler divides by eleven 1094 times and divides by ten 159 times. The frequency input to prescaler required to receive 146.94 MHz would be

table 2. Example of programming for the i-f offset counters. The BCD 800 and 400 are preprogrammed.

program terminal number	none	none	B	A	E	D	F	C	J	H	K	G
terminal BCD value	800	400	200	100	80	40	20	10	8	4	2	1
transmit value 1360	1	1	0	1	0	1	1	0	0	0	0	0
receive value 1467	1	1	1	0	0	1	1	0	0	1	1	1
your program	+5V	+5V	LT	HT	GND	+5V	+5V	GND	GND	LT	LT	LT

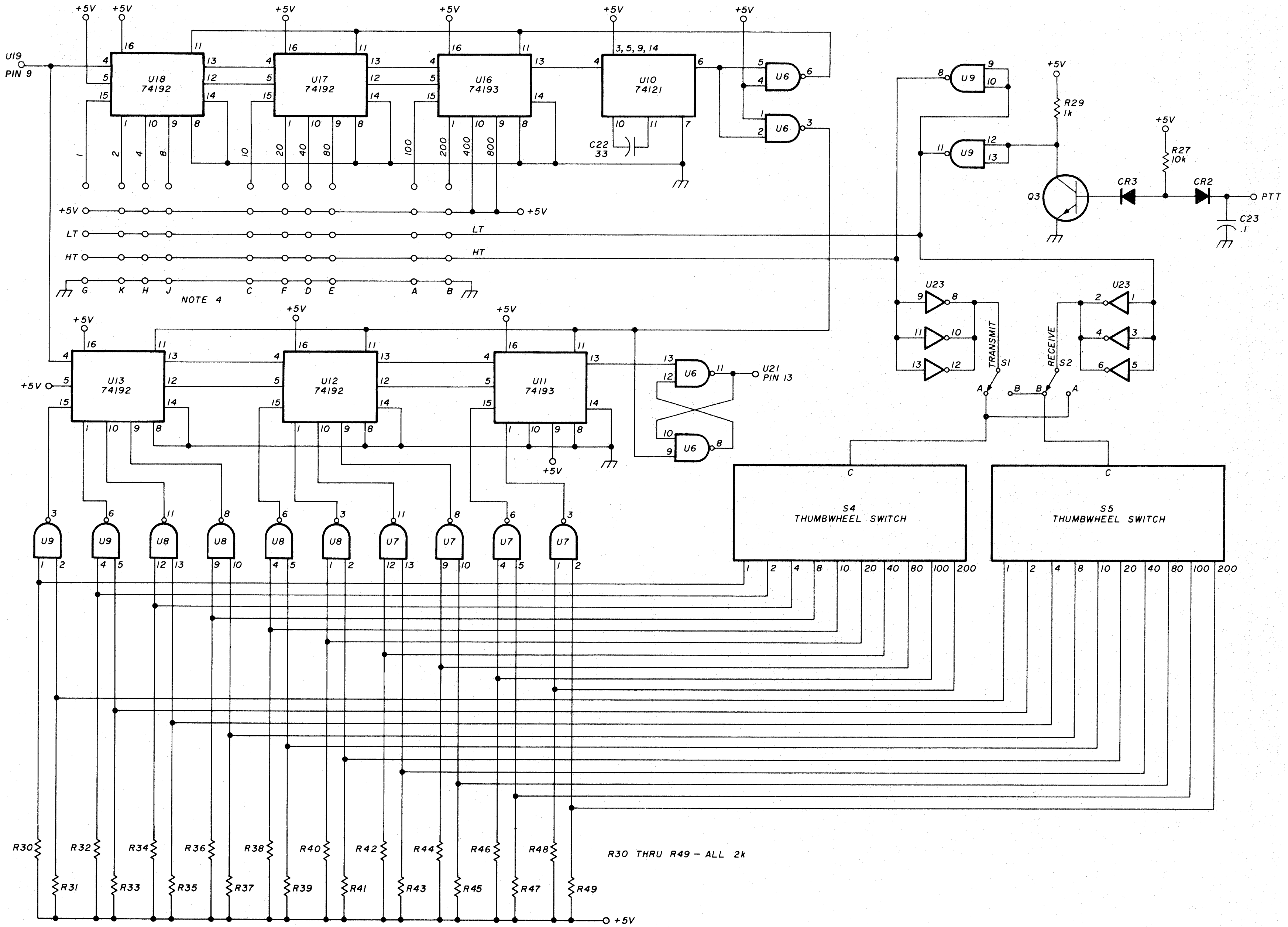


fig. 2. Diagram of the counter portion of the synthesizer. **U11, U12,** and U13 make up the channel select dividers, dividing by 400 plus the frequency selected. For example, $400 + 625 = 1052$ for 146.52 MHz. The i-f program dividers, **U16, U17,** and **U18,** are normally programmed to divide by 1360 when in the transmit mode. When receiving. the dividers are programmed by jumpers to change the divide number according to the receiver i-f offset used.

$(146.94 \times 10^6 - 10.7 \times 10^6)/6 = 22.7066 \times 10^6$. Divided by $11 \cdot (1094) + 10 \cdot (159) = 1.66666 \times 10^3$, or the same as the output of the reference frequency divider.

These two signals, when compared by the phase detector, generate an error voltage which controls CR1. CR1, C18, and L1 make up the VCO tank circuit. The capacitance of CR1 can be varied from 30 to 80 pF, depending on the bias voltage. L1 is adjustable from 0.5 to 1 uH. L1 and C18 are adjusted to center the VCO frequency. Before adjusting L1 and C18, the i-f program counter chain should be programmed for the i-f frequency of the receiver and the channel select divider should be set to 146.00 MHz. For receivers with low-side injection, adjust L1 in transmit and C18 in receive mode for minimum ac voltage at pin 7 of U22. This can be done with an oscilloscope or ac voltmeter. Adjust L1 in receive and C18 in transmit for receivers with high-side injection. This adjustment is not critical and will affect only the frequency deviation of the VCO. Terminal X

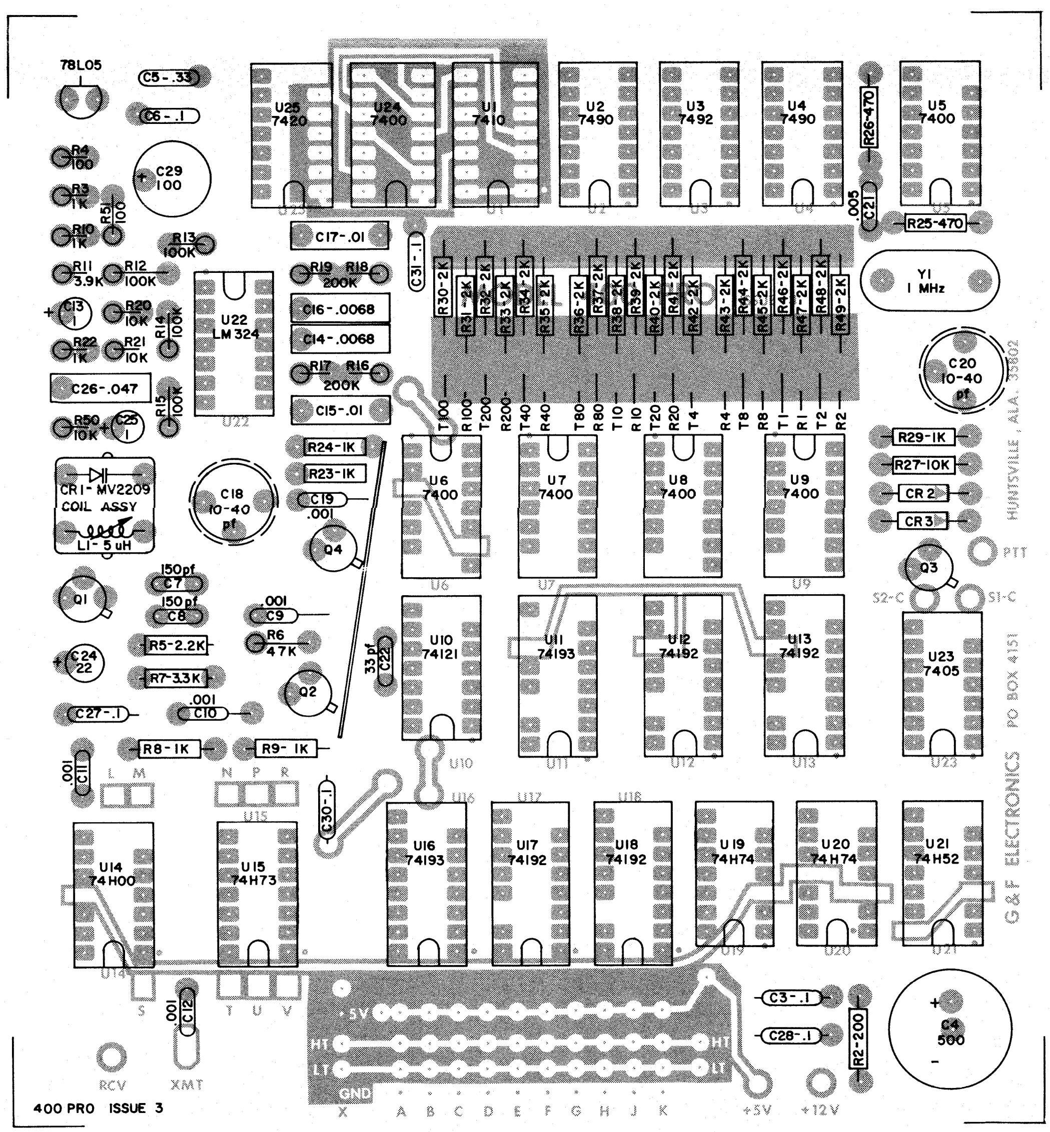


fig. 3. Parts placement diagram for the circuit board of the two-meter synthesizer. The jumpers for the output and i-f offset frequencies are explained in the tables.

the VCO. There are two outputs from the phase detector which drive pins 3 and 5 of U22. If no error voltage is generated, the level at pin 7 of U22 is 4 Vdc. If an error voltage is present, negative or positive pulses will be present at pin 7 of U22. The remainder of U22 is a four-pole, lowpass filter. At the output of the filter, the 1.666-kHz signal is 85 dB below the input.

The dc voltage from the filter biases the varactor, is strapped to HT or LT and adds C18 to the tank circuit when X is high. Q2 and U14 buffer the VCO and drive the two-modulus prescaler and U15, a dual J-K flip-flop which may be programmed to divide by two, three, or four. The output is buffered by U14 to drive the transmitter and receiver crystal osciliators.

The PTT input should be grounded during transmit. This turns Q3 off. Q3 is buffered by U9 and provides the HT and LT levels to drive the thumbwheels used for transmit and receive. This provides any split or simplex operation on any set of thumbwheels.

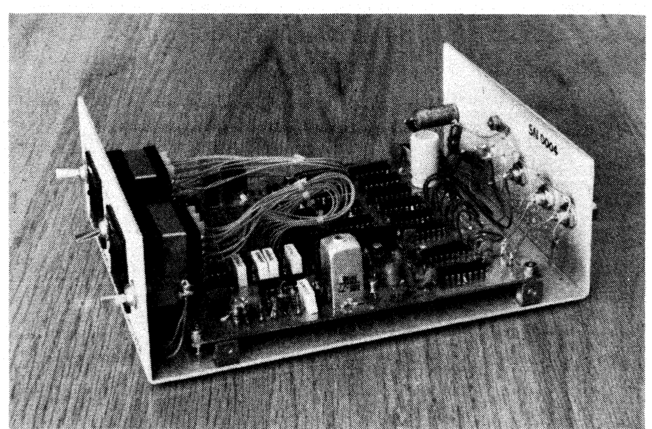
assembly

Assembly of the printed circuit board is straightforward and requires no special tools or techniques (it does require good grounding, however). Boards are available from the author* and will be by far the easiest route to go. If you prefer to lay out your own circuit board. care should be taken in the placement of components and traces to prevent the introduction of VCO whine and switching transients onto the signal lines. It is strongly recommended that a double-sided printed circuit board by used, as this is largely the secret to success in the virtual elimination of VCO shielding, the use of only one board, and the absence of "trash" on the output signal. Vector-type boards should not be used, because of the difficulty in attaining the high degree of shielding necessary. The very first 400 PRO built was on vector board. It was tough to get the output clean and it required two boards, one for the VCO and another for the rest of the synthesizer. It was also necessary to cover the entire VCO with a metal can and place a third copper-clad board between the two main boards, with everything securely grounded.

table 3. Connection to U15, the output divider, for different transmitter/receiver multiplication factors.

receiver multiplier	transmitter multiplier	jumpers
6	6	M-S, M-L
6	12	M-L, P-S, V-GND, R-HT
6	18	M-L, N-S, V-P, T-HT
6	24	M-L, N-S, T-HT
12	6	M-S, P-N, V-GND, R-LT
12	12	L-P, N-S, V-GND
12	18	U-L, V-N, P-S, R-HT
12	24	U-L, P-S, R-HT
18	6	M-S, L-N, V-P, R-HT
18	12	L-P, U-S, V-N, R-LT
18	18	L-N, V-P, N-S
24	6	M-S, L-N, T-LT
24	12	L-N, U-S, R-LT
24	24	L-N, P-S

Component placement and sizes should be kept as near as possible to those shown in the component placement drawing (**fig. 3**). All parts used, with the exception of the VCO coil and cover and copper pipes over C18 and C20, are available from most parts houses advertising in this magazine. For best results, IC sockets should not be used. However, if they are used, a good quality socket *is* a *must.* First-run ICs should be used if at all possible, as problems can be encountered with "discount house" ICs. As a general rule, most discount houses will quickly replace any bad IC. Nevertheless, replacement of bad ICs is little compensation, in many cases, for the misery encountered in finding them in a circuit.



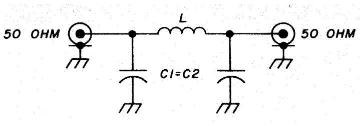
In this view you can see the i-f transformer can, which is used to shield the varactor and coil. Also notice that the board mounting technique provides four secure grounds.

A suitable source for the VCO coil and cover is the 6.5-mm (1/4-inch), four-terminal transformer used in the i-f of commercial fm receivers. Strip off the existing coil and wind 17 turns (close wound) of number 32 (0.2-mm) AWG enamel wire, terminating on two of the four terminals. (See **fig.** 3 for the correct terminals.) The MV-2209 varactor should be connected to the other two terminals, and both the coil and varactor should be covered with Q-dope and then placed back in the can. The can mounting tabs should be soldered to the ground point on the printed circuit board. If the VCO coil slug is not tight, the VCO will become sensitive to mechanical shock. If this occurs, the output of the VCO will sound just like a microphonic tube on both transmit and receive. The solution to this problem is fairly simple: after final VCO adjustment, apply a drop of candle wax or

*Double-sided, plated-through, G10 printed circuit boards with complete instructions ($18.75), completely assembled and tested boards ($89.00), and coil assembly with MV2209 ($3.50)are available from G&F Electronics, P.O. Box 4151, Huntsville, Alabama 35802.

similar material onto the slug. Remember, if you should later change to a rig that has a different i-f offset, the VCO may require a touchup. Don't lock the slug down too tightly.

The covers over C18 and C20 were made from 9.5-mm (3/8 inch) copper tubing cut in 12.5-mm (1/2-inch) lengths and soldered to the shield side of the printed circuit board. Care should be taken to prevent the inside of the copper tube from shorting



FREQUENCY	L	C1=C2
6	3.3 µH	220 pF
8	3.3 µH	180 pF
12	2.2 µH	100 pF
24	1.0 µH	47 pF
48	.47 µH	22 pF
72	.33 µH	15 pF

fig. 4. Lowpass filter that can be used on the output of the synthesizer.

the capacitor to ground. A piece of thin Mylar or Teflon sheet wrapped on the inside of the tubing will serve as a spacer. The Mylar used in drafting departments for taping printed circuit board masters is excellent for this purpose. The tubing should be quickly soldered in place using a hot iron. A thin strip of copper 1 mm (0.03 inch) thick should be soldered in place on the component side between Q2, Q4, and U6, U10 as shown in **fig.** 3. This 5 cm x 1.3 cm (2 x 0.5 inch) shield is sometimes needed to prevent VCO whine. Since it is cheap and easy to install, one should be used as a preventive measure.

The enclosure shown in the photos was handmade from 1.5-mm (0.062-inch) aluminum sheet and painted with enamel paint. The lettering was applied using a Leroy lettering set with black India ink. Press-type, dry-transfer decals will also work quite well. The lettered surface should be protected with a clear Krylon spray.

The enclosure shown in the photographs has no means of ventilation. Ventilation is recommended to reduce heat build-up, thus improving frequency stability. The unit shown was accidentally left on inside a closed vehicle on a 38°C (100°F) day with no ill effects, except for the fact that the cover became hot enough to burn your hand.

Most importantly, the box used should be one with good rf integrity. This usually means all metal with good metal-to-metal contact between pieces. Plastic panels should not be used, nor should metal boxes with vinyl contact coating. The Radio Shack box 270-254 is the most suitable commercially available one I found. It costs about $5.00 and has only one shortcoming: The front panel is thin-gauge aluminum which is a little too flimsy and thin for Digitran 2300-series thumbwheels. This problem can be corrected by gluing strips of aluminum behind the front panel and on each side of the thumbwheel switches to act as shims.

The printed circuit board should be mounted at all four corners with a good solid ground connection to the box. Make sure that your spacers do not extend beyond the copper pads provided on the circuit side of the printed circuit board, as this can cause portions of the circuit to short to ground.

The printed circuit board should be positioned such that the transmit and receive outputs are just below the rf output connections on back of the box. Lowpass filters should be used between the printed circuit board and rf output connectors (see **fig. 4** for values). If lowpass filters are not used, short pieces of buss wire (2.5 mm [1 inch] or less) should be used.

The 3-ohm, 3-watt resistor (R1) should be connected directly to the LM309K input pin and clamped against the rear panel for mechanical strength and heat transfer. C2 and C3 should be connected to the LM309K terminals and grounded to a lug under one of the 309K mounting screws (C2 and C3 should be ceramic). The +12 volt input should enter the box through a 0.01-µF feedthrough capacitor.

If needed, the PTT line can enter through a feedthrough capacitor. However, no situation has been encountered where an RCA phono jack with a 0.1-µF capacitor to ground would not suffice. All rf output connectors are RCA phono-type, but any good quality rf connector with a good ground connection will work well.

The thumbwheel switches used are the Digitran 2300-series. Any thumbwheel or rotary switch providing BCD output can be used. Lever-type thumbwheels were not used for fear of unintentional frequency changes caused by the microphone cord. Rotary switches were not used because the larger front panel required was not desirable.

programming the synthesizer

Jumper wires should be used to connect terminals

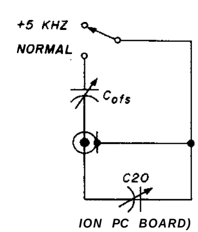


fig. 5. To use the synthesizer on repeaters that are on 15-kHz splits, the time base is shifted in frequency by the additional capacitor, C_{ofs}. This capacitor generates a time base error that shifts the output frequency the necessary 15 kHz.

A through K to the HT, LT, +5 volts, or GND buss to program the synthesizer to match the i-f frequency of your receiver. Terminal X should be connected to HT for receivers that use high-side injection, and connected to LT for receivers that use low-side injection. **Table 1** lists several i-f frequencies and the proper connections for programming. Receivers with i-f frequencies, other than those listed in **table 1,** can be programmed using the following example. **Table 2** lists the BCD values that correspond to terminals A through K. The left-hand column lists the transmit and receive values that should be programmed. The transmit value is always 1360. The receive value is determined by the receiver i-f frequency. If the receiver i-f frequency is 10.7 MHz with high-side injection, 107 should be added to 1360 for the receive value. The 800 and 400 values are pre-wired on the printed circuit board and do not have to be programmed. In the transmit frequency row, write the BCD value which equals 1360. In the receive row, write the BCD equivalent of the receive value. Under the receive row, write HT for all values which are 0 in transmit and 1 in receive. Write +5 volts for all values which are 1 in transmit and 1 in receive. Write GND for all values which are 0 in transmit and receive. These are the values which should be programmed to terminals A through K.

Output divider program is given in **table** 3. Transmit and receive divider should be the same as the crystal multiplier for your transceiver. Output divider programming terminals L through V are the square pads located at either end of U14 and U15.

interfacing

A proper interface between the 400 PRO and your transceiver is a must for satisfactory performance. With the 400 PRO, the transceiver's oscillator (transmit and receive) must operate as a buffer amplifier instead of an oscillator. The drive level at the transceiver must be the same as if a crystal were still being used. Drive level from the 400 PRO is in most cases higher than that of a crystal. It should be reduced, as required, by the addition of series resistance at the point of interface. Any additional interfacing components should be as close to the interfacing point (usually the crystal socket) as possible. Optimum results are obtained if connections are soldered at the interface point. Satisfactory results should be obtained by plugging into an unused crystal socket. If this method is used, insure that all mechanical connections are solid.

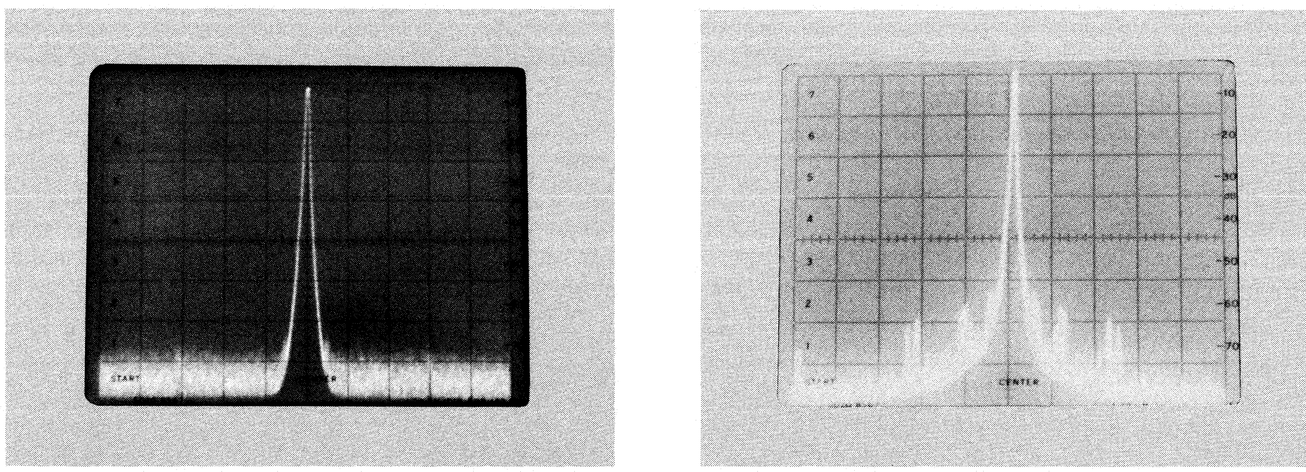


fig. 6. Output of the synthesizer as seen on the display of a spectrum analyzer. At left, the horizontal scale is 2 MHz/division, while the vertical is 10 dB/division; right, the horizontal scale calibration has been shifted to200 kHz/division.

With some transceivers, one side of the crystal is switched with the channel selector while the other side is connected to ground through a trimmer capacitor. Interfacing with this type of transceiver is best accomplished by adding a solid chassis ground to which the 400 PRO ground (shield side of coax) should be connected. The center conductor is then connected to the switched side of the crystal socket (through the proper interfacing components).

The basic requirement for operation is adequate drive at the appropriate frequency, which need not necessarily be exactly the same as the crystal it replaces. For example, if your receiver requires a 48-MHz crystal (as does the HW202), it will work quite well with a 24-MHz signal. The oscillator will then act as a doubler. Other transceivers require a 15-MHz crystal, which is multiplied by nine. These oscillators will operate with a 23-MHz input multiplied by six.

The 400 PRO, as wired in this article, will work the entire two-meter band. Few transceivers are broad enough to allow full power output or maximum sensitivity across the entire band. Don't be alarmed if a couple of MHz is your limit. This problem is very pronounced with commercial-band radios such as Motorola and GE.

When interfaced properly, the transceiver will work just the same with the 400 PRO as with a crystal. If it does not, it is either not interfaced properly or other problems exist. The most common other problem is rf leakage into the 400 PRO. The symptoms of rf leakage are distorted or low frequency audio on transmit. This distortion may come and go when the equipment is moved. In some cases, movement of the microphone, rf or audio cables, nearby objects, or even people can cause the intermittent distortion. In severe cases, the movement of one's hand near the equipment can cause distortion. The most common cause is poor or missing ground connections. Signals can also be coupled in via either the transmitter, receiver, PTT, or the +12 volt line. Any problems entering on the +12 volt or PTT line can be eliminated by the use of feedthrough capacitors. Coupling through the transmitter or receiver lines can be eliminated by the addition of a lowpass filter inside the 400 PRO. In extreme cases, an additional lowpass filter may be required at the transmitter end of the coaxial cable. The filter can be the same as in the 400 PRO and should be mounted inside the transceiver at the point of entry; ground it well. If filtering is done after interfacing, recheck the drive levels.

The 400 PRO can be used in one crystal position, with crystals used in others. With this set-up, the 400 PRO should be turned off any time the transceiver is used in a crystal mode. Failure to do so can result in spurious output in the transmit mode and unwanted birdies in the receive mode.

Interconnecting cables between the 400 PRO and your transceiver should be a good quality 50-ohm coax; the miniature RG-174 works very well. *Never* use audio cable, shielded or unshielded. Good grounding is an absolute must. Short interconnecting cables are obviously best, but several 400 PRO's are now being used with trunk-mounted, commercial-band equipment. (I use one with a Motrac U43MHT.) Miniature coax is used between the dash-mounted 400 PRO and the trunk-mounted radio. One potential problem with commercial-band radios is high resistance in the tensile cord push-to-talk line. This manifests itself in the inability to switch the 400 PRO from receive to transmit. The problem can be corrected by either replacing the microphone cord or reducing the resistance such that Q3 will switch when the PTT button is pressed.

direct fm

Most transmitters are phase modulated instead of true fm. However, for those that are true fm, the 400 PRO will not simply plug into the crystal socket and function. The phase lock action of the 400 PRO will not allow the output frequency to be shifted. To solve this problem, the VCO in the 400 PRO must be direct-fm modulated. This is accomplished by applying audio from the transceiver directly to the 400 PRO VCO. Fm produced in this manner is of superior quality. The 400 PRO can be frequency modulated when used with either fm or pm rigs.

5-kHz offset

The 400 PRO was not designed for 5-kHz output steps. A very simple modification, however, can be made that will provide 5-kHz output increments. This is accomplished by the addition of a toggle-switch selectable capacitor (see **fig. 5**) which will alter the divide ratios. The change will result in an error of 34 Hz/MHz at the two-meter output frequency. For example, if the 400 PRO is set up for no error at 146.005 MHz, the transceiver will exhibit a 67-Hz error at each end of the band (144.000 to 147.995). When not in the 5-kHz mode, the 400 PRO will operate normally and will have no error caused by this modification.

The off-set capacitor shown in **fig.** 5 can be mounted on the printed circuit board or at the toggle switch. Interconnecting cabling (RG-174) should be kept as short as is practical (15 cm [6 inches] or less).

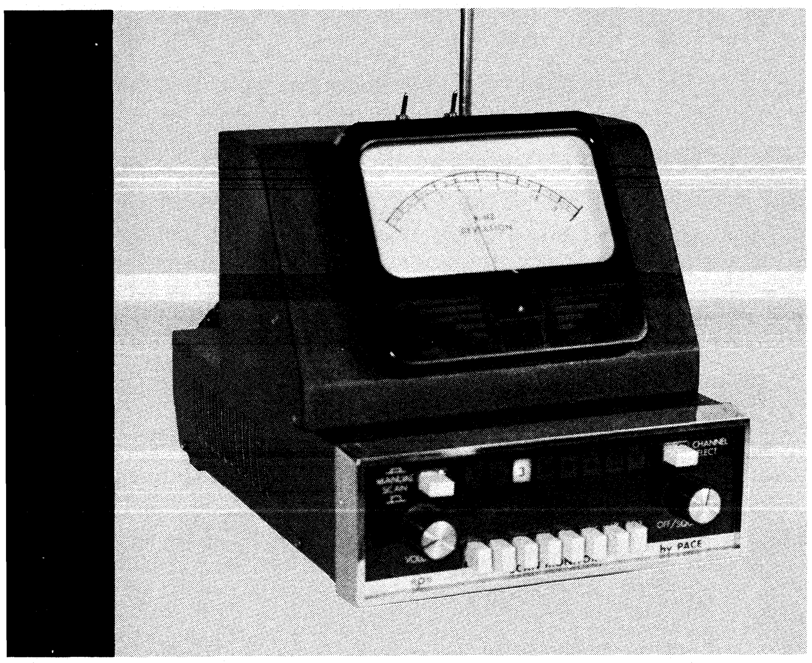
The easiest method found for initial setup is to calculate the required transmit frequency, with and without +5 kHz off-set. With the off-set selected, adjust C20 for the required frequency. Switch back to the normal position and pull the frequency back down as required with C_{ofs}. You may have to repeat this procedure a couple of times. The transmit mode was chosen for initial setup because of the calculation — no i-f off-set is involved.

After the 400 PRO is interfaced to your rig and working properly, the same adjustment procedure can be used while monitoring your transmitter output frequency with a frequency counter. Once properly adjusted in the transmit mode, the receive frequency is automatically set and will require no further adjustment.

conclusion

Although many 400 PROs have been built and used with a variety of rigs, additional interfacing information is always welcome. If you encounter any unusual interfacing problems to which you find a solution, it would be greatly appreciated if you would jot down the details and mail them to me. By the same token, if you encounter problems you cannot solve, I will be glad to try to help you. The following information must be furnished: transmit and receive crystal formulas, multiplier arrangement, and a schematic. If you are going to have a problem with the 400 PRO, it will most likely be in interfacing it to your rig.

ham radio



optimizing and measuring fm deviation

Louder isn't necessarily better in an fm transmitter — intelligent use of the deviation control will produce maximum talk power per watt

In fm radio transmission, the louder the modulating signal into the microphone the greater the variation of the carrier from its center, or resting, frequency. Generally this also means a stronger audio signal at the detector of a properly tuned fm receiver. The amount of the plus or minus frequency swing is called *deviation.* Deviation is talk power, and that's good – most of the time but not always.

There are three conditions under which increased deviation is not to the user's advantage. First, FCC Regulation Part 97.65 limits the deviation to ±3 kHz on frequencies below 29.0 MHz and between 50.1 and 52.5 MHz. Deviation is limited to ±20 kHz for *all other* authorized Amateur Radio frequencies. Continued violation of the regulation could result in zero talk power. Get the message?

Second, particularly in the 2-meter band, interference between stations operating on different frequencies is minimized if the portions of the frequency spectrum used by each of the stations don't overlap. This is shown graphically in **fig. 1.** The amateur community long ago recognized the potential for a problem and used some of its better thinking to find ways for avoiding the interference. The solution is best seen in the band plans for 2-meter repeater operation, where the generally used frequencies are 10 kHz apart, corresponding approximately to a ±5 kHz deviation limit.

The decision to self-impose a 5-kHz deviation limit when the FCC was offering up almost four times that much wasn't entirely magnanimous. (Required band-

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width is approximately 2 x deviation + maximum modulating frequency. Thus, for a 3-kHz modulating frequency, and FCC permitted maximum deviation, the required bandwidth would be ~~10 3~~ = *18.5 kHz.)* This becomes evident in an examination of the third condition, under which it's to the user's advantage to limit the deviation.

modulation index

In fm radio transmission, the total power radiated is independent of the modulation. When the carrier is modulated (deviated), power is transferred from the carrier to the intelligence sidebands. The amount of power that's transferred depends on the modulating-signal amplitude and something called the modulation index (MI). We noted previously that the achieved deviation varies with the modulating-signal amplitude. The modulation index is the ratio of the frequency deviation to the modulating-signal fre- ***quency.***

For example, if a transmitter deviates 5 kHz with a modulating signal of 2 kHz, at some amplitude the modulation index would be $5/2 = 2.5$. The relationship between the relative amount of power in the carrier and sidebands and the modulation index is shown in fig. 2. It's evident that the maximum amount of power is transferred to the sidebands when the modulation index is approximately 2.4; that is,

$$MI = \frac{Deu}{F_m} = 2.4 \qquad \textbf{(1)}$$

audio bandwidth

Students of audio-frequency phenomena know that, although the full spectrum of human speech is between approximately 100 and 8000 Hz, only a small fraction of that range is normally used. If everything below 1000 Hz is filtered, comprehension is not affected, but the result is a mechanical, unnatural sound. If frequencies above 1000 Hz are filtered, the result is a varying amplitude mumble that's almost devoid of intelligence. As more and more of the frequencies above 1 kHz are permitted to pass, comprehension increases rapidly until about 1800 or 2000 Hz is reached. Comprehension then increases less rapidly until, by 3000 Hz, almost nothing is added to comprehensibility by increasing bandwidth. If **eq. 1** is examined in light of this:

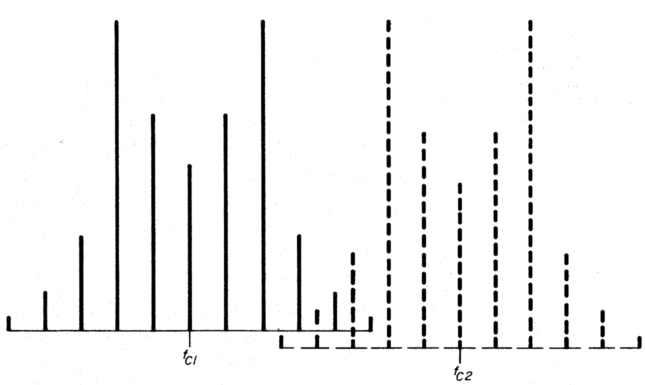


fig. 1. With center frequencies 10 kHz apart, two fm stations will still experience some mutual interference if each modulates to 4.5-kHz deviation with a 1.5-kHz tone.

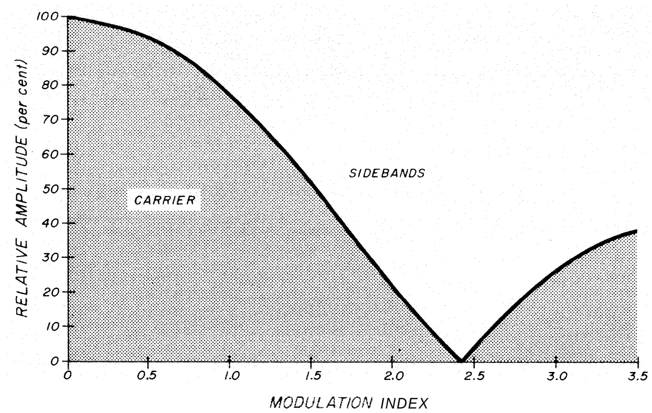


fig. 2. Relative amplitude of the rf power remaining in the carrier as a function of modulation index.

$$MI = \frac{Dev}{F_m} = 2.4$$

$$Deu = F_m \times 2.4 = (1\ kHz\ to\ 2\ kHz) \times 2.4 \qquad \textbf{(2)}$$

$$= 2.4\ kHz\ to\ 4.8\ kHz$$

Thus, the most effective way to use the rf power and obtain maximum readability is to hold the achieved deviation between about 2.5 and 4.5 kHz; the lower number for bass-voiced males, the higher number for tenors and sopranos. Thus it's possible at one and the same time to keep the FCC and fellow hams happy – or, at least, off your back – and to make efficient use of the rf power from the transmitter. Adjusting the transmitter to achieve the desired and maximum deviation isn't difficult. It consists solely of tweaking a couple of potentiometers while making a few measurements.

audio gain and deviation controls

Typically, the signal from the microphone is ac-coupled to a one- or two-stage audio amplifier; then to a clipper and filter, then to a deviation-level control stage, and finally to the modulator. Most, but not all, transmitters have a means of gain adjustment in the audio amplifier stage. All fm transmitters should have a potentiometer for controlling deviation-level set, or, maximum deviation. Really good matching of the microphone, the user's speech characteristics, and the transmitter can be accomplished only when both controls are present. If only a deviation-level set control is available a useful degree of optimization is still possible, although the user may have to experiment with different microphones to get the best match.

an analogy

To visualize the interaction between audio gain control and deviation level set, it's convenient to use a simple analogy. Picture a small system made up of a water well, an electric pump, a faucet, and the connecting pipes. It's clear that until the pump is turned on no water will flow, faucet open or closed. Now, turn on the pump with the faucet closed and still no water flows. As the faucet is gradually opened, more and more water will flow until the maximum flow rate of the pump is reached. After that, further opening of the faucet results in no further flow increase.

Thus we see that the pump really determines the flow rate; the faucet can only limit it. And so it is with the audio amplifier and deviation-level controls. The amplifier is the pump; the deviation control is the faucet. The point is that the amplifier establishes the average deviation achieved during transmission. The deviation control provides a high side limit to the deviation.

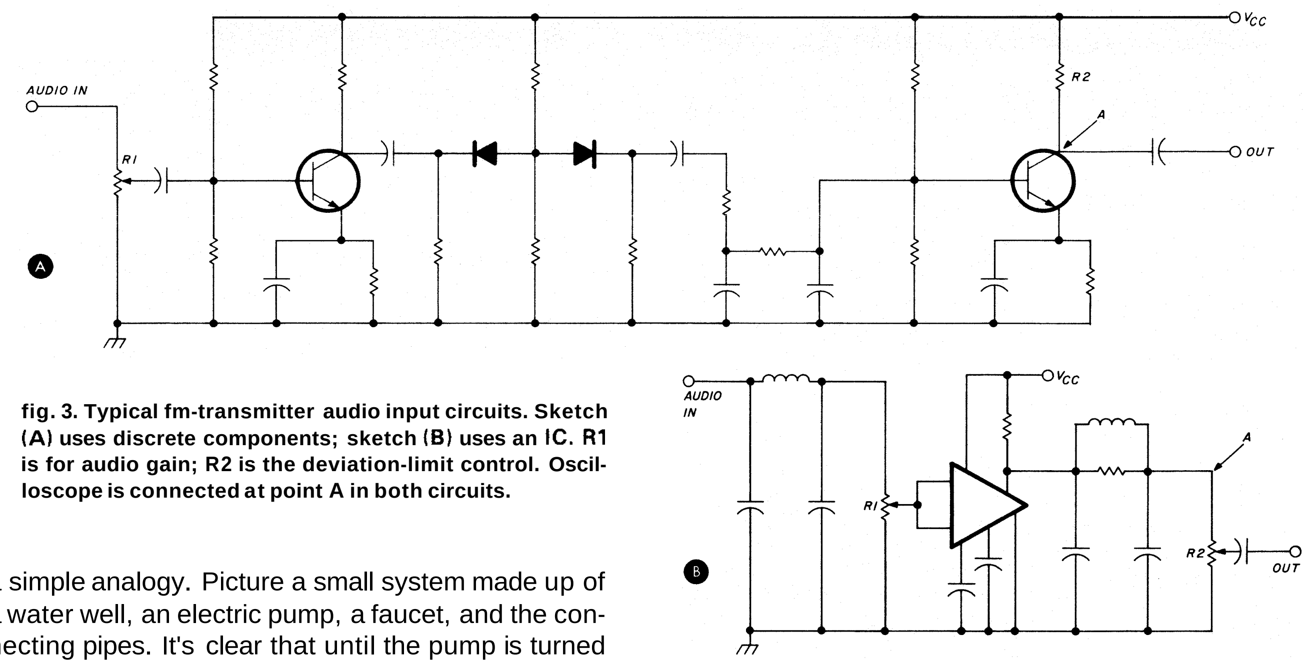


fig. 3. Typical fm-transmitter audio input circuits. Sketch (A) uses discrete components; sketch (B) uses an IC. R1 is for audio gain; R2 is the deviation-limit control. Oscilloscope is connected at point A in both circuits.

setting the deviation

For best results, audio amplifier gain should be set to place the achieved deviation somewhere between 2.5 and 4.5 kHz for normal speech, as previously described; and the deviation level should be set to limit the deviation to 5 kHz on very loud sounds into the microphone. At the same time, care must be taken to avoid overdriving the audio amplifier so that speech fidelity will be maintained. Audio gain can be set only with an oscilloscope. A deviation meter is required for the remaining adjustments. The circuit for a simple, inexpensive deviation meter is described later.

Two representative audio input circuits are shown in **fig.** 3. One consists of discrete components; the other, an integrated circuit. The oscilloscope is connected to point **A,** and audio gain is varied with the setting of R1. Excite the microphone with a steady tone in a normal speaking voice. This can be done, for example, by saying "fo-o-o-ore," as on a golf course, while varying R1 until the audio peaks are ***barely*** flattened as seen on the scope. Even in the absence of an audio-gain control this is a good check. If it's necessary to holler into the microphone to get limiting, a new mike may be in order.

simple deviation meter

A minimal deviation meter consists of a suitable radio receiver and a calibrated, peak-reading voltmeter. The accompanying photograph shows such a device. It consists of the metering circuit shown in **fig. 4** and a commercial scanner. Most of the circuit is mounted on a 25.4 x 38 mm (1 x 1.5 inch) printed circuit board fitted into the scanner. Voltage is taken from the scanner power supply. The only modification to the scanner is the removal of a capacitor, as described later. The scanner function is not affected. Contained in the meter housing are the two switches, the calibration pot, and the smoothing capacitor. The exact layout of the printed circuit board depends on the transformer and the meter-input pot used, so it's not shown here. The layout isn't at all critical. Any fm receiver capable of precise tuning to the carrier frequency of the transmitter to be tested can be used. In many cases, no modification to the receiver need be made. At most, as discussed later, it may be necessary to lift, temporarily, one end of a capacitor to use the receiver.

component selection

None of the voltmeter component values are critical. The diagram represents what I had in my junk box. I salvaged the diodes from surplus computer boards. Those selected had the lowest reverse current (cataloged, not measured). If you don't have a 100-microampere meter, the 50-microampere VU meter, 5E3705, currently offered by Polypaks for just over $1, is most acceptable. Transformer T1 was salvaged from a junked, transistorized a-m broadcast receiver. Anything with a 1k - 10k input impedance and with a transformation of two to three times that impedance in the center-tapped secondary will do. If all else fails, and your junk box is as barren as Mother Hubbard's cupboard, a Triad A31X, Stancor TA31, or Allied interstage 6T14PC would be ideal.

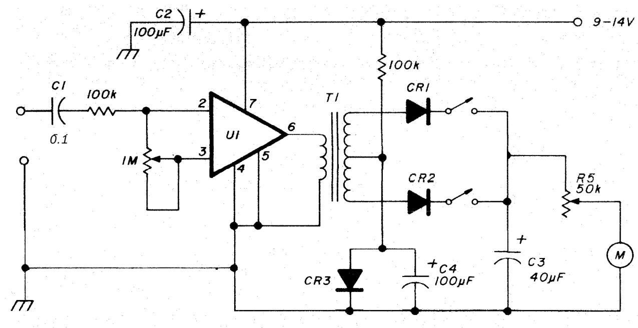
Referring to **fig. 4,** C1 is an input coupling capacitor for U1. Any value between 10 and 100 microfarads will do for capacitor C4; its function is to stabilize the voltage across CR3. The 40μF capacitor (C3), shunting pot R5, and the meter provide meter damping. The optimum value of C3 depends on the internal resistance of the meter and the magnitude of the calibration resistance. I tried several values between 10 and 200 microfarads before deciding that, for my setup, the 40-microfarad capacitor was adequate. The meter movement was damped but not to the point of being sluggish. It's a matter of personal preference.

precalibration notes

Before discussing meter calibration, I should comment on a possible need for temporary modification of the receiver. To reduce the consequences of thermally induced noise in the a-f amplifiers, it's customary to use frequency pre-emphasis and de-emphasis. The fm transmitter emphasizes the higher audio frequencies by shaping the amplifier-response curve. The receiver uses a de-emphasis circuit following the discriminator or detector to re-establish balanced levels.

To measure deviation the signal must be picked off ahead of the de-emphasis circuit. When discrete components are used, as in the circuit of **fig. 3A,** the deviation meter can be connected at point A with no circuit change.

If the receiver uses an IC for the detector, it will be necessary to defeat temporarily the de-emphasis circuit by lifting the ground side of a capacitor. If the IC



NOTES UNLESS OTHERWISE SPECIFIED
1 ALL RESISTORS ARE 1/4 WATT
2 CR1, CR2, CR3 ARE 1N459
3 M1 IS 100µA (SEE TEXT)
4 T1 (SEE TEXT)
5 U1 IS LM380

fig. 4. Schematic of a simple deviation meter discussed in the text. Component values are not critical.

is a type CA3065, MC1358, μA3065, ULN2165, C6063P, or SK3072, lift the ground side of the capacitor connected to pin 7 and connect the meter at pin 14. If it is a type CA2111A, MC1357P, C6062P, or SK3135, lift the ground side of the capacitor connected at pin 14 and pick up the signal on pin 1. For other chips consult the manufacturer's data or perhaps the receiver circuit diagram. Look for a capacitor in the range 0.005 - 0.1 microfarad from a pin to ground and not by-passing a resistor. I removed the 0.005-microfarad capacitor from the de-emphasis circuit in my Pace scanner and found that it made so little difference in the audio quality that it was never replaced.

meter calibration

Having ensured the readings will be free of de-emphasis effects, meter calibration is straightforward. The easiest way to do this is to use a digitally synthesized transmitter as a signal source.

First, disconnect the transmitting antenna and connect a suitable dummy load. Tune the transmitter and the receiver to be used with the deviation meter to the same frequency. Ctose the transmit switch *with no modulation present* and measure the dc voltage at the discriminator or detector output. Be sure to use a dc-coupled oscilloscope or a high-impedance voltmeter. If both rigs are tuned to frequency and the discriminator is properly balanced, there will be zero volts out. If some other voltage is sensed, find out why and correct it before proceding.

Next, shift the transmitter frequency up 5 kHz by using the **PLUS** 5 KHZ switch on the transmitter. Read and note the voltage at the discriminator output. It will be on the order of 1 volt. Now, leave the 5 KHZ switch alone but decrease the transmitter frequency 10 kHz. Voltage at the discriminator should be of the same magnitude but of opposite polarity to that previously measured. That is, if it was +1 volt, it

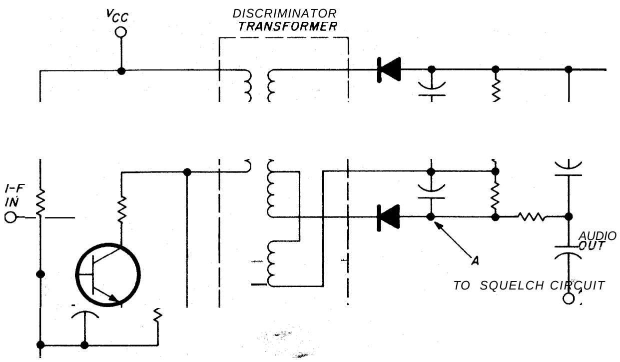


fig. 5. Typical fm discriminator using discrete components. The deviation meter can be connected at point A with no circuit changes. For circuits using ICs, see text.

should now be – 1 volt. If necessary, retrim the discriminator until positive and negative frequency offsets yield the *same* voltage amplitude at the discriminator output. Having achieved that, the receiver is known to be balanced, and the detector output for ± 5 kHz is known.

A decision is now in order. What is the desired full-scale reading for your deviation meter, and over what range is it to be linear? In the prototype unit I opted for, a 10-kHz full scale was necessary to have the desired 5 kHz at center scale. I chose, rather arbitrarily, to make the meter linear up to 7.5 kHz to allow for some padding beyond the region of primary interest. You may want to use some other numbers. Changes from the following procedure to yield numbers of your choice should be obvious.

meter linearity

Assuming the 7.5-kHz linear range is acceptable, set the output of an audio-frequency generator so that its peak-to-peak value is three times the previously measured 5-kHz deviation output voltage. (Note that for 10-kHz linearity, the p-p reading should be four times the reading previously measured.)

The frequency isn't critical. If nothing else is available, use a 60-Hz signal. A 1000 - 3000 Hz signal is preferred. Feed the signal to the deviation-meter input while observing the output waveform (pin 6 of the LM380) on your oscilloscope. Adjust the potentiometer connecting pins 2 and 3 of the IC for maximum signal output without limiting or other distortion. This establishes meter linearity.

Decrease the injected-signal amplitude until it's twice the 5-kHz deviation voltage. Again, this is peak-to-peak voltage. Adjust the meter series pot until the needle points where desired (center scale on the prototype). The deviation meter is now calibrated and can be used to set the deviation-limit level of the transmitter being adjusted.

calibration using a frequency counter

In the absence of a digitally synthesized transmitter, the deviation meter can be calibrated using an audio signal generator and a frequency counter. The technique involved derives from further application of **eq. 1**. At a modulation index of 2.4 there's no power in the carrier, all of it having been transferred to the sidebands. For a deviation of 5 kHz, a modulating frequency of 2080 Hz is necessary to get an MI of 2.4. ($5000/2080 = 2.4$).

Set up a transmitter and the receiver to be used with the deviation meter on a common frequency, as before. This time, however, adjust the discriminator very slightly so that a dc voltage is measured in the presence of unmodulated rf. Now inject the 2080-Hz signal into the audio input of the transmitter. It can be audio coupled through the microphone or directly coupled, whichever is convenient. Increase the injected-signal amplitude from a very low level while observing the dc voltage at the discriminator output. This voitage wiii drop to zero when the injected-signal amplitude corresponds to a 5-kHz deviation. Holding the amplitude constant, drop the frequency to about 1000 Hz and tweak the series pot used for calibrating the meter until the latter indicates 5 kHz. Check pin 6 of the LM380 to make sure that the sine wave at that point isn't distorted. If it is, adjust the pot across pins 2 and 3 of that chip to remove the distortion, and readjust the meter calibrating resistor as necessary.

setting deviation

Excite the microphone with any loud, high-frequency tone — a loud sustained whistle is adequate. Adjust the deviation level control so that the deviation meter reads no more than 5 kHz. Talking into the microphone in normal tones should yield an average deviation of 2.5 kHz to 3.5 or 4 kHz — again depending on individual voice pitch. If it's much less than this, and if the rig has an audio gain control, try increasing the gain slightly. Recall, however, that the gain control was previously set at the upper level of the amplifier's linear range. Therefore a tradeoff between amplitude and amplitude linearity will have to be made.

If a *TOUCH TONE™* pad is used, the high tone group should cause a deviation meter reading of 4 - 4.5 kHz; for the low tone group, it should be about 3 kHz.

You've now properly set the average and maximum deviation on your fm transmitter. Stand by to pride yourself on the fact that your rig is giving you *maximum* talk power per watt input, and that you're one of the good guys when it comes to bandwidth conservation.

ham radio

10-GHz Gunnplexer transceivers — construction and practice

Discussion of the various aspects of Gunnplexer transceiver construction and operation, including two practical transceiver designs

Microwave communications is one of the last frontiers of Amateur Radio; thanks to the Gunnplexer transceiver module offered by Microwave Associates, it's now easier than ever to put together a practical station for amateur operation on the 10-GHz band. Only a few years ago a "simple" microwave setup required a rack-full of equipment and friends in the industry who could provide hard-to-find parts. With the Gunnplexer, an entire microwave system can now be held in one hand. It can be easily backpacked to the highest mountain tops, and it can be operated from a single 12-volt battery.

As you receive it, the Gunnplexer module is not a complete transceiver; to put it on the air you need a dc power supply, a simple speech amplifier, and an fm receiver. You can put together a working system in one evening. To build a complete transceiver like that described in this article will take a little longer.

what is a gunnplexer?

The heart of the Gunnplexer is the Gunn diode oscillator, named after the IBM engineer who invented it in 1963, John Gunn. While measuring the resis-

By James R. Fisk, W1HR, ham *radio,* Greenville, New Hampshire 03048

tance of gallium arsenide (GaAs), Gunn found that when the voltage across a thin wafer of the material was increased above a certain point, the current fluctuated at microwave frequencies. The mechanism which caused this was a mystery at first, but Gunn suspected a negative resistance due to electron movements within the gallium arsenide. This eventually proved to be the case (a detailed description of the Gunn diode phenomenon is contained in reference 1).

When a Gunn diode is placed in a resonant microwave cavity, small amounts of power can be obtained at the desired frequency. The cavity can be tuned mechanically, or a voltage-variable capacitor (varactor) may be used to change the resonant frequency of the cavity. The Gunnplexer (fig. 1) uses both a mechanical tuning screw and a varactor diode; frequency modulation is obtained by placing a small modulating voltage across the varactor. Power is coupled out of the cavity through an iris. The size of the iris must be determined experimentally for the best compromise between maximum power output and isolation from changes in diode impedance and load.

In the Gunnplexer the Gunn oscillator provides both the transmit power and the local-oscillator injection for the mixer diode. The ferrite circulator couples a small amount of energy into the low-noise Schottky mixer diode and isolates the transmit and receive functions. Since the Gunn oscillator functions as both the transmitter and receive local oscillator, the i-f receiver at each end of the communications link must be tuned to the same frequency, and the frequencies of the Gunn oscillators at each end of the link must be separated by the i-f.

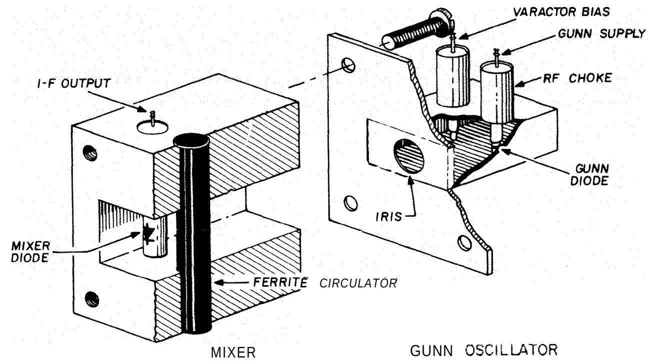
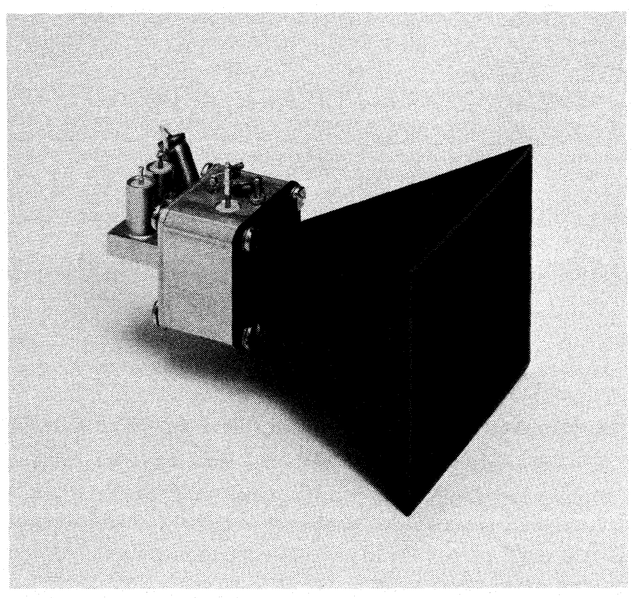


fig. 1. Cutaway view of the Microwave Associates Gunnplexer. The Gunn diode is mounted in a resonant cavity which is tuned by a tuning screw (coarse tuning) and a varactor (fine tuning). Microwave energy is coupled out of the cavity through an iris. The ferrite-rod circulator couples a small amount of rf energy into the Schottky mixer diode; the circulator also isolates the transmit and receive functions and allows full-duplex operation.

Confused? Take a look at fig. 2. Here Gunnplexer 1 is tuned to the center of the 10-GHz amateur band at 10250 MHz. If a 30-MHz i-f receiver is used, Gunnplexer 2 must be tuned either 30 MHz higher or lower than Gunnplexer 1. Assume it's tuned 30 MHz higher at 10280 MHz; its signal will mix with the 10250-MHz LO in Gunnplexer 1 to provide an output to the receiver at 30 MHz. Conversely, the 10250-MHz transmit signal from Gunnplexer 1 will mix with the 10280-MHz LO in Gunnplexer 2 to provide an output at 30 MHz.



Microwave Associates 10-GHz Gunnplexer.

gunnplexer communications range

One of the first and most asked questions about Gunnplexers is, what is their maximum range? Since most microwave communications systems are based on line-of-sight transmission, it's a relatively easy matter to determine the effective communications range of any Gunnplexer system. When the distance between the two stations is known, path loss in dB is given by:

$$92.5\ dB + 20 \log f\ (GHz) + 20 \log D\ (kilometers)$$

$$96.6\ dB + 20 \log f\ (GHz) + 20 \log D\ (miles)$$

where f is the operating frequency and D is the distance between transmitting sites. Note that each time the frequency is doubled, the path loss increases by 6 dB. This is shown graphically in fig. 3, which shows the path loss vs distance for each of the amateur bands above 1000 MHz. At 10250 MHz, the center of the 10-GHz amateur band, the path loss equation can be simplified to:

$$112.7 + 20 \log D\ (kilometers)$$

$$116.8 + 20 \log D\ (miles)$$

In other words, the path loss over a distance of 1 km is 112.7 dB; it increases 6 dB each time the path length is doubled. The objective is to build enough gain and sensitivity into the microwave system to overcome the loss over the desired path. This is a function of transmitter power, antenna gain, receiver sensitivity (noise figure), and receiver bandwidth, but it's not as complicated as it sounds because all these factors can be easily translated into dB.

The graph of **fig. 4** has been designed to simplify the calculation of communications range at 10250 MHz and is normalized to a power output of 15 mW, receiver bandwidth of 200 kHz, 12-dB noise figure, and 17-dB gain antennas at each end of the link. This is what I consider a minimal Gunnplexer system. The horizontal line labelled THRESHOLD is the beginning of reception of intelligible speech and allows no margin for fading due to rainiaii, multipath propagation, or other environmental effects. With the minimal Gunnplexer system, threshold occurs at a distance of about 127 kiliometers (76 miles). Since a carrier-to-noise ratio of 8-10 dB is recommended for reliable communications, about one-third this distance could be used for successful communications.

There are four major things which can be done to increase range: use higher transmitter power, reduce receiver bandwidth, improve receiver sensitivity, or increase antenna gain. The effects of power output and receiver bandwidth are shown in the chart on **fig. 4.** Each improvement in system performance adds the stated number of dB to the carrier-to-noise ratio. A 40-mW Gunnplexer with a receiver bandwidth of 25 kHz, for example, improves the carrier-to-noise ratio by 13.3 dB (4.3 dB for 40 mW transmitter power, plus 9.0 dB for reduced bandwidth). This

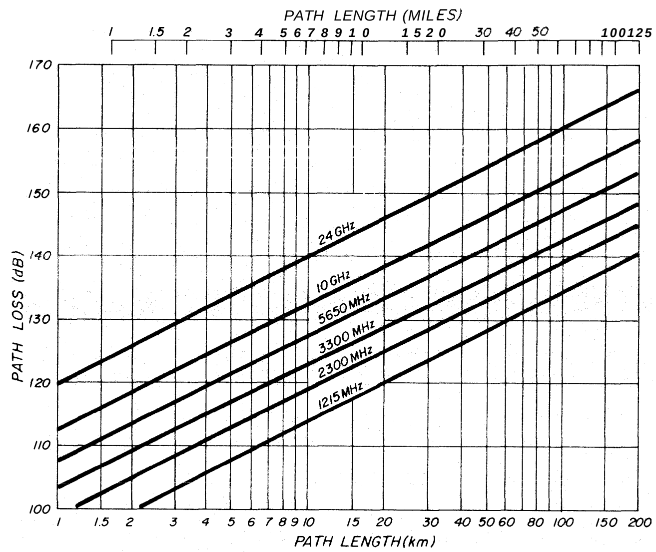


fig. 3. Path loss vs distance for each of the six amateur microwave bands. Note that path loss increases 6 dB each time the frequency or path length is doubled. Loss over a 61.7-km path at 1215 MHz, for example, is 130 **dB;** at twice the frequency (2430 MHz) the loss is 6 dB greater; at 8 times the frequency, near the 10 **GHz** band, loss is more than 18 dB greater. This graph assumes line of sight with no obstructions of any kind.

the path loss is 146.7 dB. The other item which is fixed is the thermal noise floor, at – 144 dBm, which is set by the laws of nature and determines the ultimate sensitivity of the receiver.2

Beginning at the transmitting end of the link, we have 15 mW power output (+11.8 dBm). To this is added the 17-dB gain of the antenna. When the path loss is subtracted, the signal level at the receiving site is – 117.9 dBm. The 17-dB-gain receiving antenna

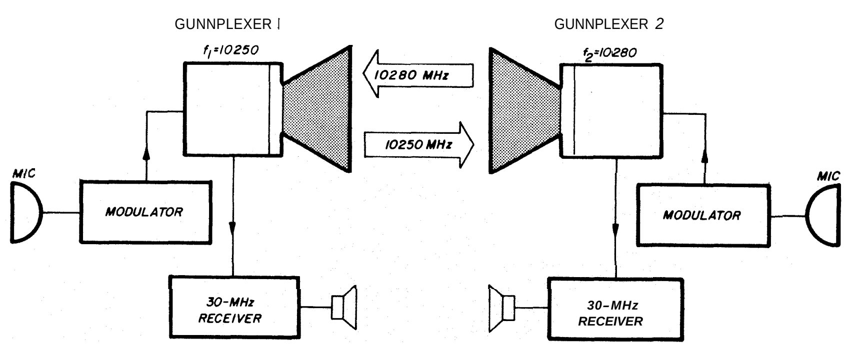


fig. 2. Gunnplexer operation. Since the same oscillator is used as both a transmitter and local oscillator for the mixer, the i-f at each end of the link must be at the same frequency, and the Gunn oscillator frequencies must be separated by the i-f. In the example shown here, Gunnplexer 1 is tuned to 10250 MHz; 30-MHz receivers are used, so Gunnplexer 2 at the other end of the link must be tuned either exactly 30 MHz higher or lower (to 10280 or 10220 MHz).

system would provide a carrier-to-noise ratio of 8 dB at a line-of-sight distance of 233 kilometers (145 miles).

When calculating the communications range of a microwave system, all the gain and loss components of the system must be considered, as shown in **fig. 5.** Here a distance of 50 km (30 miles) is assumed, so

increases the signal to – 100.9 dBm. From this must be subtracted the 12-dB noise figure and the 200-kHz bandwidth factor (23 dB), for a signal level of – 135.9 dBm. The difference between this and the thermal noise floor at – 144 dBm is the carrier-to-noise ratio. For this link, +8.1 dB.

The easiest way to improve range is to use a higher

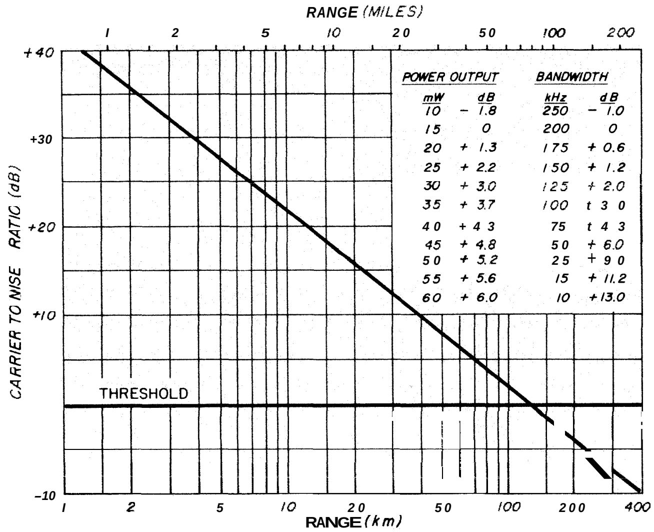


fig. 4. Carrier-to-noise ratio vs distance for two **15-mW** Gunnplexers at **10250** MHz, equipped with 17-dB-gain horn antennas; receiver noise figure is assumed to be 12 dB with 200 kHz bandwidth. The THRESHOLD line is the beginning of the reception of intelligence. At a distance of 127 km (**76** miles) the carrier is at the noise level or threshold; at a distance of about 40 km (**25** miles) the carrier-to-noise ratio is 10 dB, the minimum signal level recommended for reliable voice communications. Range can be lengthened by increasing transmit power, decreasing bandwidth, or adding antenna gain (see text). Improvements in dB for increased output and narrower bandwidth are shown.

gain antenna. Unlike the lower frequencies, where antenna gain is hard to come by, on the microwave bands it's relatively easy. A 24-inch (61-cm) parabolic reflector, for example, yields 32 dBi gain. If used at only one end of the system shown in **fig. 5,** this would have the effect of increasing the range to

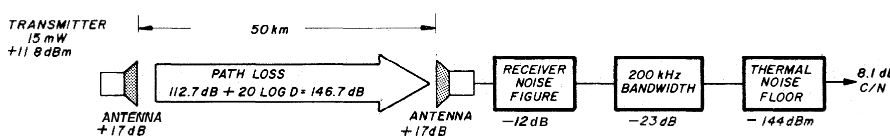


fig. 5. Example of a path-loss calculation at **10250** MHz. With two **15-mW** Gunnplexers equipped with 17-dB gain antennas, and 12-dB noise figure in 200-kHz bandwidth, the carrier-to-noise ratio at **50** km is 8.1 dB.

Transmitter power, 15 mW	+ 11.8 dBm	+ 11.8 dBm
Add transmitter antenna gain	+ 17.0 dBi	+ 28.8 dBm
Subtract path loss	+ 146.7 dB	− 117.9 dBm
Add receiver antenna gain	+ 17.0 dBi	− 100.9 dBm
Subtract receiver noise figure	− 12.0 dB	− 112.9 dBm
Subtract 200 kHz bandwidth factor	− 23.0 dB	− 135.9 dBm
Thermal noise floor		− 144.0 dBm
	Carrier-to-noise ratio:	+ 8.1 dB

nearly 2000 km (1200 miles) for the same 8.1 dB carrier-to-noise level! That's obviously well beyond line-of-sight for any two earth-based locations. One disadvantage of high antenna gain is antenna alignment; the 4-degree beamwidth of a 24-inch dish leaves little room for error when pointing the antenna at another station.

You can also increase range by reducing the bandwidth of your receiving system, but because of the thermal drift of the Gunnplexer, this requires a system which phase locks the Gunn oscillator to a crystal-controlled reference oscillator. The cost of a phase-locked system is somewhat less than that of a commercial parabolic reflector, but system gain is only on the order of 12-13 dB when compared with a system with 200-kHz receiving bandwidth. On the other hand, a phase-locked system permits the use of ***CW,*** which provides reliable communications with lower carrier-to-noise ratios than voice, so there may be the equivalent of an additional 4-5 dB gain available.

For greater range you can also increase transmitter output or improve receiver noise figure, but both are expensive and limited to a certain extent by the present state of the art.

gunnplexer performance

The Gunnplexer performance measurements discussed here were made by B. Chambers, G8AGN, of the Department of Electronic and Electrical Engineering at the University of Sheffield, England, who is also a member of the Microwave Committee of the Radio Society of Great Britain. Front-end performance was not measured because, in practice, receiver sensitivity and noise figure are highly dependent on the operator's choice of i-f strip and the degree of matching between the mixer diode and the i-f preamplifier. Therefore G8AGN made measurements only to check the performance of the Gunn oscillator.

When a variable voltage was connected to the Gunn diode, it was found that rf power was produced with an applied voltage as low as 5 volts. Most of the tests, however, were accomplished with +10 volts applied to the Gunn diode and +4 volts bias on the varactor diode.

Using a Systron-Donner model 6057 frequency counter with an upper frequency limit of 18 GHz, G8AGN found that the Gunn oscillator drifted down in frequency by about 3 MHz during the initial one-hour warm-up period. A further frequency check 15

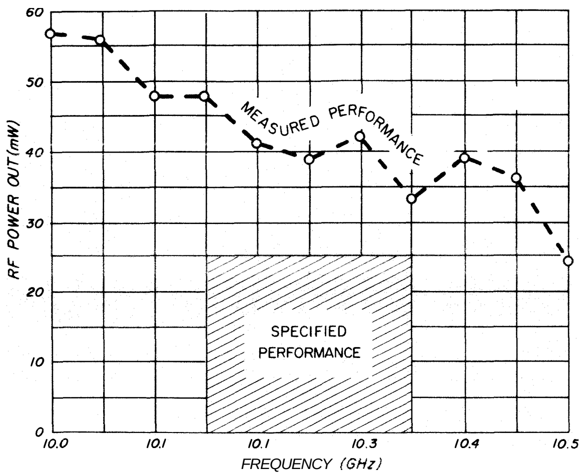


fig. 6. Typical variation of Gunnplexer output power as the frequency is tuned mechanically through the 10-GHz amateur band, as measured by G8AGN. At ali frequencies the output was well above the rated 25 mW.

minutes later showed that the oscillator was drifting down in frequency by about 28 kHz per minute. This rate of drift is quite acceptable in practice unless a narrow-band system is being used and there is no provision for AFC.

The mechanical tuning range of the oscillator was checked next and found to extend from 9641 MHz up to 10764 MHz. The rf power output over this frequency range was measured with a Marconi model 6460 power meter with a coaxial head, buffered by a fixed 20-dB pad. This was preceded by a coax-to-waveguide transition and slide-screw tuner which was adjusted before each reading to ensure that the oscillator was delivering power into a matched load. **Fig. 6** shows the variation of rf output power over the frequency range from 10.0-10.5 GHz. Rf power measurements at the extremities of the tuning range showed 39 mW at 9641 MHz and 24 mW at 10764 MHz.

For a given setting of the mechanical tuning screw, the frequency of the Gunn oscillator may be tuned electrically by changing the voltage applied to either the Gunn diode or the varactor. Varying the voltage of the Gunn diode over the range from +5 to +11 volts produced a frequency change of 13.3 MHz about a preset value of 10250 MHz. This represents approximately 2.2 MHz per volt for frequency pushing and is well within the quoted specification.

With the Gunn diode held at +10 volts, the varactor bias was varied from +1 to +12 volts and measurements were made of both frequency and rf power output. Although up to +20 volts bias may be used on the varactor, measurements were made only to +12 volts because this is the maximum voltage usually available for portable operation. **Fig. 7** shows the result of these measurements. It can be seen that the maximum electronic tuning range was approximately 100 MHz, and that over this range the rf power output varied by about 3.5 dB.

The final set of Gunnpiexer measurements made by G8AGN were concerned with the frequency-pulling performance of the Gunn oscillator. To make these measurements, the Gunnplexer was set up to deliver power to a load consisting of an adjustable short circuit; therefore, by varying the axial position of the short-circuit plane within the waveguide, a wide range of load impedances would be seen by the oscillator. For an axial variation of the short-circuit plane of 20 mm (0.8 inch), corresponding to a distance just greater than $\lambda_2/2$ at 10250 MHz, the total frequency variation was found to be 12 MHz.

The result of this test suggested that the ferrite circulator should be "transparent" enough for the Gunnplexer to be frequency locked using a cavity wavemeter, and this, in fact, proved to be the case. A TE_{011} mode transmission-type cavity wavemeter with a quoted Q factor of 8000 was available. This was simply bolted to the Gunnplexer assembly — the resulting separation between the wavemeter and the coupling iris to the Gunn oscillator being about 6.5 cm (2.6 inches). The wavemeter cavity had provision for attaching a waveguide diode holder, and this was

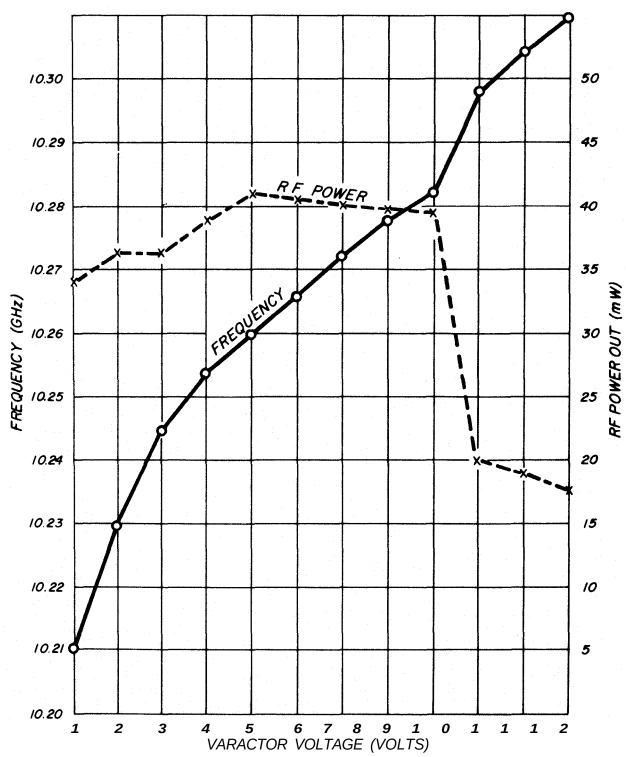


fig. 7. Frequency variation and measured power output with changes in varactor bias, as measured by G8AGN. Power output varies less than 3.5 dB over the nearly 100-MHz tuning range.

used in conjunction with a sensitive milliammeter to detect when the cavity was tuned near resonance.

With a little practice, G8AGN found that it was possible to hold the frequency of the Gunn oscillator to within 1 kHz for periods of minutes at a time. In view of this, it seems probable that the Gunnplexer could also be injection locked using a crystal controlled source, although this was not tried.

power supply

The first requirement for a Gunnplexer system is a regulated +10 volt power supply. Unfortunately, there aren't any readily available, high-current, three-terminal IC regulators with a 10-volt output (the Lambda LAS1510 meets these requirements, but is difficult to purchase in small quantities).* The answer is the Fairchild μA78MG 4-terminal regulator, which requires only two external resistors to set the regulated output voltage (see **fig. 8**). This regulator will provide in excess of 500 mA output, so it's adequate for most Gunnplexer systems.

For precise voltage adjustments, I have included a miniature 500-ohm pot between the two 4700-ohm resistors; this allows the output voltage to be set within a few millivolts of +10 volts. If you're not this fussy, you can connect the IC's control terminal (pin 4) directly to the junction of two 4700-ohm resistors — the output voltage should be within 5 per cent of the required 10 volts. This is probably close enough for most applications.

In many circuits using the μA78MG regulator the bypass capacitors may not be required. However, for stable operation of the regulator IC over all voltage and current input ranges, bypassing is recommended by the manufacturer (0.33 μF at the input and 0.1 μF at the output). The input bypass is necessary if the regulator is located far from the filter capacitor in the power supply; bypassing the output improves the transient response of the regulator.

tuning range

The frequency of the Gunnplexer is controlled by

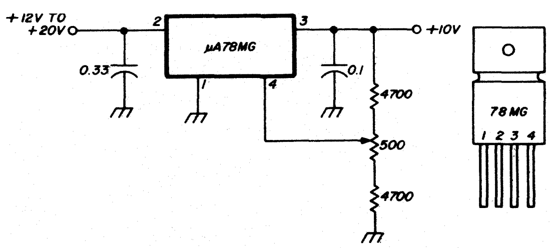


fig. 8. Regulated Gunnplexer power supply can be adjusted to exactly +10 volts; with proper heatsinking, this circuit will provide current in excess of 500 mA. The 0.33-μF capacitor at the input and 0.1 μF at the output improve circuit performance.

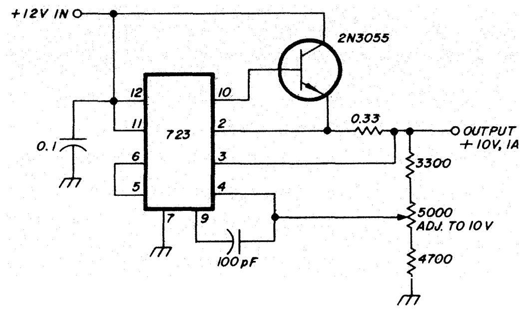


fig. 9. Simple +10 volt regulator using readily available parts, designed by W1SL, will supply up to 1 ampere. The 723 regulator is available from Radio Shack (part number 276-1740) as is the 2N3055 transistor (Radio Shack 276-1634).

the voltage on the built-in tuning varactor, the setting of the mechanical tuning screw, and the supply voltage to the Gunn diode. Unless otherwise specified, the Gunnplexer is mechanically tuned to 10250 MHz at the factory with 10.0 volts on the Gunn diode and 4.0 volts across the varactor. The output frequency of the Gunnplexer can be adjusted ±100 MHz with the tuning screw, but I don't recommend touching the mechanical tuner unless you have access to a microwave frequency counter; you will find it difficult to accurately retune the unit to 10250, the center of most amateur activity on this band.

The Gunnplexer can also be electronically tuned by varying the voltage across the varactor from 1 to 20 volts; this is the preferred method, and a tuning range of 60 MHz is guaranteed. Electronic tuning range varies from unit to unit, but data is furnished with each Gunnplexer so you can easily estimate frequency output vs varactor voltage. Many units have an electronic tuning range of 100 MHz or more, so it's not necessary to touch the mechanical tuning screw for most amateur applications.

Shown in **fig. 10** is a plot of frequency output *vs* varactor tuning voltage for a 40-mW Gunnplexer that I am using at my station. The tuning curve is quite nonlinear, with the greatest frequency change — 50 MHz — occurring as the varactor voltage is increased from 1 volt to 4 volts. An increase from 4 volts to 10 volts moves the output frequency up 40 MHz, and a change from 10 volts to 20 volts increases the output frequency 46 MHz. The total frequency change is 136 MHz. The tuning range of other Gunnplexers won't exactly follow this curve, but it gives you an idea of what you can expect.

The varactor also provides a way of frequency modulating the unit. If a small modulating voltage is

***Shortly after this article was wrinen, Fairchild Semiconductor announced the μA78C00 series of 3-terminal voltage regulators which have rated output current greater than 500 mA. A 10-volt regulator, the μA78C10C, is included in the series.**

impressed on the varactor bias, the frequency will be varied at an audio (or video) rate. Because of the wide electronic tuning range, the required modulation voltage is very small; 10 mV or so for 75 kHz deviation, or less than 1 mV for 5 kHz deviation. However, don't plan on using narrowband deviation unless you have a crystal-controlled, phase-locked system for stabilizing the Gunnplexer frequency.

The output frequency also varies with changes in the Gunn diode supply voltage — 15 MHz per volt maximum — but this isn't recommended as a tuning method. In addition, the power output and efficiency of the Gunnplexer has been optimized for a 10.0-volt supply, and you don't want to risk damaging the expensive Gunn diode.

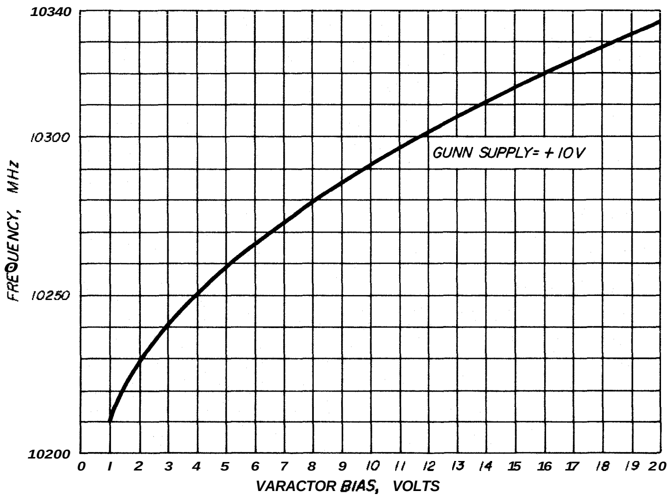


fig. 10. Output frequency vs varactor bias for a 40-mW Gunnplexer used by W1HR. The tuning range of other Gunnplexers won't exactly follow this curve, but can be estimated from the data furnished by Microwave Associates with each Gunnplexer.

In portable systems designed to operate from +12 volts, it's convenient to set the maximum varactor voltage at the +10-volt Gunn diode supply. This provides more than enough tuning range if you use a 30-MHz i-f system. Amateurs in Europe usually transmit voice on 10250 MHz and receive on 10280 or 10220; in the United States many stations have standardized 10250 for transmitting and 10280 for receiving (for full duplex operation one station transmits on 10250 and the other transmits on 10280).

If you use fm broadcast receivers at each end of the link with a 10-volt varactor supply, you may not be able to obtain sufficient tuning range to cover the required 100 MHz. However, if you use an auxiliary varactor supply that will provide up to 20 volts, you should have no difficulty obtaining the required range. In many cases the nominal 12 volts available

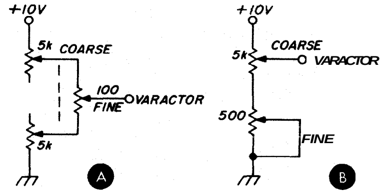


fig. 11. If a multi-turn potentiometer is unavailable, varactor bias can be controlled by one of the methods shown here, The resolution of the circuit at (A) is about four times better than with a 10-turn pot; the circuit at (B) has somewhat less resolution but is less expensive.

from an automobile battery will be sufficient. If you use an ac-powered dc supply for the varactor, however, be sure it's weii regulated and filtered. Any ripple on the supply line will result in hum modulation.

varactor bias control

Since small changes in varactor bias have a large effect on the output frequency, a multi-turn potentiometer should be used for the tuning control (with a conventional 270° pot, the frequency can change 300 kHz or more for each 1 degree rotation of the pot's shaft). Sometimes you can find precision 10-turn pots on the surplus market, but, if not, there are several alternatives. One is to use a single-turn pot with a reduction unit like the Jackson Brothers 6:1 planetary drive. This may not be completely satisfactory, however, because resolution may be limited by man-

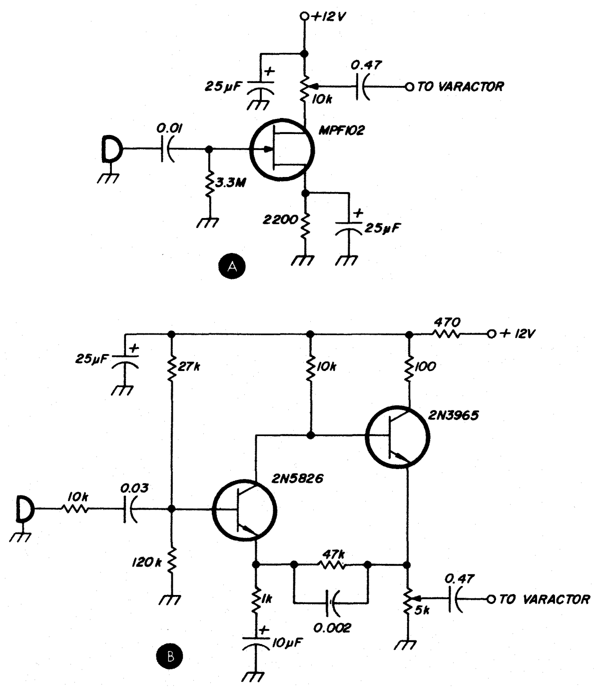
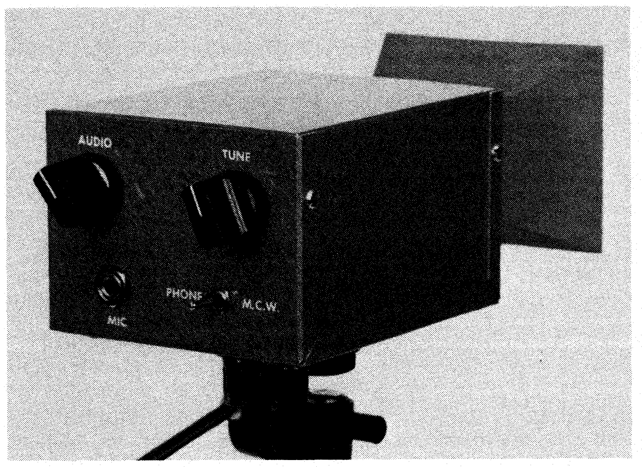
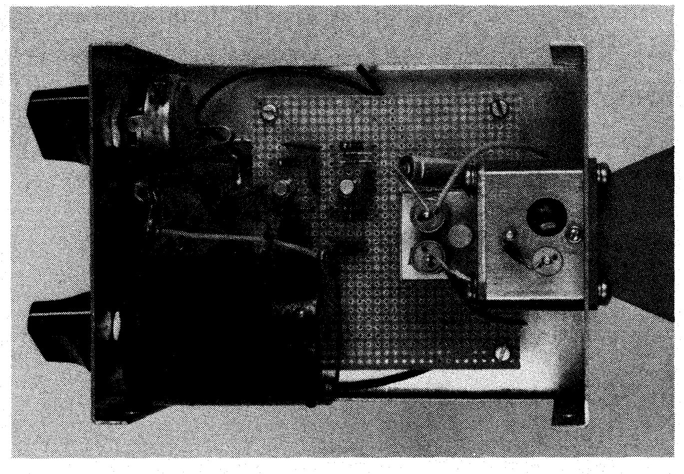
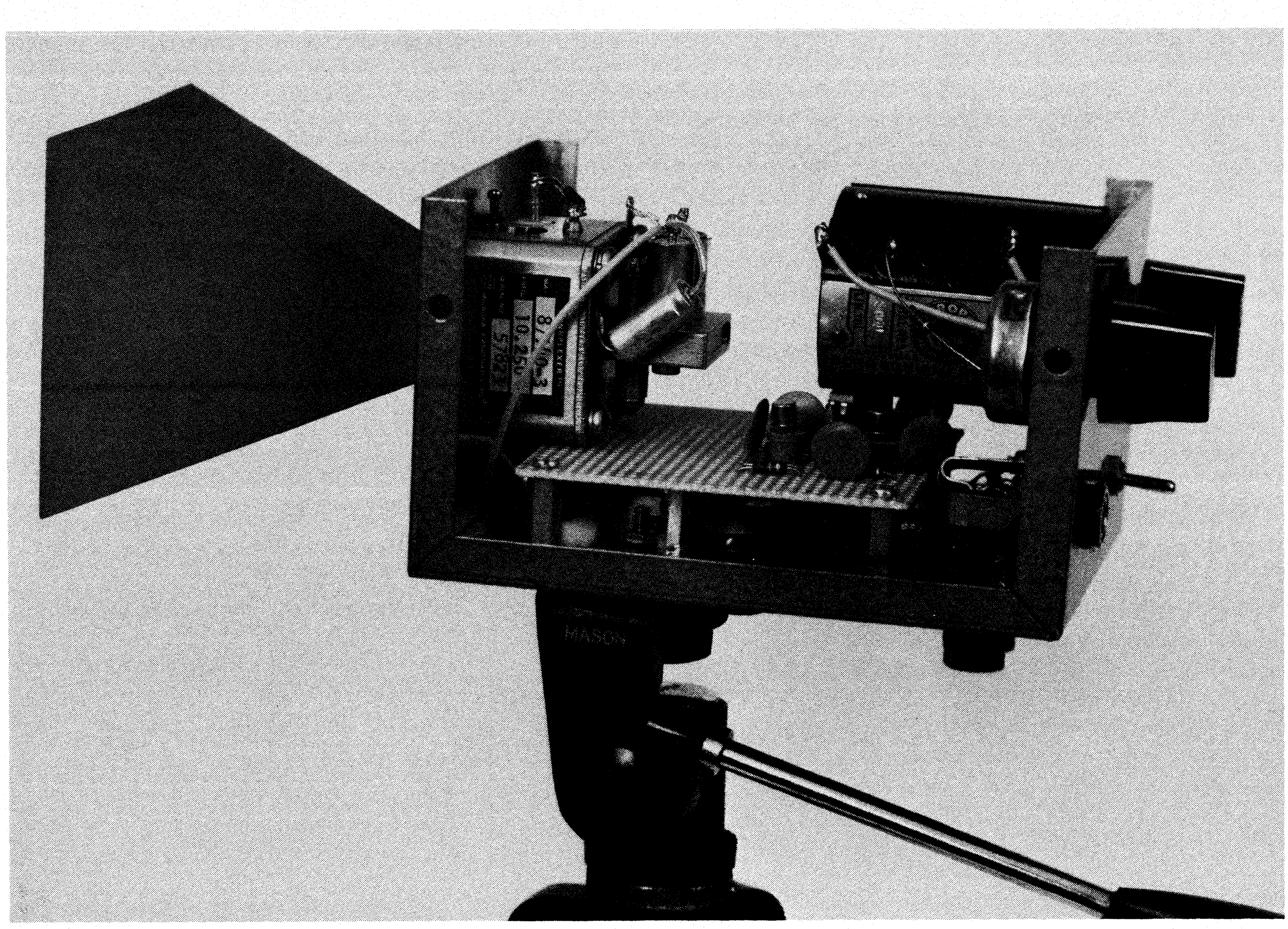


fig. 12. Two simple speech amplifiers which are suitable for use with Gunnplexers. The two-transistor circuit at (B) includes limited filtering for shaping audio bandwidth.







Minimal Gunnplexer system used by W1HR includes a 10-volt IC voltage regulator, simple speech amplifier, and tone oscillator. A phono connector on the bottom of the chassis is provided for the fm receiver. A 10-turn precision potentiometer found on the surplus market is used for frequency control.

ufacturing tolerances in the potentiometer's resistance element.

Two other possibilities for varactor control are shown in **fig. 11.** The system in **fig. 11A** uses one dual potentiometer for coarse adjustments and a single-turn pot for fine tuning. Resolution of this system is about four times better than with a 10-turn pot and is suitable for the most demanding requirements. The bias control arrangement in **fig. 11B** does not provide as much resolution but is more economical. A disadvantage is that the resolution of the fine adjustment varies, and depends upon the setting of the coarse control; when the coarse potentiometer is in the center of its range, resolution approximates that of a 10-turn pot.

speech amplifiers

Because of the high sensitivity of the varactor, a very small modulation voltage (on the order of 10 mV p-p) is required to obtain 75-kHz deviation for wide-band frequency modulation of the Gunnplexer; this greatly simplifies the design of a suitable speech

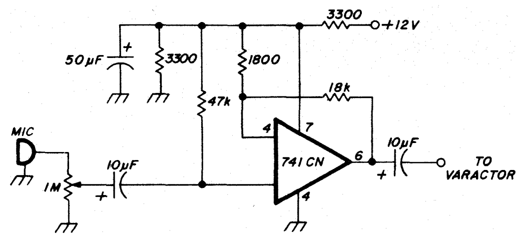


fig. 13. Speech amplifier circuit designed by G8AGN for Gunnplexer operation uses a low-cost 741 op amp and offers high input impedance.

amplifier. In its simplest form, the Gunnplexer speech amplifier requires only one transistor, as shown in **fig. 12A.** In this circuit the MPF102 fet exhibits high input impedance for a crystal, ceramic, or dynamic microphone, and provides more than enough voltage gain for full 75-kHz deviation at 10.25 GHz. Deviation is adjusted with the 10k potentiometer in the drain circuit.

The two-transistor speech amplifier in **fig. 12B** has an input impedance of about 20 kilohms and includes filtering to limit the speech bandwidth. For those who prefer to use ICs the circuit in **fig. 13** is recommended. This 741 speech amplifier was designed by G8AGN/G8CZO for use with a Gunn diode transmitter.3

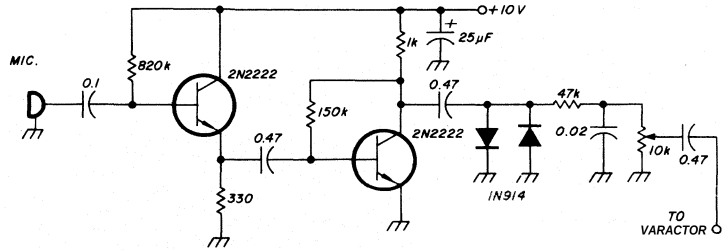


fig. 14. Two-transistor speech amplifier for Gunnplexers features high input impedance, good gain, clipping for improved audio punch, and a lowpass filter for limited audio bandwidth.

In any frequency-modulation system the speech amplifier should, in addition to providing audio gain, include some form of speech processing to limit dynamic range so the audio signal doesn't exceed the maximum frequency deviation. This can be done with audio compression or by using a simple diode clipper to limit the audio peaks. The two-stage speech amplifier shown in **fig. 14** includes a clipper and lowpass RC filter (47k resistor and 0.02-µF capacitor) to reduce the harmonics produced by clipping. I used 2N2222 transistors in this circuit because I had them in my junk box, but most high-gain NPN transistors should work. If you wish, the same diode clipper and RC filter can be added to the circuits of **figs. 12** and **13.**

For most effective fm communications, the speech system should include a system for limiting bandwidth to 300-3000 Hz, and de-emphasis to correct the speech frequency characteristic. A circuit which

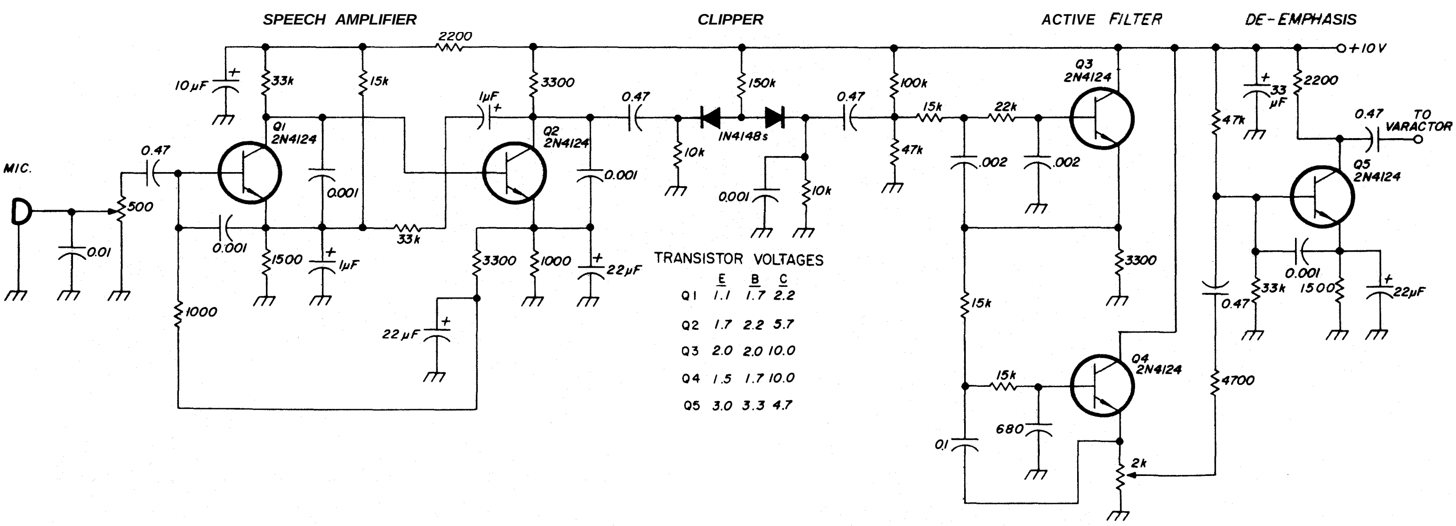


fig. 15. High-performance fm speech amplifier uses heavy feedback for reduced audio distortion. It also includes an audio clipper, 300-3000 Hz active audio filter. and de-emphasis stage. Both input and output controls are provided. Circuit board for this circuit is shown in fig. 22.

has a complete speech amplifier, clipper, active filter, and de-emphasis stage is shown in **fig. 15.4** The first two stages use heavy feedback to reduce distortion and improve frequency response. These stages are followed by a double-diode clipper and a two-stage active filter that has a 500-3000 Hz passband. The last stage provides de-emphasis. This amplifier gives an output of 100 mV for an input of 2 mV across 300 ohms; both input and output controls are provided. I have used this amplifier with good success at one end of a wideband Gunnplexer link.

tone oscillator

When lining up two Gunnplexer systems, particularly if you're using high-gain parabolic reflectors, it's helpful to continuously tone modulate your transmitter. There are several ways to generate an audio tone, but for minimum parts count I prefer the circuit of **fig. 16,** which uses a 555 timer IC. Total current drain with a 70-volt supply is only 10 mA. The I-kHz squarewave output swings from ground to +10 volts; this is reduced to manageable levels for Gunnplexer use with the series 100k resistor and 200-ohm pot. The 10k resistor and 0.1-μF capacitor form a lowpass filter; in some applications the filter may not be required.

If you have a memory keyer, it can be plugged into the key jack and used to send your call sign, a series of vees, or your location. If you wish to send only your call sign, you might consider the automatic CW ID unit manufactured by Autocode.* Although this unit was designed for automatically sending CW identification for RTTY or vhf-fm transmissions, it is ideal for Gunnplexer systems.

i-f receiver

Although a 30-MHz i-f receiver is recommended if you want to work reliably over long distances, to get started with a Gunnplexer system many amateurs have used low-cost fm broadcast receivers tuned around 100 MHz. One popular unit is the Audiovox fm converter; this receiver sells for less than $20, can be completely converted to Gunnplexer use in one evening, and is a good compromise unit for getting started on 10 GHz with Gunnplexers. Complete conversion information is available from G. R. Whitehouse & Company.† The main disadvantage of an fm broadcast receiver is i-f feedthrough. For best results you must pick a frequency that is clear of local fm broadcasters. If you take this system mountain-topping, your problems with i-f feedthrough will increase dramatically, but it is still a good way to get started. Also, it's a simple matter to add a better receiver to your system later — no other parts of the set-up will have to be changed.

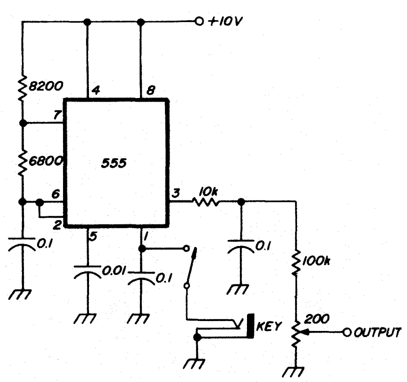


fig. 16. A 1000-Hz tone oscillator is very helpful when setting up a Gunnplexer link. It can also be used for MCW under weak-signal conditions.

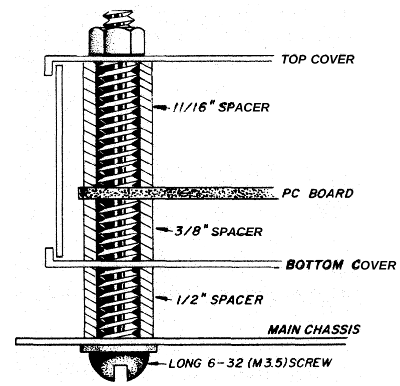


fig. 17. Method of mounting the DJ700 fm receiver with spacers and a long 6-32 (M3.5) screw. Similar mounting arrangements are used at the four corners of the fm receiver PC board.

Another low-cost approach to the i-f receiver can be found in the used two-way equipment market. Many of the fire, police, and public-service fm receivers built 10 or more years ago can now be purchased for a few dollars. The receiver you want for Gunnplexer use was originally designed to tune from 30 to 50 MHz and is built for wideband fm. Many of the newer fm receivers for this band are for narrow-band fm, so they are not suitable for Gunnplexer use. A number of companies marketed solid-state receivers of this type in the 1960s, including Lafayette, Radio Shack, and Regency. Some had provisions for crystal control; this, if you can find one, is the type most suited to Gunnplexer communications. Price for a receiver of this type is typically around $5; most users have switched to more portable narrow-band receivers with scanners, so the older, tunable receivers have practically no commercial value.

*Autocode, 8116 Glider Avenue, Dept. H, Los Angeles, California 90045.

†G. R. Whitehouse stocks 15-, 25-, and 40-mW Gunnplexers; his address is Newbury Drive, Amherst, New Hampshire 03031.

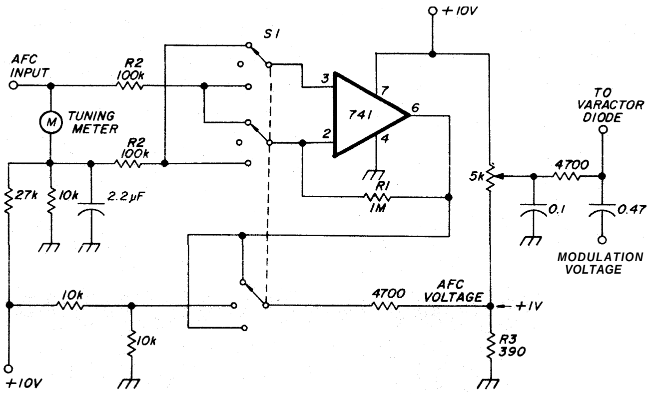


fig. 18. Gunnplexer AFC system designed by DJ7OO for use with his 30-MHz fm receiver. Circuit may be adapted to other fm receivers by changing the ratio of resistors R1 and R2. It may be desirable in some cases to replace R3 with a trimmer pot. In the center position of switch S1, the AFC is turned off, the two outer positions provide positive- and negative-going AFC voltage with increasing frequency (see text).

The choice of 30 MHz for the i-f receiver means that you can set up your Gunnplexer on 10250 MHz and tune in stations either 30 MHz above or below your center frequency. Many Gunnplexers don't have sufficient electronic tuning range to handle an i-f at 100 MHz with only + 10 volts of varactor bias. If you have a +20 volt bias supply available, many Gunnplexers will tune the required 100 MHz, but that precludes most portable operation unless you provide additional batteries for the bias supply. For reasons mentioned previously, I don't recommend touching the mechanical tuning screw.

I have used both tunable and crystal-controlled 30-MHz i-f receivers in Gunnplexer links, and the difference is like night and day. Tunable receivers are fine if you're interested .ıly in working over short distances, but if you \ /ant to communicate farther than you can shout, you have to use a crystal-controlled i-f receiver. Remember that the local oscillator for your receiver is the Gunn oscillator at the other station; for communications, the receivers at both ends of the link must be tuned to ***exactly*** the same frequency. Even at 30 MHz, a tunable receiver that is off frequency by only 1 per cent will be completely out of the passband of a wideband fm signal.

Automatic Frequency Control (AFC) is helpful when you first turn on your Gunnplexer, but if both stations are operating in essentially the same environment, I've found that frequency drift during warm-up is slow enough that it's an easy matter to keep the other station tuned in. Once the two Gunnplexers have reached thermal equilibrium, they'll sit on frequency for hours at a time.

The receiver I'm using in my Gunnplexer station was described by DJ700.5 In addition to being crystal controlled, it has built-in provisions for a tuning meter *and* signal strength meter; both are extremely useful in setting up Gunnplexer links over marginal paths. Also, the output from the discriminator is available for AFC purposes. If you're interested in serious microwave communications, I highly recommend this receiver.

As supplied, the DJ7OO receiver is built into a tin-plated enclosure with no mounting tabs. If you wish, small L-shaped brackets could be soldered to the enclosure, or the receiver could be clamped into place, In my Gunnplexer transceiver I mounted the DJ7OO receiver with spacers and long screws; this seems to be more rugged than brackets or clamps, and, since the transceiver is designed for portable use, I wanted something that would stand up to unintentional abuse (see **fig. 17).**

If you purchase a DJ7OO i-f receiver, the only problem you may have is obtaining knobs to fit the potentiometer shafts. The diameter of these shafts is 4 mm – too large for 1/8 inch shafts, and too small for 114 inch! The best solution is to purchase knobs for 1/8 inch shafts and drill them out with a no. 22 or 4 mm drill. You can also wind tape around the shafts to build them up to 114 inch, but the knob will tend to feel sloppy and will probably be eccentric.

automatic frequency control

After a Gunnplexer is initially turned on, its output frequency drifts rapidly as the unit warms up. The typical drift rate is about 300 kHz per degree Celsius, and since the Gunnplexer temperature may go up 10 degrees per minute after it's first turned on, total frequency drift is 3 MHz or more. As the unit reaches thermal equilibrium, however, frequency drift slows, and, if the unit is shielded from wind currents, the output frequency is quite stable. If the Gunnplexers at opposite ends of a wideband fm communications link ($\Delta f = 200$ *kHz*) are in similar environments, they can be used for voice communications over long periods of time without any frequency adjustments.

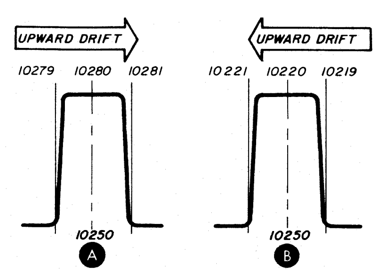


fig. 19. Receiver passband showing upward frequency drift of Gunnplexers operating above (A) and below (B) a Gunnplexer with AFC. To maintain the received signal in the center of the passband requires AFC with *positive* sense in (A) and *negative* sense in (B).

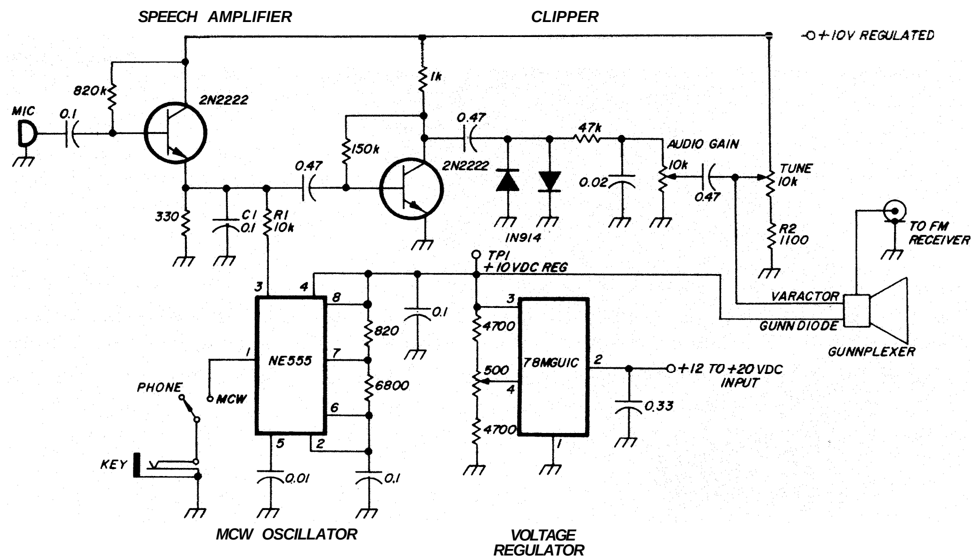


fig. 20. Circuit for a Gunnplexer transceiver without a built-in i-f receiver. In the original model, this circuit was built on perf board. Resistor R1 is adjusted for the desired tone level; R2 is set for a 1-volt drop.

(After an initial warmup of 30 minutes, two enclosed Gunnplexers in my shop remained on channel for more than a day.)

For closer frequency control you can either preheat the Gunnplexer (or use a proportional temperature control system as I suggested in an earlier article?) or use automatic frequency control (AFC). Gunnplexer temperature control would probably be a good choice for use at a base station, but because of

can be used with other fm receivers by simply changing the values of resistors R1 and R2. In some cases the circuit will work as shown, but others will require more (or less) gain — which is set by the ratio of R1 to R2. The only other adjustment is R3, which should be set for a voltage drop of 1 volt.

In the center position of switch S1 the AFC is turned off; the two outer positions provide positive- and negative-going AFC voltage with increasing frequen-

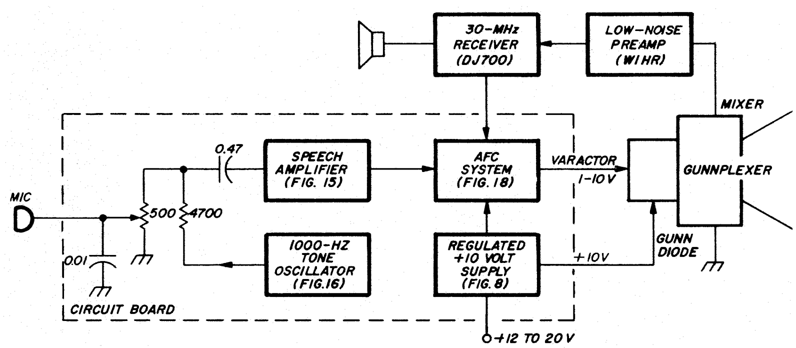


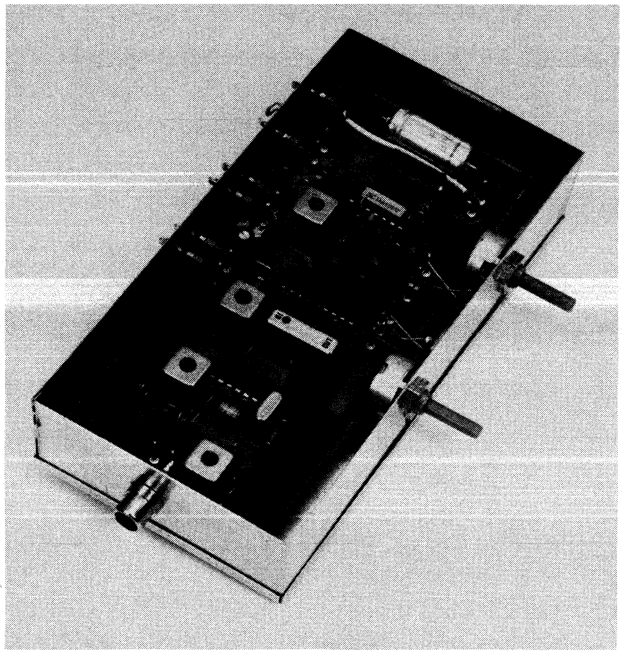
fig. 21. Complete Gunnplexer transceiver featuring high-performance speech amplifier with clipping and de-emphasis, crystal-controlled 30-MHz receiver, and low-noise preamplifier. A circuit board for the speech amplifier, tone oscillator, AFC system, and regulated power supply is shown in fig. 22.

the huge current drain of any heating system, AFC is better for portable use. In an AFC system any deviation in the average value of the i-f from the center frequency of the discriminator in the receiver will produce a dc voltage determined by the direction of the frequency deviation. This dc voltage is applied to the varactor in the Gunnplexer to bring it back on frequency. Note that the use of AFC must be limited to one end of a Gunnplexer link; the other end is allowed to run free.

In many cases, the AFC voltages for the Gunnplexer can be obtained from the i-f receiver. The AFC system shown in **fig. 18** was designed by DJ7OO for use with his 30-MHz receiver.[5] This same basic circuit

cy. Which is chosen depends upon whether the frequency of the Gunnplexer with AFC is above or below the free-running Gunnplexer without AFC. Assume the Gunnplexer with AFC is set to 10250 MHz (see **fig. 19**). If the free-running Gunnplexer is at 10280 MHz and drifting higher, the incoming signal is moving upward through the receiver passband. Therefore, a positive AFC voltage is required to shift the 10250-MHz LO up to recenter the 10280-MHz signal on the middle of the passband. If the free-running Gunnplexer drifts downward, the opposite occurs. In either case, however, the sense of the AFC voltage is the same (positive) as the necessary frequency shift.

Now consider what happens if the free-running



Wideband 30-MHz fm receiver designed by DJ7OO for use with Gunnplexer systems (described in the August, 1978, issue of *ham radio*). At the left is the mosfet input stage, followed by the SO42P crystal-controlled local oscillator and mixer, TDA1047 i-f strip, and TAA611 audio power amplifier. The two controls are for squelch and audio gain.

Gunnplexer is *below* the one with AFC at 10220 MHz. If it is drifting higher, the incoming signal is moving *down* through the receiver passband, and a *negative* AFC voltage is required to move the 10250-MHz LO down to shift the 10220-MHz signal to the center of the passband. Therefore, if the frequency of the Gunnplexer with AFC is above that of the free-running Gunnplexer, the sense of the AFC voltage is opposite (negative) to the necessary frequency shift.

Obviously, the *sense* of the AFC voltage is extremely important. If the AFC sense is incorrect, it tends to chase the received signal out of the passband. In fig. 19B, for example, if positive AFC is used, upward drift toward 10221 MHz will reduce the AFC voltage, moving the LO toward 10249 MHz — the wrong direction! If the AFC has the wrong sense, you'll find it almost impossible to tune in the signal; in many cases the LO will actually oscillate back and forth across the receiver passband several times per second. If you've built a Gunnplexer system with AFC and have experienced this problem, now you know what caused it.

gunnplexer transceivers

To build a complete Gunnplexer transceiver, all you have to do is combine some of the previous circuits and build them into a single enclosure. Two examples are shown in the accompanying photographs. The first, which I call the "minimal" Gunnplexer system, is built into a 125 x 100 x 75 mm (5 x 4 x 3 inch) Minibox and doesn't include the receiver (a phono jack is provided so it can be used with a variety of external receivers). The other transceiver, which is built into a 225 × 150 × 125 mm (9 x 6 x 5 inch) aluminum utility box, includes a built-in 30-MHz receiver with a low-noise preamp and speaker.

The circuit of the minimal Gunnplexer transceiver is shown in fig. *20.* Basically, it consists of the two-transistor speech amplifier (fig. 14), 1000-Hz tone oscillator (fig. 16), and regulated dc power supply. Note that the lowpass filter at the output of the tone oscillator is combined into the speech amplifier. No receiver was included because at the time I built this transceiver I was still undecided about a receiver and wanted to try several options. Since it was built it has been used successfully with a variety of i-f receivers at 30 MHz, 100 MHz, and, more recently, 111 MHz (the New England spot for retuned fm broadcast receivers).

The Gunnplexer transceiver shown in fig. 21 might be called the "deluxe" model. In addition to the built-in 30-MHz receiver and low-noise preamps, it features the high-performance speech amplifier with clipping, audio shaping, and de-emphasis (fig. 15),

10-GHz Gunnplexer system setup by the W1FC group on Pack Monadnock in New Hampshire during the September vhf/uhf contest. Two-way communications were established with Gunnplexer-equipped stations in Maine, New Hampshire, and Vermont.

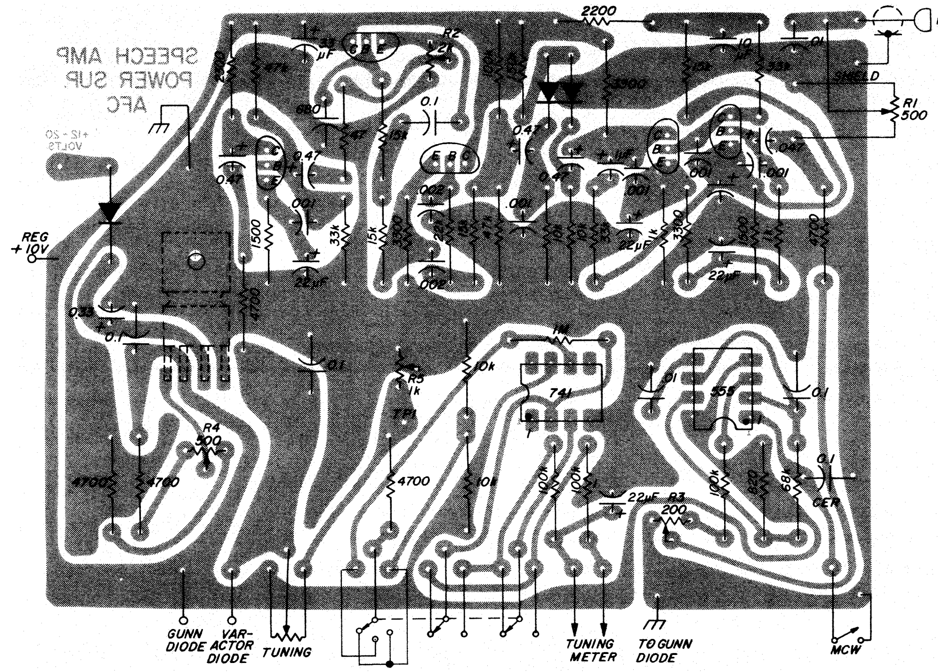


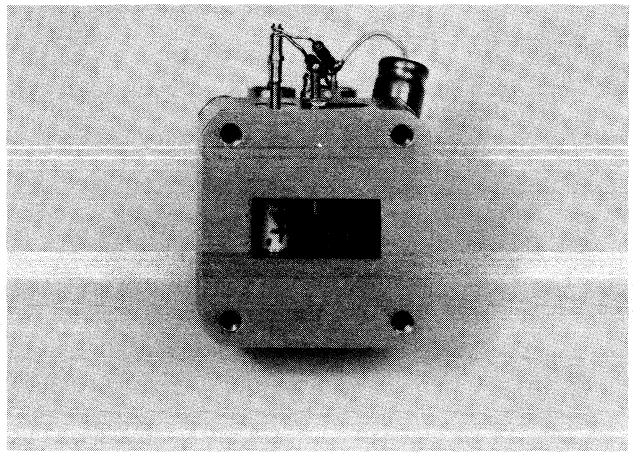
fig. 22. Component layout for the printed-circuit board for the Gunnplexer transceiver. At the top of the board is the speech amplifier with clipping and de-emphasis. Below, right to left, are the 555 tone oscillator, 741 AFC amplifier, and **78GU1** voltage regulator. (Note that the **78GU1** is mounted on the foil side of the board.) In the speech amplifier, pot **R1** sets the microphone gain; pot **R2** is used to set maximum deviation. Pot R3 sets the tone voltage level into the speech amplifier; R4 is the **+10** volt adjust, and **R5** is set for **+1** vott at **TP1**. The capacitors in the audio amplifier are tantalum types.*

an AFC system, tone oscillator, and regulated power supply. To save space and improve reliability, these circuits are built on a printed-circuit board **(fig. 22)**. For improved heatsinking of the 78GU1 voltage regulator, this IC is mounted on the foil side of the circuit board. In addition, an aluminum mounting spacer is used to conduct heat to the chassis. The result is a very cool-running voltage regulator, even with 500 mA of output current. In the transceiver this circuit board is mounted on the rear wall of the utility box. A 100 x 100 mm (4 x 4 inch) aluminum plate, 6 mm (1/4 inch) thick, is mounted in the bottom of the enclosure and tapped for a 1/4-20 screw. This is the standard thread for camera tripods sold in the United States.

When setting up this transceiver, first set the 500-ohm voltage adjust potentiometer, R4, for +10.0 volts at the Gunn diode, then adjust R5 for 1.0 volt at test point 1 (TP1). The tone output level adjustment, R3, is set for the same deviation as the microphone; this and the other audio adjustments are discussed later.

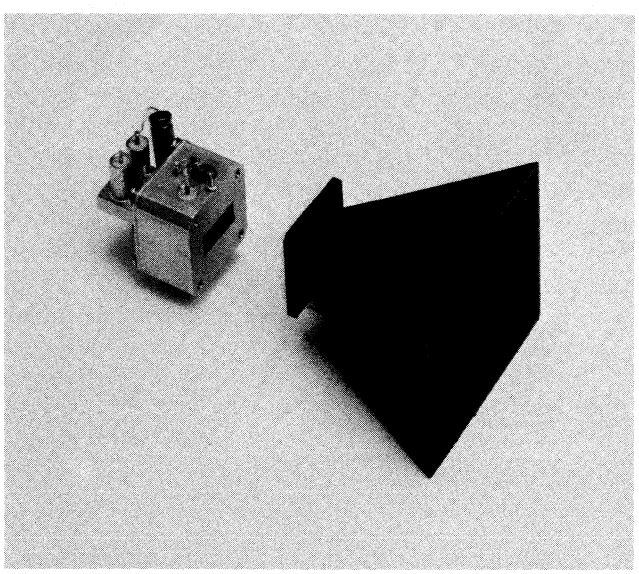
fig. 23. Full-size printed-circuit layout for the Gunnplexer speech amplifier. tone oscillator, AFC amplifier, and voltage regulator. Component layout is shown in fig. 22.

"Complete parts kits, including PC board, are available from G. R. Whitehouse, 10 Newbury Drive, Amherst, New Hampshire 03031.



Head-on view of the Gunnplexer showing the mixer diode, left. and ferrite circulator (black cylinder to right). The small screw which protrudes through the top of the waveguide is used to adjust mixer injection.

One feature of the transceiver which is not shown on the schematic should be mentioned: a small relay to turn off the speaker during voice transmissions. When communicating with a Gunnplexer system, the receiver detects both the signal from the distant station and the local transmitted signal. In addition to being annoying, this is sometimes the cause of unwanted howls and squeals because of audio feedback. To solve this problem, some builders have installed a spdt switch to transfer the audio output to a 4.7-ohm resistor. In my transceiver I installed a miniature spdt relay which is operated by the PTT switch on the microphone (most 12-volt relays work quite well on +10 volts). The speaker circuit isn't affected when the tone oscillator is used for CW, so I have a built-in CW sidetone system.



Microwave Associates 10-GHz Gunnplexer and 17-dB horn antenna. Receiver section is housed in waveguide section machined from large block of metal. This improves thermal stability of the unit.

waveguide flange layout

If you wish to mount your Gunnplexer inside an aluminum Minibox, you must match the waveguide and mounting screws to a cutout in the enclosure. There are feedthrough waveguide flanges on the market, but they're expensive and seldom make their way into the surplus market. The only alternative is to carefully lay out the mounting holes for the UG-39/U waveguide flange and then hand file a cutout to match the interior dimensions of the waveguide. This is difficult if you don't have access to a waveguide handbook because the screw holes are not at the corners of a square, as you might suppose, but are slightly offset as shown in fig. 24. This is done intentionally so it is impossible for a technician to cross polarize sections of waveguide.

To locate the mounting holes for the UG-39/U flange, prick punch the center and use a compass to swing an arc with a radius of 15.5 mm (0.61 inch) as shown in fig. 24B. Now use a carpenter's or machinist's square to draw two vertical lines which are tangent to the arc (fig. **24C**). Using the same center point, swing another arc with a radius of 16.3 mm (0.64 inch) and use the square to draw two horizontal lines (fig. **24D**). The screw mounting holes are located at the intersections of the straight lines. To check their location, swing an arc with a radius of 22.5 mm (0.884 inch); it should cross the center point of each of the hole locations (fig. **24E**). When you are satisfied that the mounting holes are correctly located, drill the holes with a number 18 (4.3 mm) drill for the 8-32 mounting screws. Temporarily mount the Gunnplexer to make sure the holes mate with the tapped holes in the Gunnplexer flange.

After the screw holes have been located you can make the rectangular cutout for the waveguide. This cutout measures exactly 0.4 to 0.9 inch and is centered on the same point as the mounting holes. After scribing the outline with a square, I found the best approach was to drill out the center point with an 8 mm (5/16 inch) drill. This provides clearance for an Adel nibbling tool.

A word of warning: don't try to make the ***finished*** cutout with the nibbler; you're sure to botch the job. Use the nibbler only to make the rough cutout — within about 1 mm (1/32 inch) of the finished edge. Then carefully hand file the edges of the opening so they match the waveguide.

Temporarily install the Gunnplexer to check progress, but carefully wipe off the metal filings first so they don't get into the Gunnplexer. And don't leave the Gunnplexer in place while you're filing the opening — that's an open invitation to disaster!

setup and test

The easiest way to set up a Gunnplexer system is to get together with a friend and set up your 10-GHz stations at the same time. With two Gunnplexers running on the bench, it takes only a few minutes to adjust the speech amplifier and tone oscillator for best performance. Tests and adjustment of an AFC system take longer, but most work can be done in one evening.

The only problem you're apt to encounter with two Gunnplexers in the same workshop is high signal strength — if you have an S-meter, you can be sure it will be against the pin, regardless of the direction you point your Gunnplexers. However, wood makes an excellent microwave absorber, as does the conductive black plastic foam which is often used to protect CMOS integrated circuits. A small section of black foam placed across the waveguide will reduce the radiated signal by 30 dB or more; a section of 2 x 4 pine lumber in front of the Gunnplexer reduces signal strength by 10 or 12 dB. Once you have reduced signal strength to manageable levels, you can make the necessary adjustments. First set one Gunnplexer up with +10 volts on the Gunn diode and +4 volts on the varactor; unless tuned specially by the manufacturer, the operating frequency will be close to 10250 MHz. Now tune the other Gunnplexer to a frequency below the first where you hear the carrier, and carefully adjust the varactor bias for a zero reading on the carrier meter (if your receiver doesn't have a zeroing meter, adjust for maximum signal strength). Make a note of the varactor voltage; this Gunnplexer will now be tuned to 10220 MHz if a 30-MHz i-f is being used.

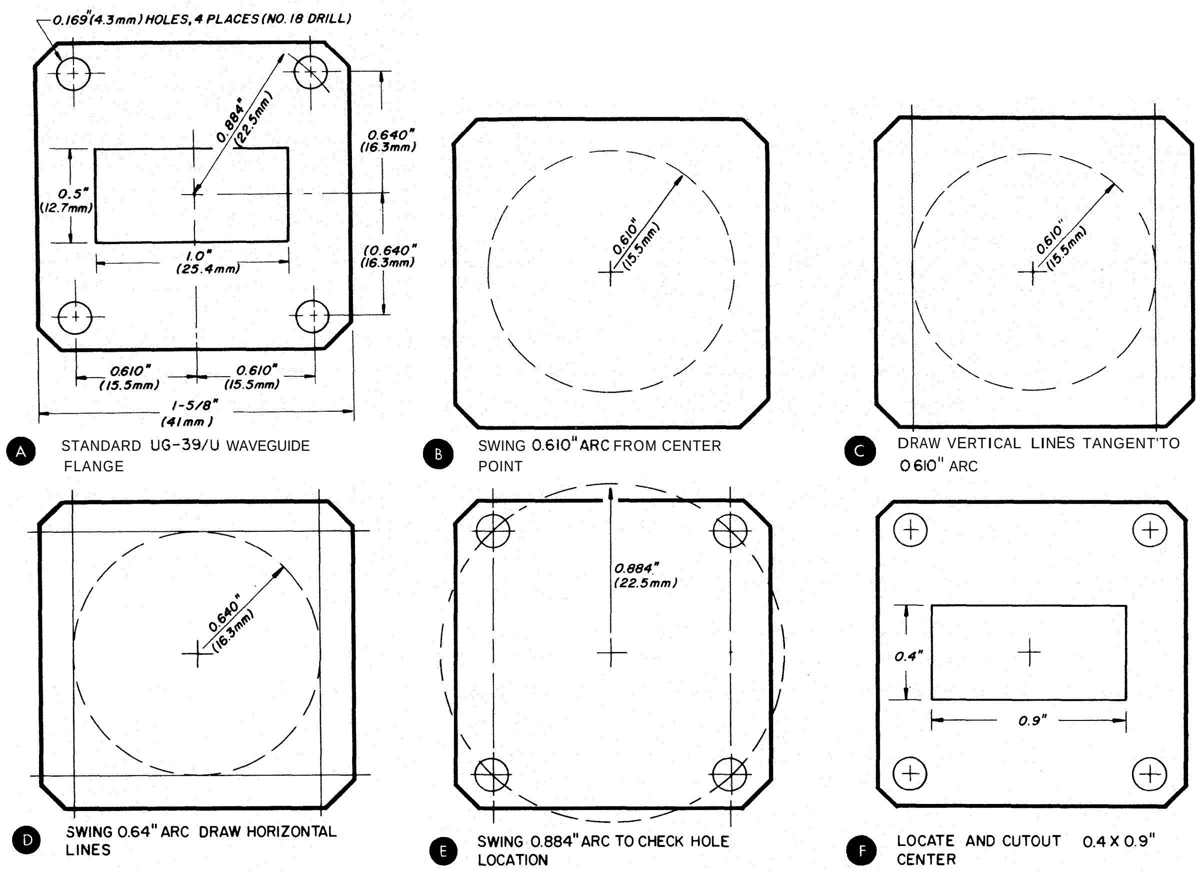


fig. 24. How to lay out the chassis to match a UG-39/U waveguide flange (used on all Microwave Associates Gunnplexers). Note that the flange holes are not symmetrical. Step-by-step instructions are given in the text.

Now tune the second Gunnplexer above the first until you hear the carrier and carefully center the carrier in the passband of the receiver. Make a note of the varactor voltage (Gunnplexer tuned to 10280 MHz with a 30-MHz i-f). If you wish, you can now set the varactor voltage on this Gunnplexer to +4 volts and make similar measurements on the other unit. If you use a turns-counting dial on the varactor bias potentiometer, it's helpful when mountain-topping to know which dial settings correspond to the operating frequencies of 10220, 10250, and 10280 MHz.

Now tune the Gunnplexers to one another and carefully center the carriers, Plug in your microphone and increase the speech gain control. You will note that the received audio signal will have excellent fidelity at a certain setting of the gain control, but, as gain is increased beyond that point, the signal becomes distorted. Back down the gain control to a setting slightly below that which causes audible distortion.

If you wish to measure the actual deviation of your signal, you can use the Bessel function relationship to determine the audio input frequencies at which the fm carrier will completely disappear; this technique is discussed in most of the popular vhf-fm books. Table **1** lists the audio frequencies for carrier nulls at several popular deviations (use 75-kHz deviation for wideband fm receivers); it is not practical to use carrier nulls beyond the third.

In most cases it's not necessary to make an actual deviation measurement; reliable fm microwave communications can be obtained by a simple adjustment of the speech gain control for no audible distortion. Once the speech gain had been adjusted, turn on the tone oscillator and adjust the tone signal level for a signal strength approximately the same as for voice. If you have both input and output controls in your speech amplifier, set the output control for full deviation or minimum audio distortion with the microphone gain control set at about one-half full gain – this will leave plenty of leeway for microphones with higher or lower output. Set the tone oscillator level as before.

If you don't have a friend with a Gunnplexer, you can use the *Boorr .ng* system shown in **fig. 25,** which was origina.ed by the San Bernadino Microwave Society. All you need is an X-band crystal mixer and a 1 to 2 mW local-oscillator source at 30 MHz (if you're using a 30-MHz i-f receiver). When setting up the mixer, be sure to provide a dc return (rf choke) for the mixer diode. Place the mixer 100 meters (300 feet) or so from the Gunnplexer. The transmitted signal from the Gunnplexer will mix with the 30-MHz

table 1. Audio frequencies which will produce a carrier null for various amounts of frequency deviation (use 75 kHz deviation for wideband fm receivers).

modulation frequency	deviation produced 1st null	2nd null	3rd null
2717.3 Hz	k6.53 kHz	515.00 kHz	±23.52 kHz
4528.9 Hz	±10.89 kHz	±25.00 kHz	k39.19 kHz
5000.0 Hz	±12.02 kHz	527.60 kHz	±43.27 kHz
8666.8 Hz	±20.84 kHz	±47.84 kHz	±75.00 kHz
10000.0 Hz	524.05 kHz	±55.20 kHz	±86.54 kHz
13586.7 Hz	±32.67 kHz	k75.00 kHz	±117.58 kHz

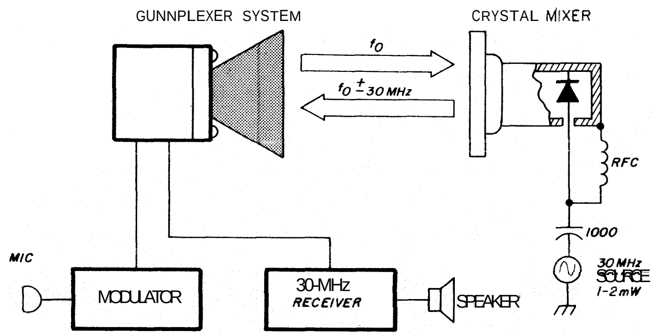


fig. 25. Boomerang system devised by the San Bernadino Microwave Society for testing microwave systems. It requires only an X-band diode mixer and 1 to 2 mW at 30 MHz. The mixer should be placed 100 meters or so from the Gunnplexer to eliminate i-f feedthrough; if the X-band mixer is too close to the Gunnplexer. radiation from the 30-MHz signal source will completely block the i-f receiver.

LO, be re-radiated, and picked up by the Gunnplexer receiver. With this system you can make all the adjustments discussed previously.

When using the Boomerang system don't place the X-band mixer too close to the Gunnplexer. If it is too close, primary 30-MHz radiation from the LO will feed directly through to the i-f receiver. You can tell very quickly if this is happening because the receiver will be completely blocked.

radiation hazard

Although 20 mW isn't usually considered to be very much rf power, in the Gunnplexer it's concentrated at the small, open end of the waveguide, so power density is about 6.2 mW per square cm (up to 19 mW/cm^2 for higher-powered Gunnplexers). This is considerably above OSHA's 10 mW/cm^2 safety limit. Fortunately, rf power density falls off to safe levels with a few feet (2 meters), but remember that your eyes are especially susceptible to damage from rf radiation – never look into the open end of a Gunnplexer while it's operating.

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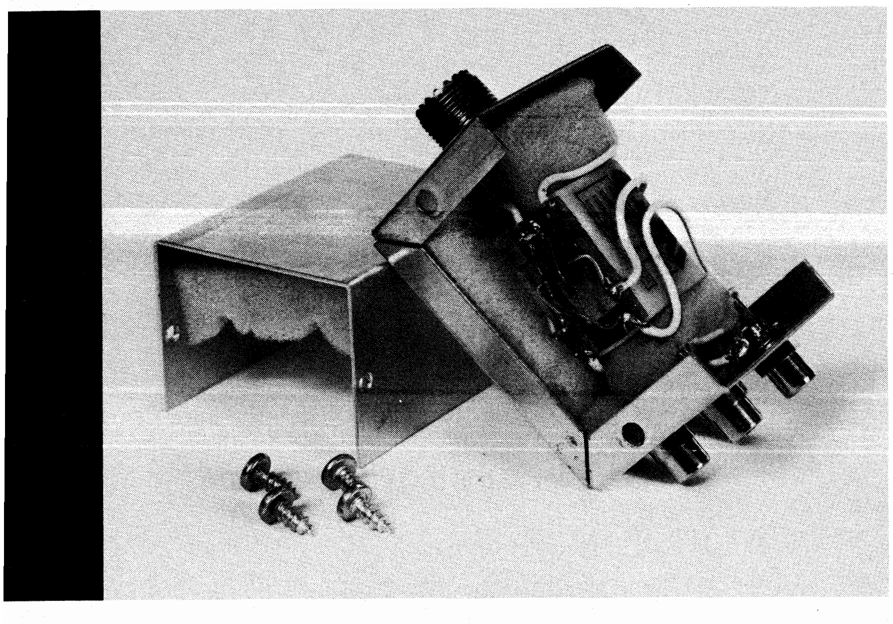
ham radio



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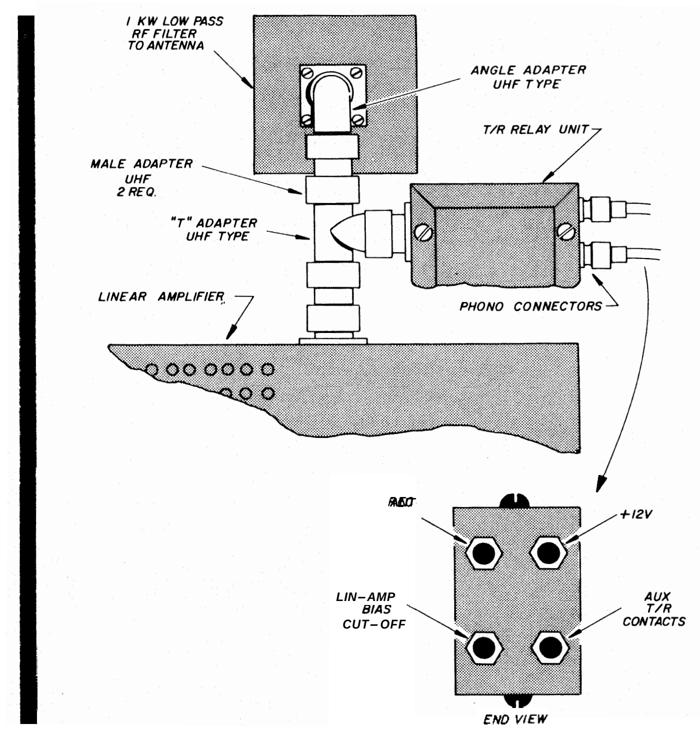

fig. 1. Full-scale drawing of the interconnections between the T/R relay unit, amplifier, and antennas. The four phono connectors, mounted on the small Minibox, are used for the connections to the receiver and control circuits.

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Observations made over a period of years indicate the majority of amateurs still cling to the push-to-talk method of operating their linear amplifiers during CW and voice operation. Continued use (or abuse) of push-to-talk operation can be blamed, in a large part, for the loud clacking noise generated by transfer (exciter/final) relays installed in most linear amplifiers. In an otherwise quiet ham shack, this loud and rapid clacking can become very annoying during both CW and phone operation.

The T/R relay unit described in this article, and linear amplifier modification, will go a long way in reducing the noise generated by the linear amplifier transfer relay; it will also provide fast CW break-in and VOX operation.

Basically, the T/R relay unit and system is a free adaptation of the ideas suggested by Dick Frey, K4XU.[1] The T/R relay unit consists of permanently connecting the operating antenna to the linear amplifier rf output connector through a UHF T connector. As seen in **fig. 1,** the T/R relay unit acts as the interconnection between the exciter/amplifier, receiver, and antenna. When the amplifier is being used, the STANDBY/OPERATE switch, having been rewired, holds in the amplifier's internal transfer relay, with the T/R unit controlling the operating bias. During exciter-only operation, placing the switch in STANDBY will bypass the rf around the amplifier.

construction

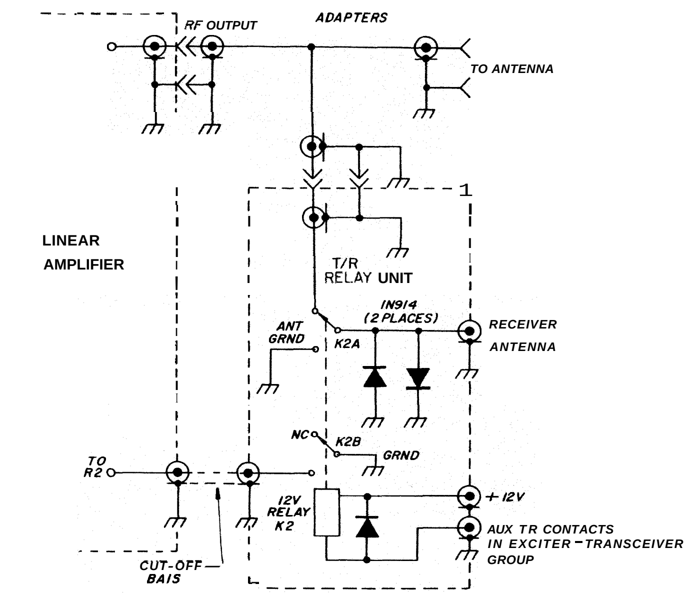
The T/R relay unit consists of a small, aluminum utility box, approximately 7.6 x 7.6 x 5.1 cm (3 x 3 x 2 inches [Radio Shack 270-2351), UHF and phono connectors, and a miniature dpdt 12 Vdc relay [Radio Shack 275-2061. The miniature relay is wired between the connectors using no. 22 (0.6mm) AWG wire. In addition, it's supported by small urethane pads which also serve as sound absorbers.

operation

When K2 (see **fig.** 2A) is operated by the exciter/transceiver auxiliary T/R contacts, K2A transfers the receiver's antenna input from the antenna to ground. K2B shorts the amplifier cutoff bias resistor (R2, **fig.** 2B) to ground, thereby placing the proper operating bias on the amplifier tube. The 1N914 diodes are installed for receiver input protection. The 1N4006 diode, installed across relay coil K2, is used for transient switching protection. This diode has a tendency to delay the release time of K2, however this delay is not noticeable, even at high keying speeds. The modifications, as outlined, are for a TENTEC "Triton IV" transceiver and a DenTron "MLA-1200" linear amplifier. However the principles can be applied to other linear amplifier-receiver/transceiver combinations.

results

Although the response time (T/R switching) and quieting does not approach that outlined by Dick Frey's article, the results have been quite satisfactory – and less expensive. Fast CW break-in at the 1-kW input level has been retained, with no transients present on either transmitting or receiving. A gratifying improvement is the absence of the noise gen-



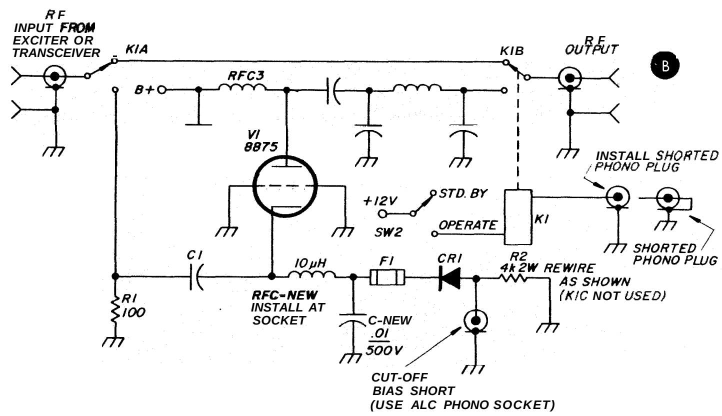


fig. 2. Schematic diagram of the T/R relay and modifications to the DenTron MLA-1200 amplifier. The wiring within the Minibox containing the double-pole, double-throw is at (A). In (B), the MLA-1200 has been modified to provide remote switching of the amplifier's bias. Additional components have been installed to prevent rf from getting into the bias line.

erated by the amplifier transfer relay. Note that this system of break-in can be applied only to receiver/exciter combinations and transceivers having a separate receiver antenna input.

I wish to acknowledge the helpful suggestions of Milt Hirsch, W1AUB.

references

1. Dick Frey, K4XU, "How to Modify Linear Amplifiers for Full Break-In Operation," *ham radio,* April, 1978, page 38.

ham radio

adjustable-voltage
5-ampere power supply

High output current, adjustable voltage, and a low parts count highlight the benefits possible with the Fairchild μA78/79 hybrid voltage regulators

Fairchild has recently introduced its μA78HGC (positive) and μA79HGC (negative) "hybrid" voltage regulators, which should find wide application among amateurs. These regulators are capable of supplying current in excess of 5 amperes over a 5 to 24 volt output range (– 24 to – 2.2 volts for the negative regulator). Load and line regulation is better than 1 per cent. The hybrid nature of these regulators is the result of mating a low-current voltage regulator IC, a Darlington series pass transistor, and two short-circuit detection transistors in one 4-pin TO-3 package. A block diagram of the device is shown in **fig. 1.** The output voltage is set by two external resistors. This design greatly simplifies the construction of relatively high-current power supplies.

The cost of the device is competitive with that of the discrete components. The parts count is a mere eight, including power transformer, rectifier, and filter capacitor, and the numerous possible applications include solid-state power amplifiers, vhf rigs, large digital projects, repeater supplies, audio equipment, and variable bench supplies.

table **1.** Characteristics **of** the Fairchild hybrid. voltage regulators

Input Voltage Maximum	**40** volts
Output Current	5 amps
Minimum Input-Output Differential	3 volts
Maximum Input-Output Differential	25 volts
Line Regulation	1 per cent V_{out}
Load Regulation	1 per cent V_{out}
Control Pin Voltage	4.8-5.2 volts
Short Circuit Current Limit	7 amps

Table 1 summarizes the electrical characteristics of this device family, which also includes three fixed-output devices with otherwise identical specifications: μA78H05C (5 volts), μA78H12C (12 volts), and μA78H15C (15 volts). The fixed-output regulators come in a standard 2-pin TO-3 case and include internal voltage-set resistors.

All of these regulators have thermal overload protection against excessive dissipation or current drain, along with internal short-circuit protection to limit current output. Safe area protection is provided for

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the output transistors. When a short circuit is seen by the regulator, the rise in internal temperature puts it into thermal overload, shutting down the device for as long as the current demand generates excessive heat. The short circuit current limit is 7 amperes.

The basic positive regulator circuit (78HGC) is shown in **fig. 2.** R1 and R2 may be determined by the simple equations shown with the circuit. The nominal reference voltage on the control terminal is **5.0** volts (4.8 to 5.2 volts). To produce the recommended **1.0** mA current flow in the control string would require making $R2 = 5k$ *ohms*. With $R2 = 5k$ *ohms*, the output voltage becomes $V_{out} = [(R1 + R2)/R2] \bullet$ *control voltage* (where R1 and R2 are in k-ohms). For example, if the supply is to provide 13.8 volts and R2 = **5k** ohms, then R1 must equal 8.8k ohms. Precise setting would require trim pots.

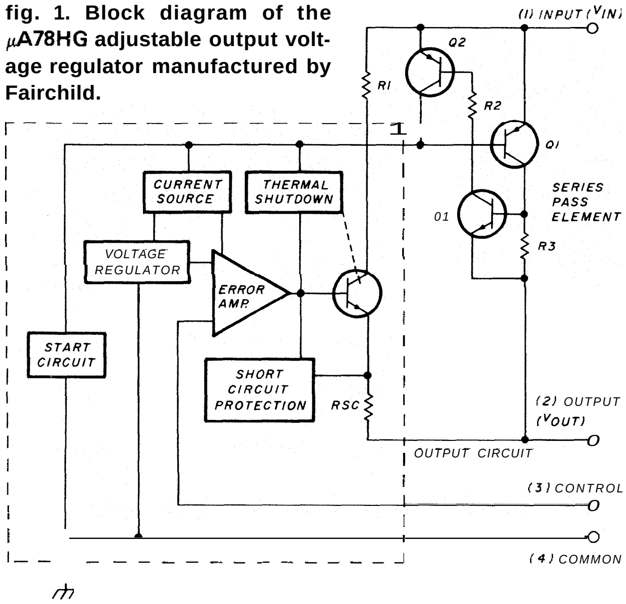


fig. 1. Block diagram of the μA78HG adjustable output voltage regulator manufactured by Fairchild.

As with virtually all such regulators, input and output capacitors should be used to improve transient response and to prevent oscillation of the regulator under certain feedback conditions. These capacitors also provide rf-field protection. Tantalum capacitors are preferred, but good quality ceramic discs may be used. Mounting should be as close to the device as possible.

The four-pin base diagrams (top view) for the regulators are shown in **fig.** 3. Note that the pin-outs for the two devices are different. The case is electrically isolated from the internal circuitry in the four-pin adjustable devices, but is the common in the fixed-output regulators.

Mounting may be accomplished with or without a socket. I have used a modified TO-3 socket by removing the center (collector) pin and drilling two additional holes for pins 2 and 3. Mounting R1 and R2 directly at the regulator will significantly improve the load regulation of the device.

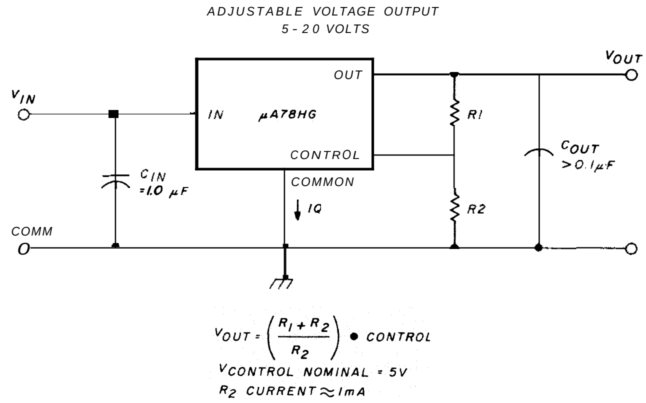


fig. 2. Typical circuit configuration to provide an adjustable output voltage. As discussed in the text, the resistor values are selected for approximately 1 mA of current flow in the divider and also to provide a control voltage input of 5 volts.

This series of regulators is rated for **50** watts of internal power dissipation at a case temperature of 25°C. Increased case temperature, of course, reduces this rating. A graph of maximum power dissipation versus case temperature is shown in **fig. 4.**

To achieve rated performance, attention must be paid to both heatsinking and input voltage, which are interrelated. Under normal operation the regulator will see some input voltages greater than that demanded as its output. The minimum input-output differential should be approximately 3 volts for proper regulation. The greater the difference between the input to the regulator and its output, the greater the dissipation required by the device (actually, by its internal pass transistors). By tailoring the input voltage to the output voltage, heatsinking requirements are reduced, *i.e.*, less heat must be dissipated in normal operation. To draw 5 amps from the regulator would set a maximum limit on the input-output differential of 10 volts. A lower differential reduces the heat to

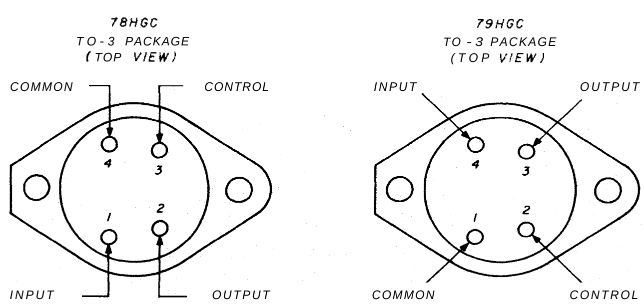


fig. 3. Pinout diagrams for the positive and negative voltage regulators (78HGC and 79HGC, respectively).

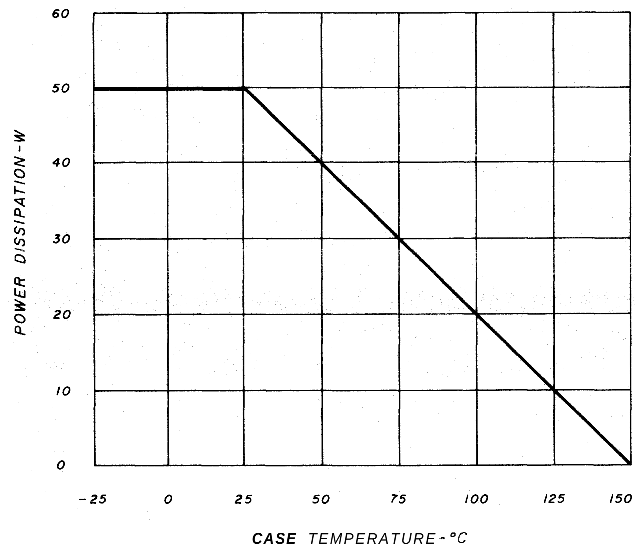


fig. 4. Graph of the maximum power dissipation for different case temperatures.

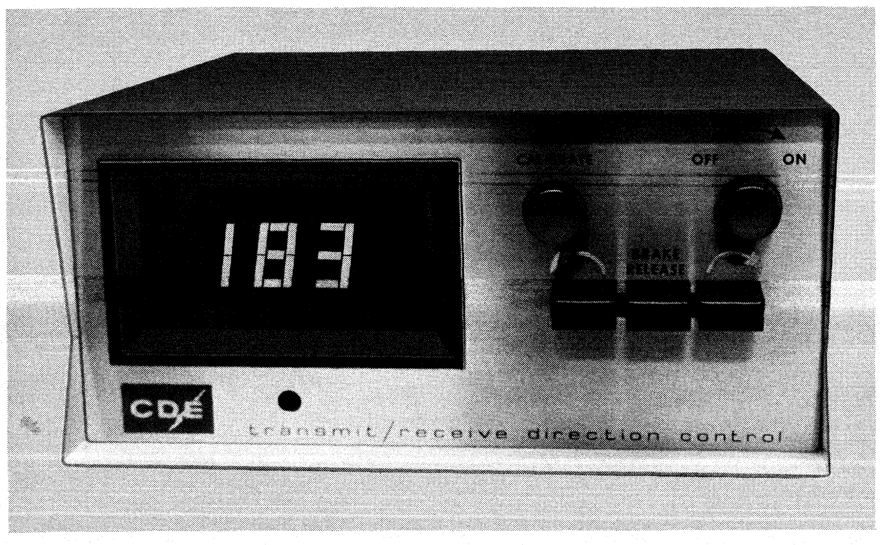
be dissipated. For example, a 13.8-volt supply drawing 5 amps and fed by a 24-volt input would require heatsinking sufficient to dissipate approximately 50 watts in normal use. This same supply fed by an 18-volt input must accommodate only slightly over 20 watts.

Many transformers may be easily modified to adjust their output voltage. Generally, the secondaries of these low-voltage transformers are on the outside and are readily reached after the laminations are removed. A count of the secondary turns will yield the voltage-turn ratio, making it a simple matter to remove (or, for that matter, add) the necessary turns. What you are looking for finally is an output voltage (after rectification and filtering) that, with full load, is only slightly above the 3-volt minimum input-output differential.

Heatsinking must keep the junction temperature below 125°C to meet specifications. Typically, a sink with a thermal resistance of approximately 1.5°C/watt would be adequate. Proper mounting, along with the use of a good thermal compound, is required.

This series of hybrid regulators from Fairchild offers a significant reduction in the parts count and complexity of power supplies. Its substantial curent capacity, along with regulation quality and device protection, make it an economical solution to a wide variety of amateur power-supply applications. The zener diode is increasingly being moved out of the power supply business as integrated regulators become more diverse. These Fairchild regulators are one more step in that evolution.

ham radio



digital readout for the Ham-3 rotator

Add a digital readout to the Ham-M rotator series by incorporating this simple but accurate analog-to-digital converter

It appears that Amateur Radio has gone digital. The signs of change are all around the shack. First came the digital IC keyer, which later had a digital memory added. Digital frequency readout was once a luxury; now, it's a standard feature of most new transceivers. Every shack, of course, now sports a digital clock built from a $3 clock IC. With the exception of the keyer, all the "digital" applications are merely conversions of analog data to digital *readout.* Since this trend is likely to continue, it's worth looking into.

Electronic analog-to-digital conversion can be accomplished using several different techniques." These include parallel (or "flash") tracking, successive approximation, or single- and dual-ramp converters. The flash converter is often used in extremely high-speed applications. Successive approximation is a general-purpose, medium-speed approach, while the dual-ramp or dual-slope is suited to low-speed applications. Integrated circuits are now available at reasonable cost to perform each of these conversions. In choosing a converter, you must consider several aspects: speed, accuracy, resolution, and output format. The actual analog signal being converted must also be considered. Is it a voltage or a current? Perhaps the best way to illustrate the reasoning behind a data-conversion project is by example.

The largest analog indicator in my shack is the meter on my HAM-3 control box, an obvious choice for digital readout. First, consider the A-D converter. Speed is not essential. Conversion times of a few hundred milliseconds are acceptable, so a dual-slope converter is adequate. The accuracy of this system is likely to be limited by the linearity of the Ham-3 indicator system, since most converter products are within 0.1 per cent accuracy. The Ham-3 indicator has an accuracy of about 5 degrees.[1] Resolution is a term describing the number of discrete values that can be recognized, in the same way that digital frequency readouts have resolution limitations. It seems foolish to have rotator readouts to tenths of a degree. A resolution of 1 part in 360 is adequate.

Since converter manufacturers produce both binary- and BCD-output devices, output format must be chosen. In this application, since the readout is a seven-segment visual display, the BCD output is the

"For detailed discussions of conversion techniques an excellent text, *Analog-Digital Conversion Notes,* is available from Analog Devices Incorporated, Norwood, Massachusetts. for $5.95.

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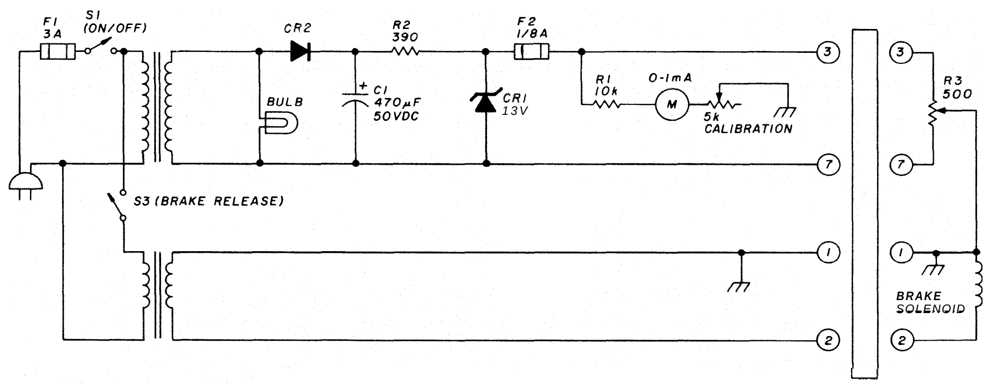
best choice. If you wanted to control the station from the shack computer (no doubt many hams are already considering this), a binary output would be better since computers tend to think in straight binary.

The Analog Devices AD2020, a 3-digit BCD output A-D converter, fits the requirements. It is a low-cost, low-speed, 3-digit BCD output converter widely used in digital panel-meter applications. It is an excellent choice for the HAM-3 digital display because it requires a minimum of external components; the readout is in millivolts. If a 0-359 millivolt signal representing the antenna heading is generated, the IC receives a convenient scale requiring no special conditioning.



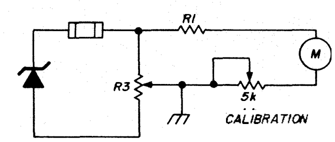


fig. 1. Schematic diagram of the original Ham-3 control circuit. The metering circuit has been redrawn for clarity. Note that the zener reference is floating; the bottom end is not grounded.

signal conditioning

First, consider the original indicator circuit. Referring to **fig. 1,** the 13-volt zener reference is applied directly across the 500-ohm slide wire potentiometer inside the rotator. The wiper of the pot is connected through a 10k-ohm fixed resistor, the 5k-ohm calibration pot, and the meter to the plus side of the 13 volts. The current flowing through the meter is 1 mA at full scale, representing full clockwise rotation. When the rotator is moved to the full counterclockwise position, no current flows. Unfortunately, CDE chose to return the wiper of the indicator pot through the "ground" line, which also carries the ac for the brake and motor circuits. This causes problems with the digital display that aren't apparent when only the original analog meter is used. More on this later.

If you consider the voltage at the wiper of the pot with respect to the minus side of the regulated 13-volt supply, it varies linearly from 0 volts at full clockwise to 13 volts at full counterclockwise. A resistive voltage divider must be used to reduce this to a usable 0-360 millivolt range. In addition, this divider must have a high enough input impedance that it doesn't load down the pot. At mid scale, the pot represents a source impedance of 250 ohms, decreasing to zero at either extreme of its travel. The resistive divider shown in **fig. 2,** 100k-ohms and 2.49k-ohms, will not cause loading problems. In addition, the 100 meg-ohm input impedance of the AD2020 will not cause errors due to loading effects on the equivalent source impedance of the voltage divider.

circuit description

Now that the required signal conditioning has been accomplished, support for the AD2020 chip must be examined. Very little additional circuitry is required. The displays can be any common-anode LEDs. Liquid crystals can be used, at the added expense of some additional circuitry. Personally, I like the LEDs, since they are more visible under the cowl of the control-box cover. Driver transistors for the LEDs can be any pnp transistor capable of delivering about 100 mA. The AD2020 is designed to mate with the Fairchild 9374 decoder/driver chip. This chip differs from the commonly used 7447 in that it has on-board current limiting and requires no resistors. Also, the displays for numbers greater than BCD 9 are different. When used as a pair, the decoding provides EEE for positive overload and --- for negative overload. If a 7447 is used, the – sign decodes as a C. The blanking inputs use different logic on each of these chips, so use caution is you substitute. Current-limiting resistors of 330 ohms should also be used with the 7447. The integrating capacitor is shown on the AD2020 data sheet at 0.27 µF. I used 0.1 µF with good success, so this value doesn't appear too critical.

Power for the readout circuit is derived from available voltages in the control box (see **fig.** 3). Since the services of the indicator lamp are no longer required, it can be removed. Once the lamp is removed, the instrument transformer in the control box is capable of furnishing the approximately 150 mA required by the digital readout circuit. A separate regulator, such as the LM309 or a 7805-type, is used.

calibration and antenna positions

In order to keep the circuit simple, the readout is based on a south-centered scale. This way, one rotation extreme represents 0 degrees and the other represents 360 degrees. In a north-centered system, full clockwide represents 180 degrees, midscale is

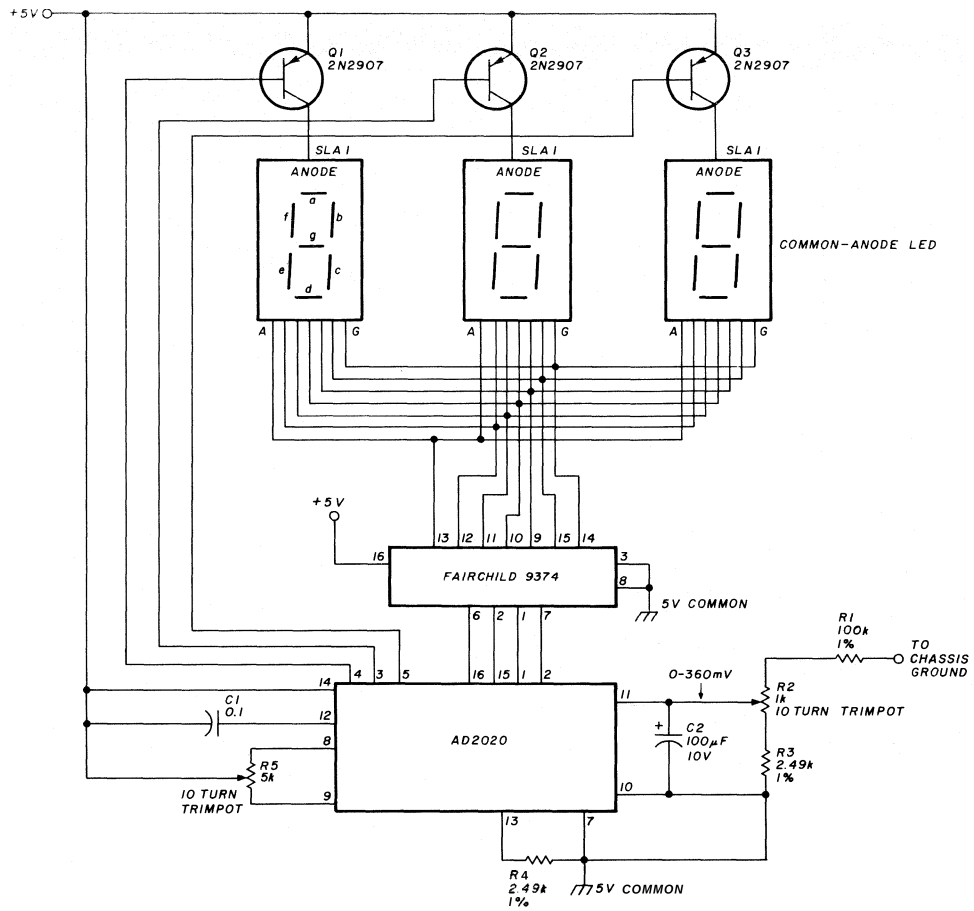


fig. 2. Diagram of the digital readout system. The 5-volt common line is not chassis grounded; it is tied to the bottom of the zener reference. C2, across the input of the A-D converter, is used to prevent the ac voltage developed across the cable resistance from producing an erroneous readout.

This regulator is supplied from the half-wave rectified and filtered dc from the transformer, tapped off points 11 and 13 on the power supply board of the Ham-3. Point 13 is treated as the "ground" for this supply. A heat sink *must* be used on the regulator, or separate regulators should be used for display power and for power to the AD2020 chip.

Consturction of the circuit is not terribly critical. I built the prototype on perforated board cut to fit behind the original meter bezel. A scrap of rubylith was used as a red filter to conceal the other components on the board and to yield a nice-looking front panel.

360 degrees or 0 degrees, and full counterclockwise represents 180 degrees again. This presents a more complex problem. A 180-millivolt offset must now be switched in and out, requiring a comparator, relay, and reference supply. In my opinion, the headaches of trimming such a circuit outweigh the hassle of climbing the tower to turn the beam 180 degrees inside the rotator. In addition, changing the stop to north from south has another hidden advantage. In general, propagation follows a clockwise route. This means that conditions tend to favor Europe, then Africa, swinging south through the Americas, on into the Pacific, and finally to Japan. In the original

configuration, having the stop at south made you swing your beam almost a complete revolution in order to follow the propagation as the peak passes due south.

When you change antenna position, first turn the rotator full counterclockwise. Then turn the antenna north with enough feedline slack to allow a full clockwide rotation. Finally, reverse the connections to terminals 3 and 7 on the back of the control box. This is done to provide a full-scale voltage at full clockwise rotation. As originally configured by CDE, full clockwise provides full-scale *current* and zero voltage at the wiper of the pot relative to pin 7.

Calibration of the circuit is fairly simple. The original CALIBRATE button on the control box now serves no function. First apply power to the unit with the rotor in the full counterclockwise position. After a few minutes' warmup, adjust the zero pot, R5, for a reading of 000. Now rotate the antenna full clockwise. Then adjust the voltage divider trimpot, R2, for a reading of 360. The readout is now fully calibrated.

When I first installed the readout in my Ham-3, the display was rock stable until the brake release switch was pressed. As **fig.** 4 shows, a large amount of brake and motor ac current flows through the common ground connection, causing the display to jump wildly. The problem was solved by placing C2, a 100-μF electrolytic capacitor, across the inputs to the AD2020. This capacitance represents a low impedance to 60 Hz ac and reduces the effects of the ground currents to less than 1 count. In some cases, rf bypassing might also be required. A .001-μF ceramic capacitor should provide adequate rf filtering.

summary

Before long, dozens of other analog functions around the ham shack will be converted to digital, either for readout or for computer control. The same considerations discussed in this article will arise in any situation requiring analog-to-digital conversion. As converter ICs become increasingly available to the amateur, an understanding of the underlying principles will become important.

Consider the following scenario. The station digital clock indicates 0000 UTC, signaling the station computer that the contest has begun. The receiver tuning algorithm is initiated, applying voltage through a digital-to-analog converter to a varactor in the receiver vfo. A signal is found, and the program jumps to the identification routine, including the Morse code translator. The station is identified and found to be calling CQ TEST. A quick check through memory shows that the station is not a duplicate. Elsewhere in memory, the correct beam heading is found. The rotator position is read in from its analog-to-digital converter, and the computer determines in which direction to begin turning the antenna. At the correct heading, signal strength is read from another A-to-D, and the RST for the exahcnge is computed. The keyer speed is adjusted and the computer now calls the other station. When the QSO is completed, the computer logs the contact and the sequence repeats. With an advanced system like this, a contest operator can relax and watch the football game on TV while his station operates itself and prepares a printed, duplicated log within minutes of the end of the contest.

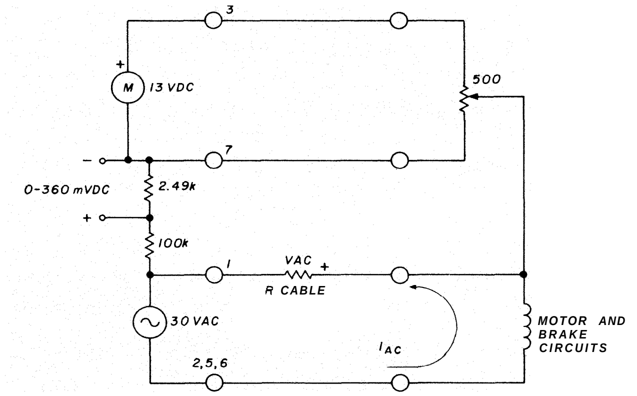


fig. 4. Equivalent circuit diagram of the control box used to illustrate the ac ground loop problem. The ac current flowing through the cable resistance causes a large ac voltage to be added to the direction indicating voltage at pin 1. Filtering across the 2.49k resistor is more effective, since a capacitor across terminals 1 and 7 is effectively shorted out when the pot reaches either extreme of its travel.

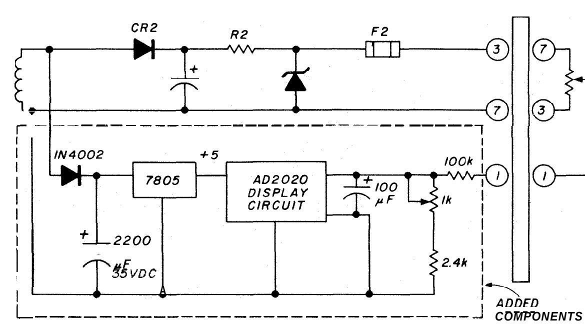


fig. 3. Wiring diagram of the modified indicator circuit. Wires on pins 3 and 7 of the terminal strip have been reversed, causing the readout to increase as the antenna is turned in a clockwise direction.

references

1. Richard Klinman, W3RJ, "How to Update Your Ham-3 rotator," *CQ*, June, 1978, page 34.

ham radio

anodizing aluminum in the amateur workshop

Complex chemical processes for treating aluminum are translated into simple procedures for your home lab

Aluminum is used in many construction projects. But how do you decide on which type of aluminum to use? If you're interested in making panels, chassis, or boxes to house equipment, there's a right way to process the metal for durability and appearance. This article gives some pointers on how to process aluminum by anodizing, a chemical process that can be used in your workshop. Also included is information on how to apply colored dye to aluminum parts using simple procedures.

Aluminum is one of the most abundant elements on earth, forming about 8 per cent of the earth's crust. It's relatively inexpensive, easily machined and worked, lightweight yet strong, and an excellent electrical conductor. Its disadvantage, when uncontrolled, is its pronounced affinity for oxygen: a process called corrosion.

Aluminum oxidizes rapidly. Its natural surface breaks down, causing it to be unsuitable for applications where a long-term stable surface is needed. In ordinary atmospheric environments, even when few pollutants are present, alloyed aluminum surfaces oxidize within moments. The oxide is invisible to the naked eye; even the apparently bare surface of a recently machined aluminum part is immediately coated upon contact with atmospheric oxygen.

controlling oxidation

The formation of surface aluminum oxide can be controlled by anodizing. An electrochemical process is used to form the crystalline structure known as gamma aluminum oxide (γ-Al_2O_3) in an electrolytic cell, with the item to be anodized becoming the anode in the cell. The nature of the anodized metal has particularly significant physical characteristics:

1. Extreme hardness, approaching that of diamond
2. Electrical nonconductivity
3. Extreme porosity on a molecular scale

The gamma aluminum oxide film is closely related structurally to other oxides of aluminum, such as those used in manufacturing synthetic grinding wheels, and commonly substituted for natural corundum. Synthetic sapphires and rubies are, in fact, oxides of aluminum. Even though several different compounds are designated aluminum oxide (Al_2O_3), the crystal lattice structure takes on many quite

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peculiar variations, hence the designation of the anodic film as the *gamma* aluminum oxide.

Aluminum oxide coatings, or films, vary greatly from transparent to opaque, depending on film thickness and also on the alloying elements present in the aluminum alloy used. The film thickness is controllable, and can be from one to twenty microns.* The fact that the film thus generated is molecularly bonded to the aluminum and has a porosity that can be dyed makes an anodized film the most durable and useful of all possible finishes for this metal.

The porous surface of the gamma oxide film, whether dyed or left clear, is easily converted by immersion in boiling water to the closed, or sealed, crystalline state of the monohydrate of aluminum oxide, known as boehmite, designated $Al_2O_3.H_2O$ (or more correctly *A100H*). Boehmite has a large volume/area ratio of aluminum; therefore, the volume of anodized film is increased to close the pores, or *seal* the film. The conversion of the simpler gamma aluminum oxide to the boehmite structure makes the anodized surface unstainable, the dye unleachable, and the item so treated remains permanently dyed in the chosen color.

the anodizing process

Aluminum is anodized by immersing it in an aqueous electrolytic solution in which the aluminum item to be anodized becomes the anode (positive pole). A direct current is passed to it from the cathode (negative pole). Oxygen released from the water combines with surface aluminum molecules to form aluminum oxide; the crystalline lattice is of the gamma form. Although the acid electrolyte is not *used* in the *oxide-forming* reaction, it influences the characteristics of the formed film.

Two common methods of electrolytically producing an anodic film on aluminum offer different properties in the formed anodic film. The two methods use different acids in the electrolyte, chromic or sulphuric. These methods are discussed at length below.

alodizing

An alternative method of protecting aluminum is called alodizing. This method provides no color choice and results in only about a 3-micron (3×10^{-3} mm or 1.2 x 10-4 inch) film thickness. This process is a chemical-dip treatment, which produces an electrically conductive coating, usually of a smutty, mustard-like color. After buffing, alodizing does offer a degree of protection to the otherwise easily corrodible metal. An unbuffed alodized finish accepts primer and finish painting.

anodizing by the chromic-acid process

The chromic-acid process is not suitable for aluminum alloys that contain more than 5 per cent copper, but it is fine for all other aluminum alloys. The chromic-acid process is especially recommended for anodizing assembly parts, particularly where inadequate flushing and rinsing of trapped sulphuric acid could lead to later problems. The films generated by the chromic-acid process are thinner than those obtained with the sulphuric-acid process, but despite their relative thinness, they are durable and offer a highly stable protective coating — from a corrosion standpoint especially.

The thinness of chromic-acid coatings is sometimes of value in manufacturing procedures, especially where ultra-close fits are involved and where matching is to be within sub-mill tolerances. The U.S. government specification MIL-A-8625A (December 14, 1954), which calls for 250-hour salt-spray resistance, authorizes the chromic-acid process for all aluminum alloys except those bearing more than 5 per cent copper. Although the chromic-acid-generated anodized finish is much harder than the untreated metal itself, only limited abrasion resistance is afforded by this process because of the extreme thinness of the film. Also, with the sulphuric-acid anodizing process, a greater porosity occurs, with increased dye take-up in the thicker coating.

For the purposes of the average amateur requiring the anodized film characteristics of hardness, durability, and acceptance of dyes, the sulphuric-acid process is preferable.

Three practical references,'-3 supply details on the chromic-acid process for those rare applications where an amateur may need it.

aluminum alloys for effective anodizing

As mentioned, the qualities of the anodic film produced are affected significantly by the presence of other metals alloyed with the aluminum. Depending on the intended use of the aluminum, cost and availability may dictate which aluminum alloy is used, rather than the precise and sometimes subtle differences between the various alloys. The chemical differences between the various aluminum alloys can

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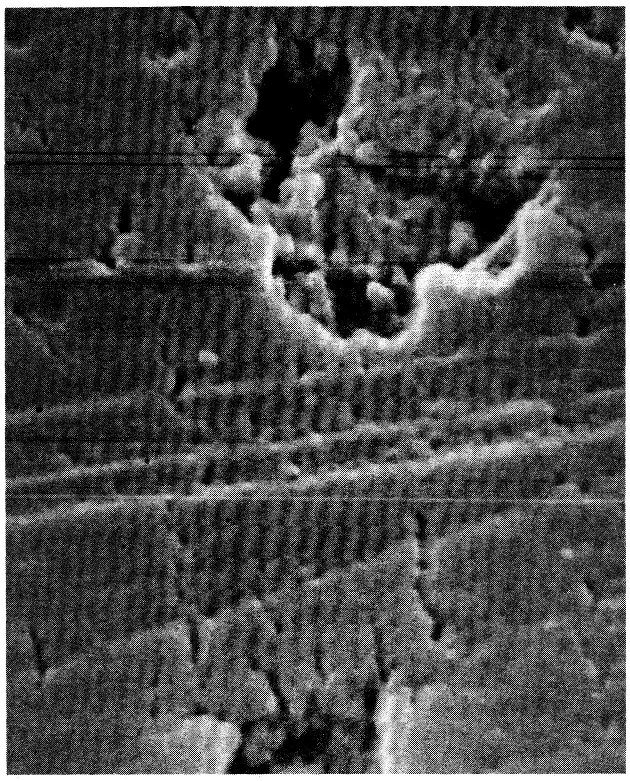
"To get an idea of the magnitude of the dimensions involved, the following conversions are given:

I micron	=	10^{-6} meter (3.94 x 10^{-5} inch)
1 Angstrom	=	10^{-1} millimicron or 10^{-7} mm (3.9 x 10-$inch)
25.4 microns	=	2.54 (10^{-2}) mm (10^{-3} inch)
1 micron	=	10^4 Angstroms



Scanning electron micrograph (x 12,000). showing anomaly in anodic film probably caused by a carbon speck or other alloying constituent. Surface appears extremely smooth and lustrous, even under optical microscope inspection. Note that, even at 12,000 x magnification, molecular scale porosity can't be seen. Porosity is important in dye takeup of anodized parts. (Photo courtesy Dept. of Metallurgical Engineering, University of British Columbia, Vancouver, B.C.)

usually be found in the handbooks of large industrial suppliers of nonferrous metals.

In its purest form, aluminum is very soft and quite ductile. For most purposes, however, greater strength and hardness are required; thus high-purity aluminum is seldom used. Greater strength is achieved in two ways. Usually, the pure metal is alloyed with other metal elements, such as manganese, copper, silicon, iron, and magnesium. In addition, the alloy may be heat treated to give it even greater strength. The designation T plus a number after the 4-digit alloy number indicates a heat-treated alloy.

Aluminum of high purity, when anodized by the sulphuric-acid process, yields a completely colorless anodic film. This film can be left clear and sealed off, or dyed and then sealed off. The paler colors are more readily dyed into the anodic films generated on the surface of pure, nonalloyed aluminum. The more alloying impurities present, the greater the tendency toward a pale-green or pale-brown cast to the undyed anodic film. This is seldom a problem, unless matching of the various parts of a structure made from different alloys is desired.

When pale dyes are to be used, the original alloy must be considered, and the acid concentration as well as current density should be controlled for a thick film formation; *e.g.,* 15-29 microns or about 1.5×10^{-2} mm (6×10^{-4} inch), thus permitting greater dye take-up, The more concentrated the acid electrolyte solution, the softer and more porous the anodic film.

Experimentation is often required to achieve the required anodic film properties of a particular dye. A sample scrap piece of the alloy to be used can be processed in a trial run, thereby assuring more predictable results.

Of course, more predictable results can be obtained when alloys of known composition are used. Some alloys are better candidates for taking an anodic finish than others. For example, the well-known Alcan 6061 T4 and 6061 T6 take an excellent anodic finish. The finish will reflect absolutely and exactly the smoothness or roughness of the final machining operation.

Anodizing makes no noticeable difference to the texture of the aluminum. To see the finish (except for the color), even the strongest optical microscopes are useless. As discussed earlier, the scale being dealt with here is molecular; the transmission electron microscope is required to reveal the anodic film texture. Even with a scanning electron microscope (SEM) the porosity is invisible.

the sulphuric-acid process

First of all, aluminum anodizing should not be done indoors unless special ventilating equipment can be installed. Ideally, the anodizing workshop should be outdoors with plenty of air circulation; the ideal outdoor workshop is a home carport or garage with all doors open. If something approaching a chemistry laboratory fume hood with a spark-free extractor fan can be placed over the anodizing tank to exhaust the gaseous hydrogen emitted at the cathode, indoor anodizing may be possible. Remember that electrolytic dissociation breaks water down into two atoms of hydrogen for each atom of oxygen. Oxygen reacting at the aluminum-anode surface — allowing aluminum-oxide formation — causes the liberation of hydrogen gas at the cathode. This gas is emitted with a small amount of acid vapor from the electrolyte and is best vented outdoors, where it will be rendered harmless by mixing with air.

safety precautions

Small anodizing jobs in the home workshop can be done safely. The degree of hazard is similar to that of quick-charging an automobile lead-acid storage battery. The acid concentrations are roughly the same and the amount of discharged gaseous hydrogen is similar.

The major hazards are the effects of acid on skin or eye tissue and the risk of a spark's igniting hydrogen gas. Both hazards are avoidable. The golden rule of mixing acid is always *pour concentrated acid into the* water — *and* slowly! This allows the heat of the chemical reaction between the water and acid to be absorbed by the larger volume of water. If a sudden expansion of the smaller amount of acid should occur due to the rapid temperature increase that occurs on contact, it is water that is present in quantity, rather than the more dangerous acid. Once the 15-25 per cent solution of sulphuric acid is mixed, it then becomes the working electrolyte, in which the anodizing process takes place. After evaporation has occurred the tank can be topped-up safely by carefully pouring more water into the dilute solution.

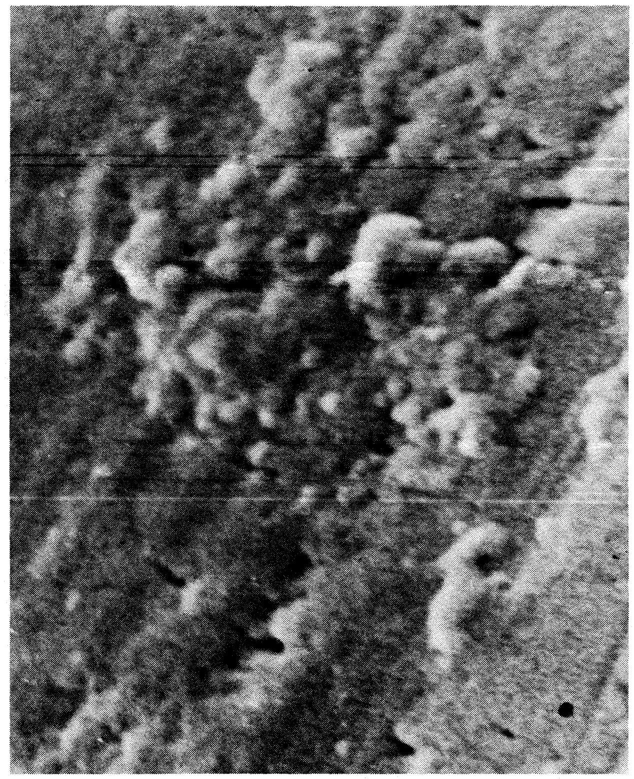
When mixing acid, immersing an item to be anodized, or removing it, wear protective clothing such as an apron made from a heavy fabric (canvas or rubber).

Use large rubber gloves to protect the hands and wrists, and acid-proof safety goggles over the eyes. Even a tiny splash of only a few milliliters of acid can cause serious damage to the eyes. If appropriate precautions are taken and the working area is clear and safe, the degree of risk is minimized. As with other procedures, human error, misjudgement, and carelessness (including too much speed) are most dangerous. Keep a pail of water handy!

the anodizing tank

The anodizing tank must be large enough to accommodate the lead cathode (which takes up very little space) and the largest article to be anodized. (If nothing larger than a thimble is to be anodized, the tank could be a plastic coffee cup, and the process could be done indoors with minimal ventilation.) Most of the items anodized in my setup were small — seldom larger than a dinner plate. The tank can be any container that's nonconducting and impervious to dilute sulphuric acid. A plastic pail, a hard rubber vessel, a glass tank (such as may be salvaged from a large lead-acid storage cell) or even a heavy plastic kitchen dish pan can be used.

The cathode must be constructed of lead. If the tank is a polyethylene or hard-rubber pail, round or square, the cathode can be easily fitted from a sheet of plumber's lead. Cut a 1-3 mm (0.04-0.1 in.) sheet of lead so it can be rolled into a liner in the shape of an open-ended cylinder that can be placed inside the



Scanning electron micrograph photo (x 16,000). Although porosity isn't visible, the crystalline texture of the gamma aluminum oxide can be seen. The surface appears lustrous and smooth and is highly reflective. (Photo courtesy Dept. of Metallurgical Engineering, University of British Columbia, Vancouver, B.C.)

walls of the pail. A cathode termination can be made from a 20-30 mm (0.8-1 in.) wide strip of the same lead sheet, soldered to the upper edge of the cylindrical cathode, and extended up to the top of the tank. At that point, clear of acid contact, the lead can be soldered to a flexible length of 3.3 or 2.6 mm (no. 8 or 10) copper wire for connection to the negative terminal of the anodizing power supply. The size of the cathode, in surface area, must be at least equal to the area of the surface being anodized. The cathode in my tank covers the interior walls of the pail (20 liters, or 5-1/2 gallons) and extends the full depth of the acid contained in it when about two-thirds full. The bottom of the pail remains uncovered by the cathode so that some items being anodized may be set on the nonconducting tank bottom. The tank bottom could also be covered with lead, offering a larger surface-area cathode, but it would then be difficult to avoid contact with the bottom.

The electrolytic-tank anode pole is formed by using only aluminum, including aluminum screws, bolts, or other connectors, except for parts which no acid will contact. The item to be anodized can be fastened either by friction fit or by aluminum fasteners to an aluminum rod or strip and hung into the central area of the tank. A wooden slat or two across the tank top serves as a stabilizer for the central anode fixtures. Connection by aluminum fasteners to the item to be anodized should be made on some part of the item where it won't matter. The point of contact – where the anode connection is made – obscures a small area that remains unanodized.

acid concentration

The sulphuric-acid electrolyte used for anodizing should be between 15 and 25 per cent concentration by weight. The table below shows appropriate quantities of concentrated sulphuric acid for dilutions between 15 and 25 per cent by weight.

acid dilution (per cent)	concentration per liter (quart) of water
15	173 ml (5.9 oz)
18	212 ml (7.2 oz)
20	240 ml (8.2 oz)
25	310 ml (10.5 oz)

The plastic containers of sulphuric acid sold by automotive parts stores for filling new automobile storage batteries make an excellent source of acid for anodizing. Simply mixing the acid with water in a 1:1 ratio makes a good dilution for a working anodizing solution.

Minor impurities that occur in drinking water, mainly small concentrations of minerals and alkali, will have little effect on the anodizing results. The other metals in the aluminum alloy appear to play a more important role in determining the undyed color of the anodic film. If the water is especially alkaline, the resultant acid concentration obtained by the table may give lower actual concentrations due to neutralization.

power supply

The power-supply capacity that will be needed is determined by the size of the aluminum items to be anodized. The current density required in the sulphuric-acid process averages 1.5 amperes per decimeter2 (3.4 inch^2) of surface area of the item to be anodized. However, the current density will vary in relation to several factors:

1. Acid concentration
2. Voltage potential between anode and cathode
3. Electrolyte temperature

The voltage potential is not critical, although the softer and more porous films are generated at the higher voltages and current densities. Any voltage between 6 and 20 volts will generate an anodic film, but 16-18 volts appears to be optimum when working with electrolyte temperatures between 18 and 22 C (66 - 72 F). The power supply must be able to produce full-wave direct current, not necessarily filtered, of 18 volts and 30-50 amperes for periods of about an hour without overheating.

Ordinarily the voltage is preset. Current flow will then be determined by acid concentration, temperature, and the surface area of the item to be anodized. A voltmeter and ammeter are helpful. Current flow does not decrease with anodic film buildup as in the chromic acid process.

operating conditions

Anodizing can be done with the electrolyte at a number of different temperatures.

When the electrolyte is started at room temperature ($\approx$20 C, or 68 F), after several hours of anodizing a rise in electrolyte temperature can occur. This increase varies with the size of the items being anodized and the current flow through the electrolyte. As the temperature rises, the anodic film will be softer and more porous – which makes for better dye takeup – but the film will have a reduced hardness. The tank can be left to cool at this point. Sometimes the rise in electrolyte temperature is acceptable, especially when extreme hardness is not necessary and when the dark-colored dyes, such as black or deep blue, are being used.

If accelerated electrolyte cooling is required (seldom necessary in most amateur setups), the cathode could be constructed of lead tubing, with a suitable coolant pumped through it, thus permitting the electrolyte temperature to be thermostatically controlled.

Anodic-film porosity is controlled by acid concentration, current flow, and voltage, all of which are interrelated. Film thickness, however, is controlled by the length of time of film generation. Some experimentation will demonstrate more exact times and there is some latitude in this variable; but generally two categories of time length apply: if the anodic film is not to be dyed, about 15-25 minutes is usually ample. If the film is to be dyed, however, and especially with the dyes requiring a high degree of film takeup, periods in the range of 45-60 minutes should be used. Different alloys will require different times, even when all other variables are held constant, including voltage, current density, electrolyte temperature, and electrolyte concentration.

sealing

The anodic film generated in the electrolyte is gamma aluminum oxide. The molecular porosity of this oxide and its extreme hardness are desired characteristics. If the surface is not to be dyed, however, it will offer greater permanence to its uniform coloration if it is converted to the nonporous boehmite. This conversion is easily made by immersing the rinsed and clean anodized item in boiling water for about20 minutes.

If the anodic film has been dyed, the sealing process of simple immersion in boiling water can cause dye leaching. To avoid this, chemicals are added to the sealing solution, usually a low concentration of nickel acetate in water held at 95 - 98 C (203 - 208 F). The sealant chemical may vary depending on the dye. I've never required an antileaching agent, since a small amount of leaching has been tolerable.

dyeing

Commercial procedures for the uniform dyeing of anodized aluminum can be complex and expensive. One of the most critical factors is the *pH* of the dye solutions and sealant solutions.

The purpose of anodizing an aluminum surface, aside from increased durability and hardness, is to produce a porosity that will allow dye to penetrate. As mentioned, however, the porosity formed from the anodizing process is on a molecular scale. Unlike dyes used for cloth, where absorption of dye is an easy matter because of the large pores in the fabric, dyes capable of takeup by an anodized surface must have molecular constituents small enough to fit into the pores in the film surface. Many ordinary dyes that permanently stain ordinary fabrics have no effect whatsoever on the more subtle porosity of anodic films. For this reason, special dyes have been developed.

In North America, two large suppliers of commercial dyes and supplies for the anodizing industry are Sandoz* and the Allied Chemical and Dye Corporation.† Chemical dyes available from them come in about 50 different colors. If a more limited selection of dyes can be accepted, some ordinary, inexpensive fabric dyes sold in drugstores will prove satisfactory for anodic film takeup if a few special measures are taken. Wool dyes must be selected, rather than those only for cotton or other fibers. The *pH* must be controlled and the concentration must be higher. Usually about 25 grams/liter (0.9 oz./qt.) of the solute, with the addition of about 1 ml/liter (0.03 oz./qt.) of acetic acid (vinegar), will yield reasonable coloring results. Certain dyes, because of their large molecular size, will be unusable. However, after experimentation, you may find that many different dye colors can be used at a fraction of the price of commercial anodizing dyes. The golds and blues tend to be most effective, some without the addition of acetic acid.

For the most effective dye takeup by a well-generated anodic film, heating the dye solution to 55 - 75 C (131 - 167 F) is required. The dye takeup won't increase after about 10-15 minutes immersion in the heated dye. Different dyes take up at different temperatures; experiment to find the optimum values.

Very dark black anodic dye will probably have to be purchased from one of the commercial dye sources or from an anodizing shop. Commercial anodizing dyes are extremely powerful, so only a small amount will be required.

After repeated use the dye will become gradually acidic from acid leaching out of the anodized surfaces, even though these surfaces have been carefully rinsed. At this stage the *pH* of the dye solution must be restored by adding small amounts of alkali, usually lye solution. If the inexpensive drugstore fabric dyes are used, an alternative to fussy pH control is to replace the acidic dye with a fresh mix of new dye, a practice that has been acceptable in my experience.

summary

Anodizing aluminum is an exact science. For the amateur in the home workshop it may be an art that requires much experimentation before you develop consistent results. But anodizing offers many advantages over other protective coatings and yields a permanent and stable finish for aluminum.

*Canada: Sandoz Colors and Chemicals, Box 385, Dorval, Quebec H9R 4P5. U.S.A.: Sandoz, Inc., 608 5th Avenue, New York, New York 10020.

†Allied Chemical and Dye Corporation, Industrial Division, 1348 Block Street, Baltimore. Maryland 21231.

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ham radio

simple CMOS keyer

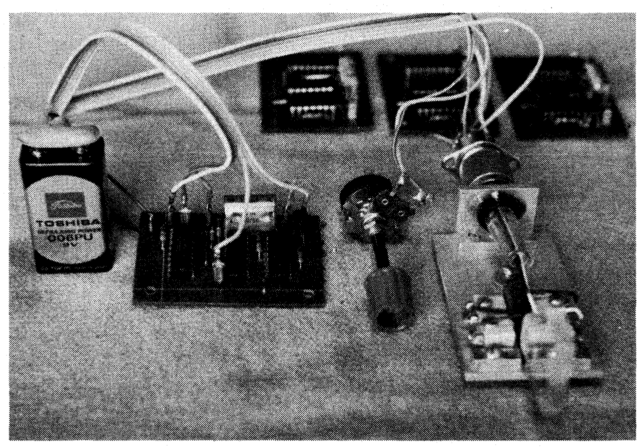
Build this
simple, low-cost
CMOS keyer
for inclusion in any
battery-powered equipment

The construction of this keyer is a result of an effort to reduce the overall weight of my "Mountain Day" contest transceiver. The circuit was developed from a proven RTL design. Although it does not offer "squeeze keying" or dot/dash storage, it fits the needs of the beginner as well as of the high-speed brass pounder.

The dash/dot ratio remains exactly 3:1 over the whole speed range. After each dot or dash, a pause of exactly one dot length is inserted. When both dot and dash contacts are closed, dashes are sent. With a 9-volt power supply, keying current is about 2 mA. When the keyer is not being operated, only about 10 nA is drawn from the supply; you may therefore connect the battery at all times and forget about the ON/OFF switch. The keying transistor switches positive voltages to ground. Changes in supply voltage have no appreciable effect on the speed. When the circuit is mounted and adequately shielded, it is not susceptible to rf pickup – even without rf chokes and bypass capacitors.

circuit description

The schematic diagram shown in fig. 1 is divided into two main parts, the time base and dot/dash generator. The time base, a stable RC oscillator is composed of gates U1A, U2A, and U1B, plus the associated components. Dot flip-flop U3A, dash flip-flop U3B, and the summing gate U2C form the dot/dash generator. In the quiescent state, these are the logic levels: logic 1 on pin 9 of U2D (due to the AND gate formed by R5, R6, CR1, and CR2) and both flip-flops reset, providing a 1 on pin 10 of U2D. U2A and U2B form the control flip-flop for the oscillator, which is blocked by the zero from pin 3 of U2B. After a short closure of either the dot or dash contact, U2B



Complete keyer including battery, paddle, and speed control.

enables the RC oscillator. (Time t_0 on the timing diagram, fig. 2).

The first half cycle of the oscillator places a 0 on pin 2 of U2B, thus keeping the oscillation even when the keyer lever is released. The rising edge of the first clock pulse clocks U3A to the SET state. If it was the dash contact that caused the start of the time base, a 1 from U1C would release the J input of U3B, allowing this flip-flop to be triggered by the rising edge of

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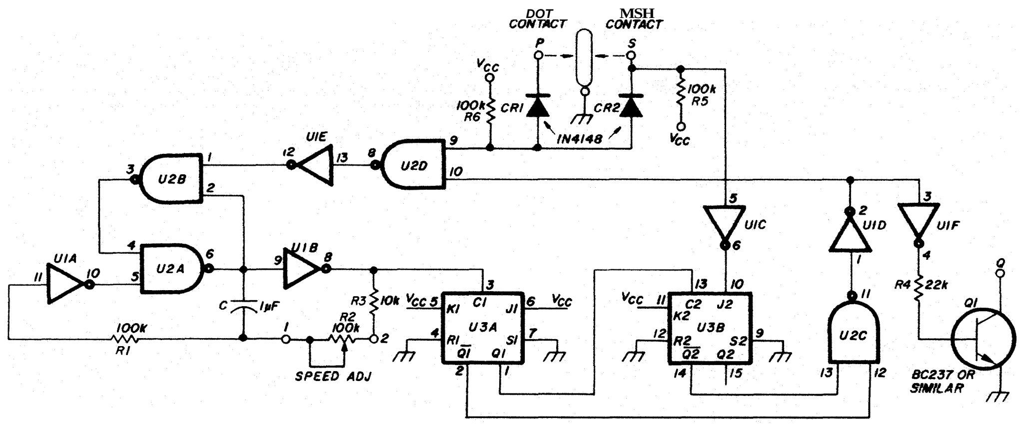


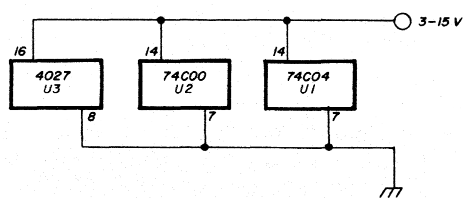


fig. 1. Schematic diagram of the simple CMOS keyer. The speed control, R2, should have a reverse, log taper. The keyer will work with any battery supply between 3 and 15 volts. External connections are denoted by the circled terminals.

U3A's output. In case of a dot contact closure, U3B remains reset because the zero on its J input prevents it from toggling.

The outputs of the flip-flops are summed by U2C, and via the inverters drive the keying transistor. The keying signal is fed back via U2D and U1E to the control flip-flop. As long as a dot or dash is being sent, this flip-flop maintains the oscillator in the operating state. When U3A toggles back to the RESET state, the dot or dash is terminated. At this time (for dots t_1; for dashes, t_2) the clock signal is a one, maintaining oscillation for another half clock period. At this time (t_3) the voltage at terminal 1 is **1.5** V_{batt}, which via U1A places a zero on U2A, thereby preventing the control flip-flop from reacting to premature trigger signals. After another half dot length, the voltage

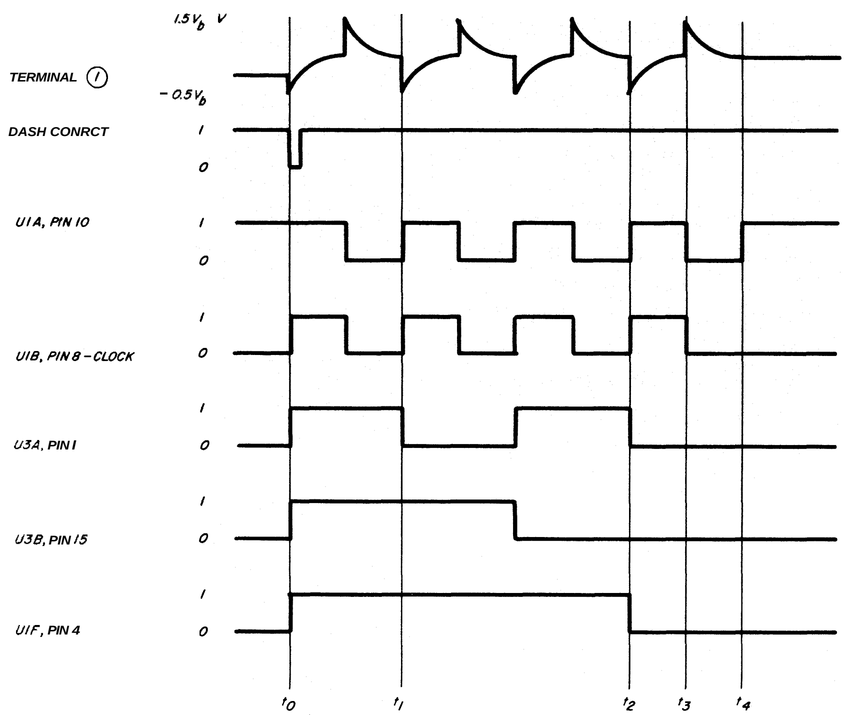


fig. 2. Timing diagram showing the levels within the keyer during the generation of a dash.

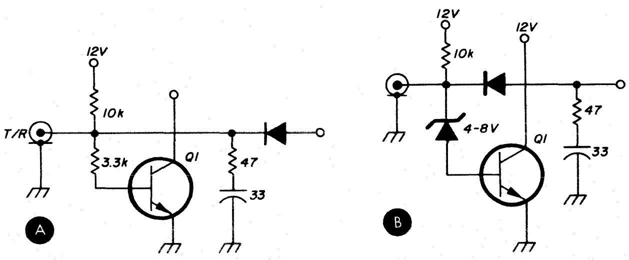


fig. 3. The diagram of the original keying circuit in the Ten-Tec Argonaut is shown in A. To handle the saturation voltage of semiconductor keyers, the circuit was changed to the configuration shown in B. The value of the zener diode can be between 4 and 8 volts.

at terminal 1 of C has discharged to almost 0 volts. This level is transferred by U1A, as a logic 1 to the control flip-flop, which, while maintaining state, can now be triggered again by signals from the keying contacts. After this pause of one dot length, (t_2-t_4), the circuit is again in the quiescent state and ready for another dot or dash.

transmitter connections

Due to its small size, the keyer circuit can easily be built into virtually any transmitter or transceiver. However a word must be said concerning the keying circuit involved. The voltage to be keyed must be positive with respect to ground. It must not exceed the voltage blocking capabilities of the keying transistor and the keyed current must be within the limits of this transistor. The keying circuit should support keying by semiconductors; with a voltage drop of up to 1 volt across the KEY terminals, the circuit must still operate properly. With a TenTec Argonaut, this was not the case, although a minor modification according to **fig.** 3 solved the problem.

Transmitters with a negative-keying voltage must be modified to have a positive keying voltage. Compatibility with straight keys or relay keyers is, of course, not impaired by such a modification. **Fig. 4**

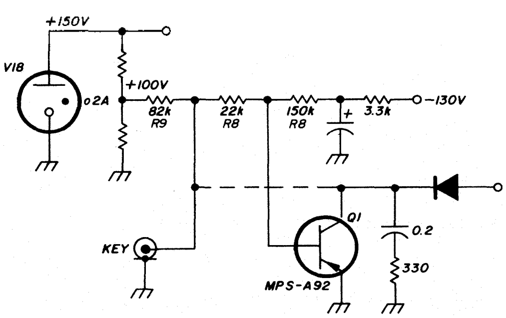
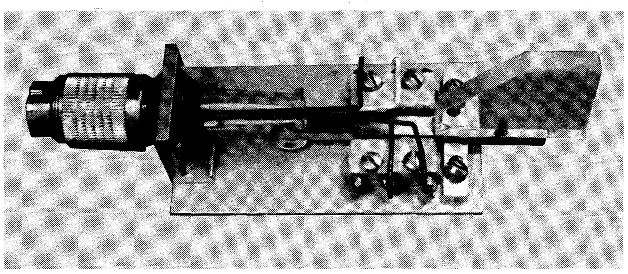


fig. 4. Diagram of the keying circuit of an HW101 that has been modified for this keyer. Other than R7, R8, R9, and Q1, all other components are from the original circuit.

shows the modification of a Heath HW101 as a representative of the tube transmitter family. Here are some general hints for this kind of modification: the voltage divider, R7, R8, and R9, must be set up to accept a current in the range of 0.5 to 10 mA. The internal resistance of the positive and negative sources must be taken into account when the values of the resistors are determined. With the key open, the voltage at the base of Q1 may not exceed V_{EB} maximum (4 to 8 volts, depending on Q1). With the key down, the voltage between R7 and R8 (base of Q1 not yet connected) should be substantially more negative than the 0.7 volts needed to completely drive Q1 into saturation.



Paddle mechanism used by the author with his Mountain-Day transceiver. Microswitches are used as the contacts.

Although the original purpose of this keyer circuit was incorporation into small transceivers, nothing prevents you from using it as an external electronic key. The use of a reed relay or an opto coupler in the output circuit would render the keyer more versatile (at the expense, however, of considerably higher power consumption).

construction

Circuit layout is not critical. An example of a printed circuit board layout is shown in **fig. 5***. Be careful to use a polyester (or equivalent) timing capacitor. The leakage current of tantalum and aluminum electrolytic capacitors is not compatible with the high-impedance CMOS logic. In the most commonly used speed range, R2 has a value of between 3 and 30 kohms. A potentiometer with a negative logarithmic characteristic would therefore be ideal. A standard 100-k logarithmic pot may be used instead, but the turning direction for an increase in speed would be counter-clockwise. If you insist on clockwise direction, a 100-k linear pot will do the job even if speed adjustment isn't best.

*An etched, drilled, and plated printed circuit board is available (air-mailed to the USA and Canada) from the author for 10 sFr (USA $5.00).

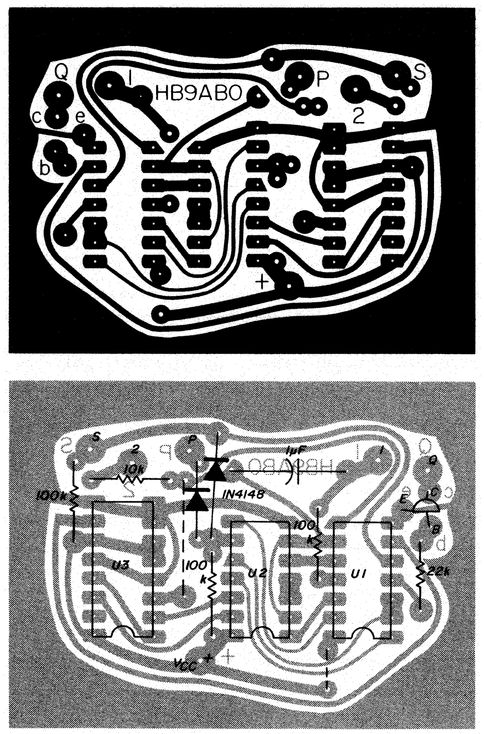


fig. 5. Printed circuit board layout of the simple CMOS keyer, with the parts placement diagram shown below. The two dashed lines indicate jumpers that must be installed.

the paddle

An example of a lightweight portable paddle is shown in the photograph. The unit is connected to the transceiver by a 3-pin DIN plug whose shell is screwed to the socket in the rig. The brass lever measures 60 x *3 x 5* mm (2-3/8 x *118 x 3/16* inches) and is centered by the buttons of two micro switches. It is fixed to the bottom plate at the pivoting point. The lateral movement is limited by two small aluminum blocks whose distance from the lever is adjusted to allow the lever to move just slightly beyond the switching points. Although there might be more elaborate solutions to this problem, the unit has worked nicely for over a year.

Further ideas for a portable keyer paddle can be found in **ref.** 3 and **4,** the latter being more promising according to the rule *the simpler the better.*

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ham radio

IGITAL TECHNIQUES

digital techniques basic rules and gates

Digital circuits are a useful and fascinating part of today's electronics. Devices and their applications have increased by such a proportion that an amateur who is not employed in the electronics industry may be confused by the jargon surrounding the technology. This series of articles will present the basics and, it's hoped, give you an insight into practical applications.

We are familiar with linear or "analog" circuitry, but what is a digital circuit? It is simply a decision-making device based on two voltage levels per input. The output also has two voltage levels. A two-valued input and output is called *binary.*

Digital circuitry (or, *digitallogic)* is made from simple building blocks which obey specific logical rules. Interconnection of many simple blocks is possible, whether on a circuit board or a single chip of silicon. Modern technology allows an almost unlimited combination on a single chip, spawning hundreds of different digital devices. Despite their complexity of function, all digital devices are made from the basic blocks.

Several digital families exist. Differences are internal and have an effect on interfacing. The two largest families will be described: TTL or Transistor-Transistor-Logic, the *bipolar* family branch, and CMOS or Complementary-MOS, the fet branch. Interconnection between families is possible within certain rules.

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logic level reference

Binary levels must be defined. A *low* level is near ground. A *high* level is close to the supply voltage. Some fet digital devices have more than one supply, so these refer the high level to the "V_{cc}" supply. The high level is assumed more positive, relative to ground or common.

Logic levels may be *positive* or *negative.* Positive logic is most common and retained throughout this series. Levels have different descriptions, so it might be well to memorize the following:

Low = 0 = near-ground = logic 0 = false

High = 1 = positive = logic 1 = true

Low, high, 0, and 1 are the most common terms. True and false may apply to devices with double outputs, one being an inverted level of the other.

A few data sheets refer to negative logic. This is generally taken as just a voltage reversal, although low and high are still the same.

basic building block

This is the gate, the fundamental decision maker. Each gate may have any number of inputs, but only one output. The six basic gates are shown in fig. **1** along with an inverter. The latter has only one input and is used mainly for level inversion.

Input states, for a specific output, will determine the type of gate. Note the small tables of I s and 0s for each gate. These are *truth tables* and tell the most about a particular function. Truth tables exist for circuits and all device types; some are time dependent.

The AND gate output will be a 1 when both A *and* B inputs are 1. All inputs of a multiple-input gate would have to be 1 for a 1 output. The OR gate output will be 1 when *any* input is a 1; output is 0 only when all inputs are 0. Exclusive-OR gates have only two inputs; a 1 on either input will produce a 1 output. But a 1 on both inputs will output a 0.

NOT, NAND and NOR

Compare the truth tables of the AND and NAND, OR and NOR, and the Exclusive gates of **fig. 1.** Each pair, of the three types, wilt have opposite output states. All six types are needed for design flexibility, but the NAND, NOR and Exclusive-NOR may be confusing.

Digital technology uses the term "not" when a desired signal is low; *i.e.,* it is not high. A line named SIGNAL would be considered active (desired) when high. Renaming it $\overline{\text{SIGNAL}}$ with the overbar means it is active when low. The name $\overline{\text{SIGNAL}}$ is pronounced "signal not:' or "signal bar," and either form is used.

A NAND gate output is active low. Its name comes from "not-AND." A NOR gate output is active low; its name is "not-OR." Similarly, an Exclusive-NOR is active low.

Symbol shape and little circles describe the type. Shape denotes general function while the circle or "inversion bubble" indicates an active low input or output. The bubble isn't always shown on spec sheets, so check for a device pin, name overbar, or the truth table.

The uses of NAND and NOR gates may not be apparent, so let's examine some simple gate arrays. The function of the array in **fig. 2** is to provide a high output when the inputs from either A and B or C and D are high. For ease of illustration, C and D are kept low. Intermediate AND output states E and F may not be used but are good for following an array function.

the ubiquitous NAND

The array in **fig.** 3 performs the same function as

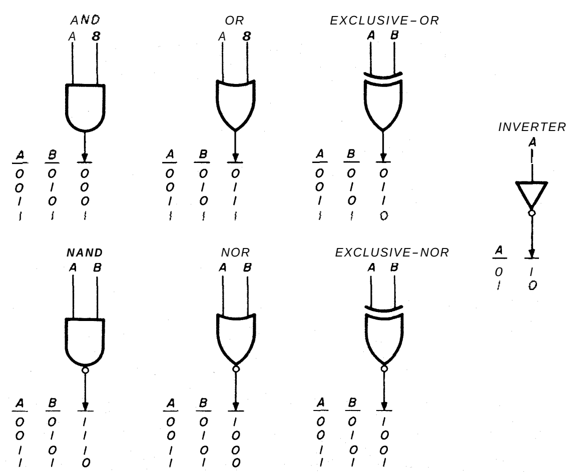


fig. 1. Schematic symbols and truth tables for six **basic gates.** AND, NAND, OR, NOR, Ex-OR, **and** Ex-NOR, **and the inverter.**

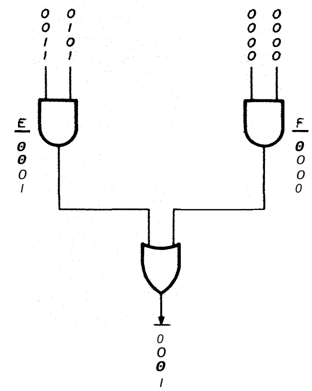


fig. 2. Conventional AND/OR **gates can be combined to produce the desired outputs. In this example, a 1 output will be present when A and B, or C and D are both ls.**

the one in **fig. 2,** even though all gates are NANDs. The equivalent OR function has bubble inputs, matching the active low NAND outputs. If this is confusing, go back to **fig. 1** and check the state conditions of NAND input versus output. Intermediate states E and F are useful here.

NAND gates can be used for any equivalent AND-OR array cascade. Most TTL gate arrays are built up entirely of NANDs and came about through early all-transistor logic circuits. Economy in earlier days dictated a minimum number of discrete devices and resulted in inverted outputs. Designers found that all-NAND gate array cascades worked just as well as older diode gates. The first integrated circuit gates used equivalent NAND structures.

NANDs are now so numerous that an unofficial "NAND RULE" is used to analyze and design gate arrays.

> THE NAND RULE: *Any* low input will cause a high output state; *All* inputs must be high to cause a low output state.

NAND gates used for an AND function will have active high inputs, just like an AND. The equivalent OR function requires active low inputs. Direct equivalents are shown in **fig. 4. Fig. 4B** is the same as an AND, while **fig. 4A** is the same as an OR. Inverters take care of necessary input and output state changes.

Fig. 5 is a simple array which produces a high output when either A and B are high or $\overline{C}$ input is low. Note that the overbar indicates $\overline{C}$ is active low. If conventional AND-OR gates were used, you would have C with an active high. This array shows an interesting input control condition.

Holding $\overline{C}$ low will prevent both A and B from affecting the output. Inputs A and B could then be in any state combination and the output truth table would indicate them as *don't* care states. Since they cannot affect the output, you don't care what states

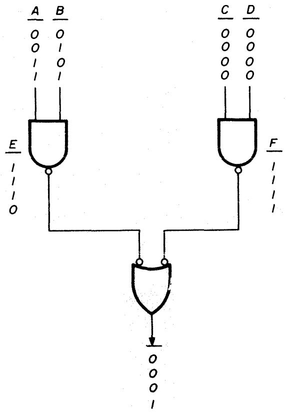


fig. 3. In this example, the AND/OR gates of fig. 2 have been replaced by NANDs and NORs. Even with the change, a 1 will be present at the output for the same input conditions.

they are in. An "X" on the truth table indicates the don't care condition.

When low, the $\overline{C}$ input can be considered an inhibit for A and B. Conversely, it could be an enable input when high. Many multifunction devices have inhibit and enable inputs. A word of caution: Inhibit and enable controls may be active high or low; check the device spec sheet for bubbles and overbars.

TTL and CMOS families

TTL is the most common family. It was pioneered by Texas Instruments, and wide industry acceptance prompted all major semiconductor manufacturers to "second source" (make the same product under license) most or all devices in the original family. Their popularity resulted in other IC makers' designing their own TTL devices; TI "second sources" many of these.

TTL is sometimes referred to as the 54/74 Series, after TI's original numbering scheme. TI now uses an

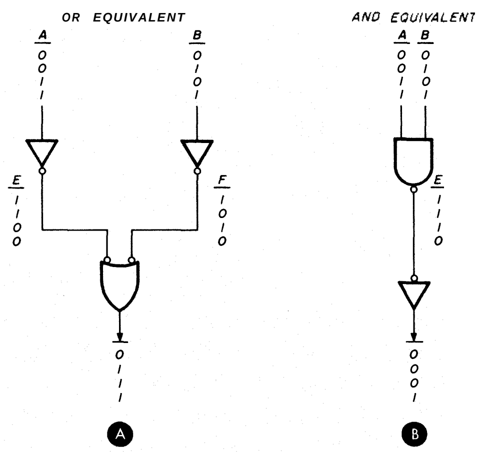


fig. 4. In A, since the NAND gate can be represented by an OR gate with inverters or the input, the complete OR function can be duplicated by using a NAND gate with an inverter in each input. B shows the AND equivalent by using a NAND gate with an inverter on the output.

SN prefix, while other makers have different prefixes. The "54" or "74" number identifies the device. A 7400 package is a quadruple two-input NAND gate, regardless of source. Many parts lists omit prefixes, since second source devices have identical characteristics.

A "54" part is miiitary temperature grade, – 55° to +125°C. A "74" part is commercial or industrial grade, with an operating range of 0° to +70°C. There is a slight difference in operating characteristics, but this would rarely affect amateur equipment.

CMOS is the most common fet family and is ideal for low-power applications. Where TTL has internal bipolar transistors, CMOS has N-type and P-type MOSFETs in complementary arrangements. The MOSFETs are insulated-gate types with extremely high input impedance.

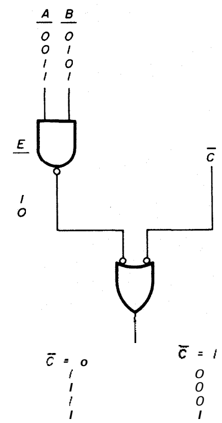


fig. 5. The C input, in this case, can be used as a circuit inhibit. If low, the output will always be high. regardless of the A or B inputs. When C is high, then the output will be enabled and will respond to the other inputs.

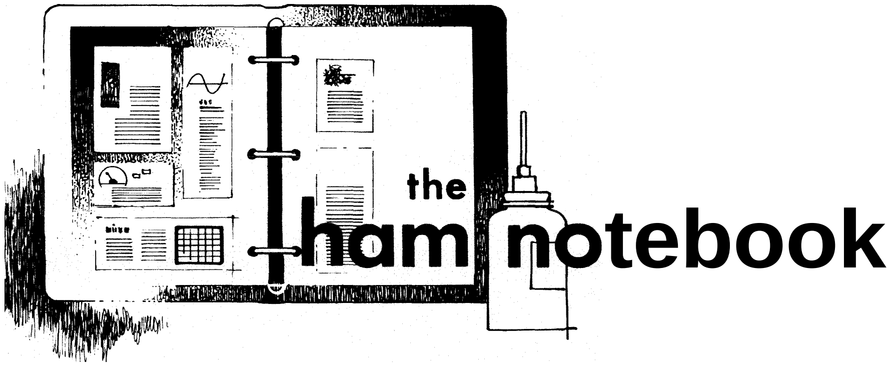
RCA developed CMOS and uses a 4000 Series numbering system with a CA prefix. CMOS is also second-sourced, but part numbers for equivalents vary; cross-reference tables are required for most second sources.

CMOS military temperature is the same as TTL, but CMOS industrial-grade temperature is –40° to +85°C. CMOS is also more lenient in power-supply voltage. TTL requires +5 volts ±5 per cent, while CMOS supply voltages can vary from 3 to 15 volts! One pays a price for such tolerances; the same device will be slower at lower voltages.

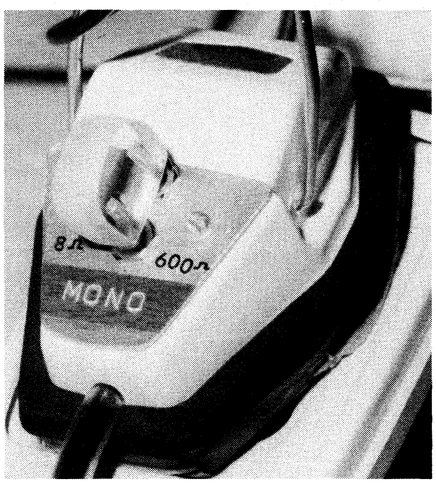
RCA also introduced "B" series CMOS (suffix to number) as an improved version of their original "A" series. The B series incorporates output buffers for driving lower loads and is characterized at 5-, 10-, and 15-volt supplies. All new designs are in the B series.

The next article in this series will take a *detailed* look inside the devices to point up differences between TTL and CMOS.

ham radio



dual-impedance headphones



The switching arrangement is shown in this view of the dual-impedance headphones. Photo by WD9CXG.

Dual impedance headphones offer versatility, convenience, and private listening pleasure. With different types of receivers available on the market today, it is not unrealistic to have a ham-bands-only and a general-coverage receiver in the shack.

In some instances, especially with older military equipment and newer ham gear, different impedances will exist at the headphone jack. This being the case, in order to have private listening, two headphones with different impedances were required (one for each rig). Thus, the need for a single pair of dual-impedance headphones.

Taking into account that the main source of audio was the amateur receiver (Kenwood TS-820), monaural 8-ohm headphones were purchased. The desire to listen to a military general coverage receiver meant switching headphones to provide the 600-ohm load specified in the manual.

It was not totally unrealistic to employ one pair of phones and change the impedance as desired. This was accomplished by placing a 600-to-8 ohm audio transformer in the line, switching it out for the 8-ohm load. If the components are carefully chosen (for size) they will all fit neatly into the housing of one of the phones, eliminating any external boxes. To prevent possible trouble, two double-pole, double-throw miniature toggle switches were used to completely isolate the transformer from the line.

A plastic bar was used to throw both switches at the same time. Shaped into a design of your own choosing and drilled for clearance of the bats on the switches, a drop of epoxy will hold the "throw bar" in place. Before final insertion of the bar onto the switch bats, make sure both switches are thrown in the same direction.

Jim DiSpirito, AB9Q

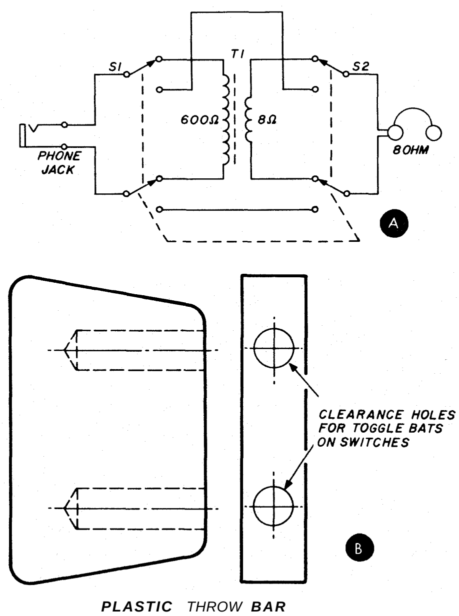


fig. 1. The schematic diagram in (A) shows the connections between the switches and audio transformer. T1 is a Calectro DI-724, having a 1200-ohm, center-tapped primary with an 8-ohm secondary. The switches can be ordinary miniature double-pole, double-throw switches. The small plastic throw bar is shown in (B). This bar can be shaped to fit any particular pair of headphones.

HW-2036 antenna socket

The antenna socket on the rear of the Heath HW-2036 2-meter transceiver is directly in line with a trace on the final amplifier printed circuit board. This line, which connects pin 10 on the relay and pin 2 on the plug P301, carries 13.8 volts. When using a phono plug with the long center pin, this pin will touch the board, shorting the 13.8 volts to ground. It's best to use either the RCA or Motorola style plug, since they have a shorter center pin. The phono plug/PL-259 adapter sold by Radio Shack also has the long center pin, requiring that part of it be cut off to prevent it from contacting the circuit board. However, if you wish to take the time to disassemble the transceiver, a small piece of electrical tape can be placed over the trace to prevent accidental contact.

Jim Conner, W3HCE

improved tuning on 160 meters with the T-4X transmitters

When using either a T-4XB or T-4XC transmitter below 1850 kHz, a true dip could never be obtained and loading was difficult, even when using a 50-ohm dummy load. Through discussions with other Drake owners, I found that this frustrating problem was shared by other T-4X-series transmitter owners. Being curious about this strange behaviour, we called Drake only to find that their low-end cutoff frequency is 1840 kHz. With this news, we decided to optimize the output network for the 1800-1850 kHz band, since that's all we have in New England. The modifications are almost trivial, requiring only two capacitors in the output network and a third for the driver tank circuit, but the results are excellent. The transmitter can be loaded and controlled on the low end of 160 just the same as on any other band.

As shown in **fig.** 3, the modification to the output network requires the addition of a capacitor on each side of the pi network. The part numbers in parentheses apply to the T-4XC; the others, to the T-4XB. Using the pictorials provided in the Drake manual, locate S4H and C65 (C86). Add a 120-pF, 2000-volt capacitor in parallel with C65 (C86). Next, locate C67 (C89), an 865-pF capacitor on S4G, and add a 680-pF capacitor in parallel.

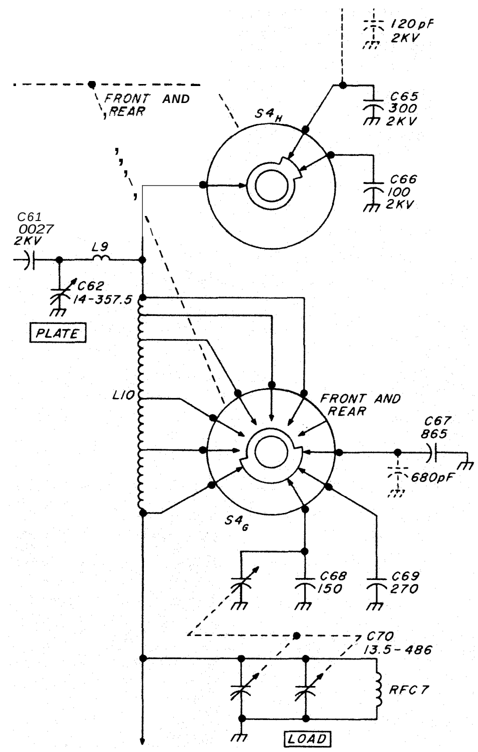


fig. 2. Schematic diagram of the changes made to the pi network to enable it to cover the low end of 160 meters. The numbers in parentheses refer to the components in the T4XC. The first designation is for the T4XB.

In addition to the output pi network, the driver tank circuit also required padding, since the driver control had to be rotated fully counterclockwise. This modification is depicted in **fig. 4.** A 36-pF capacitor was connected in parallel with C39 (C54), located on the rear of S4F.

With the implementation of these simple and inexpensive modifications, our Drake transmitters will load very nicely in the 1800-1840 kHz region, with the driver control showing a nice peak rather than being fully against the stop.

Steven E. Holzman, W1IBI
John D. Adamson, W1HZH

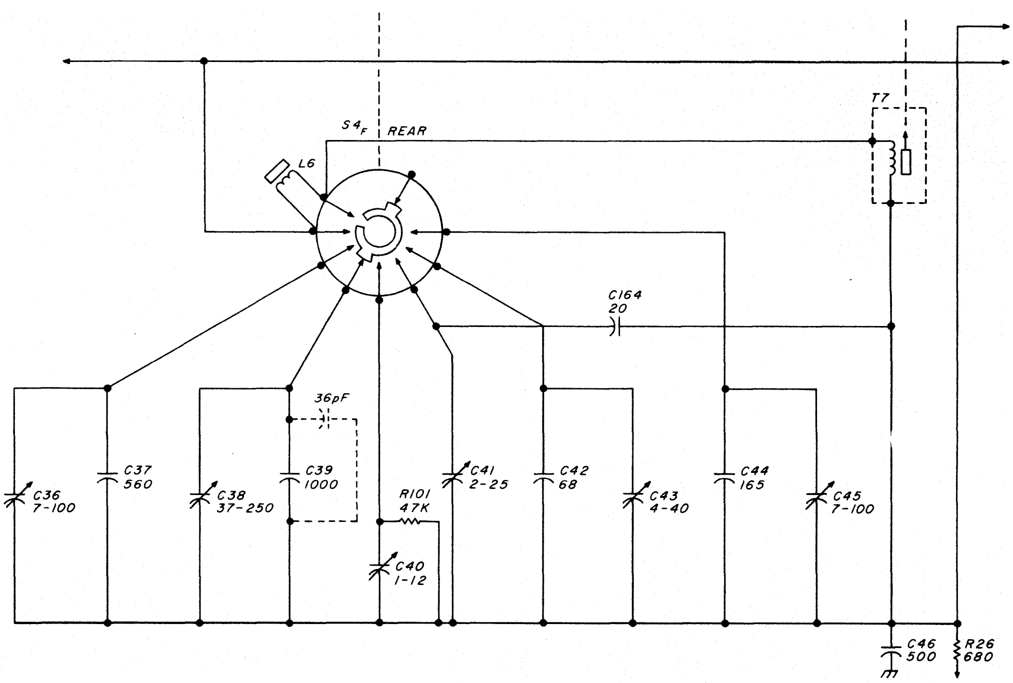


fig. 3. Changes made to the driver network for low end coverage of 160 meters.

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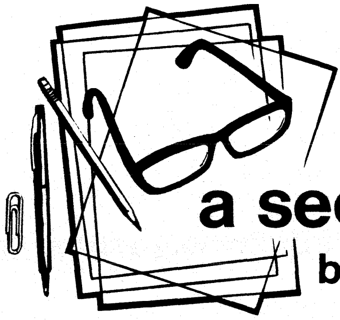


a second look

by Jim Fisk

This year Amateur Radio may be facing one of its biggest challenges in 20 years. What I'm referring to, of course, is the World Administrative Radio Conference which will convene this fall in Geneva, Switzerland. Better known in ham circles as WARC 79, this conference of ITU member nations will consider all the high-frequency allocations, including those of Amateur Radio, broadcasting, maritime mobile, and the other radio services which require operating frequencies.

There are some Amateurs who would like you to believe that the high-frequency Amateur bands will be completely decimated at this conference — ravaged by the greedy big-money interests who covet our bands for their own selfish purposes — but I think that the people who are all too eager to promote that turn of events are either alarmists or poorly informed; probably both. Obviously, it's impossible to forecast the outcome of WARC 79 at this point in time, but the Amateurs I've talked to who are officially involved with preparations for the conference are all cautiously optimistic that the high-frequency Amateur bands after WARC will be pretty much the same as they are now. And they are the ones who should know, not the purveyors of gloom and doom who apparently get their information from the Wizard of Oz — or some other equally unlikely source.

In the past, dozens of magazine articles have been written about the "terrible drubbing the Amateur Radio service has taken at every international frequency allocation conference." If you carefully review the record, however, you'll find that exactly the reverse is true; in every case American Amateurs have actually gained more high-frequency spectrum than they lost.

It is generally believed, for example, that at one time Amateurs had exclusive use of all wavelengths below 200 meters. That's a fable which has been quoted so often it's now accepted as fact. Actually, the 1912 regulation in question restricted all stations not involved in commercial traffic from going above 200 meters. That included *all* private, commercial, and experimental stations not transacting business or developing equipment for business purposes. Amateurs had no exclusive claim on "200 meters and down" — they shared that spectrum with virtually every other radio service. In fact, Amateur Radio stations at that time were required to specify their operating wavelengths, which were invariably 150, 175, or 200 meters — three spot frequencies below 200 meters.

In the early 1920s it became apparent that the 1912 law was hopelessly inadequate for the then existing conditions. More stations were on the air than ever before, the broadcast boom was well underway, and Amateurs had demonstrated the long-distance capabilities of the short waves. The scramble for short-wave territory was on, and every service was pushing for all the high-frequency spectrum it could get. To bring order to the ensuing chaos, a domestic radio conference was held in Washington in 1924; part of the outcome was the establishment of four harmonically-related Amateur bands: 160, 80, 40, and 20 meters. It's important to remember that this was not an international agreement, however, nor in fact did it have the authority of law — it was purely a mutual agreement between the various radio services in the United States.

The 1927 International Radio Conference in Washington saw precious kHz shaved off the American 160, 40, and 20 meter bands, but in return we received an exclusive new band at 10 meters. Amateurs in Europe fared less well, and some will argue that American Amateurs now had to share 40 and 80 meters with the foreign broadcasters, but that was true before the 1927 conference convened. Compared with the spot frequencies given to Amateurs in 1912, the new international allocations were a vast improvement.

There was no change in the Amateur bands at the Madrid conference in 1932, nor at Cairo in 1938. The next conference was scheduled for Rome in the spring of 1942, but because of the war, the next International Radio Conference was not held until 1947, in Atlantic City. Amateurs lost some space on 160, 20, and 10 meters at Atlantic City, but we received a nice bonus in return: a brand new band at 15 meters. Hence, there was not a net loss at all, but a gain! Those are the same bands we are still using today.

In reviewing the record of high-frequency Amateur allocations, we have progressed from what was essentially spot-frequency operation in 1912, to 3485 kHz of high-frequency operating space in 1927, to 3500 kHz today. Based upon past performance, and the proven service of Amateur Radio to the public, I believe we have every reason to be optimistic about the future.

Jim Fisk, W1HR
editor-in-chief

comments

zip-cord feedlines

Dear HR:

The article on "Zip-Cord Feedlines" published in the April 1978 issue prompts me to record, somewhat belatedly, my own experience on the subject. I have, for many years, happily used zip-cord feedlines particularly for temporary or semi-temporary antennas for use up to 21 MHz. I have always used the type with clear insulation, thinking that there was less chance of fillers absorbing the rf energy. I have to admit, though, that I have no concrete reason for this feeling, and I could be quite wrong. I have always used a simple antenna tuner and balun to couple the transmitter to the line, and, without exception, I have found the results to be gratifying. For example, I have quite recently used a 20-meter dipole slung between a tree and the chimney of my apartment building, about 20 feet high, fed by about 30 feet of Radio Shack loudspeaker wire (the heavy-duty type with stranded conductors) and driven by a Heath HW101. I have always been able to raise interesting stations, and have had a good number of enjoyable ragchews with European operators on single sideband.

The characteristic impedance of zip-cord is, of course, unknown. In my particular example I measured and, from standard formulae, estimated its impedance to be around 100 ohms. I think that this impedance presents as good a match to any "real" antenna at modest height with nearby structures as does any well-characterized coaxial cable. The unknown impedance presents a problem, however, when attempting to measure the line loss, for the line must (usually) be terminated. To get some idea of the loss on the 30-foot length, I tried the following experiments:

First, I terminated the line with a 110-ohm carbon resistor combination. I fed the line through my antenna tuner, and, in the coax feed between the transmitter and the tuner, I inserted my SWR bridge. I then loaded the transmitter on 14 MHz and adjusted the antenna tuner for minimum SWR on the coax; I was able to get it down to an indicated value of 1.1:1. Then I removed the terminating resistor and measured the SWR both with the far end of the line open and shorted. I tweaked the tuning for minimum SWR before taking the reading. The reverse power was so high in each case that it was not possible to get reliable readings. I estimated the SWR to be at least 10:1 (neglecting possible error in the bridge – the forward and reverse powers are about the same with an open or short circuit on the output of the device). This indicates that the total loss, one-way, in the tuner and feeder is on the order of 0.9 dB, and this is a worst-case figure. This is not comparable with RG-8/U coax, but for a simple system it is a figure that I can certainly live with.

Tony Garratt-Reed, ex-G3VBZ/W1
Malden, Massachusetts

manned free-aircraft

Dear HR:

The Presstop of the October, 1978 issue is in error; the Double Eagle II was not the first free-aircraft to cross the Atlantic – it was the first *manned* free-aircraft to do so. The first free-aircraft to make the crossing was a high-altitude research balloon flown from Trapani, Sicily, in July, 1975, and brought down near Lexington, Kentucky after about 84 hours. Another balloon was flown from Sicily to Massachusetts in July, 1976. These balloons flew at an altitude of approximately 125,000 feet and had about 20 million cubic feet of capacity; they were launched and flown by National Scientific Balloon Facility personnel from Palestine, Texas for scientists from Italy, England, and Germany.

Spencer Petri, WA5JCI
Palestine, Texas

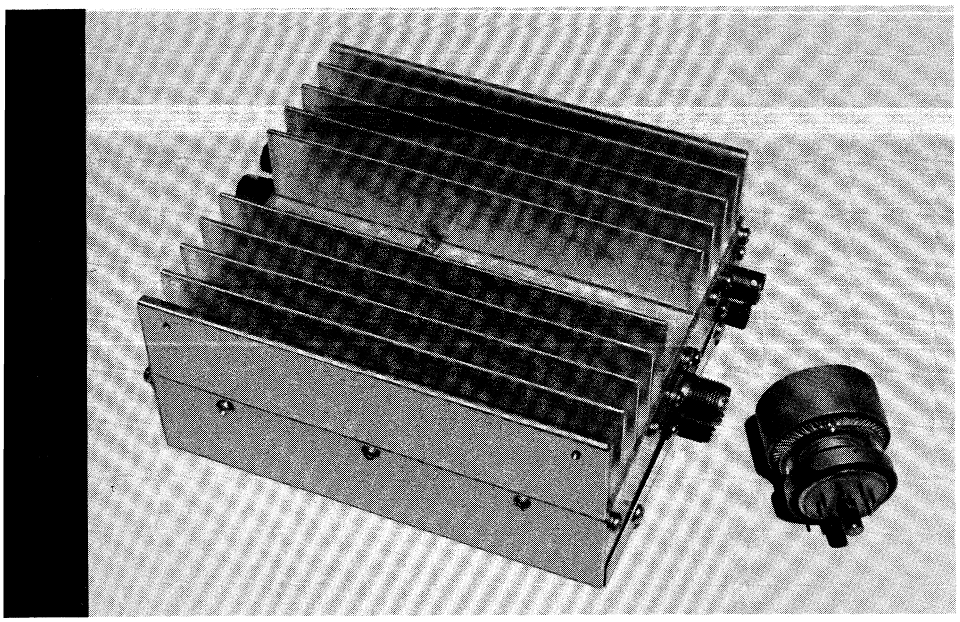
keyer memory

Dear HR:

I read with interest the Ham Notebook correspondence from K9WGN in the August 1978 issue of *ham radio* concerning my programmable keyer memory accessory. While it may be true that his unit programs the memory chips properly by simply pulling the RW pin to +5 volts, this is not good design practice. If you examine the data sheets for the 1101 (or any other static MOS RAM chip, for that matter) you will see that the RW pin should be pulsed only after the address is stable. The manufacturer's information for the 1101, in fact, recommends both the data and address inputs remain stable for at least 100 ns after the falling edge of the WRITE pulse. If the RW pin is held near +5 volts during programming, there is a chance some undesired bits may be altered while the input memory address is changing. For these reasons I feel the 74121 monostable, U8, is justified.

K9WGN is correct in stating that you can substitute the 7493 for the 74193s I used in that design. However, since the 7493 is a negative edge triggered ripple counter, the inverter, U11B, should be eliminated.

Andrew B. White, K9CW
Urbana, Illinois



10/80-watt amplifier for 2 meters

Design of an rf-activated, power-selectable, 2-meter amplifier intended for use with a 2-watt fm handie-talkie

It all began rather innocently. My good friend Jim Fisk, WB6YED (no relation to ham radio's editor), won a pair of Motorola HEP S3041 40-watt vhf power transistors at a prize drawing during the 1975 Dayton Hamvention. Several of us who were with him envied his good fortune, and we wondered what he would do with his new treasures. Months later, during a visit to Jim's home, I got my answer.

Jim had a 2-watt, hand-held fm transceiver he wanted to use in his car with a separate microphone. He envisioned a package that would produce either 10 or 80 watts output for 2 watts of drive, depending on his distance from the station he wanted to talk to. In addition, T/R switching had to be accomplished without a separate control line between the transceiver and the power amplifier. As Jim described what he was thinking of, I had the feeling he was going to ask me to tackle the job, which he did. Since he had been more than generous with his time and resources when I needed them, I was glad to help.

Putting together his requirements, I came up with the block diagram shown in **fig. 1.** The power amplifier had to have enough gain to produce roughly 80 watts output from 2-watts input; this amounts to 16 dB gain. A little more gain wouldn't hurt, because there are sure to be losses between the input to the entire circuit and the input to the power amp. The attenuator is switched into the circuit when low-power operation is desired. As it turned out, the attenuator wasn't that simple.

The requirement that transmit-receive switching

By Edward J. Paragi, WB9RMA, 14539 U.S. Highway 24 East, New Haven, Indiana 46774

be done without control lines necessitated the use of some type of rf-actuated T/R switch to put the power amp in the circuit during transmit operation. During receive the entire amplifier is simply bypassed with a coaxial line.

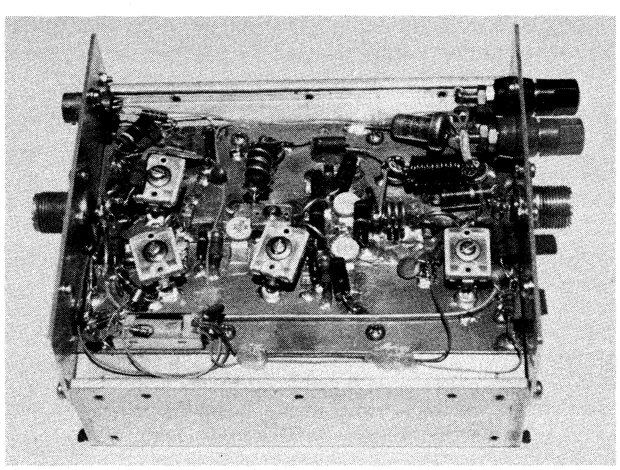
circuit description

I started the project by building the basic power amplifier. The HEP S3041 is equivalent to a 2N6084. A pair of 2N6084s at 150 MHz can be driven to 80 watts output with 25 watts input. This amounts to 5 dB gain; a driver capable of at least 11 dB gain is necessary to ensure a full 80-watt output from 2 watts of input. Since it suited my needs so well, I used the two-stage amplifier circuit and layout described in the Motorola Application Note AN-585, with as few differences as possible.[1] A 2N6083 was used for the driver in the amplifier described in the application note. Lacking a 2N6083 I used a 2N6136, a 25-watt uhf device useful to over 500 MHz and readily available to me.

The initial attempt at the amplifier was a copy of the layout described in the application note (see **fig. 2**). It was unstable for all but low drive levels. Any output greater than a few watts was accompanied by a healthy spur approximately 5 MHz from the desired frequency and several other spurs of lesser amplitude, including one at 5 MHz. I had used a uhf device for the driver, and so this was not entirely unexpected.

The chokelbead combinations in the transistor base circuits are used to suppress spurious oscillations as well as to provide dc return paths for the bases. Ferrite beads are low-Q inductors at vhf and, as such, make excellent broadband chokes. At lower frequencies, such as the high-frequency bands, the

Amplifier with the dust cover removed. The rf input is on the left side, the 10180-watt control connector is in the upper left, and the 12-volt connections are in the upper right. The TIR circuitry is in the lower left portion of the chassis.



resistance of the beads decreases while the inductance increases. The result is an increased Q. Uhf transistors exhibit tremendous gain in the high frequency region, and high Q circuits are an invitation for them to take off. The solution consists of adding 10-ohm resistors in shunt with the chokelbead combinations to guarantee a low Q at lower frequencies.

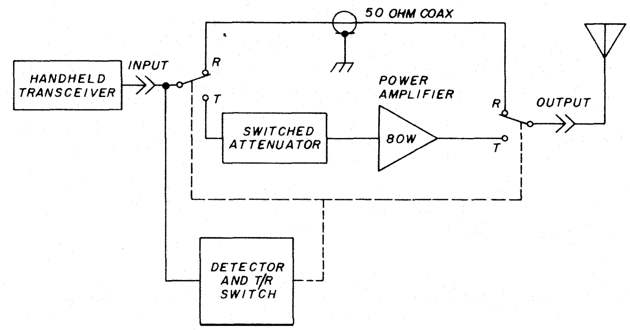


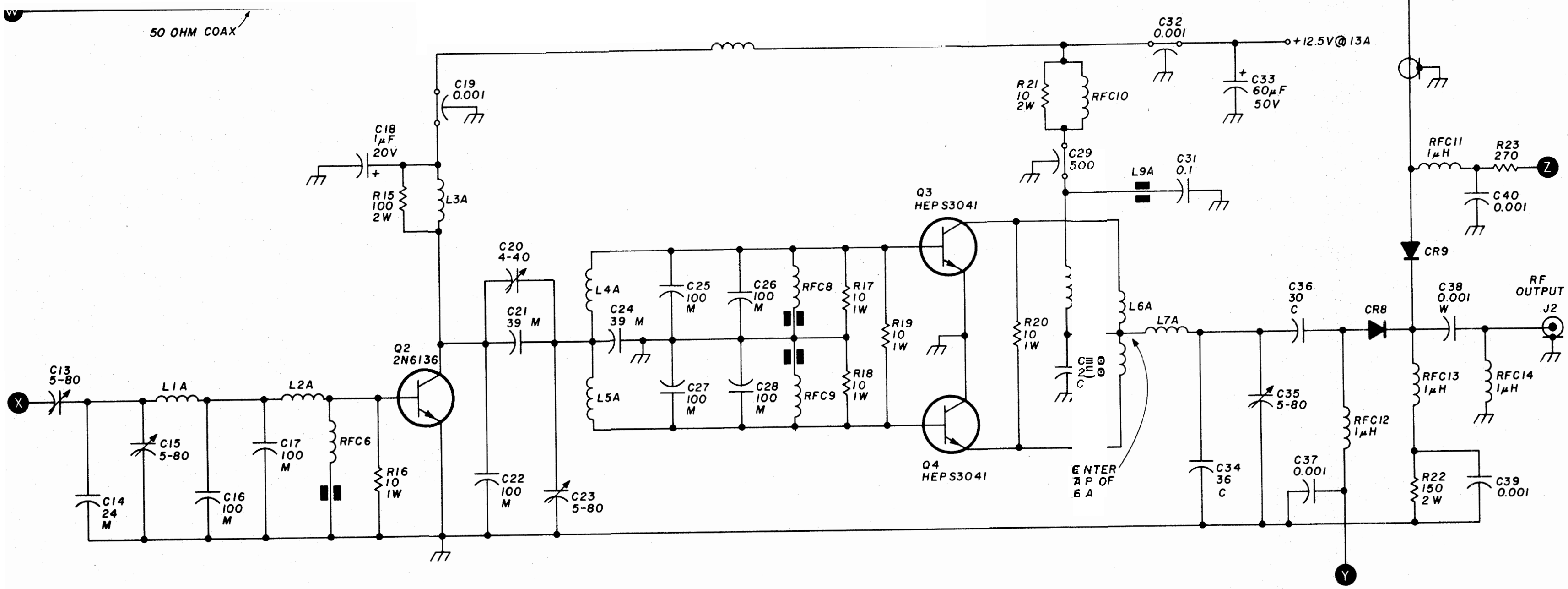
fig. 1. Block diagram of the 10180-watt amplifier for 2 meters. The switched attenuator is used to vary the drive level into the amplifier to switch between 10- and 80-watt power levels.

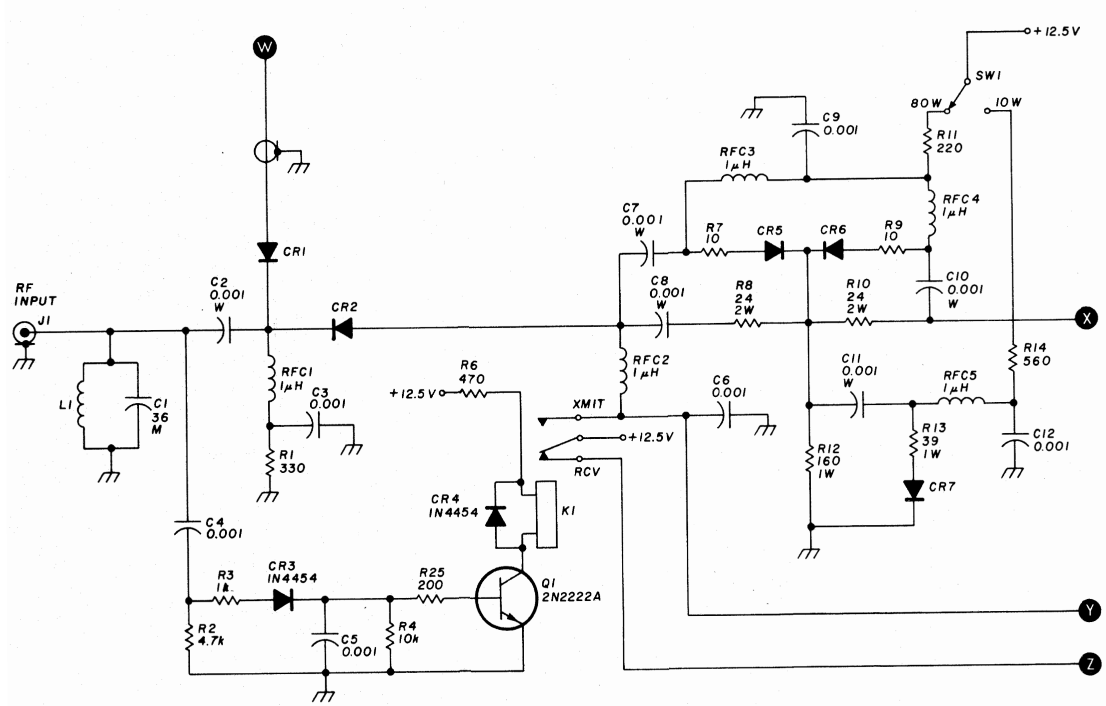
Another addition to the circuit in the Motorola Application Note is the parallel tank circuit (L1 and C1) at the input to the amplifier. It helps reduce any second- and third-harmonic energy coming from the hand-held transceiver.

After tuning up the completed amplifier, I had 80 watts output for 1 watt of drive. This was more gain than necessary, since 2 watts were available. Rather than detune the amplifier, I decided to put some attenuation in the circuit for high-level operation. Using an input of 2 watts, 2.6 dB of attenuation results in an output of approximately 80 watts, while 10 dB of attenuation provides an output of approximately 10 watts.

The switched attenuator shown in **fig. 1** is drawn schematically in **fig. 3(A)**. It is a tee configuration in which PIN diodes are used to add or remove resistors that are in parallel with the elements of the attenuator. PIN stands for "P-intrinsic-N" and describes regions of P and N semiconductor material that have a piece of undoped or "intrinsic" silicon sandwiched between them. This construction gives the PIN diode its unique properties. For low-frequency signals, the device behaves like a conventional PN-junction diode. At uhf, the diode is a current-controlled resistor. For high-forward current, the series resistance is low. For low-forward current, the series resistance is high. Reverse biasing the PIN diode further raises its series resistance. An ideal PIN diode should not rectify rf, but in the real world these diodes do rectify and this must be dealt with in rf-switching applications.

For 80 watts of output, the series PIN diodes, CR5





Addresses of Suppliers

PIN diodes Microwave Associates, Northwest Industrial Park, Burlington, Massachusetts 01803; G. R. Whitehouse and Company, 10 Newbury Drive, Amherst, New Hampshire 03031.

Ceramic Wafer Capacitors Vitramon Incorporated, Post Office Box 544, Bridgeport, Connecticut 06601; Erie Technological Products, 644 West 12th Street, Erie, Pennsylvania 16512.

Ceramic Chip Capacitors American Technical Ceramics, 1 Norden Lane, Huntington Station, New York 11746; Vitramon Incorporated — see above; JFD Electronic Components Corporation, 15th Avenue at 52nd Street, Brooklyn, New York 11219.

Unelco J-101 series Capacitors Underwood Electric and Manufacturing Company, Post Office Box 188, Maywood, Illinois 60153.

fig. 2. Complete schematic diagram of the 10/80-watt amplifier including the basic power amplifier (with modifications) described in *Motorola Application Note AN-585*. Inductors with the suffix A are described in the Application Note. L1 is four turns of no. 36 (0.13-mm) AWG enameled wire wound on a 100k-ohm, 1/4-watt carbon resistor. CR1, CR2, CR5, CR6, CR7, and CR9 are Microwave Associates MA47047 PIN diodes, while CR8 is an MA47080. All resistors are 1/2-watt, 5 per cent tolerance carbon composition resistors, unless otherwise specified. For the capacitors M = mica, W = ceramic wafer, and C = ceramic chip. RFC6, RFC8, and RFC9 are 0.15-μH molded rf chokes with a Ferroxcube 56 590 65/3B bead on the ground lead. RFC7 is a Ferroxcube VK200 19/4B wideband ferrite choke, and RFC10 is ten turns of no. 14 (1.6-mm) AWG enameled wire wound on R21. K1 is a 6-Vdc, 500-ohm coil dry reed-type relay.

and CR6, are forward biased, and shunt PIN diode, CR7, is left off. For 10 watts of output the series diodes are left off while the shunt diode is forward biased. The calculated equivalents of the two conditions are shown in **figs. 3(B)** and **3(C)**. The resistor values for the attenuatoi were chosen so the impedance looking in either direction would be close to 50 ohms. It is important that a good grade of carbon-composition resistor be used for the elements in the attenuator that carry rf current, and also that leads are kept as short as possible. Because of stray elements, this type of attenuator becomes more difficult to construct as frequency is increased. Two meters is probably getting near the practical limit.

The last block in **fig. 1** is the rf detector and T/R switch. The detector, CR3, is simply a rectifier with a low enough reverse recovery time to be effective at 150 MHz. The rectified rf signal is used to forward bias the relay driver, Q1, which in turn energizes K1 during transmit operation. The relay closes when approximately 0.7 watt of power is fed to the input jack of the amplifier. The relay contacts are used to switch the bias to the PIN diodes in the actual T/R switch, These PINs are CR1 and CR2 at the amplifier input, and CR8 and CR9 at the output.

For simplicity, a relay was used instead of solid-state bias switching. W9KHC pointed out in his article that the dc reverse bias on a PIN diode should be equal to the peak rf voltage across the PIN in question.2 For CR9, which is reverse biased during transmit operation, this would amount to 88 volts, assuming 80 watts is being delivered to a 50-ohm load. A higher load impedance would require a higher bias voltage. If a PIN diode rectifies when the reverse bias is not great enough, it will conduct and be destroyed due to excessive power dissipation. However, if the PIN diode bias line is just left open, any rectified current will tend to reverse bias the PIN and protect it. Open is the key word here. A set of open relay contacts is much better than a reverse biased switching transistor, which has some measurable leakage. Since the relay was employed, it eliminated the need for a high-voltage supply which would require an inverter for 12-volt mobile operation. With the arrangement shown, T/R switching is accomplished using the same dc supply that powers the amplifier.

component selection

Since the construction of this power amplifier involves relatively high power levels and frequencies, a discussion about the components used is in order. The PIN diodes used in the T/R switch and attenuator are general-purpose diodes for uhf work. They should have a low series resistance when forward biased, since they contribute directly to insertion loss. The MA-47047 typically has 1.5 ohms of resistance at 30 mA of forward bias, while the MA-47080 has a resistance of 0.45 ohms at 100 mA. In addition, the minority carrier lifetime should be roughly 10 times longer than the period of the lowest frequency in use. This prevents rectification if leakage occurs. The low-power PIN diodes used in the construction

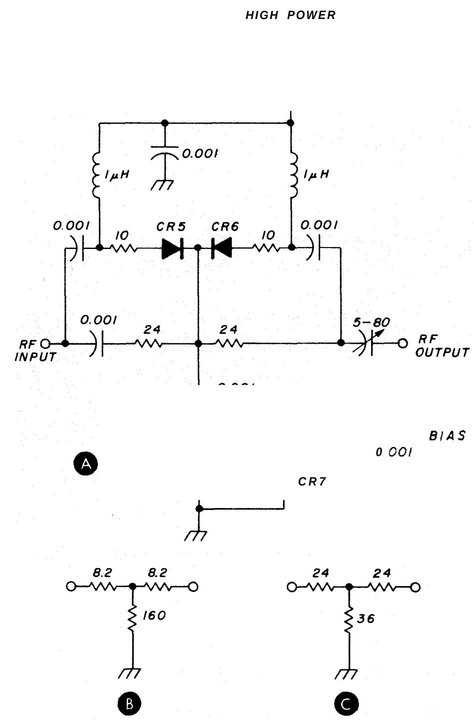
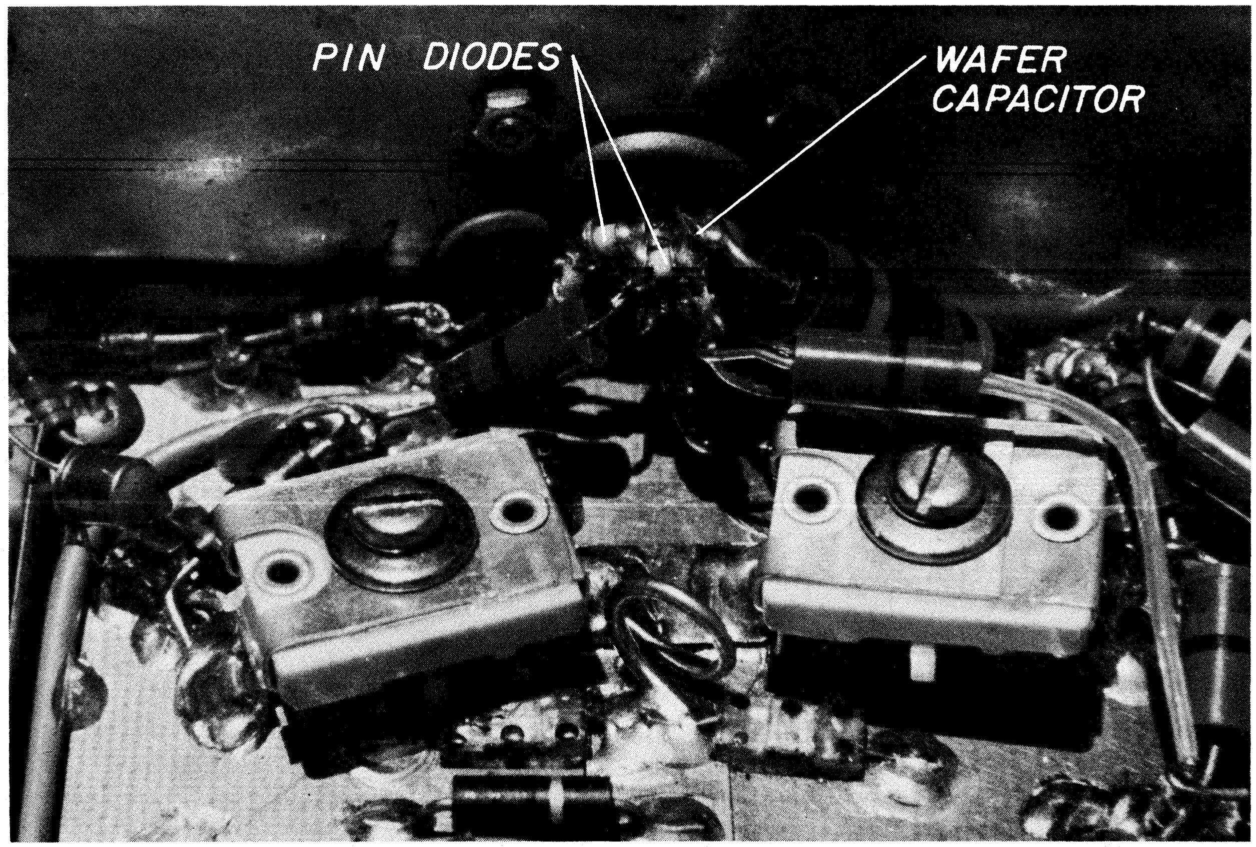


fig. 3. Schematic diagram of the switchable tee attenuator is shown at (A), while (B) and (C) respectively, show the effective attenuatorsfor high power (2.6 dB) and low power (10 dB).

of this project were in a pellet-style case. The MA-47047 comes in a lead-style case similar to that of a signal diode. Either type will work – just keep lead lengths short. The MA-47110 could be substituted for the MA-47047.

PIN diode CR8 must have a high current rating because it carries the output power of the amplifier. It will get fairly warm and should be heat sinked directly to the output connector via C38. Transferring heat through a capacitor may seem unconventional, but it is accomplished with a wafer-style capacitor. Wafer capacitors are a truly leadless capacitor with a ceramic dielectric. They are especially well suited for bypass and coupling applications where tolerance is not critical. The 1000-pF units used in this project are little squares about 4.1 mm (0.16 inch) on a side and 0.76 mm (0.03 inch) thick. The two large surfaces are metallized with a silver compound so they can be sol-



Detail of the circuitry near the rf input connector showing PIN diodes and ceramic wafer capacitors attached to the connector.

dered. It is a good idea to handle the capacitor with fine tweezers so that oil from your hands doesn't make soldering difficult. For best results, the wafers should be soldered using a silver-bearing solder and a very small iron tip. Ordinary tin-lead solder is likely to leach away the silver and make soldering impossible. Alpha Metals is one company that produces a 62 per cent tin, 36 per cent lead, and 2 per cent silver solder.

Another uncommon capacitor used in the amplifier is the ceramic chip. Chip capacitors are made from a material similar to the wafer capacitor dielectric. They are small cubes approximately 2.5 mm (0.10 inch) on a side with a pair of opposite sides metallized to accept solder. These parts are available in tolerances as fine as ± 1 per cent. In addition, they are also leadless and have very low loss. Due to their high Q (low loss), they are capable of carrying several amps of rf current where other types of capacitors would quickly go up in smoke. Since all of this comes at a price, chips were used only at the higher power levels in the amplifier.

A less expensive part is the capacitor made by Unelco which Motorola used in the application note. This is a silvered-mica type and is suitable for use at high power levels. It is physically quite large compared with the ceramic chip, but that is not a drawback for most applications. To reduce the effect of lead inductance of higher value capacitors (which have low reactance), it is often possible to parallel two capacitors, as in the case of C25 and C26.

The coaxial cable used to bypass the amplifier on receive is a solid-jacket type used because it was available. Any good 50-ohm cable can be substituted.

protection circuitry

Many high-power, solid-state rf amplifiers have protection circuitry to reduce rf drive, or completely shut down the amplifier, when a high vswr is sensed. Since this amplifier was designed for a specific installation, the extra circuitry was not included. A load vswr of 2:1 or less should be adequate for the design.

Another circuit commonly associated with high-power amplifers is a thermal protection circuit. These circuits sense high temperature, usually near the output stage devices, and act to reduce drive or shut the amplifier off. Since a substantial heatsink is used and two-way operation is always intermittent, this type of circuit is not justified. If the amplifier is to be operated for more than short periods of time and in a place where air circulation is restricted, some thought should be given to a thermal protection scheme.

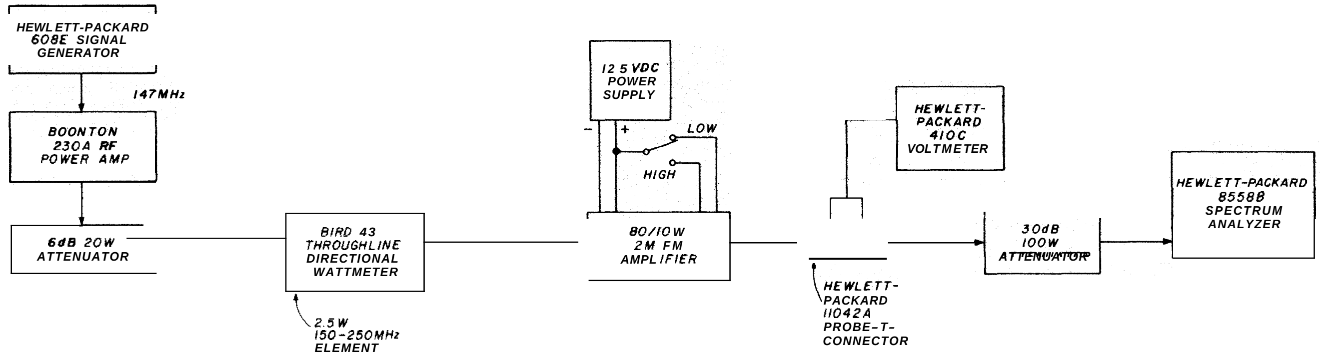


fig. 4. Amplifier test set-up.

Additional comments on circuit protection are given in **ref.** 3.

construction

No attempt was made to miniaturize the amplifier. With the cover in place, the assembly measures 7.6 x 14.0 x 15.9 cm (3 x 5.5 x 6.3 inches). The amplifier was built in "bread-board" fashion on a 10.2 x 15.2 cm (4 x 6 inch) piece of 1.6 mm (0.062 inch) thick epoxy-fiberglass, copper-clad board. Connections, which are isolated from ground, are made on miniature standoff insulators. This construction lends itself well to building power amplifiers, as it nearly eliminates rf ground problems. The heatsink may be a little larger than necessary, at 3.8 x 14.0 x 15.2 cm (1.5 x 5.5 x 6.0 inches), but since size was not important, the extra area is cheap protection. Long pieces of wire and ferrite beads are held in place with RTV-type adhesive.

Parts placement follows the general layout of the amplifier pictured in Motorola Application Note AN-585.

The amplifier is easily installed in the trunk of a car and controlled from the driver's seat. A pair of wires and a switch are used to turn the power on. Use at least no. 14 (1.6-mm) AWG wire for the dc connection to reduce the voltage drop from battery to amplifier. The output power level is controlled by three wires and a single-pole, double throw switch. One of the three wires can be the +12 volt line that supplies the amplifier. Adding a coaxial cable to carry rf from the hand-held transceiver to the amplifier completes the installation.

Placing the cover on the amplifier slightly detunes it. This is compensated for by slightly mistuning the trimmer capacitors by trial and error until the performance with the cover on is optimized.

table 1. Performance data for the 80110-watt amplifier at 147 Mhz with the cover installed.

receive insertion loss — 1.1 dB
drive level to accomplish T/R switching -- 0.7watt

	low power	high power
drive power	2 watts	2 watts
output power	7.2 watts into 50 ohms	74 watts into 50 ohms
current drain	4.0 amps	12.0 amps
2nd harmonic	40 dB	50 dB
3rd harmonic	44 dB	47 dB
input swr	1.6:1	1.4:1

purity of emissions

Since the completion of this project, the FCC has issued new requirements on transmitter spurious radiation. For transmitters and amplifiers operating between 30 and 235 MHz and having more than 25 watts of output power, all spurious emissions, including harmonics, must be at least 60 dB below the mean carrier level. The second harmonic of the carrier is the most important spurious output to be dealt with in this amplifier. Since at 80 watts output the second harmonic is already 50 dB below the fundamental, a simple pi or tee network after the amplifier will provide enough attenuation to meet the 60 dB requirement. A summary of the performance data from the amplifier test circuit (see **fig. 4**) is given in **table 1.**

acknowledgments

I would like to thank Bob Yankowiak for his patient technical assistance during the construction of the amplifier and the writing of this article. I would also like to thank George Johnson, K9ODF, for converting my color negatives to black and white prints for use in this article, and my wife Karen for typing the manuscript.

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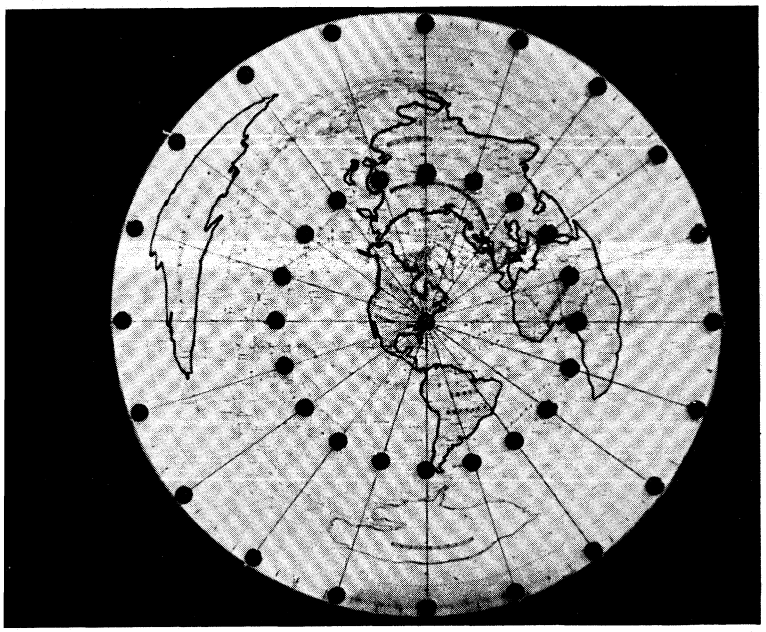
1. "VHF Power Amplifiers Using Parallel Output Transistors." from *Motorola Application Note AN-585*, Motorola Semiconductor Products, Phoenix, Arizona, 1973.
2. James K. Boomer, W9KHC, "PIN Diode Transmit/Receive Switch for 80-10 Meters," *ham radio*, May 1976, page 10.
3. "Design Techniques for an 80-Watt, 175-MHz Transmitter for 12.5-Volt Operation," from *Motorola Application Note AN-577*, Motorola Semiconductor Products, Phoenix, Arizona, 1972.

ham radio



solid-state antenna position display

Another approach for converting an antenna rotator to digitized readout, using discrete LEDs to show bearing segments

The need for an improved indication of antenna heading is most apparent during contesting and DX chasing. Many operators have considerable difficulty relating the compass heading of the beam, as indicated on the typical antenna-rotator control box, to the location of a particular country on the globe.

Normally, a table, slide rule, or a chart is used to find the correlation between the prefix of a call and the beam heading. People have a general sense of where a country or area is, but they are less than adept at translating this sense into an angular heading.

I once saw a global display scheme using a balanced pointer driven by a pair of selsyns, one coupled to the mast on the tower and the other driving the pointer on a wall-mounted map. My goal was then established, to construct an electronic display using the existing analog voltage at the rotor control box.

A completely solid-state design using available low-cost components was a must, and the display had to be suitable for mounting at my operating desk. I considered several options for the construction and presentation of the display. Personal taste and the builder's skill are involved in making the choice; a simple and effective version is described which is suitable for anv home craftsman with moderate skills.

principles of operation

The Ham II provides a 13-volt signal swing which drives a 1-mA meter, calibrated in degrees, to indicate antenna position. This voltage provides an ideal analog signal source for digitizing with a high-impedance CMOS A/D converter.

The Motorola MC14433 DVM chip was judged as most suitable for this display. It will accept a 0 to ±1.999 volt analog input voltage swing, and provides a binary-coded-decimal (BCD)/TTL compatible output. In addition, it has a self-contained clock and provides timing pulses to drive the TTL control logic and decoder drivers that make up the remainder of the display circuitry. Low-cost BCD-to-decimal drivers were chosen because of the display format.

The display background itself is a polar great-circle map of the globe divided into twenty 18-degree sectors. With a 0 to 1.99 volt input signal swing and a maximum 360-degree rotation, each sector corresponds to an increment of 100 mV. Each hemisphere corresponds to an increment of 1 volt of analog input signal. For example, as seen in **fig. 1**, sector 1 is displayed when the input potential to the MC14433 has any value from 0.0 to 0.1 volt. This corresponds on the map to the sector between 180 and 162 degrees in the S-SE quadrant of the Ham II meter scale. Likewise, sector II would be displayed when the input voltage is between 1.0 and 1.1 volts,

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which corresponds on the map to the sector between 360 and 342 degrees in the N-NW quadrant of the scale. My rationale for choosing twenty 18-degree sectors was

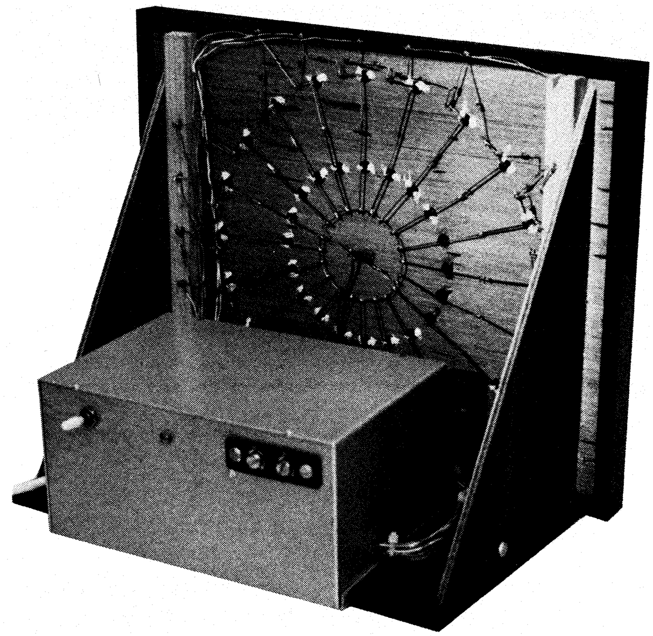
1. the system fits the decimal system;

2. 18 degrees is roughly the beam width of many rotary antennas;

3. 18-degree sectors provide enough room for a beam to coast to a stop once the motor drive is stopped.

Fig. 1, the block diagram, and **fig. 2,** the schematic, show the signal and logic flow from the input of the A/D converter to the sector indicator. The analog signal between 0 and 2 volts is applied to pin 3, the analog input, from the voltage divider, R5, which accepts the 13-volt signal from the rotator. C3 and R7 are used to provide RFI immunity.

The encoded TTL data from U1 is available in a multiplexed form at the Q_0 through Q_3 outputs. The A/D converter used in this indicator normally drives a four-digit multiplexed display, with the outputs Q_0 through Q_3 acting as the data lines, while DS1 through DS4 are the corresponding digit-select lines. For example, when the right-most digit (LSB) is to be displayed, the DS4 line is high and the data appears in a BCD format on the Q lines. As the display is



Rear view of the display showing the LEDs, diode-AND gates, and electronics enclosure.

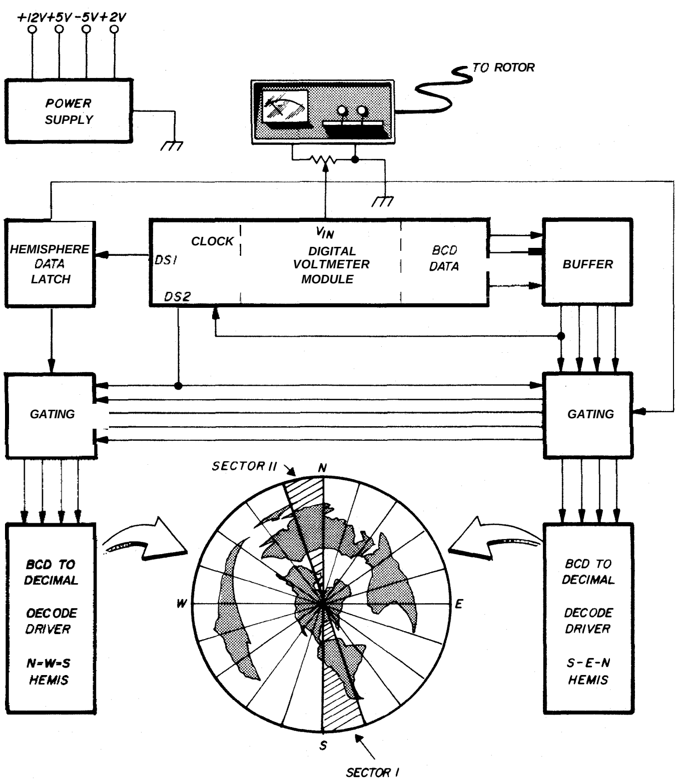


fig. 1. Functional block diagram of the solid-state antenna position display. The entire system is based on a 3-1/2-digit DVM IC manufactured by Motorola.

scanned, the appropriate digit-select line goes high with the correct BCD code for that digit appearing on the data lines.

The analog input to this IC is in the range of 0 to 1.999 volts. Therefore, the left-most digit will change between only 0 and 1. Or, it can be thought of as breaking the input voltage range into two segments, 0 to 0.999 and 1.000 to 1.999 volts.

In my display, the left-most digit, or MSB, provides the hemisphere data. That is, pin *6* of U2C (or Q_3) is low for an input voltage to U1 with any value between 0.0 and +0.99 volt. This corresponds to an antenna heading anywhere in the S-E-N hemisphere. Conversely, pin *6* of U2C is high when the input voltage to U1 is any value between 1.0 volt and 1.99 volts. This corresponds to an antenna heading anywhere in the N-W-S hemisphere.

The hemisphere data from U6 is stored in latch U8B by clocking the data in during DS-1 time, and holding it through the complete scan cycle of the A/D converter. The latch outputs, pins 5 and 6 of U8B, gate the appropriate hemisphere decoder/driver, U6 or U7, through NAND gates U3, U4, U5B, and U5C. U8B is reset after the end of each A/D converter scan cycle by an "EOC" pulse from pin 14 of U1.

To derive the appropriate 18-degree sector within the hemisphere, the next most significant digit of the A/D converter is used. As previously mentioned, each sector has an incremental voltage width of approximately 100 mV, thus allowing ten sectors in a 1-volt increment. The next most significant digit represents the correct 100-mV segment with the

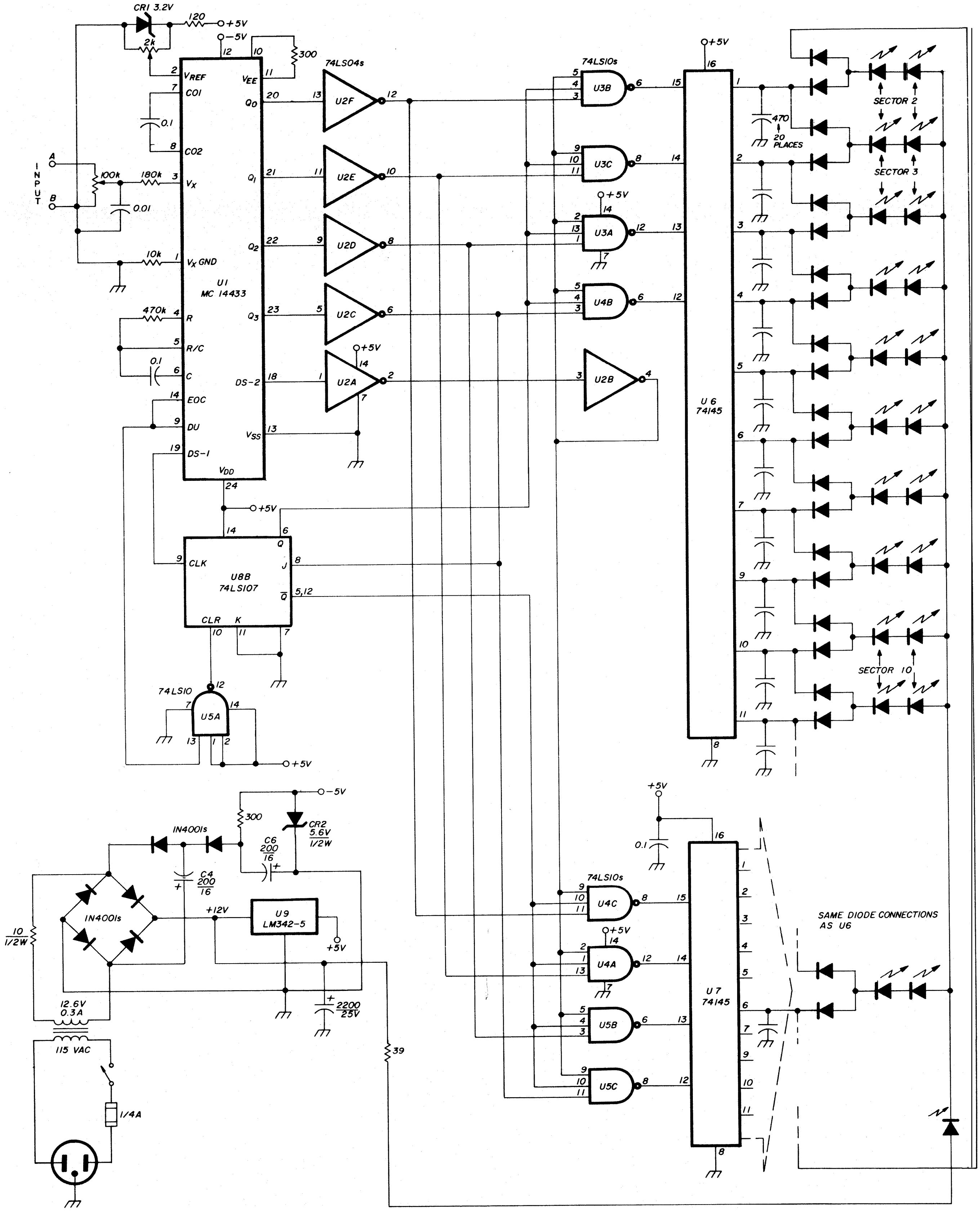


fig. 2. Schematic diagram of the solid-state antenna position indicator. Each output from the decoder is bypassed with a small disk ceramic capacitor to help eliminate noise problems (see text). C1 and C2 are mylar capacitors.

BCD value appearing during DS2 "time" at the output terminals.

During DS2 pulse, pin 18 of U1 is high. U2A and U2B act as a noninverting buffer which drives gates U3, U4, U5B, and U5C. These gates in turn drive U6 or U7 to display the appropriate sector during DS2 time. Valid input data to U6 or U7 is in a low-level state. This is provided when all three inputs to the NAND gate driver are high. The outputs of U1 require the buffering because of the limited source/sink current capability of the CMOS circuitry.

With this scheme, any one of twenty sectors can be displayed. The DVM chip can provide A-to-D conversion on the hundredths and thousandths decimal value of input voltage and are presented to the multiplexed outputs during DS3 and DS4 time respectively. However, this capability is not used in the display application.

The frequency of the internal clock in the DVM module is determined by R1 and C1. With the values shown in **fig.** 2, the clock runs at approximately 66 kHz, giving a conversion time of 250 ms. In simple terms, the conversion time is the interval required for the DVM circuitry to measure the analog input,

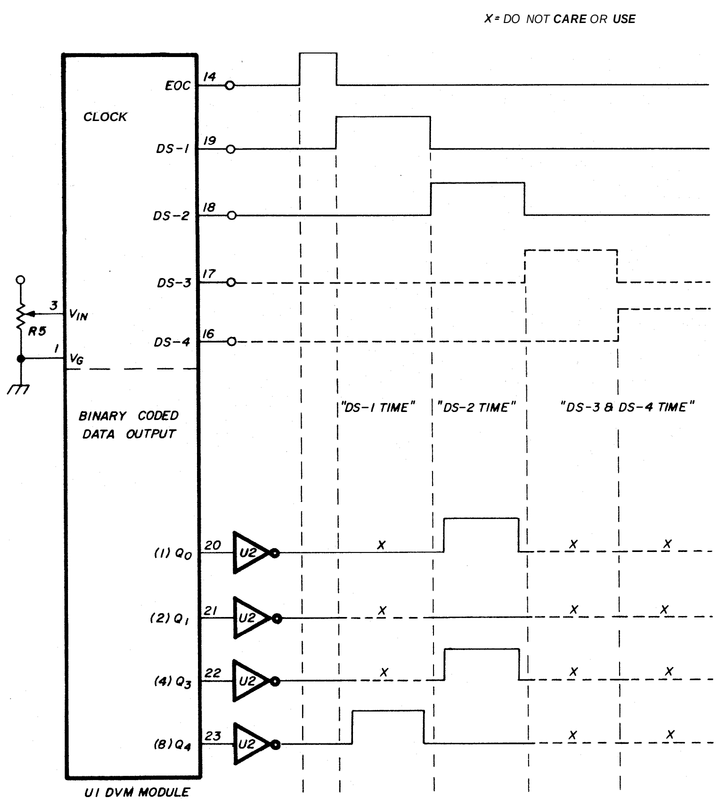
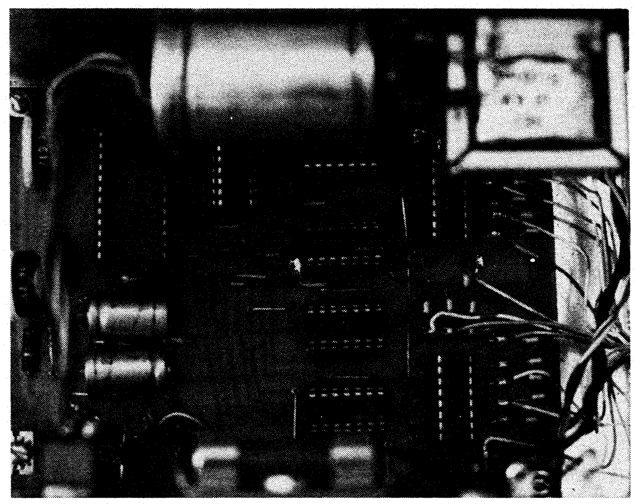


fig. 3. Timing diagram of the digital voltmeter IC. The MSB data output appears on Q_3 during DS1 time. In this example, Q_3 is a one for any input voltage between 0.000 and 0.999 volt, indicating that the N-E-S hemisphere is selected. The next most significant digit appears during DS2 time. As indicated, DS2 shows data for a value of 0.50 to 0.59 volt, indicating that the antenna is within sector 5.



Printed circuit board within the electronics enclosure.

compare references, compensate, integrate, encode data, and develop the output signals for the "3-1/2" digits. In this display application only "1-1/2" digits are used. An individual sector is displayed with five LEDs. The LED in series with R8 is positioned at the center of the display, Kansas City, and is continuously illuminated. The remaining four LEDs are selected from the display electronics and illuminate the perimeter of the sector.

LEDs DA1/DB1, DA2/DB2, etc., are positioned on radial lines drawn on the polar map, which is the background of the display panel. These radial vectors start at the center of the map and are displaced by 18 degrees.

The LEDs designated DA are mounted at the midpoint of the radial, while the LEDs designated DB are positioned at the ends of the radials. Diodes designated DC are low-cost silicon switching or low-PIV rectifier diodes (1N4001s) grouped in pairs to form a negative-OR type gating circuit.

To illustrate, refer to **fig. 4,** where sector A is to be displayed. Either U6 or U7 has been gated on during DS2 time, depending on which hemisphere has been selected, to display one of the possible twenty sectors. Only one of the twenty output pins from U6 and U7 will be in a low-voltage or current-sink condition, as represented by the closed switch at output terminal B in the driver. Current flows through R8, the LED in the center of the map, the two parallel branches formed by DA1, DB1, DC2, and DA2, DB2, DC3, and the low-impedance path (closed switch) in the display driver. With all other driver outputs in a high-impedance state, no current will flow in any of the other radial branch circuits.

Since the LEDs are illuminated only during DS2 time (approximately 60 ms out of each 250-ms conversion period), the LED supply voltage had to be raised to a level of approximately 12 volts. The value

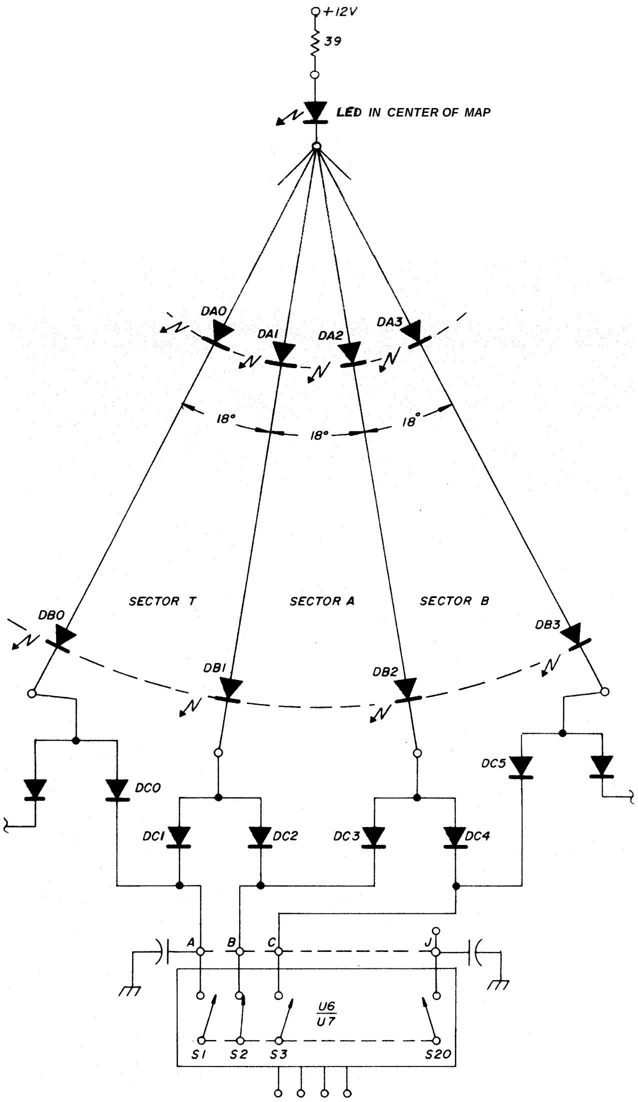


fig. 4. Diagram illustrating the current flow when sector 1 is selected. In this case, the appropriate output of the decimal driver is low, to sink current. The OR gates then determine which LEDS are conducting.

for R8 was chosen to give an average LED current of 25 to 30 mA, providing adequate illumination. Diode matching, to achieve uniform brilliance, has not been a problem. However, I suggest you order more display LEDs than needed and select the best ones.

As seen in **fig. 2,** bypass capacitors are shunted across each driver output. Any disk ceramic from 470 pF to 0.02 μF will work. These capacitors reduce the slope of the driver output signal during the switching transient. The cable between the display electronics and display panel acts as an antenna, and hiss was detected in my SB303 on 10 meters before the output lines were bypassed.

Except for the 12-volt transformer, C5, and R9, all components are mounted on the printed-circuit board. The transformer should be able to supply 300 mA. U1 requires a negative 5-volt supply, which is provided by C4, C6, R10, CR2, and the two-additional 1N4001s. The LM 342-5 provides a regulated, 5-volt supply for logic circuitry. The DVM module requires an external reference voltage of 2.0 volts which is provided by CR1, R3, and R4.

display panel construction

Many options are open in the construction and layout of the display panel, depending upon the creativity, skill, taste, and resources of the builder. A rather straightforward approach yielded the display shown in **fig. 5.** The printed-circuit board is mounted inside a 15 x 7.5 x 10 cm (6 x 3 x 4 inch) box-type interlocking chassis. The box should be mounted on the rear of the base to provide stability for the display panel and frame.

adjustment

Before inserting any of the ICs (especially U1), the +12, +5, −5, and the 3.2-volt reference should be checked. Then adjust R3 and R5 to the extreme counterclockwise position. Use a high-impedance voltmeter, 1 megohm or greater input (probe) impedance, when checking any terminal voltages on U1. Lower impedance voltmeters present a significant load to the circuitry resulting in erroneous readings. With U1 in the circuit, adjust R3 to provide +2.00 volts at pin 2 of U1. Temporarily jump a wire from a +5 volt supply point to the A side of R5, the non-grounded rotator input terminal. Only sector 1 should be illuminated. Turn R5 clockwise about half way. The display should step through the 20 sectors as this is done, and only one sector should be illuminated at any one time. If the LEDs light out of sequence, check the cable wiring between the driver outputs and diode gates in the display panel. If a

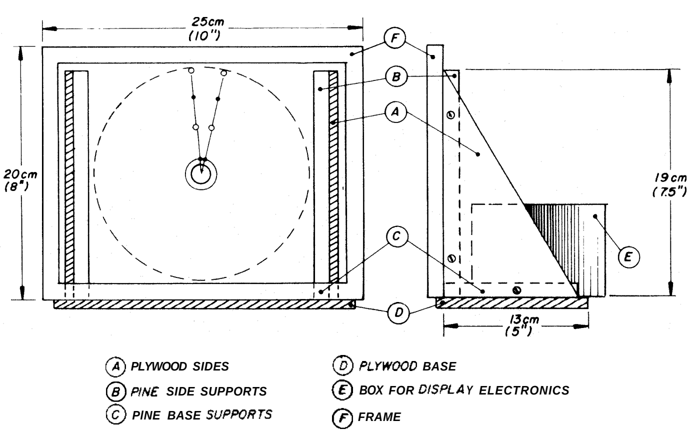


fig. 5. General construction plan for the display panel. The outline dimensions are approximate, depending on the frame used. Exact details of the map and framework have been omitted since they will depend upon the builder's resources and abilities.

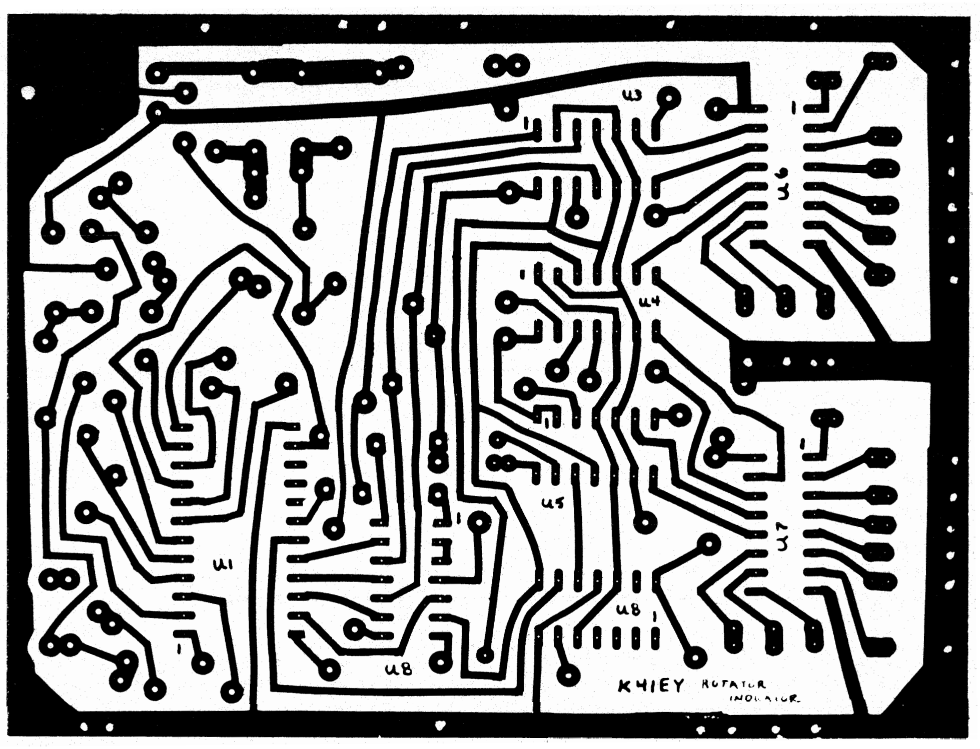
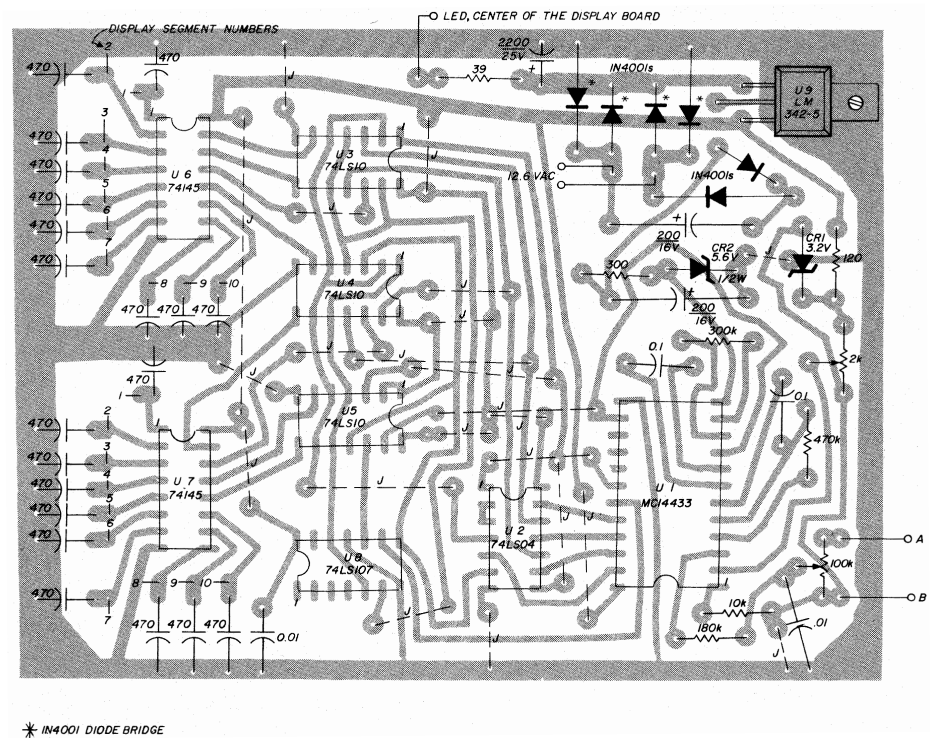


fig. 6. Printed circuit board pattern for the antenna position display (above) and parts placement diagram (below). A printed circuit board can be obtained from the author.



series diode string on a radial vector does not light, check the diode gates, LEDs, and U6 or U7. If one hemisphere does not light, check U8, the appropriate input gates to the drivers U6 or U7, as well as the appropriate display driver.

With the above check-out completed, return R3 to the extreme counterclockwise position and disconnect the +5 volt jumper to R3. Attach a cable from the grounded terminal of R5 on the display electronics board to terminal 1 on the Ham II control box terminal strip, and another lead from the other side of R5 to terminal 3. Rotate the antenna to 198 degrees in the S-SW quadrant. Approximately 13 volts should appear across R5 on the display input terminals. Carefully rotate R5 clockwise until just the 198 to 180 degree sector illuminates. Rotate the antenna back through a clockwise rotation and watch how the sectors on the map light up as the appropriate compass headings on the control box are passed. With minor adjustments of R3 and R5, tracking to within 10 degrees can be maintained throughout the entire antenna rotation. Linearity of the potentiometer in the rotator can cause minor variations.

I have not attempted to connect this display to makes of antenna rotors other than the Ham II. However, with the circuit explanation given, it should be easy to adapt this system to other rotators as long as there is more than a 0 to 2 volt analog signal swing available to drive the display electronics.

component procurement

All the ICs, with the exception of U1, the Motorola MC14433, are readily available from most supply houses. U1 was purchased through Circuit Specialists, Box 3047, Scottsdale, Arizona 85257, for under $15.00. With the influx of low-cost DVM kits, there should be lower costs for this item in the future. The total cost of the electronic components and printed circuit board came to less than $45.00. LEDs can be obtained at good prices when ordered in a quantity of 100 rather than on a per-diode basis.

I wish to express my appreciation to Dick Keil, N4JU, Bob Winter, WB4AYW, and Walt Short, N4SW, for their advice, and particularly to N4SW and K4GOK for their assistance in artwork preparation.

ham radio

great-circle maps

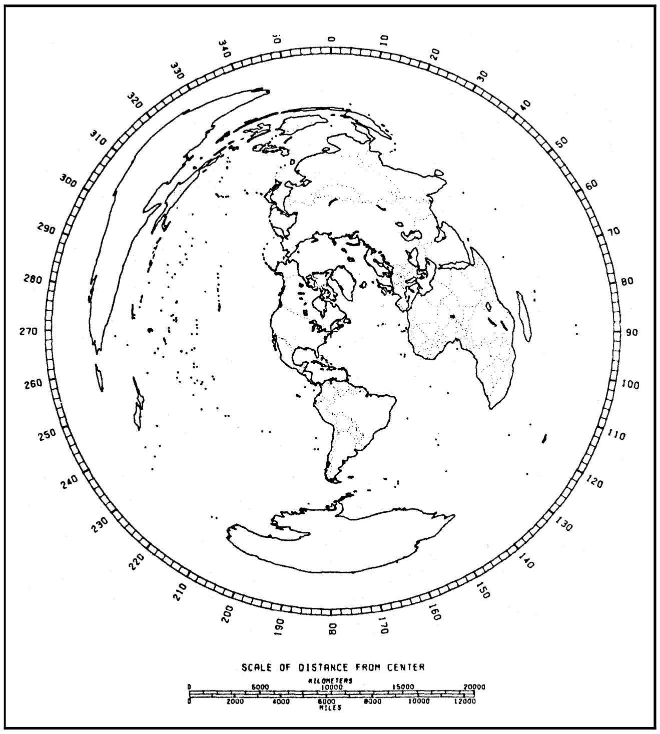
For the past several years I have been offering computer generated great-circle bearing printouts which have been extremely popular with DX operators.* In a recent article in *Ham Radio Horizons* I discussed the use of these charts and also showed several great-circle maps. The great number of readers who inquired about obtaining azimuthal equidistant maps prompted me to complete work on a computer program I started several years ago to draw such maps.

The program itself is straightforward, but the data base associated with it is truly staggering, consisting of almost 20,000 data elements. This is why I put off completing it for so long. The computer time required to process and draw each map is much more than that required for the standard great-circle printout, so the cost is slightly greater. The maps are printed on 11 x 14 inch paper; in addition to geographical data, all major political boundaries are shown, but no attempt has been made to label individual countries because of the enormous programming complexities it would entail, not to mention the additional cost.

I will supply custom-made azimuthal equidistant maps to interested *Ham Radio* readers according to the following price schedule:

- $5.00 postpaid via 3rd class mail, worldwide
- $5.75 postpaid via 1st class mail, USA, Canada, Mexico
- **$6.50** postpaid via Air Mail, worldwide.

*Computer generated charts for your station location are priced at $1.00 for surface mail or $2.00 for air mail, and list 660 distant locations along with bearings, distances, and return bearings.



Computer-drawn great-circle map centered on Greenville, New Hampshire.

When ordering your map, be sure to include your mailing address and the location for which the chart is to be made. If you live in a rural area or a town of less than 10,000 population, carefully describe your location with respect to other nearby towns so your latitude and longitude can be determined.

Bill Johnston, N5KR
1808 Pomona Drive
Las Cruces, New Mexico 88001

phase coherent RTTY modulator

Discussion of the need for using a phase-coherent AFSK system to generate FSK with a single-sideband transmitter

With amateur activity increasing on RTTY there is a growing need for a quality RTTY modulator and demodulator to interface between the station receiver and transmitter. Although some transmitters have provisions for a frequency shift key (FSK) input to the VFO circuitry, most do not. One of the standard means of producing an FSK carrier on the high-frequency bands is to insert pure tones into the audio input of an ssb transmitter. An audio FSK generator (AFSK) will produce an FSK signal when it is applied to a single-sideband, suppressed-carrier transmitter. However, there are limitations to this technique.

FSK problems and approaches

In RTTY circuits, there are basically two frequency shifts used, a narrow shift (170 Hz) and a wide shift (850 Hz!. The only FCC requirement is that the shift be less than 900 Hz. The reason for a shift at all, of course, is to distinguish a mark from a space, thus conveying information. Using digital language, the mark and space may be redefined as a logic one or a logic zero. The definition of which frequency will be used as a mark (one) and a space (zero) must be compatible between all communicators, otherwise the shift will be inverted from the one expected. On the 20-meter band, for example, the mark frequency is normally defined as the higher of the two frequencies, while the space is the lower of the two. The mark and space frequencies are offset by 170 Hz.

The narrow-shift mode is almost always used on the high-frequency bands, while some amateurs use the wide shift on the vhf bands. **Fig. 1A** shows the ideal frequency spectrum of a narrow-shift FSK signal. The bandwidth due to the information rate is not shown, but it will be centered on each carrier frequency and will have the effect of widening the FSK signal spectrum. The information bandwidth depends on the speed at which the RTTY is sent, and it will not be considered here. Note in this figure that the mark frequency is exactly 14.097875 MHz, while the space frequency is exactly 14.097705 MHz. This would be a narrow-shift FSK signal, since the difference is 170 Hz.

If the frequency determining unit in the transmitter is appropriately modified, a frequency shift of 170 Hz is easily attained. As an example, an oscillator may be modulated by using a varicap to generate the required FSK signal. An alternate method, however, would be to modulate a single-sideband transmitter with pure sine wave tones. Theoretically, in an ssb, suppressed-carrier transmitter, (SSB-SC), if a single frequency is used to modulate the transmitter, a single frequency appears at the output of the transmitter. When in the upper-sideband mode, the frequency that appears is the sum of the modulating frequency and the suppressed-carrier frequency, or the suppressed-carrier frequency minus the modulating frequency when in the lower-sideband mode. By changing the audio frequency, a signal can be generated which is FSK modulated in step with the audio signal. Although the signal fed into the transmitter is AFSK, the signal generated by the transmitter is a true FSK signal; only one rf frequency exists at a time at the output.

There are certain demands put on the transmitter

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when either FSK or AFSK is used. If FSK is implemented by modifying the VFO, then the short-term and long-term stability of the shift circuits must be considered. Since the frequency-determining section of a transmitter is being modified, the circuits must not cause drift in the oscillator that will make the selected frequency unstable. For instance, if a dc potential is used to vary the capacitance of a varicap diode, the dc may have an ac ripple component. This ripple, from a poorly filtered dc power supply, will cause fm modulation to appear on the shifted-carrier frequency. This is certainly not desirable, Also, any temperature drifts in the dc control circuits or in the varicap diode response will cause a frequency change to occur in the fsk carrier frequencies. Care must be exercised to ensure that any modifications do not influence the stability of the oscillator.

With AFSK modulation applied to a transmitter, another set of problems appear. Fortunately, these problems are not associated with the stability of the VFO, since it is not modified. However, the quality of the ssb generation (carrier suppression, sideband suppression, and sideband filter response) is important. In a SSB-SC system, a double-sideband signal is first generated with the carrier suppressed. Normally, a sharp filter is used to pass only the desired sideband, either the upper or the lower. Because of this method, the unwanted sideband will be present, along with the carrier, although suppressed to a large degree (see **fig. 1B**). When operating an SSB-SC transmitter as an FSK generator, care must be exercised to ensure that the carrier is properly balanced out, and, in addition, the unwanted side-

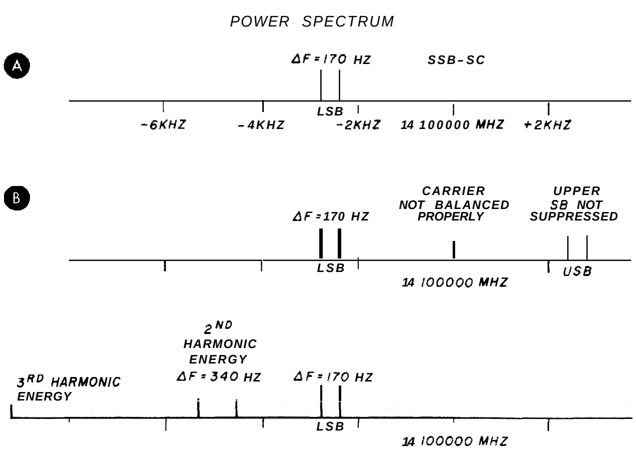


fig. 1. Frequency spectrum of an FSK system on 20 meters. (A) shows the ideal spectrum, with the mark being the higher of the two frequencies. In all cases, the information bandwidth is not considered. In (B), you can see the consequences of improper carrier balance and poor upper-sideband suppression when using an audio tone to generate an FSK signal. The spectrum when the sideband filter does not properly suppress harmonics is seen in (C).

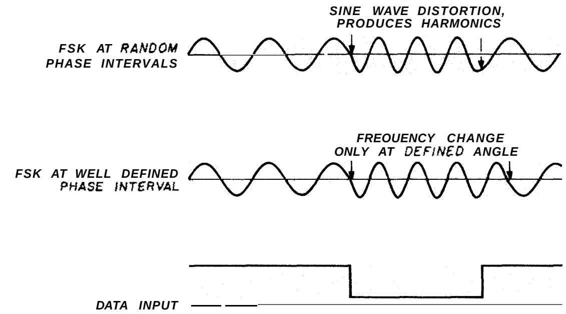


fig. 2. Diagram of coherent and incoherent frequency shifts. The incoherent shift results in sinewave distortion during different parts of the cycle, while a coherent frequency shift has a smooth phase transition, resulting in less distortion.

band is adequately filtered. Carrier null is probably the worst culprit, since balance circuits will change with time. The balance should be periodically checked, otherwise a continuous carrier will be present along with the FSK information.

The sideband filter response is important because of the distortion present in the modulating frequency. Although a sine wave modulating frequency is prescribed, it is extremely difficult to attain. Even if a perfect (no harmonics) sine wave were used, any nonlinearities in the audio stage would introduce some distortion into the modulation. For example, if a frequency of 2125 Hz is used to modulate the transmitter, the sideband filter should suppress the second, third, and other high-order harmonics. **Fig. 1C** shows the effect of insufficient harmonic suppression when a mark or space frequency is being transmitted. As was mentioned earlier, even a pure sine wave injected into the transmitter audio input will end up distorted because of preamplifier nonlinearities. This preamplifier induced distortion will be reduced somewhat due to the response of the sideband filter. If a low audio frequency is used for the mark or space, the second and third harmonic may fall into the passband of the sideband filter. If, on the other hand, the mark and space frequencies are chosen to be high enough in the response band of the filter, the sideband filter will reject the harmonic energy created by signal distortion.

The importance of undistorted wave forms is clear when considering the frequency spectrum of an FSK signal. Too much distortion and the harmonic content is more than the filters can adequately remove. Thus, steps should be taken to ensure that a reasonably clean modulating waveform is applied to the audio input of the transmitter. At the time of the frequency shifts, the phase transitions from one frequency to the other should not contain discontinuities which would contain energy at frequencies other than the space or mark frequencies. This basically means that all space and mark frequency changes should be done smoothly (see **fig. 2**). One way to guarantee this is to always change phases at a zero-crossing of the signal. Using this approach, the audio modulating signal will always change frequency at the same phase point in all information transitions.

and higher-order odd harmonics cause distortion. The lowpass filters are designed to attenuate the third harmonics by 40 dB. The third harmonic of a squarewave is normally 10 dB lower than the fundamental, so this additional 40 dB should put the third harmonic 50 dB below the fundamental. Of course, in practice, this may not be obtained, but the low-

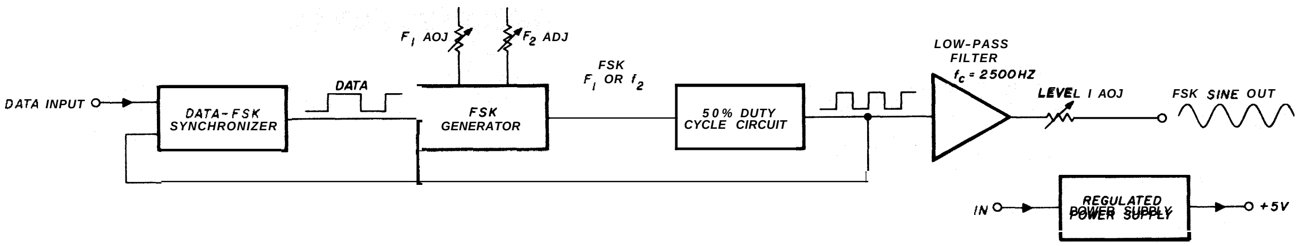


fig. 3. Block diagram of an AFSK generator capable of coherent frequency shifts. Lowpass filters reduce the harmonic content to acceptable levels.

A block diagram of an AFSK generator is shown in **fig.** 3. The generator begins with an oscillator set to twice the needed frequency. When a mark or space is needed, the oscillator's frequency is adjusted for the proper.frequency change. A divide-by-two circuit generates the correct frequencies for the AFSK output waveform, with a lowpass filter removing objectional harmonic energy from the square wave – resulting in a near sine wave output.

Because a square wave is used, the even harmonics theoretically do not exist. Thus, only the third

pass filter response coupled with the bandpass response of the sideband filter in the transmitter should give satisfactory suppression of all harmonics. It should be noted that the mark and space frequencies chosen (2125 and 2295 Hz) must be within the response of the sideband filter in the transmitter. Otherwise, these frequencies will be attenuated.

There is nothing magic about these modulating frequencies. Frequencies of 1800 and 1970 (170 Hz shift) could have been used. However, the receiver demodulator must be matched to these same fre-

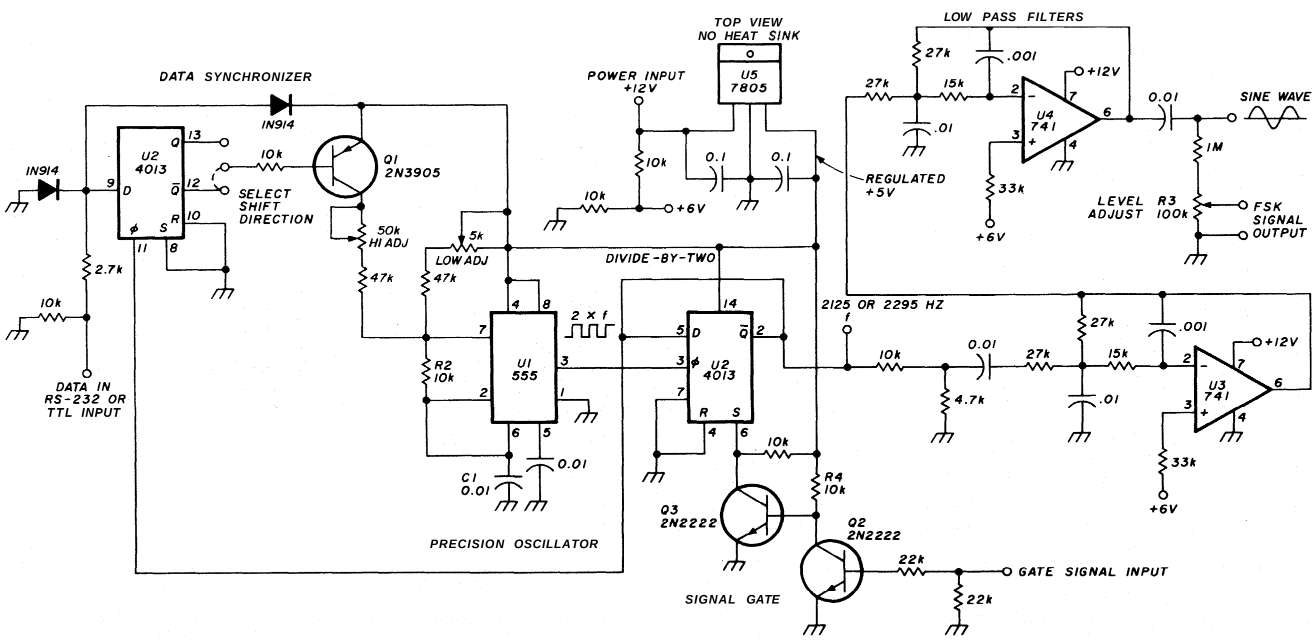


fig. 4. Schematic diagram of the AFSK generator described in the text. The active filter removes harmonic energy while the flip-flop synchronizer only allows phase coherent frequency shifts to occur. All parts associated with the 555 oscillator should have a low temperature coefficient to reduce frequency drift due to temperature change.

quency pairs. This assumes that the receiver-transmitter combination operates on the identical frequency.

circuit description

A circuit which reflects the block diagram just discussed is shown in **fig. 4.** The ubiquitous 555 astable oscillator is used as the frequency generator. Two miniature potentiometers are used to adjust the mark and space frequencies (actually twice the required frequencies). By inputting a logical one or zero, transistor Q1 turns on, changing the RC time constant. For the 555 oscillator, the frequency of oscillation is:

$$F = \frac{1.44}{(R1 + 2R2)\,C1}$$

where R1 is the parallel combination of the resistors used for frequency setting. Normally, the frequency of the 555 is set to twice the mark and space frequencies. U2 is used as a divide-by-two circuit which will guarantee a 50 per cent duty cycle. The other half of this dual-D flip-flop is used as the input data synchronizer. The mark or space input will affect the frequency of the 555 only when the output of the divide-by-two changes state. Thus, all frequency shifts are synchronized by one well-defined phase point in the oscillator's period. The divide-by-two circuit is easily disabled, creating a convenient method of gating the oscillator. This is useful for gating the oscillator off and on to a CW identification. Q2 and Q3 simply buffer the input, which inhibits the divide-by-two. U3 and U4 are used as a dual-stage, active lowpass filter. These filters each have a two-pole Butterworth response. Each lowpass response results in a 40 dB per decade roll-off characteristic. In tandem the responses add, yielding and overall 80 dB per decade response. The lowpass filters use inexpensive 741-type operational amplifiers. The last output of the filter is attenuated by R3. This potentiometer can be adjusted to set the output drive level feeding the audio input of the transmitter.

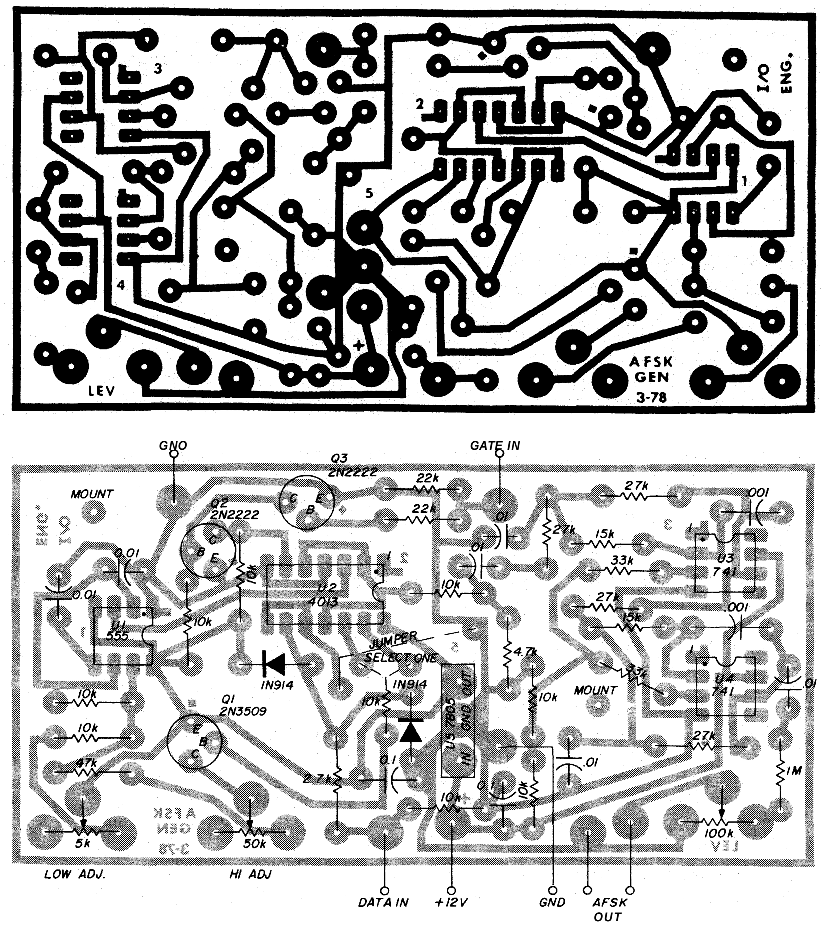


fig. 5. Foil pattern (above) and parts placement diagram (below) for the phase coherent RTTY modulator.

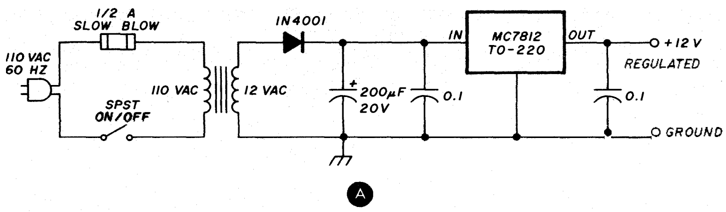
The circuits are powered from a 12- to 15-volt dc source. The voltage reference for the oscillator section, however, is regulated by a three-terminal regulator to ensure that voltage fluctuations will not influence the frequency of oscillation. Even though the 555 oscillator has a specified 0.1 per cent volt tolerance to power supply change, I have found it best to regulate the power source to remove any problems with errors due to this change. If you use a well-regulated power supply, the 7805 three-terminal regulator may be omitted and a jumper wire installed.

Layout of the circuit is not critical since only audio frequencies are generated. A printed-circuit-board foil pattern is shown in **fig. 5.** The power bus should be filtered to remove dc transients which might be propagated from other circuits attached to the same power source. A typical power supply is shown in **fig. 6A.** Also shown is a circuit which may be used to convert a TTY current loop to the proper voltage for driving the modulator input. For those who wish to duplicate this modulator, a circuit board is being made available.*

adjustment

Once the circuit is constructed, checkout is relatively straightforward. A 12-15 volt power supply should be connected between the power input and ground. Since the "shift direction" is jumper programmable, one or the other polarity for the shift direction should be selected. If needed, a single-pole double-throw switch could be used to remotely select the shift direction. Assuming pin 12 is selected for the shift direction and no "Data-in" signal is present, Q1 will be turned off. With Q1 off, the "Low-Adj" potentiometer should be adjusted for an output frequency of 2125 Hz. This frequency can be measured at the FSK signal output port, or monitored at pin 2 of U2. Once the frequency is brought into the proper range, "Data-in" should be connected to a logic one level. Q1 will now turn on, and the "High-Adj" potentiometer should be adjusted for an output frequency of 2295 Hz.

If no signal appears at the output or at pin 2 of U2, measure the level at pin 6 of U2. It should be at ground potential when no "Gate" signal is present. If it is not at ground potential, Q2 and Q3 may be the wrong type or inserted improperly. If the frequency can not be adjusted to the proper range, C1 may be at fault. Since C1 determines the timing, only a high-quality capacitor should be used. Any temperature drifts of this capacitor will create a proportional drift in the oscillation frequency. Although a 0.01-μF capacitor is shown, slight variations of this value are permissible due to the use of potentiometers for determining the frequency. However, extreme variations may not work because of the limited range of the potentiometer adjustment. Therefore, if the frequency will not adjust to the exact frequency needed, try using another capacitor for C1. If all else fails, a smaller-value capacitor could be paralleled with C1 to "fine tune" the frequency.



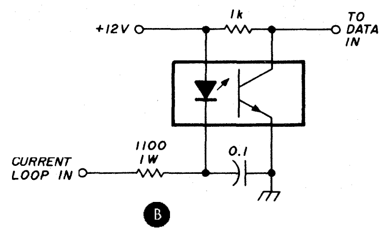


fig. 6. Typical power supply for use with the RTTY modulator is shown in (A). A TTY current loop interface, using an op-isolator, is shown in (B).

After the oscillations are set to within the tolerance wanted, go back and check the frequencies as the "Data-in" line is switched from a logic one to a logic zero. The "Gate" signal input can next be checked by connecting a logic one voltage to the input. The output oscillations should cease. Note, if the opposite logic polarity is needed for the disable gate input (logic zero for disable), Q2 and R4 can be eliminated, but be sure to jumper between the collector and base of Q2. This is why two transistors are used in the Signal Gate circuit, to allow for the option of inverting the gate signal.

I should mention that the 4013 dual flip-flop is a CMOS device. Thus, care should be exercised when handling the unit because static buildup can damage the sensitive MOS input transistors. Also, beware of bargain basement CMOS devices. I have seen some ICs purchased from outlets which by no means met specifications. All outputs should swing from the power supply potential, for a logic one, to practically ground potential, for a logic zero. This assumes no current is being "sourced" or "sinked" by the outputs. If the CMOS device does not meet this simple criterion, send it back; it is defective.

I hope this will clarify the AFSK approach for generating FSK with single-sideband equipment. The pitfalls to avoid should be recognized for compliance with FCC regulations and for reducing interference on the amateur bands.

ham radio

*A predrilled, single-sided printed circuit board is available for $5.00 postpaid from I/O Engineering, **12412** Mossy Bark, Austin, Texas 78750.

time-current charging of nickel-cadmium batteries

Time-current charging: a technique for quickly charging nickel-cadmium batteries

Want to charge sealed, nickel-cadmium batteries quickly and safely? The dump, time-current charging method is not new, but it seems little known by many people using nickel-cadmium batteries in electronic equipment.

Sealed nickel-cadmium cells may be charged and discharged at very high rates (high currents), if certain rules are observed. When discharging, overheating of the cells should be avoided. Not letting them get too hot to handle is a safe rule. (Note: Cells used in portable soldering irons are effectively short circuited by the low-resistance soldering element; for short periods, they may supply hundreds of amperes without damage.)

In the case of charging, the same rule applies – with one major limitation. This limitation is that high currents ***must*** be ***avoided*** when the cell is near or above full charge. When above full charge, the cell will produce gas if it is charged at a rate above 10 per cent of its (one hour) ampere-hour (A-H) rating. In most cases, this is the recommended slow-charge rate and is the rate that can be used for prolonged periods of overcharging without apparent damage.*

*If you are going to trickle charge batteries during idle periods, a rate of 1 per cent of ampere-hour rate would probably be more reasonable.

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Open cells are not an altogether different matter. Most of this article applies also to that type of cell. However, the following items are important if you use open cells. Always open the filler vent when charging; do not trust any automatic vent that may be provided. Unless the cell has leaked, add only distilled water to bring the electrolyte back to the proper level. Charge the cell until it freely "outgasses"; that is, until many bubbles start to rise in the electrolyte. The dump, time-current charging method can be used with open cells. However, since the gas pressure problem does not exist and a good "full

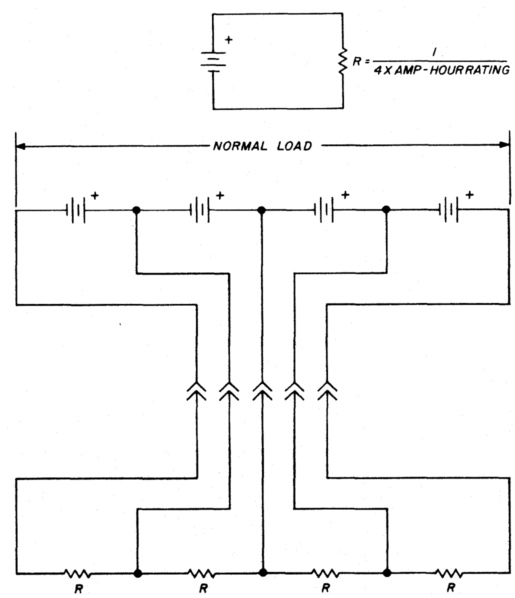

fig. 1. The dump circuits in this diagram can be varied to suit any situation. If the batteries are not soldered into the circuit, a simple battery holder, with dumping resistors soldered across the terminal, can be used. A voltmeter can be used to measure the cell voltages. Discharge each cell to about 0.5 volt; at 0.5 volt, the cell has less than 5 per cent of its full charge.

charge" indicator does exist, the method is not very useful.

time-current charging

The basics of this system are to first completely discharge the cell and then to recharge it to less than 100 per cent full charge with a known high current for a specific length of time. It may be used at this point, or, if full capacity is required, charging may continue at the normal 10 per cent rate.

To avoid reverse charging, it is very important that fast discharging be done *individually* on each cell. If you are working with a battery of cells, a jig must be made to separately discharge the cells. Even at normal discharge rates, care must be taken not to discharge cells connected in series. In some cases, the discharged cells are reverse-charged and frequently become reverse-polarized. When this occurs, the cell will not recharge in the normal manner; it will retain its reverse polarity. Sometimes the cell can be brought back to normal polarity by giving it a massive charge in the proper direction. Typically, half-ampere cells are charged at rates of several amperes for a few minutes. This "cure" works in many cases, but the reliability of the cell is questionable from that point on.

dumping

Discharging (or dumping) can be safely accomplished at four times the rated one hour A-H current. Typically, a four A-H cell can be safely discharged at 16 amperes. In this case, a full charge will take about fifteen minutes to dissipate. If the cell is less than fully charged, correspondingly less time will be required. Satisfactory values of resistance for several popular nickel-cadmium cell sizes are given in table **1.** The circuit for discharging single or multiple cells is shown in **fig. 1.**

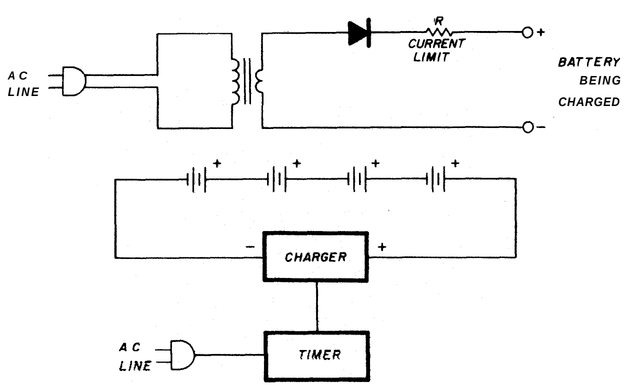

fig. 2. Diagram of a simple charging circuit. Commercial chargers can be used, or a simple transformer/diode circuit can be built.

Table 1. Resistance values needed for a cell discharging system.

cell size	A-H rating	discharge current amperes	resistance	minimum wattage
D	3.5 A-H	12.0	0.1 Ohm	25.0 Watts
C	1.5	6.0	0.2	10.0
AA	0.5	2.4	0.5	5.0
—	0.25	1.0	1.2	1.0

charging

Charging is most effectively done with the cells connected in series. This allows a single charger to put the same charge current through all of the cells simultaneously (see **fig. 2**). Charging can be done at currents as high as 50 times the one-hour ampere-hour rating of the cell; a 150-mA cell can be charged at 7.5 amperes. The charging time is calculated as follows:

$$time = \frac{A\text{-}H\ rating}{charging\ current}$$

$$= \frac{0.150\ A\text{-}H}{7.5\ A}$$

$$= 1.2\ minutes$$

This short a time, however, is an extreme that should be avoided because of the timing accuracy required. Missing by a few seconds could lead to an accident. A misrating on the cell could be equally dangerous. If you choose a rate of five times the A-H rating, the time would be 12 minutes and the time tolerances become reasonable. Plus or minus one minute will result in about 10 per cent of full charge.

At any rate of charge, the 100 per cent charge time may be calculated using the previous formula. Although the prime advantage of the dump, time-current charge method is speed, somewhat slower discharge and charge rates will tend to be safer than high rates, which may ruin a battery if care is not used. I strongly recommend that a timer be used to turn off the charger, rather than trusting the clock-watching method.

acknowledgement

The assistance of Mr. E. L. Williams in the preparation of this article is gratefully acknowledged.

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ham radio

quartz crystals — gems for frequency control

A misunderstood hero of the electronics world is the quartz crystal. Quietly it awaits your command to put your transmitter on frequency, to reject all but one sideband, or to select one rare CW signal sandwiched between adjacent kilowatt signals. What is the secret of quartz? Can an amateur operator zero-adjust his crystal oscillator, or is he stuck with a bad crystal? How do these pieces of quartz operate in an oscillatory circuit? It's hoped that this article will answer some of your questions and help you in procuring and designing circuits with that celebrated mineral.

the quartz crystal — some background

Quartz technology is based on its piezoelectric property. The application of an electric field causes certain substances to oscillate; conversely, the application of a mechanical force or vibration causes substances to generate an electric field, known as the piezoelectric effect. Quartz is useful as an electrical oscillator operating in a very narrow frequency

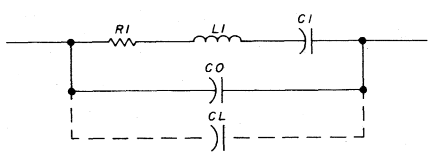

fig. 1. Equivalent circuit of the quartz crystal. L1, C1, R1 are the primary, or motional components, which determine frequency and circuit Q. Capacitance C_0 is the electrode, mounting structure, and holder capacitance. Capacitance C_L is the parallel capacitance across the circuit.

band. The precise frequency, activity, and temperature characteristics are determined by the position and angle of cut on the crystal.

The old concept that the quartz crystal is a *standard of frequency* was born in an age of less-critical applications. Old timers knew that the crystal was much more accurate and repeatable than any LC circuit. Because they didn't have to multiply 18 times and trigger a repeater, it's easy to see how the legend of quartz stability became exaggerated.

The basis for stability in quartz is its high inductance and low capacitance, resulting in extremely high Q. In an 8-MHz crystal unit, for example, the Q might be 150,000 while the Q of a typical LC combination at that frequency is about 300. Yet a crystal's frequency may be pulled; and in time, it will drift.

equivalent circuit

The simplest and most commonly used equivalent circuit of the crystal is shown in **fig. 1.** $L1$, $C1$, and R1 are the primary components which determine frequency and Q. These are referred to as the *motional components* and their parameters can't be measured directly. C_0 represents the electrode capacitance, the mounting-structure capacitance, and holder or case capacitance. C_0, the static capacitance, affects the crystal operating frequency, but to a lesser degree than $C1$. C_0 can be measured by a capacitance bridge across the terminals. As you may expect, the capacitance of the circuitry, shown as another parallel capacitance, C_L would also have an effect on the crystal working frequency. The equation for the working frequency is

$$F_W = \frac{1}{2\pi\sqrt{L1\left[\dfrac{C1(C_0+C_L)}{C1+C_0+C_L}\right]}} \qquad \textbf{(1)}$$

With the help of a calculator, you can determine how much the crystal is pulled by the oscillator circuit. By changing the circuit loading, the crystal may be pulled (within limits) for fine tuning or fm applications.

mode of operation

The classic crystal reactance curve, **(fig. 2)**, is useful in demonstrating the relationship of different operating frequencies. At two points the reactance is zero; *i.e.*, the crystal looks purely resistive. The lower of these frequencies is the series-resonance frequen-

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cy, (F_S); while the higher frequency is the anti-resonance-frequency, (F_A). The resistance is low at series resonance and high at anti-resonance. The range of frequencies between is known as the natural bandwidth of the crystai. Anti-resonance is a very unstable point, and for amateur purposes, may be forgotten.

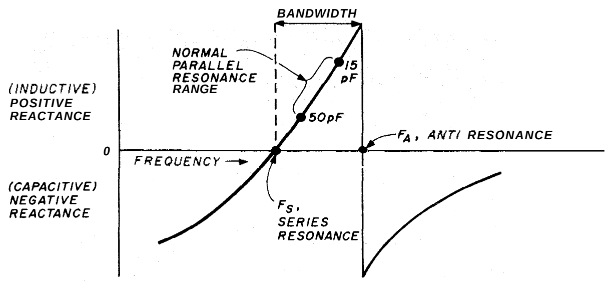


fig. 2. Classic crystal reactance curve, which is useful in demonstrating the relationship of different operating frequencies.

Parallel resonance is commonly recognized as the band of frequencies between F_S and F_A, although classic crystal theorists have another definition. You may think of this band as the range where the crystal will operate if a capacitor is placed in parallel with it. At F_S the capacitance will be infinite; at F_A, the capacitance will be zero. Practical limits are between 15 pF and 50 pF, where poor stability exists at the

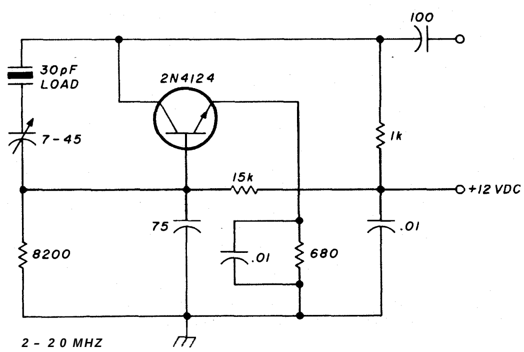


fig. 3. Example of crystal loading. The trimmer capacitor has a negative reactance, so the crystal frequency is shifted into the positive-reactance region of the reactance curve (fig. 2).

low-capacitance end and reduced activity degrades the high end. When ordering a crystal, you must specify ***series resonance*** or ***parallel resonance at a specified load capacitance.***

There's a lot of confusion about crystal loading. Responsibility for this confusion falls directly onto the quartz-crystal industry, whose members have never acted together to educate users. Guidance committees have recommended that we consider the

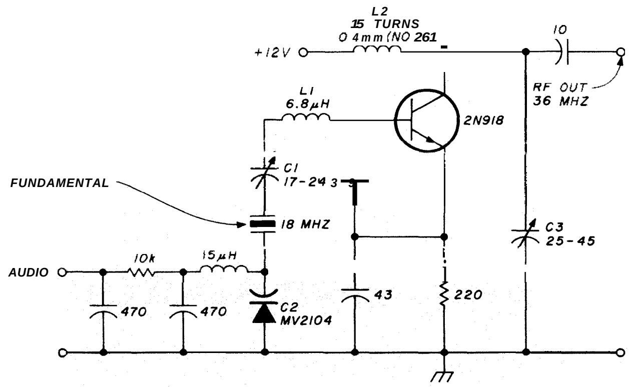


fig. 4. In this example the crystal operates into a complex load at series resonance. L1, C1, and C2 balance the crystal at zero reactance. Capacitor C1 fine tunes center frequency. Tank circuit L2, C3 doubles the output frequency. Circuit operates as an fm oscillator-doubler.

crystal operation in the ***positive-reactance*** mode when above series resonance and in the ***negative-reactance*** mode when below series resonance. This recommendation is technically correct and allows us to discuss a useful range of operation (below F_S) of the crystal, which is not usually considered.

In **fig.** 3, the crystal is in a feedback circuit from collector to base. A trimmer capacitor in series shifts the point on the reactance curve where the crystal operates, thus providing a frequency trim. The capacitor has a negative reactance so the crystal is shifted to operate in the positive reactance region of the curve **(fig. 2)**.

The series trimmer does not mean the crystal is

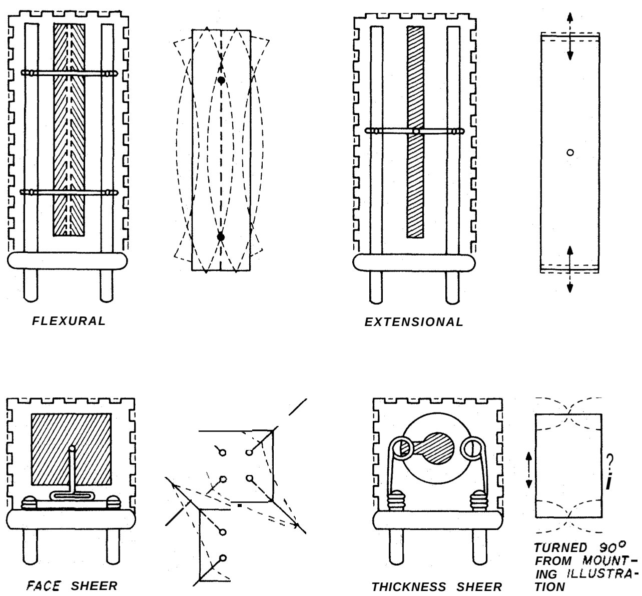
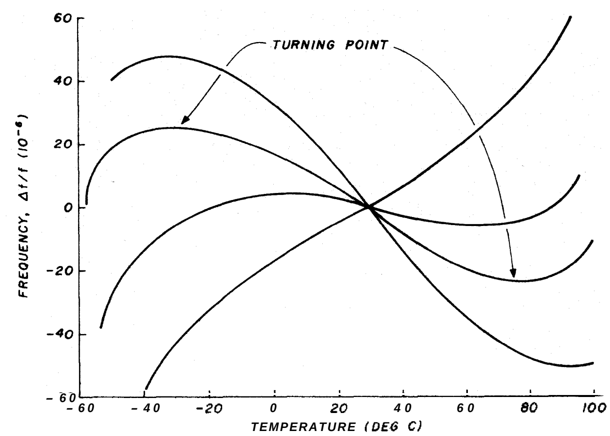


fig. 5. Examples of vibration modes and mounting **structures**. Note that the quartz supporting structures are **fastened** at points of least motion (nodes).

operating at series resonance. It is said to be operating in the parallel-resonant mode with a load capacitance approximately equal to the trimmer value. If an equivalent circuit were drawn, you could see that the trimmer would be the parallel load. By placing the capacitor in series, you isolate the crystal from other circuit reactances, enabling the trimmer to tune more effectively than in the parallel connection.

Oscillators using fundamental crystals usually operate in the positive-reactance mode with a trimmer for exact tuning. One reason for operating this way is seen when the trimmer is removed, leaving only the crystal in the feedback circuit. The crystal should operate at series resonance, but circuit reactance will usually pull it slightly off frequency.

In **fig. 4** the crystal operates at series resonance into a complex load. *L1, C1,* and *C2* balance the crystal at zero reactance. Capacitor *C1* fine-tunes center frequency. The tank circuit, *L2, C3,* doubles the output frequency. As the audio signal varies *C2* capacitance, the crystal will operate alternately in the positive, then the negative-reactance mode.



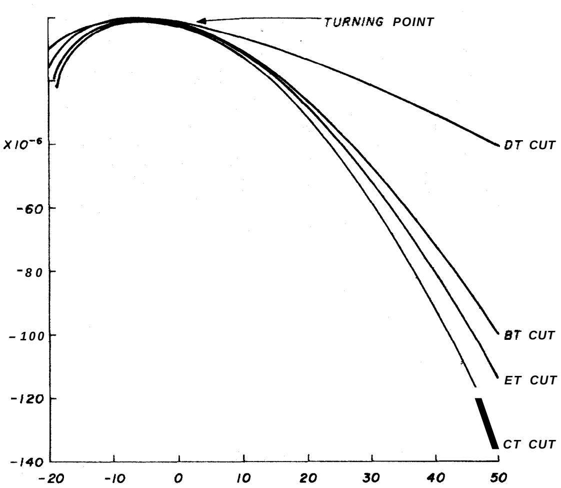


fig. 6. Typical family of AT-cut frequency/temperature variations for change of angle only (thin plates), (A). These are cubic functions centered on 27 degrees C. Sketch (B) shows frequency/temperature curves where the point of zero temperature coefficient can't be controlled. These curves are typical of low-frequency cuts.

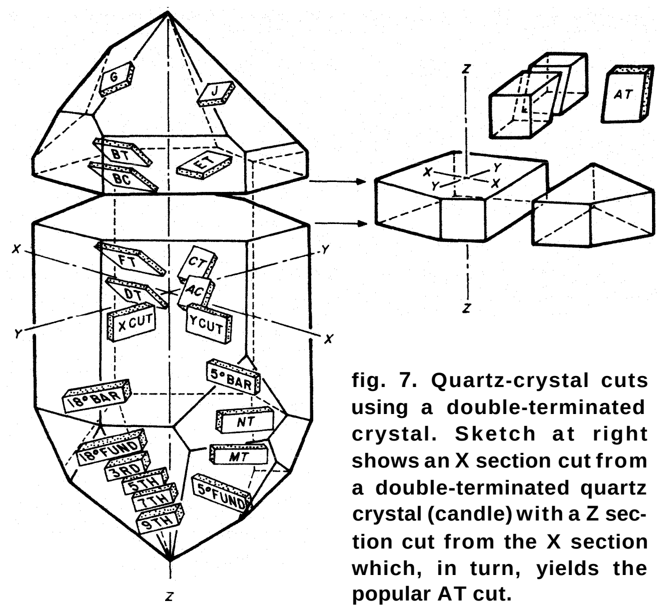


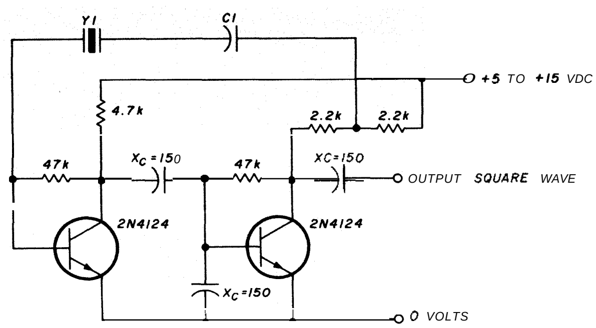
fig. 7. Quartz-crystal cuts using a double-terminated crystal. Sketch at right shows an X section cut from a double-terminated quartz crystal (candle) with a Z section cut from the X section which, in turn, yields the popular AT cut.

Deviation per volt of modulation is greater at series resonance than it would be into a capacitive load (positive-reactance operation). It would be even more desirable to operate this circuit completely in the negative-reactance mode because of more favorable deviation per volt (see **fig. 2**). The average amateur would have a problem designing a circuit for the negative-reactance mode because he must order his crystal at a higher frequency than the design frequency. Crystal manufacturers do not calibrate their crystals to tune into an inductive load. A second, and stickier, problem is that one manufacturer's crystals are more easily pulled than others. Plainly, some crystals won't work in a design acceptable for another crystal.

Using a small inductor in series with the crystal is usually a practical way to lower the frequency slightly. It's the only way to lower a crystal frequency operating at series resonance. A trimmer capacitor, also in series, can be used for fine tuning. Inductor values will depend on the crystal frequency but will be microhenries or fractional microhenries for **1-20** MHz crystals. As suggested before, not all crystals of the same frequency will shift equally. There's a limit to how much each crystal can be pulled and still operate reliably. It's good practice to see if the oscillator will start and maintain oscillation under extremes of temperature and voltage.

practical circuits using fundamental-mode crystals

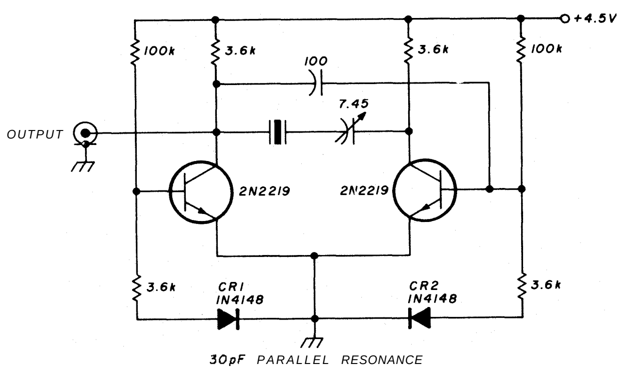
Some practical circuits using fundamental-mode crystals follow. These circuits were chosen to demonstrate a point and should be good for reference. All are believed to be workable although I have not built all of them.



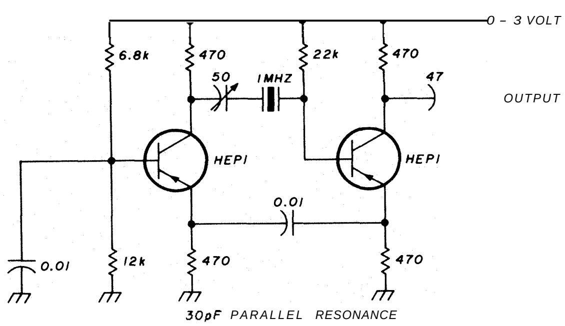
NOTES

1. Y1 IS H, NT, OR E CUT
2. C1 IN SERIES WITH THE CRYSTAL MAY BE USED TO ADJUST THE OSCILLATOR OUTPUT FREOUENCY. VALUE MAY RANGE BETWEEN 20pF AND 0.01μF, OR MAY BE A TRIMMER CAPACITOR AND WILL APPROXIMATELY EOUAL THE CRYSTAL LOAD CAPACITANCE.
3. X VALUES ARE APPROXIMATE AND CAN VARY FOR MOST CIRCUITS AND FREOUENCIES; THIS IS ALSO TRUE FOR RESISTANCE VALUES.
4. AOEOUATE POWER SUPPLY DECOUPLING IS REOUIRED. LOCAL DECOUPLING CAPACITORS NEAR THE OSCILLATOR ARE RECOMMENDED
5. ALL LEADS SHOULD BE EXTREMELY SHORT IN HIGH FREOUENCY CIRCUITS.

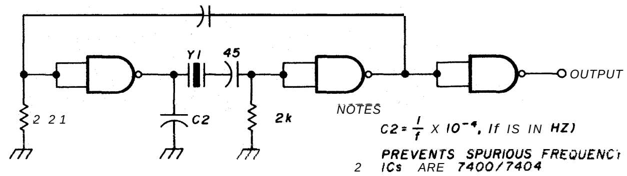
Low-frequency oscillator — 10 kHz-150 kHz.



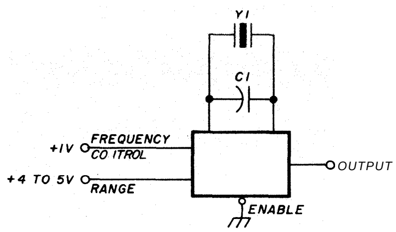
100-kHz standard oscillator. CR1, CR2 stabilize output.



Standard oscillator for 1 MHz.



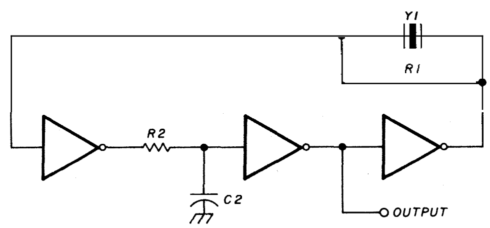
TTL oscillator for 1 MHz-10 MHz.



NOTES.

1. FOR FREQUENCIES ≤ 1 MHZ, C1 = 5 TO 15pF. FOR FREQUENCIES ≥ EMHZ, C1 MAY BE
2. IC IS SN74S124 FOR f MAX. OF 60 MHZ IC IS SN74LS124 FOR f MAX. OF 35 MHZ

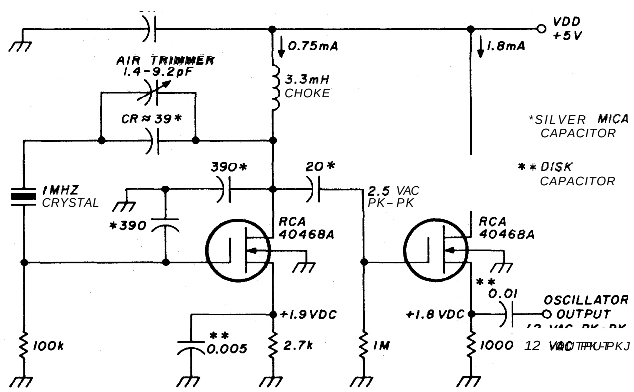
Voltage-controlled oscillator using ICs.



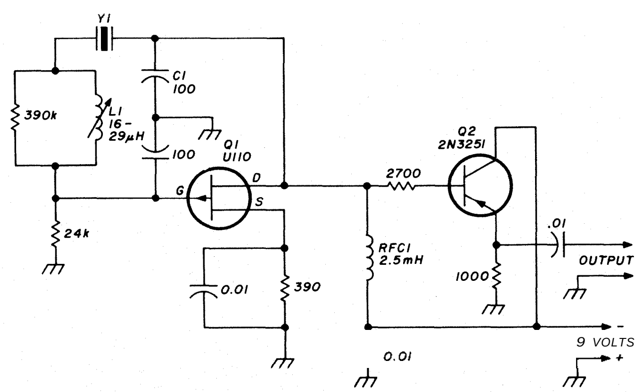
NOTES:

1. 1M < R1 < 5M
2. SELECT R2 AND C2 TO PREVENT SPURIOUS FREQUENCIES
3. ICs ARE 74C04 OR EOUIVALENT

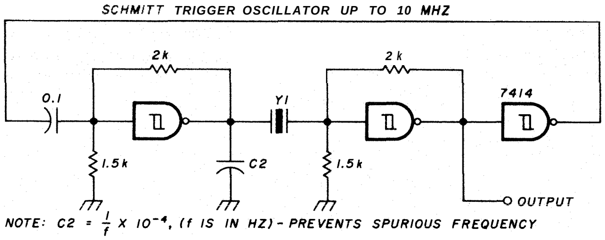
CMOS oscillator — 1 MHz-4 MHz.



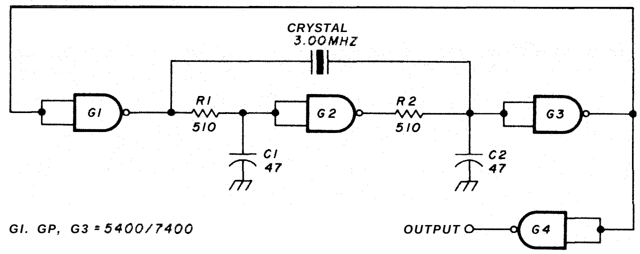
1-MHz fet oscillator and buffer. Circuit exhibits less than 1-Hz frequency change over a V_{DD} range of 3-9 volts. Stability is attributed to mosfets and caps.



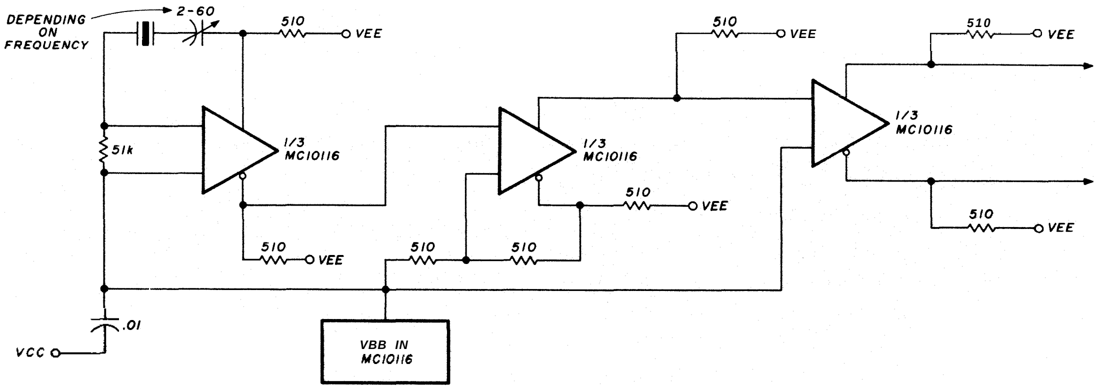
Stable VXO using 6- or 8-MHz crystals uses capacitor and inductor to achieve frequency pulling on either side of series resonance.



Schmitt trigger provides good squaring of output, sometimes eliminating need for an extra output stage.



Crystal-controlled oscillator. This circuit, described in reference 1, oscillates without the crystal. With the crystal in the circuit the frequency will be that of the crystal, Circuit has good starting characteristics even with the poorest crystals.



1-20 MHz oscillator. Circuit operates on fundamental frequency of the crystal selected without a tank circuit. It provides noninverting output. V_{BB} **is** 1.2 volts, available from the IC; V_{EE} **is** -5.2 volts. Second section of **IC** is connected as a Schmitt trigger driving the third section, connected as a buffer, to give good square-wave output suitable for use as a clock driver.

device construction

The natural classification of crystal resonators is according to frequency. The frequency range covered commercially by quartz-crystal units may be taken as a few hundred Hz to over 250 MHz. Use is made of several cuts and patterns of motion (modes). Three common modes of vibration are: flexural, extensional, and shear. **Fig. 5** illustrates these modes and typical mounting techniques.

table 1. Some quartz crystal vibrators and their principal characteristics.

vibrator designation	usual description	vibration mode	usual frequency range
J	+5° X duplex	flexural	0.2-10 kHz
K	+5° XY bar	flexural	2-16 kHz
H	+5° X plate	flexural	8-100 kHz
N	NT	flexural	8-100 kHz
E	+5° X plate	extensional	40-200 kHz
C	CT	face shear	150-750 kHz
D	DT	face shear	100-500 kHz
G	GT	extensional	90-250 kHz
S	SL	face shear	200-1000 kHz
A	AT	thickness shear (fundamental]	0.8-25 MHz
B	BT	thickness shear (fundamental)	3-40 MHz
A, B	AT or BT	thickness shear (n^{th} overtone; n = 3, 5, 7 etc.)	15-250 MHz

The designations of certain quartz-crystal vibrators with some of their principal characteristics are summarized in **table 1.** At lower frequencies there are advantages to using one vibrator design over another. Tolerance, activity, and temperature characteristics exemplify the need for choosing. Above 1 MHz most crystals are AT cuts. In general, the choice of cut is that of the manufacturer based on the specification.

temperature characteristics

Most crystals in amateur service are AT cuts, as our needs are primarily above 1 MHz. A notable exception is the 100-kHz calibrator crystal, which is likely to be an ET cut. Excellent temperature stability and aging are attributed to the AT-cut resonator because of its high Q and cubic temperature curve. In **fig. 6A** a family of AT-cut temperature curves is depicted. The difference between these curves is determined by a change in the angle of cut of the quartz of only a few minutes of arc.

A manufacturer first determines the precise angle that will give the best temperature characteristic commensurate with users' needs. He then determines the crystallographic axes of the quartz and cuts it in the appropriate orientation. Several cuts are illustrated in **fig. 7.**

The temperature characteristics of low-frequency cuts are usually parabolic, as shown in **fig. 6B.** Many of these can be adjusted with respect to the temperature of the turning point; but tolerances are poorer than the AT types. For greatest accuracy in any type of crystal, proportional control ovens, operating at the crystal's turning point, are used.

overtone crvstal units

Crystals with frequencies higher than 20 MHz are usually overtone types, although fundamental-type crystals have been made at as high as 35 MHz. Overtone crystals are distinguished from fundamental crystals by their design, which is to operate at an odd harmonic of the crystal basic frequency. It's generally practicable to excite At- and BT-cut plates into third, fifth, seventh, and ninth harmonics of the fundamental frequency; hence a 10-MHz crystal can be vibrated at approximately 30 MHz, 50 MHz, 70 MHz, and 90 MHz. The relationship between overtone and fundamental frequencies is approximately, but never exactly, equal to the integer expressing the harmonic order.

To use the overtone crystal most effectively it's important to know the following characteristics:

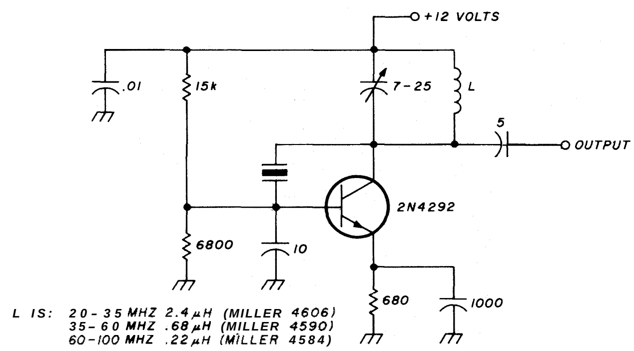
1. A tuned circuit must be used with the crystal to excite it into the desired harmonic mode. If the Q of this circuit is too low, improper operation of the crystal will result. The designer must also take care that no other resonances are present that may excite the crystal into another mode.

2. Overtone crystals are designed for operation at series resonance. Because of the narrow bandwidth and low motional capacitance, these crystals are not suitable for fm or for variable-crystal oscillators. Phase-lock operation is practical, however.

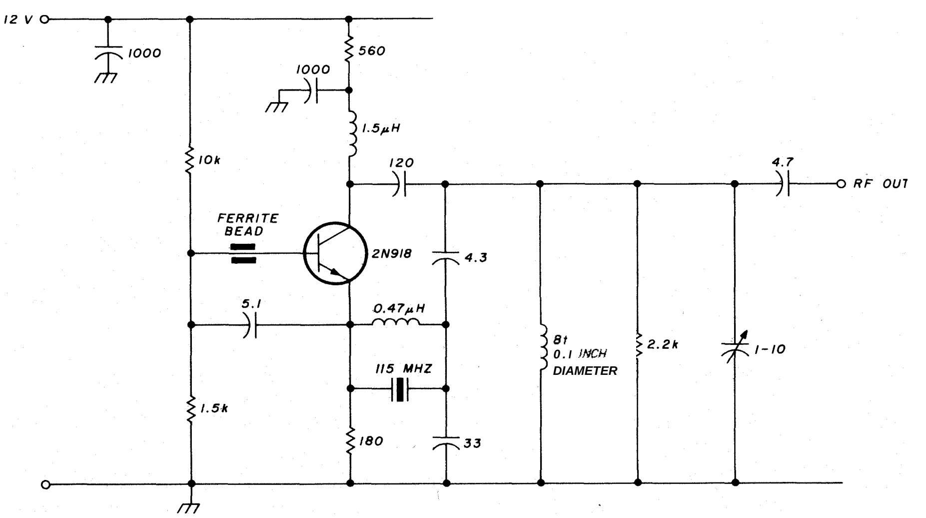
3. The characteristics of a crystal such as temperature coefficient and equivalent resistance apply only to the design frequency. These properties are different for fundamental operation or other harmonic orders.

practical overtone crystal oscillators

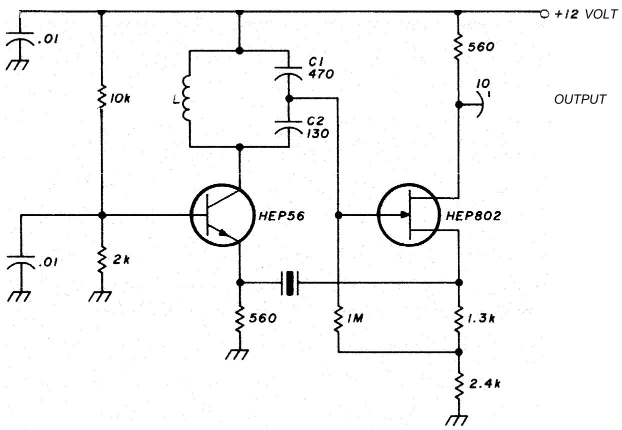
Some useful overtone oscillator designs are shown below, and on the facing page.



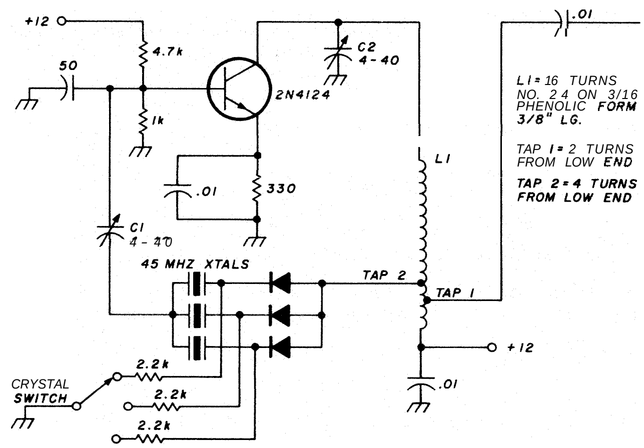
This oscillator is designed for overtone crystals in the 20-100 MHz range operating in the third and fifth mode. Operating frequency is determined by the tuned circuit.



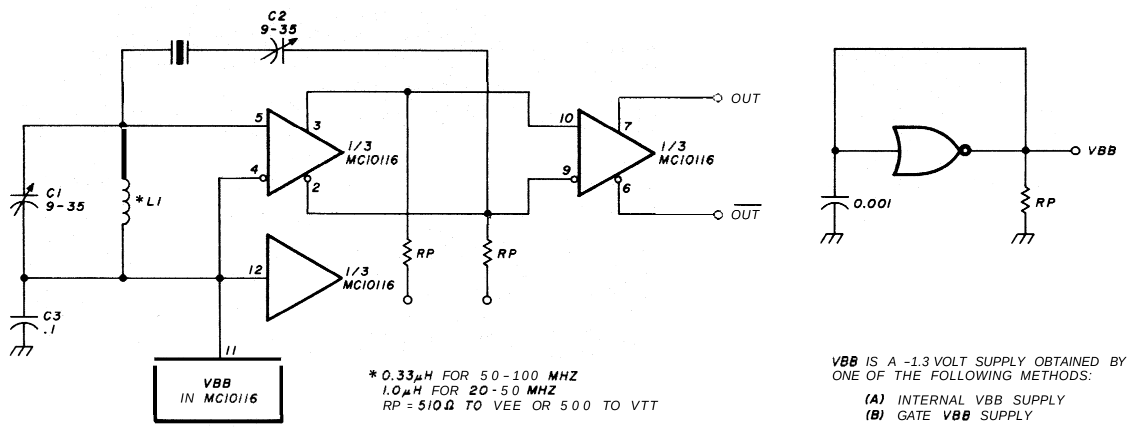
Design for high reliability over wide temperature range using fifth and seventh overtone crystals. Inductor in parallel with crystal causes antiresonance of crystal C_0 to minimize loading. Technique is commonly used with overtone crystals.



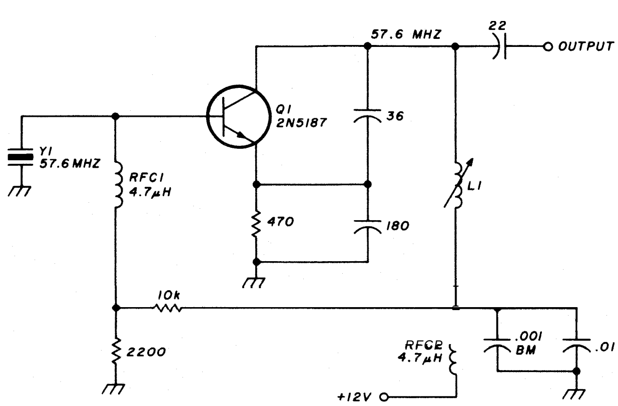
Typical Butler oscillator (20-100MHz). An fet should be used in the second stage; circuit is not reliable with two bipolars. Sometimes two fets are used. Frequency is determined by LC values.



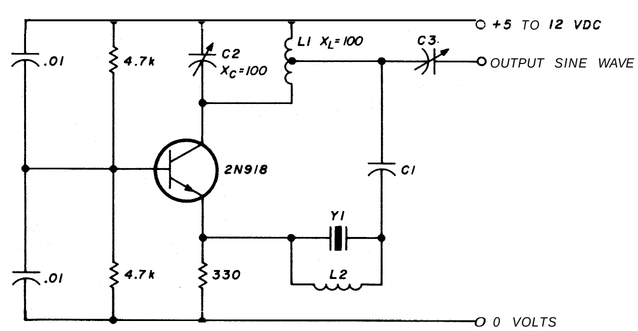
Overtone oscillator with crystal switching. Similar circuits electronically switch the crystals. The large inductive phase shift of **L1** is compensated for by **C1.** Overtone crystals have very narrow bandwidth, therefore the trimmer has a smaller effect than for fundamental-mode operation.



Overtone oscillator using Motorola MECL devices. Frequency range is20 MHz-100 MHz, depending on crystal frequency and tank-circuit tuning. The tank, **C1,** L2, is tuned to select the proper overtone mode. C2 compensates phase shift of the **IC.** More details are given in reference 2.



Fifth-overtone oscillator isolates the crystal from the dc base supply with an rf choke for better starting characteristics.



NOTES:

1. Y1 IS AT CUT OVERTONE CRYSTAL.
2. TUNE L1 AND C2 TO OPERATING FREQUENCY
3. L2 AND SHUNT CAPACITANCE, CO. OF CRYSTAL (APPROXIMATELY 6pF) SHOULD RESONATE TO OSCILLATOR OUTPUT FREQUENCY (L2 ≈ .5µH AT 90 MHZ). THIS IS NECESSARY TO TUNE OUT EFFECT OF CO.
4. C3 IS VARIED TO MATCH OUTPUT.

50 MHz-150 MHz overtone oscillator uses a **2N918.**

effects of drive level

The level of drive imposed on an oscillator crystal is usually specified in terms of the pcwer dissipated in it. Ideally, the crystal oscillator should be regarded as a source of stable frequency, but in practice it must also be considered as a source of power.

Changes in drive level will affect the resonator frequency; therefore, it's important for the manufacturer to know the drive level of the oscillator circuit for calibration of the crystal. Crystals operated at high drive levels will become unstable, sometimes jumping frequency into a spurious mode. Excessive resonator heating may cause a permanent shift in frequency or possibly facture the quartz. The NT resonator is particularly vulnerable to fracturing. A good rule is to operate the crystal at the lowest drive level compatible with good starting characteristics.

The old WWII surplus pressure-type crystals use larger pieces of quartz than their modern counterparts. As might be expected, these can withstand higher drive levels. You may also find that pressure-type crystals can be pulled in frequency more easily than modern units because their motional capacitance, $C1$, is higher.

aging

Like mountain dew, most crystals improve with time. Just after the crystal is manufactured, there are stresses, which when relieved, change the crystal frequency. Most manufacturers age the crystal by temperature cycling or high-temperature aging until the worst changes have occurred. You'll then experience slower drift. In many applications, the drift is negligible but is present. The most stable crystals are those in the 4-5 Mhz range.

Aging can be positive or negative, depending on which factors are present in a particular unit. Migration of small particles within the crystal holder is usually blamed for frequency changes. If these dirt particles land on the crystal, its frequency decreases. These particles are present despite the most rigorous cleaning procedures. Metal-cased, gas-filled crystal units will usually age negatively. Some crystals are evacuated rather than gas filled. These units are cleaner and have better aging characteristics but lower drive level ratings. Evacuated units may age higher in frequency because some of the plating is vaporized. You won't find these crystals on the surplus market; they are mentioned here as a point of interest.

tips on using crystals

The great enemy of quartz is drift. Old pressure types have been known to fail because of particles from the rubber gasket, which may have deteriorated. Careful cleaning with alcohol or similar solvent will bring these crystals back to life. The same procedure will probably increase the frequency of a unit that hasn't failed. Most certainly it will increase crystal activity.

This trick isn't practical with solder-seal holders, but then these units are much more reliable. Don't open the holders on the solder seal units or you'll find that the frequency has changed. This is because of a change in pressure and of gases surrounding the quartz element. Besides the frequency change, reliability is compromised by the introduction of dirt.

Sometimes an oscillator crystal is used in a filter application, but performance will not always be satisfactory. Special designs are used for filter crystals. These crystals have lower activity and are virtually free of spurs (unwanted modes). In oscillator service, the presence of unwanted modes is not as critical as in filter service, where broadband energy will excite all modes.

There are many uninvestigated facets of the quartz crystal. While some people still claim crystal manufacturing is akin to witchcraft, this is just not so. A few years ago, natural quartz, which was mined in Brazil, was used for all U.S. crystals. Synthetic quartz made in the U.S. has been improved to the point where it's now used in all but the most critical applications.

Natural quartz still has higher *Q*. Synthetic quartz has the advantage of perfect crystalline structure and uniform size. The use of natural quartz incurs much waste – rarely is there a fully perfect crystal, and small structures may not be practical for cutting. Crystallography is certainly a science not fully investigated but one we should study.

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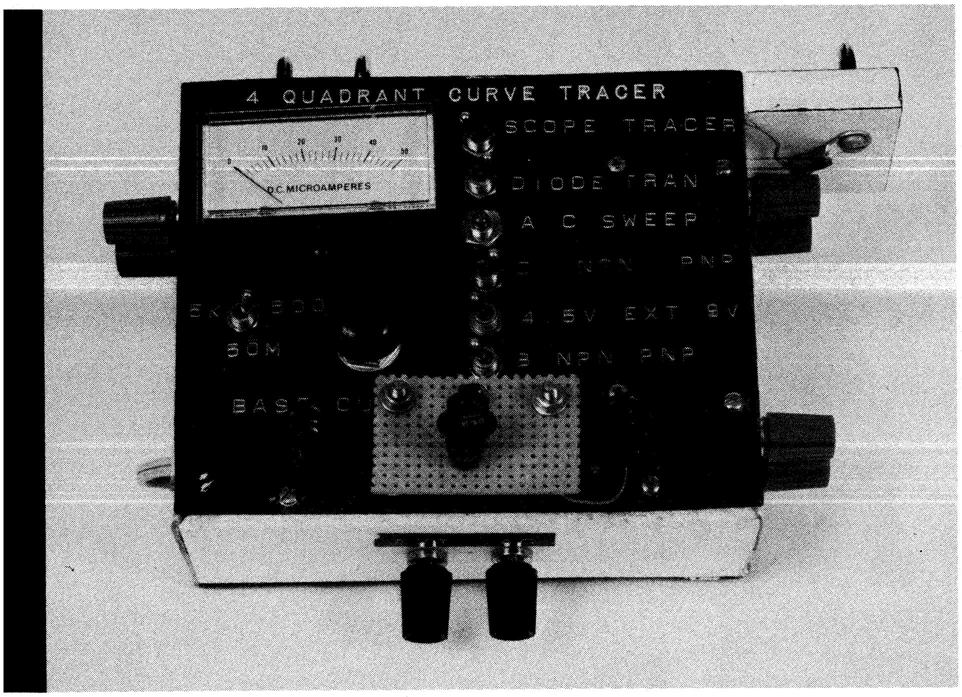
ham radio



four-quadrant semiconductor curve tracer/analyzer

It's not just
a transistor tester —
it's a versatile instrument
that can be used
for checking and
designing electronic circuits
under static
and dynamic conditions

The test equipment used in building and testing electronic circuits is still one of the most interesting parts of Amateur Radio. With the increased use of semiconductors, much more data is needed for their use and replacement than can be obtained from simple transistor testers. This is why I felt it necessary to build the instrument described here.

features

The semiconductor curve tracer/analyzer is as versatile as your imagination yet is economical and simple to build. It can be used for checking as well as designing electronic circuits under both static and dynamic conditions. It can also be used to determine parameters of signal and power transistors, unijunction transistors, field-effect transistors, silicon-controlled rectifiers, and triacs.

Most diodes can be analyzed, including signal and power devices, zeners, protection diodes, bias diodes, point-contact diodes, hot-carrier diodes, and light-emitting diodes.

By Stuart Tuma, W1QXS, 17 Briggs Street, Melrose, Massachusetts 02176

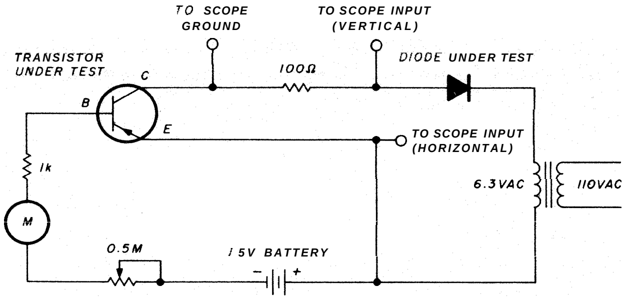


fig. 1. Basic circuit of the curve-tracerlanalyzer.

Another feature of the instrument is that of checking photocells made of cadmium sulfide or cadmium selenide. With appropriate adapters, the instrument can also be used to check integrated circuits.

The analyzer is not restricted only to seminconductor devices. It can also be used to check the piezoelectric effect of quartz crystals under various circuit conditions and to check the design of amplifier as well as oscillator circuits. In electrical circuits, the analyzer can be used to check the sensitivity and internal resistance of D'Arsonval meters and galvanometers as well as the sensitivity of relays, including the popular reed relay. With proper adapters, low-power vacuum tubes can also be checked.

theory of operation

As a transistor curve tracer, the unit is designed so that the oscilloscope vertical input measures the voltage across a 100-ohm resistor to ground, which is used to measure collector current. The oscilloscope,

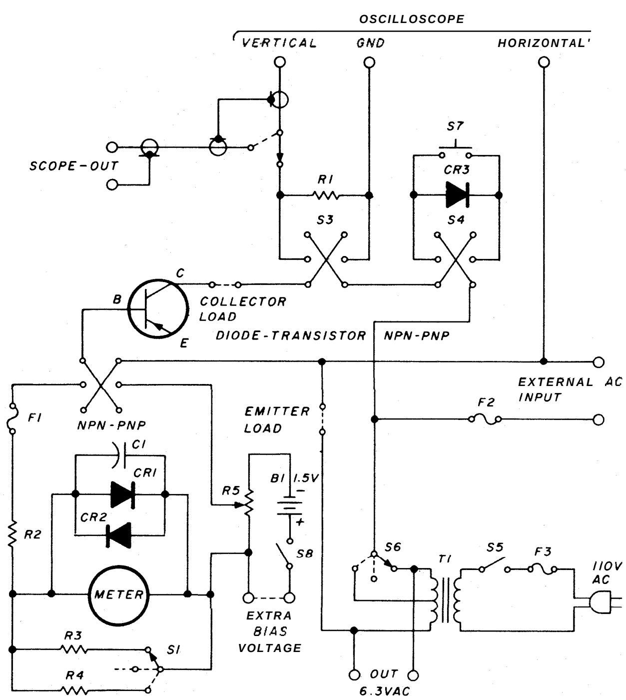


fig. 2. Schematic of the four-quadrant curve-tracerlanalyzer.

having a vertical sensitivity of 0.1 volt per division across 100 ohms, gives 1 mA per division. Should you desire to increase the current per division, you can use 1 volt per division, thus giving 10 mA per division, and so on. **Fig. 1** shows the basic circuit.

The oscilloscope horizontal deflection is used to measure collector-emitter voltage. The oscilloscope is calibrated to read 1 volt per division. Since the horizontal amplifier input is not directly calibrated, it will be necessary to use the sweep voltage, which is approximately 9 volts peak pulsating direct current. This gives a value of 1 volt per division horizontal de-

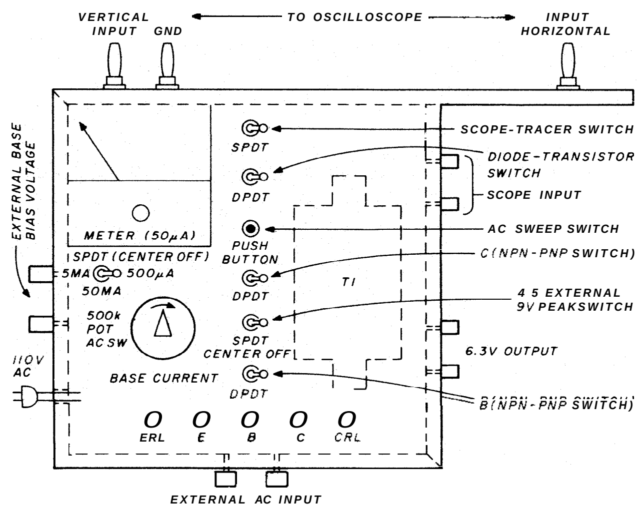


fig. 3. Sketch of front panel showing parts layout.

flection for 9 divisions. Now we have a method of checking both voltage and current from our 60-Hz sweep signal supplied by the 6.3 VAC source.

The emitter-base circuit has a separate supply — a 1.5-volt battery for the base bias. Higher bias voltage can be added to the emitter and collector circuit if desired. The base-current circuit employs a 50-microampere meter movement (which has an internal resistance of 1000 ohms). This circuit will read 50 microamperes (no shunt), 0.55 milliamperes with a 100-ohm shunt, and 5.05 milliamperes with a 10-ohm shunt.

The meter is protected by two silicon diodes **(fig. 2)**, which I found to have a forward-bias-voltage drop of 0.4 volt. The voltage drop across a 50-microampere meter, full scale, having a resistance of 1000 ohms, should be 0.05 volt. This gives good protection for the meter.

Most silicon diodes have about 0.6 volt forward bias, so other types of diodes could be used. However, check the diode's forward-bias voltage before you install it. You can do this by connecting the diode in series with a 1000-ohm resistor and a **1.5-** volt battery. Measure the voltage drop across the

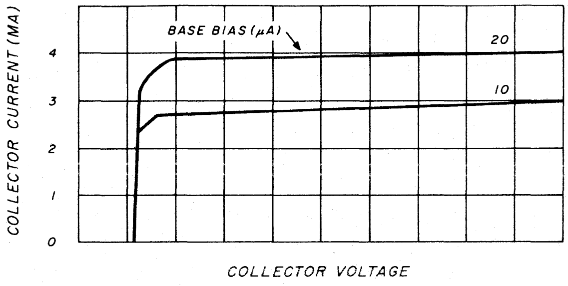


fig. **4.** Typical curve for a pnp transistor (type **2N252**) showing the relationship between base bias and collector current and voltage.

diode when the diode is conducting. This voltage should be 0.6 volt or less.

A 0.001-μF capacitor is connected across the meter for still more protection from rf coming through the line or from some other external source. The meter circuit is further protected by a 1/4-ampere fuse.

The base-emitter current of the transistor under test is controlled by a 10-k pot (R5, fig. **2**) and a voltage divider (R3, R4) in series with the 1000-ohm resistor (R2) and the meter. This gives a maximum current of about 1.5 mA.

construction

A schematic of the curve tracer/analyzer is shown in fig. **2.** A parts list is given in table **1**. The sketch of fig. 3 shows parts layout on the front panel.

wiring. Wiring was easy. I used a pencil-type soldering iron, a good grade of solder, and a clean, tinned soldering tip. The circuit was connected with 0.4-mm (no. 26 AWG) hookup wire. The reversing (inverting) switches were wired first. External connecting leads were added for future wiring. All components were mounted on the top panel and wired as shown in fig. 2. The extended leads were then soldered to the proper binding posts. I used one strand of 0.08-mm (AWG no. 40) wire for the 1-ampere fuse on the external binding post. I used shielded wire for the oscilloscope tracer switch and scope output.

The five terminal posts on the top front panel were made for plug-in adapters on which you can mount various "TO" sockets for transistors or a module-type socket for testing other devices. Note that different load resistances can be added to the emitter circuit as well as to the collector circuit. This allows you to build prototype circuits before putting them into a breadboard circuit. If desired, a solderless breadboard adapter can be used.

mechanical details. The top and bottom panels were made of two pieces of plastic 152 mm (6 inches) square. The sides were made from 6.4-mm (1/4-inch) plastic channel molding, 51 mm (2 inches) wide, obtained from a local lumberyard dealer. These pieces were cut to form the four sides. I used small metal screws to put it together. Parts were laid out in a convenient order. (Mark or etch parts locations on the top and side panels.)

The top panel required a 38-mm (1-1/2-inch) diameter hole to mount the meter. I found that an old pencil soldering iron was just the thing for this, since the plastic melts at a very low temperature. I used a pipe reamer for the finishing touches. I made a 9.5-mm (318-inch) hole for the potentiometer. I used a

table **1. Curve-tracer/analyzer** parts list.

component	description	approximate **cost**	source
B1	1.5 volt battery type AA	$0.20	Lafayette Radio
banana plugs	(screw mounting) M3.5 (6-32) (pkg of 10)	3.40	Lafayette Radio
binding posts	5 way (pkg of 6)	1.69	Lafayette Radio
C1	.001 μF 1000V ceramic	0.15	Lafayette Radio
CR1, CR2, CR3	1A 600 PIV silicon diodes (pkg of 3)	1.19	Lafayette Radio
F1, F3	114-A 3AG fuses (pkg of 5)	1.05	Lafayette Radio
F2	see text		
meter	50 μA (99PS1146V)	6.95	Lafayette Radio
R1	100 ohm 1W 10% composition	0.20	Lafayette Radio
R2	1000 ohm 1/2W 10% composition	0.15	Latayette Radio
R3	100 ohm 1/2W 10% composition	0.15	Lafayette Radio
R4	10 ohm 1/2W 10% composition	0.15	Lafayette Radio
R5, S5, S8	500-k pot with two SPST switches	2.09	Lafayette Radio
S1, S6	SPDT 3A 125V mini toggle switches (center off)	1.39 ea.	Poly Paks
S2, S3, S4	DPDT 3A 125V mini toggle switches	1.95 ea.	Poly Paks
S7	SPST momentary mini switch	0.79	Lafayette Radio
T1	6.3V ct 1A or equivalent	3.75	Lafayette Radio
		$25.25	

smaller reamer for the finishing touches. I used the same reamer for the 6.4-mm 1114-inch) holes for mounting the switches.

I painted the inside of the top and bottom panels flat black. I used 5-mm (3116-inch) holes to mount the binding-post terminals. The terminal posts were spaced 19 mm (314 inch) apart. I constructed the instrument so it would plug into my Conard oscilloscope. However, with proper external leads, this analyzer should fit into any standard oscilloscope.

There are probably a thousand and one uses for this instrument. I've listed only a few, but enough so you'll become familiar with its use, both as a curve tracer and analyzer. I'm sure you'll find other uses, and I'd like to hear from you in this regard.

testing transistors

Plug the curve tracer into the proper oscilloscope inputs. The vertical output goes to the oscilloscope vertical input, and the horizontal output goes to the oscilloscope horizontal input. Ground the curve tracer to the oscilloscope ground.

pnp transistors. Set up the oscilloscope as follows:

quantity	setting	measurement
vertical gain	0.1 V/division	1 mA/division
horizontal gain	9 divisions (with transistor in circuit)	approximately 1 V/division (peak)
horizontal sweep source	external	
intensity	normal	
focus	normal	

Set up the curve tracer as follows (see **fig. 3**):

switch positions	switch to
scope-tracer	tracer
diode-trans	trans
AC sweep	pushbutton released
C NPN-PNP	PNP
B NPN-PNP	PNP
4-1/2V-EXT-9V	9V (peak)

Switch meter to 50 microamperes or to a convenient current rating. Set **BASE CURRENT** counter clockwise (CCW). Be sure that jumpers are connected between emitter-to-emitter-load and between collector-to-collector load terminal posts **(fig. 2).**

Connect the test transistor to the proper input terminals or into a transistor plug-in adapter. Turn the **BASE CURRENT** control until the current begins to increase. Note the variation of the trace. Increase the trace two or three divisions and note the increase in the base-bias current (record this reading). Increase the collector current until one more division is obtained. Again, record the base-bias reading. From the data taken, the transistor beta and alpha gain can be determined as shown below.

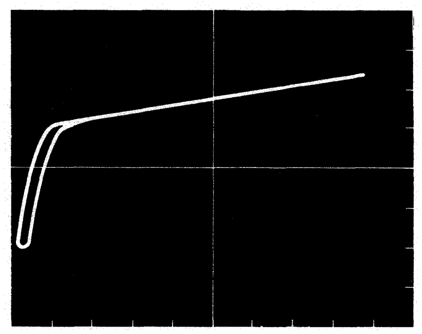
The beta gain is determined by taking the variation of the collector current and dividing it by the variation of the base bias current. Example:

beta gain = Acollector current/ Abase bias current
beta gain = Δ1 mA/ A10 microamperes = 100

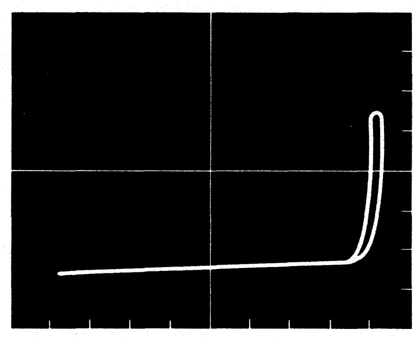
The alpha gain is equal to the beta gain divided by the beta gain plus 1. Example:

alpha gain = beta gain/beta gain + 1
alpha gain = 100/700 + 1 = 0.99

Fig. 4 shows the relationship of collector current, collector voltage, and base bias for a typical pnp transistor (type 2N252).



NPN transistors. The setup for npn transistors is the same as for pnp transistors, except that the emitter and base **NPN** and **PNP** switches are both set to the **NPN** position. The curve on the oscilloscope may have to be recentered. The beta gain can be determined as with a pnp, except that the curve will be in the opposite direction as shown below.



testing diodes

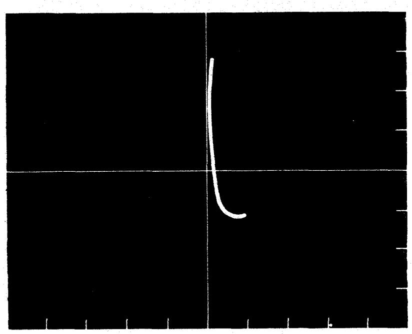
Set up the oscilloscope as follows:

quantity	setting	measurement
vertical gain	**0.1** V/division	1 mA/division
horizontal gain	9 divisions (with diode in circuit)	approximately 1 V/division (peak)
horizontal sweep source	external	

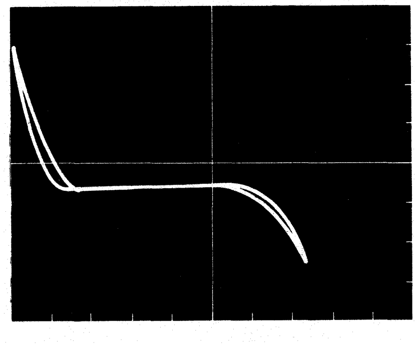
Set up the curve tracer as follows:

switch position	switch to
scope-tracer	tracer
diodetrans	diode
AC sweep	pushbutton released
C NPN-PNP	NPN
B NPN-PNP	not in circuit
4-1/2V-EXT-9V	9V (peak)
base-current meter shunt	not in circuit
base-current control	not in circuit

Connect the diode cathode to terminal post **E** and the diode anode to terminal post **C.** Note the **L-**shaped pattern. To give better diode action press AC SWEEP. This shows when the diode is not conducting.



For checking 5-volt zeners, install a 3900-ohm resistor in series with the emitter load and emitter terminal post. This allows about 10 milliamperes of current flow through the zener. (For checking other types of zeners, a different load resistor must be employed.) Connect the diode cathode to the emitter terminal and the diode anode to the collector terminal post. The oscilloscope and curve analyzer are set up exactly as in the diode setup. For higher-voltage zeners, EXT input can be used with a Variac or variable AC voltage source.



testing photocells

Photocells made from cadmium sulfide or cadmium selenide working on the principle of photo conductivity can be checked with the analyzer by using the base-bias circuit. The photocell has only two leads and can be attached to **E** and B on the curve analyzer. As the light increases on the photocell, the circuit conductivity also increases. Using the circuit as shown, the conductivity of the photo cell can be obtained.

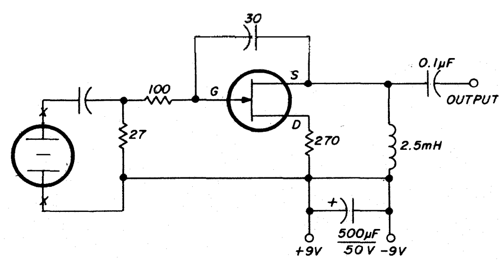
This circuit can also be used as a photographic light meter.

testing D'Arsonval meters

This circuit can measure characteristics of meters with a sensitivity of 10 microamperes to 5 milliamperes full scale. This is done by connecting the meter under test to terminal posts **E** and B **(fig.** 3). Increase the current flow with the **BASE CURRENT** control until the meter reads full scale. Note the sensitivity of the unknown meter in amperes. To determine the internal resistance, connect a decade box or a 500-ohm pot across the meter under test. When the shunt resistance decreases the full-scale reading to one-half scale, measure the resistance of the 500-ohm pot to the center arm . This should give the meter's internal resistance.

crystal-oscillator checker

The crystal-oscillator checker uses an N-type junction fet (RS2035) field-effect transistor, which can be obtained for about one dollar. An equivalent type could be used. The module is wired as shown. **It** is installed into the curve tracer across the **E** and **C** terminals. Using a 50-μF filter across the supply input to the module as shown, the dc input is about 9 volts. The circuit works very well for checking quartz crystals as low as 100 kHz to as high as 15 MHz.



Set the oscilloscope as follows:

quantity	measurement
Vertical gain	1V/division
Horizontal sweep	1 mS/division
Horizontal mode	internal

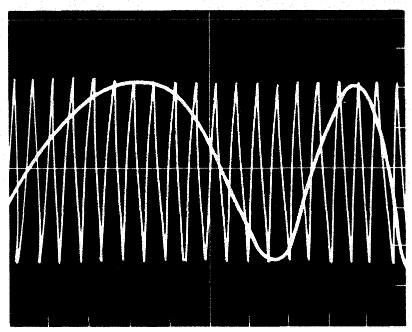
Set all other controls to normal position.

Set the curve tracer as follows:

switch position	switch to
scope-tracer	scope
diode-trans	not in circuit
AC sweep	pushbutton released
C NPN-PNP	not in circuit

switch position	switch to:
B NPN-PNP	not in circuit
4-1/2V-EXT-9V	9V (peak)

Connect the crystal to crystal terminal on the module. The module output should be connected to the



oscilloscope input. The waveform can be observed on the oscilloscope. For a more accurate frequency reading, a frequency counter should be used.

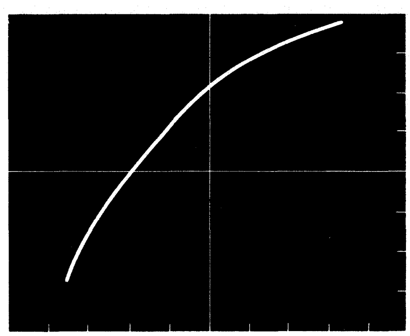
some other uses

The curve tracer/analyzer can be used for checking other devices. Presented below are the results of some tests I've run on junction fets, unijunction transistors, silicon-controlled rectifiers, and triacs. The setup instructions for the scope and analyzer are as in the previous examples.

junction fets

oscilloscope:

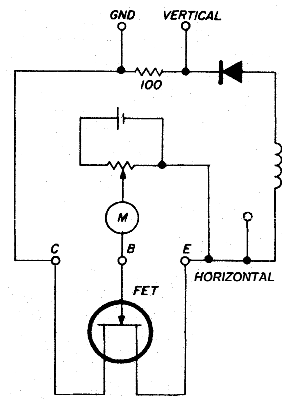
quantity	position	measurement
vertical	1V/division	1 mA/division
horizontal	1V/division	9 divisions
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton released
C NPN-PNP	NPN
4-1/2V EXT 9V	9V
base-current	5 mA
meter shunt	
base control	adjust to proper gate voltage *vs* source current

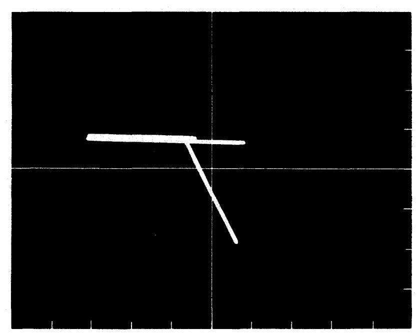
Connect fet as shown below. Use externai voltmeter to measure gate voltage *vs* source current.



unijunction transistors

oscilloscope:

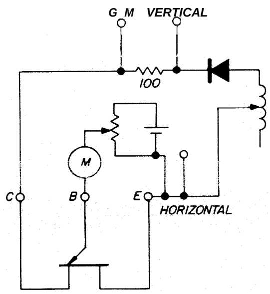
quantity	position	measurement
vertical	0.1V/division	1 mA/division
horizontal	1V/division	4-1/2 divisions
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton released
C NPN-PNP	NPN
4-1/2V EXT 9V	4-1/2V
B NPN-PNP	NPN
base-current	5 mA
meter shunt	
base control	adjust for gate and B1 voltage with external meter

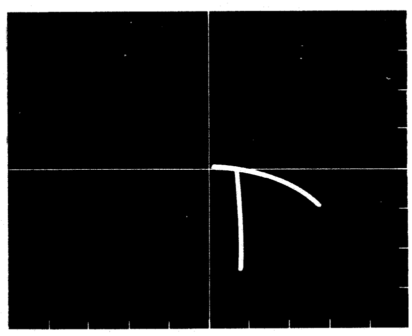
Connect unijunction device to curve tracer as shown using a 0-5-volt dc meter. Measure trigger voltage between gate and B1.



silicon-controlled rectifiers

oscilloscope:

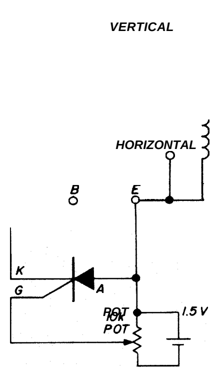
quantity	position	measurement
vertical	1V/division	10 mA/division
horizontal	3V/division	9V
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton (AC)
C NPN-PNP	PNP
B NPN-PNP	not in circuit
4-1/2V EXT 9V	9V

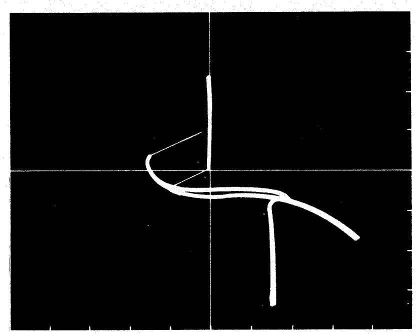
Connect scr as shown in circuit below; anode to (E) and cathode to (C). Use a 1.5-volt battery in series with a 10k-ohm variable resistor as a variable voltage source for gate and anode. With a separate voltmeter measure the gate-anode trigger voltage. Note current flow through the scr.



triac-controlled rectifiers

oscilloscope:

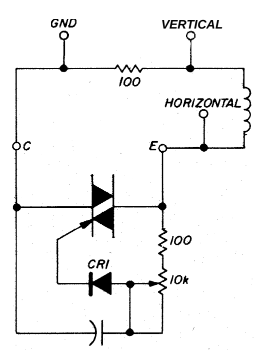
quantity	position	measurement
vertical	1V/division	10 mA/division
horizontal	3V/division	9V
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton (AC)
C NPN-PNP	PNP
B NPN-PNP	not in circuit
4-1/2V EXT 9V	9V

Connect as shown in circuit below using external R-C network. Measure gate voltage between emitter terminal and gate. Note ac current flow.



final remarks

I've presented the results of my work in trying to improve the lot of the home builder who likes to work with semiconductors. No doubt you'll come up with other uses for the basic instrument, and I'd like to hear from you. If you have any suggestions or questions, please send them to me in a self-addressed, stamped envelope, and I'll be glad to reply.

ham radio

the ultimate noise blanker

Conventional noise-blanker designs overlook the effectiveness of the blanking switch — here's a new approach to the problem using fm techniques

Noise blankers are commonly incorporated into hf communications receivers but use less than a perfect switch and switch-control timing. This article presents a new concept for a noiseless i-f switch that allows a very effective impulse noise blanker to be constructed. The fundamental concept of noise blankers is also reviewed.

conventional noise-blanker design

All noise blankers operate on the principle that a separate noise receiver listens in a no-signal portion of the spectrum (typically 30-35 MHz) especially for the purpose of receiving noise pulses. The detected pulses are processed and applied to a blanking switch in the communications receiver i-f strip to momentarily interrupt the signal path during the noise pulse period.

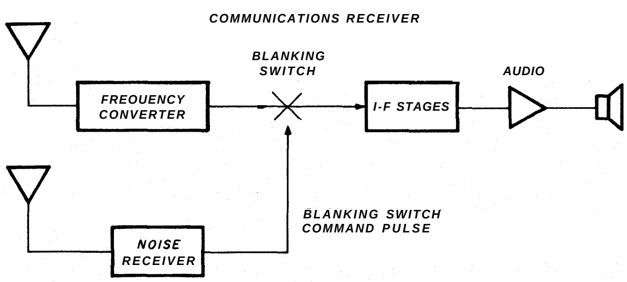


fig. 1. Block diagram of a typical noise blanker used in communications receivers.

design considerations

Typically, the noise pulse has a duration of only a few microseconds. If allowed to pass through the communications receiver, several factors cause the pulse to be stretched, including narrow filters and saturation. If the blanker operates properly, the noise pulse is removed with negligible effect on the communications signal.

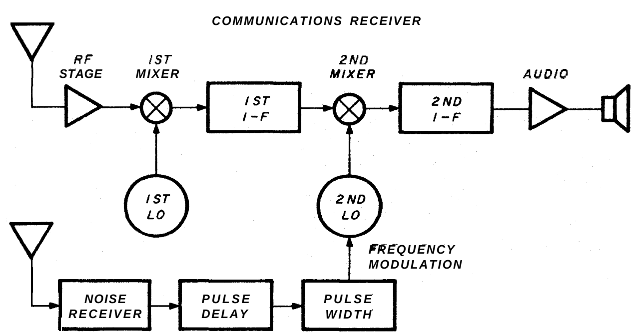


fig. 2. Block diagram of an improved noise-blanking system using a frequency-shift network ahead of the second local oscillator in a dual-conversion receiver.

The major limitation in previous designs has been the effectiveness of the blanking switch itself. Most have a limited on/off ratio and introduce switch-transient noise as well. A typical blanking system block diagram is shown in **fig. 1.**

For proper operation, timing is an important factor. The blanking switch should open the signal path for the pulse period only. If the switch is open too long, unnecessary distortion of the signal occurs. If the switch is not open during the complete pulse period, a portion of the noise pulse will leak through.

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This aspect does not receive proper attention in typical designs.

The design of the noise receiver must include adequate bandwidth to ensure proper processing of narrow pulses to enable accurate control of the blanking switch. Also, the time delay through the noise receiver must be sufficiently shorter than the delay through the communications receiver front end to enable pulse shaping and control before application of the control signal to the blanking switch. Delay through the receiver is a function of the i-f bandwidth. In practice, it has been found that the use of standard 10.7 MHz fm i-f transformers in the noise receiver is the best choice. Adequate delay in the communications receiver (ahead of the switch) is normally realized in the first i-f section.

Optimum design of a communications receiver suggests a multipole narrow bandwidth i-f filter immediately following the first mixer. Conversely, a multipole filter has significant time delay and stretches the noise pulse width ahead of the blanking switch. Although this is undesirable from a purely technical viewpoint, the effect is negligible and can be ignored.

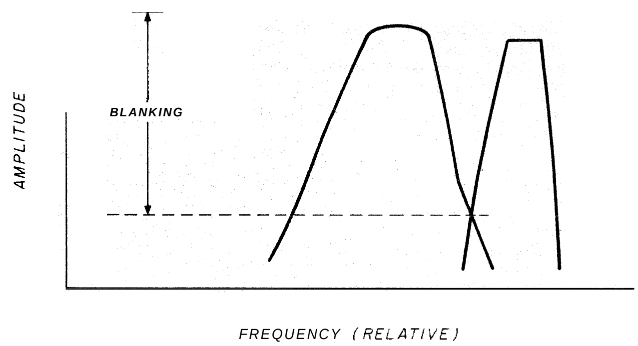

fig. 3. Blanking-switch isolation can be achieved by misaligning the conversion frequency during the blanking period. The amount of frequency shift can be determined by comparing the shape factors of the first and second i-f filters.

If the blanking is located in the receiver after a substantial amount of gain, saturation of the i-f amplifier may result (especially under weak-signal conditions

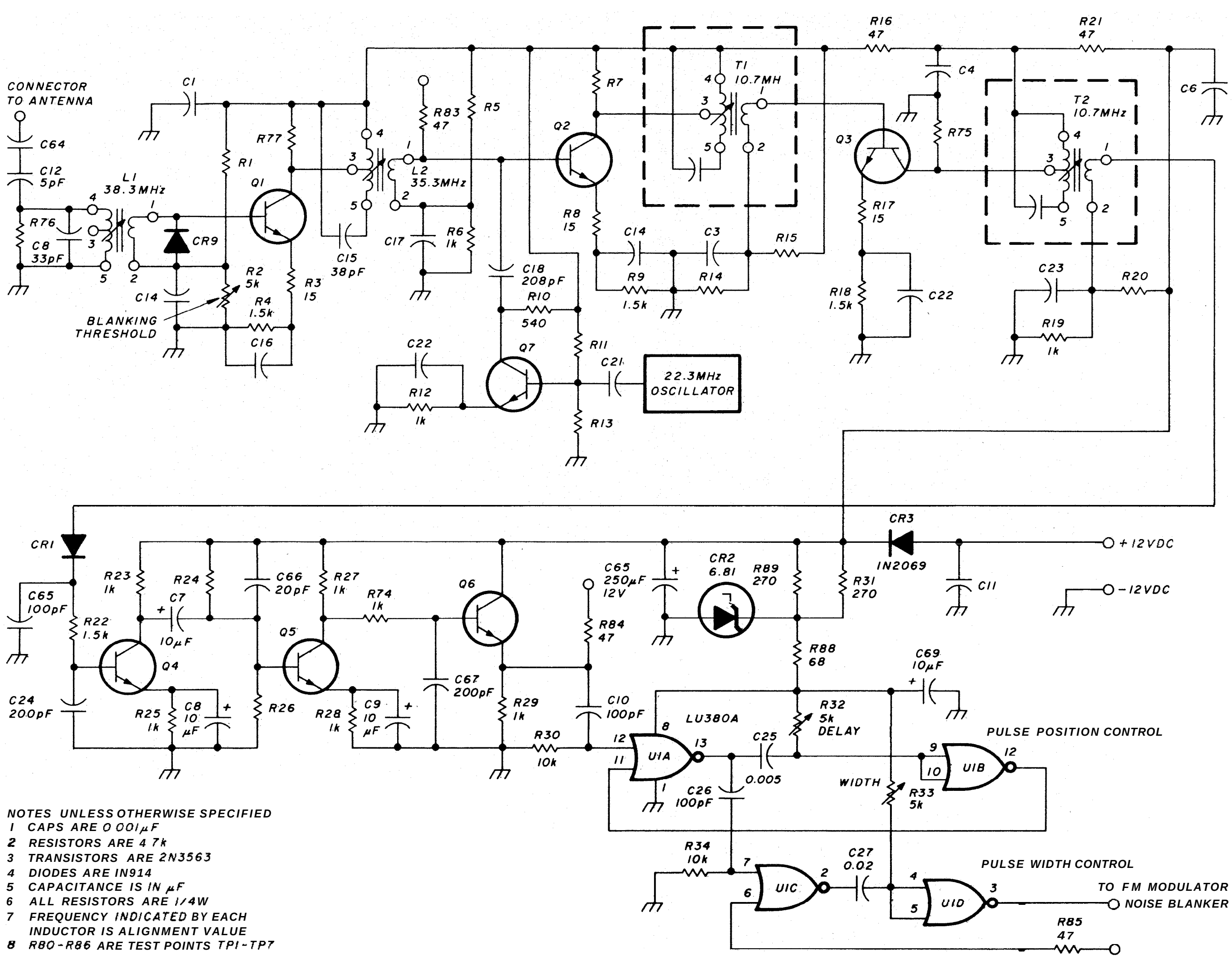

fig. 4. A practical noise-receiver schematic using the technique described for taming switch action.

with no agc). Slow recovery is normally associated with saturation, hence extreme amounts of effective pulse stretching may occur. However, typical i-f blanking switches must operate at relatively high signal levels, since the noise introduced by the switch is proportional to the signal level at which they operate. If the blanking switch is located immediately after the first i-f filter, signal levels will typically be in the microvolt region.

a new switch

Switching transients can be eliminated by using a frequency modulation technique. With reference to **fig. 2,** note that when a command signal from the noise receiver is applied to a frequency-shift network associated with the second local oscillator in a dual-conversion communications receiver, the mixer output will be at a frequency other than that required to transfer the signal from the first i-f to the second i-f. In other words, if the time of the command signal occurs at exactly the time the noise pulse propagates through the first i-f and is applied to the mixer, the noise pulse energy will leave the mixer at a frequency other than that of the second i-f.

Since there is no amplitude noise (switching-transient noise) associated with the frequency shift, no switching noise due to the blanking action will occur when the signal path is momentarily interrupted.

The amount of required frequency shift may be determined by comparing the shape of the first and second i-f filters (see **fig.** 3) and misaligning the conversion frequency during the blanking period to achieve the desired switch isolation. Typically, a few kHz will be adequate. Normally, the second local oscillator can be frequency modulated by tens of kHz with no adverse effects.

In practice, the addition of a small capacitor and switching diode to a receiver second local oscillator implements the desired switch. If the second local oscillator is crystal controlled, it must be replaced with an oscillator that can be frequency modulated.

results

Performance of the fm switch blanking system is astonishing, compared with conventional systems. An hf mobile receiver operating at a busy intersection had no noticeable ignition noise interference with the switch operating. With the switch disabled, communications was impossible.

A complete schematic of a practical noise receiver is included in **fig. 4.** For a particular application, the pulse delay network may require component value changes depending on the delay in the first i-f filter of the communications receiver.

ham radio

power-line noise — the cause and cure

Discussion of the steps necessary to locate power-line noise, starting with in-house noise sources and ending with utility-pole-generated noise

Power line noise can be one of the most exasperating forms of irritation experienced by amateurs living in or near metropolitan areas. This problem, which can drive active operators beyond the point of sanity in record time, is characterized by a long-term arcing-type sound similar to that produced by loose antenna or high-voltage connections. Usually, this noise will be apparent (in varying amounts) on the 80- through 10-meter amateur bands. Many cases of line noise arcing or radiation, however, expand this field of interference to include the frequencies associated with television and the a-m broadcast radio. The magnitude of power-line noise interference varies with each case. Sometimes this "hash noise" can be tolerated, but occasionally it approaches an S-9 level and must be eliminated. Since many amateurs find themselves in an awkward position during such times, this article will present an informal guideline which may be used to help eliminate this electrical plague.

clean house first

Many noise interference situations prove to be created by sources other than commercial power lines. Thus, your own house should be in order. Check all antenna connections and transmission lines, being highly critical of any metallic objects that could come in contact with guy wires or antennas. Remember, too, that rain may cause items like wood or cloth to act as conductors. Next, check the plugs and line cords on equipment in your house, being particularly suspicious of appliances used in the kitchen. For example, bad electrical igniters, as found in many gas ranges, can create a surprising amount of interference on the ham bands. While this type in-

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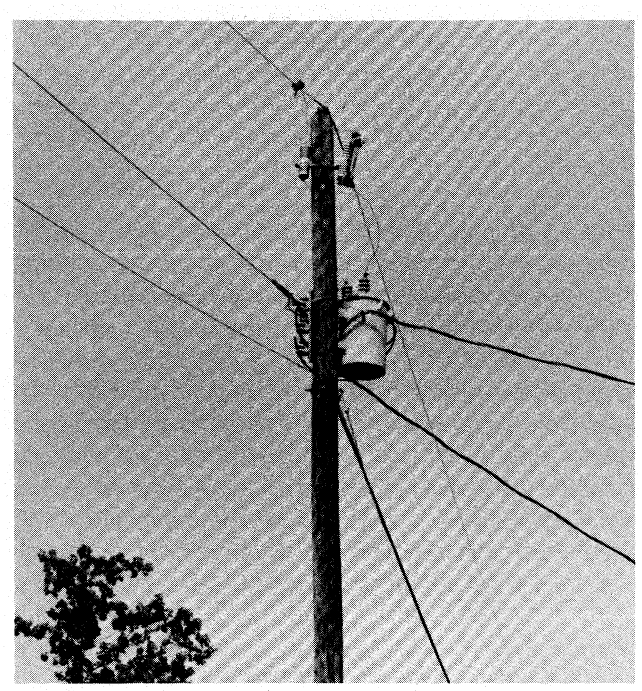
terference can be reduced by installing a bypass on each side of the ac line to ground, that doesn't eliminate the problem. The defective element should be replaced. Likewise, water heater thermostats, electric blankets, and heater tapes (used for wrapping outdoor water lines during winter months) should also be checked by temporarily disconnecting their ac power. Assuming the noise interference still exists, you are now ready for the Sherlock Holmes phase of locating this electrical villain.

locating the villain

If you have a directional antenna, it can be used to determine the approximate direction of your interference. An accurate "fix" on this noise is usually obtained by searching for a null, rather than a peak. Once the approximate direction of the noise has been determined, try mobiling in that area while listening for interference peaks on a portable rig. A two-meter sideband transceiver is particularly convenient for such ventures. Fm handi-talkies should not be used for these tests, since they should be unaffected by a-m (line noise) variations. (While a portable a-m radio could be used for location techniques, its susceptibility to normal radiation from every power line and pole makes its indications very unreliable.) When using this method, you should travel a reasonable distance beyond the point of maximum noise pickup to ensure that you've definitely located the source. While not common, it's quite possible that a remotely generated noise may be propagated along the power lines with a peak occurring right at your door. Using the S meter on your rig for indication, try walking the area of the noise source to pinpoint the interference to a specific pole, transformer, meter box, or home.

Next, study the line noise for several days, and try relating it to various weather and time situations. Noise that is more apparent on warm days than cold days, and disappears during periods of rain, is often caused by loose line clamps or cracked insulators on power poles. Noise that is more apparent during heavy-load evening hours is often caused by leaky transformers or defective heating devices. Thoroughly investigate the area around the apparent source of noise during the day and night, looking for frayed wires swinging in the breeze, small animals that may have become trapped between high tension lines, or a visible arcing near transformers or pole-top mounted insulators. However, don't climb power poles to shake wires, or viciously swing grounding lines to power poles. You could be electrocuted if a loose or corroded connection suddenly broke.

Assuming you have now determined the line-noise producing area and confined it to, say, three or four poles, you're almost ready to notify the local power company. Before doing so, however, reinvestigate the whole area and make notes on which poles have transformers and fuses. These fuses are approximately 51 cm (one foot) long, and swing into mounts near the top of the pole. The local power company



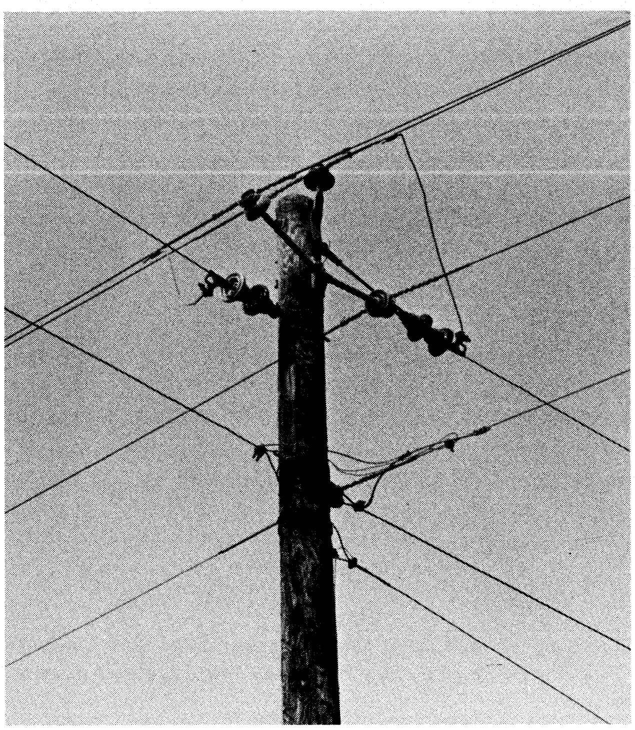
The fuse, which can be used to isolate individual lines, is located at the top of the pole. Note the large clamp on the top wire.

can disconnect these fuses ("drop lines") to locate the discrete points and elements creating interference.

call for help

Your first call to the power company will probably be stopped by the front desk, so, explain your situation, describe the noise-producing area and your detection methods. Leave you name and telephone number so their engineer can contact you before visiting the trouble area. This prescheduled meeting is particularly advantageous for tracking down periodic line-noise problems. There's nothing to be gained by looking for arcing insulators, for example, during or after a mild rain. When the power company's task force arrives, repeat the results of your investigation and monitor, with your mobile or portable gear, as each line is broken or checked. The power company will usually proceed with matters beyond this point, repairing whatever elements are found to be creating interference. Be sure to note the name of the power company employees repairing the line noise, so you can sidestep the front desk should future problems arise.

Many times, amateurs become "trapped" with front-desk executives (who've never heard of Amateur Radio, and know even less about line noise) and can't get line-noise problems resolved. Don't despair. Somewhere in their organization is a communications department – and often one or two radio amateurs. Not only are these people usually sympathetic and understanding, they usually handle special problems like line-noise interference. This group usually has its own array of elaborate noise-locating gear. Often, one or two of these men work exclusively on line-noise problems, because situations that create line noise often also creates problems that lead to power outages. Yes, you may actually be helping your community when reporting line-noise interference to the local power company. Naturally, television reception will improve when "white dots in the picture" disappear, and a-m radio reception is better without its "frying sound." But you also illustrate that neighborhood amateurs can help eliminate, rather than create, interference.



Photograph of clamps and wires which are prime sources of line-noise interference. The aluminum clamps, which hold the jumper wires, can expand and contract with temperature variations. The large insulators may crack, causing "frying" noise interference.

Many noise interference problems, when located, prove to be created by customer-use devices like doorbell transformers and thermostats. Fortunately, most power companies are quite helpful and understanding in locating these troubles. Their step-by-step method for locating these noise sources usually involves sequentially "dropping" various lines and evaluating the effect on the noise. As previously mentioned, a battery-powered, high-frequency (or 2-meter) rig is highly beneficial during these tests.

The danger of experimentation with high-tension power lines by unauthorized personnel (that's you!) cannot be overemphasized. Today's amateurs have learned to respect high voltage in the shack; the number one killer is now outdoor power lines and large antennas, Instant retirement results when these two items are placed together. Don't take chances. Stay clear of problem areas during tests, and don't stand where lines could accidentally break and fall on you.

Here is a synopsis of the more common power-line noises and their causes:

1. Constant noise during the heat of the day, intermittent noise at night, no noise during rain: Cracked insulators or loose clamps on pole. This noise is predominant during summer months.

2. Constant noise during rain, intermittent noise during light mist: Small object or animal caught between high voltage lines.

3. Constant noise in a specific direction: Power transformers or home-user devices.

checklist for line-noise corrections

1. Check your own home area by disconnecting the main circuit breaker while monitoring noise on a battery-powered rig. If noise stops, reconnect the main breaker and switch off power to each area of your home until the noise again disappears. Then pinpoint source.

2. Use directional antennas and portable/mobile gear to isolate the noise-producing area.

3. Relate noise interference to time and weather conditions.

4. Call the power company and arrange to be at home when they disconnect service to various areas.

5. As confirmation of power-line (as opposed to atmospheric) noise, the power company may elect to temporarily cut all power to an area. Monitor the results on your mobile gear.

6. Be patient and persistent. A quiet band is worth the wait and effort.

ham radio

gate structure and logic families

The first part of this series explained the basic gate as a component. This part will examine the internal gate structure of TTL and CMOS, similarities and differences, as well as loading. A typical TTL two-input NAND gate is shown in **fig. 1.**

Multiple emitters may seem strange, but they are easily made in integrated circuits. They are the key to NAND-gate operation. When any emitter of 01 is pulled down to less than +0.4 volts, its emitter-base junction is forward biased via R1. Collector-emitter junction voltage is then low enough to cut off Q2 and 04. Output voltage goes high, since 03 is conducting by forward bias from R2. R3 limits output current.

Q1 is nearly cut off when all emitters are at +2.4 volts or higher. Full cutoff is not achieved, since the internal structure is arranged to forward bias the base-collector junction; this allows Q2 to conduct by the forward bias through R1. 03 cuts off and 04 will conduct because of the Darlington connection to Q2. The collector-emitter junction of 04 will then saturate at less than 0.4 volts.

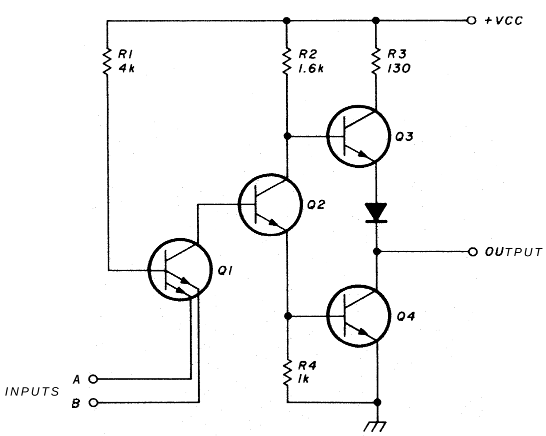


fig. 1. Simplified schematic diagram of the two-input, TTL NAND gate. In the open-collector version, R2, R3, Q3, and the diode are omitted, leaving the collector of Q4 open.

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TTL outputs will *sink* more current at low output levels (electron flow from ground to output) than they can *source* at high outputs (flow from output to supply). This fits the input requirements for devices connected to the output, Input current at high levels is one-fortieth of the low level current. Output current biasing is set for this 40:1 ratio and *fan-out* is usually set for driving ten inputs with one output.

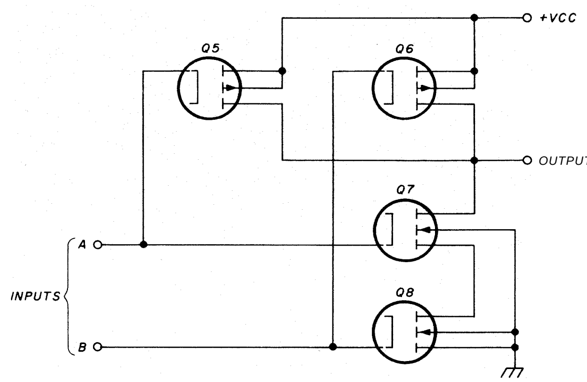


fig. 2. Simplified internal structure of the two-input, CMOS **NAND gate.**

Fig. 2 shows the equivalent gate function in a CMOS device. The insulated-gate FETs reduce input current to nanoamperes, primarily from leakage. Gate resistors are seldom internally used, allowing all inputs to present a high input impedance.

The N- and P-type FET arrangement in a CMOS gate will determine function. If one input is low, the appropriate parallel-connected P-type (Q5 or 06) will conduct and make the output high. The low input overrides any high input by cutting off one of the series-connected N-types (Q7 or 08). ***All*** inputs must be high for the series path to ground; the NAND RULE is satisfied.

Output sink and source current capability is nearly the same at either level; it varies with supply voltage. A-Series CMOS will work at 3 to 15 volt supplies; B-Series goes to 18 volts. The output FETs do not fully saturate, so exact voltage level is load dependent. Data sheets must be consulted for exact values.

Rearrangement of N- and P-types allows a greater range of functions than with TTL. Inverting the series and parallel connections will form a NOR. Other possibilities exist, and CMOS devices are available with a greater range of functions.

important differences

The high input impedance of the CMOS implies the capability of driving hundreds of inputs from one input. This would be true if it were not for package and circuit capacitance. Capacitance and output FET characteristics limit loading. Capacitance also limits TTL, but to a lesser extent. Input current at dc limits TTL.

Input and output level symmetry allows CMOS to be biased for linear, small-signal operation. This is impossible with TTL. A *Schmitt trigger* must be used with TTL for inputs having slow rise and fall times.

Input threshold voltages are fixed in TTL. This is primarily due to saturated bipolar transistor operation from a fixed supply. CMOS level thresholds are approximately one quarter of supply maximum for a low and three quarters of supply minimum for a high.

CMOS devices are more susceptible to noise pickup due to their high input impedance. Circuit layouts must be carefully done. TTL is more tolerant to noise, but high input levels are still affected; this can be seen from typical TTL characteristics given in **table 1.**

Total circuit current of CMOS is less than TTL. CMOS current demand is influenced by switching speed, *i.e.*, charging and discharge of the circuit capacitance. Supply current increases with increasing logic switching frequency. It is also true for TTL but to a lesser extent. TTL input current masks most of that effect.

Bipolar transistors in TTL allow faster switching speeds. FETs are becoming faster as discrete devices, but CMOS designs retain low power and slower transistors. CMOS is good for portable battery-powered equipment, but note that most such equipment is slow speed.

The NAND gate output of **fig. 1** is called a totem-pole, from the appearance of Q3 on top of Q4. Most outputs are of the totem-pole variety, but a few are *open-collector.* Since, in normal operation, Q3 conducts little current, this transistor, R2, R3, and the diode can be deleted, using an external *pull-up* resistor to the supply line for source current. This is the open-collector version.

open-collector applications

TTL chip transistors have relatively low breakdown voltages. Redesigning just the output transistor into an open-collector chip (Q4 in **fig. 1**) allows driving high-voltage devices such as relays and neon lamps.

table **1.** Typical **TTL** gate characteristics.

	conventional			schottky	
	medium speed	high speed **(H)**	low speed **(L)**	low **(LS)**	medium **(S)**
maximum output source current in microamperes (I_{OH}) when output is high*	400	500	200	400	1000
maximum input sink current in milliamperes (I_{OL}) when output is low*	16	20	3.6	8	20
maximum t_{pLH}, nanoseconds	22	10	60	15	4.5
maximum t_{pHL}, nanoseconds	15	10	60	15	5
maximum input source current in microamperes when input high*	40	50	10	20	50
maximum input sink current in milliamperes when input low*	1.6	2	0.18	0.4	2
nominal input pull-up resistor, kilohms	3.9	2.7	39	18	2.7

*High level output voltage minimum is 2.4 for conventional TTL, 2.7 for Schottky (V_{OH}). Low level output voltage maximum (V_{OL}) is 0.4 for conventional, 0.5 for Schottky. High-level minimums and low-level maximums apply to inputs also.

"Actual high-level circuit voltage may be any value between minimum and the supply voltage, depending on output loading.

Another function is the *wired-OR* depicted in **fig.** 3. This takes advantage of the NAND's active-low input. Several open-collector outputs can be wired to a single input, giving an equivalent OR function. The OR symbol with connecting dot is purely symbolic. No specific open-collector symbol exists.

Why use a wired-OR? One reason is economy, another the number of available packages per board. If you don't have room for a NAND equivalent OR gate, but do have an input, the wired-OR will do the job. Open-collector outputs are slightly slower than totem-poles, so be wary in high-speed circuits.

three-state wired-OR

A modification of the totem pole output was designed by National Semiconductor, called TRI-STATE by them and *three-state* by others. The design has the good points of both totem pole speed and the ability to wire-OR.

A control line is added to enable the output or outputs. An enabled output behaves like the totem pole. A disabled output appears as a high impedance. Three-state outputs can be wired in parallel and are

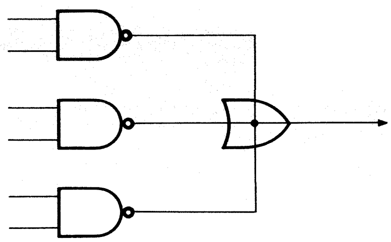


fig. 3. Symbolism used to denote a wired-OR configuration using open-collector NAND gates.

fast, but need the extra control. Such outputs are ideal for computer bus lines. Bus lines are parallel data lines carrying information in either direction.

Both three-state and open-collector output devices require some precaution during circuit design. Data books should be consulted for the fine details.*

unused inputs

All inputs must be connected somewhere. An unused TTL input will automatically assume a high state if left open (leakage in Q1 of **fig. 1**). It is poor practice to tie it high by an open connection; impedance is higher and noise might sneak in. Tie the unused input to V_{cc} or ground, depending on the desired function. Gate inputs can be paralleled. A parallel connection increases logic-I current, but logic-0 current is the same.

Unconnected CMOS inputs can assume either state due to the inherently high impedance, leakage, and construction. They *must* be tied to either V_{cc} or ground. Removable circuit boards should have potentially open inputs connected to ground through a 100-270k resistor. This trick will protect the input from possible static surges when disconnected.

CMOS handling precautions

Unmounted CMOS devices may be damaged by static electricity; an unfelt static charge is enough. Unused devices should be kept in anti-static containers or pushed into conductive plastic foam. Such foam is usually black, somewhat hard, and will read a few kilohms on an ohmmeter.

'The Texas Instruments TTL Data Book is available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048, for $4.95 plus $1.00postage.

Everything on the workbench should be grounded, including an aluminum work-area plate. Use a *grounded-tip, three-wire* soldering iron. Strap together and ground all powered test equipment; this should be done anyway, since it is possible to get a lethal shock from some test gear.

Avoid all plastic-fiber clothing if possible. Ground yourself through a one megohm resistor (two megs with 220-volt mains) and flimsy wire. The resistor is a precaution against lethal shock current. The flimsy wire should break if you slip. All this may seem overcautious, but is standard industrial practice. Ask yourself how many expensive CMOS devices you can afford to lose from a static zap.

TTL variations

There are five different versions of TTL, identifiable by one or two letters after the 54 or 74. No letter means the original and is called medium speed. An L identifies low power and low speed. An H stands for high speed and high power. However, these two versions are being phased out of new designs in favor of Schottky versions.

LS, or low-power Schottky, is as fast as medium speed, with only slightly more power demand than L versions. S is as fast or faster than H and takes no more power than medium speed. Differences are internal and you need only consult data sheets for application. Schottky inputs are to bases of grounded-emitter input stages. This and added clamping diodes make them different from ordinary TTLs.

Schottky versions are recommended for new designs. A lot of ordinary TTLs are available at low prices and should work just well, except in very fast circuits.

I^2L, or integrated-injection-logic is similar to TTL and is circuit compatible. It is appearing in many large-scale integrated circuits.

PMOS and NMOS

This family group started before CMOS. PMOS was the first, using P-type MOS (usually junction FETs). PMOS devices need two or three supply voltages, but have inputs and outputs compatible with CMOS and most TTL devices. They find use as microprocessor memory chips; CMOS handling precautions should be observed.

NMOS uses N-type MOS and most devices use only 5-volt supplies. Again, the inputs and outputs are compatible with both CMOS and TTL. The Motorola 6800 microprocessor devices are almost entirely NMOS. NMOS is less sensitive to static shock, but it won't hurt to be cautious.

The next article in this series will go into the importance of propagation delay and discuss flip-flops.

ham radio

the weekender

combination field strength meter and volt-ohmmeter

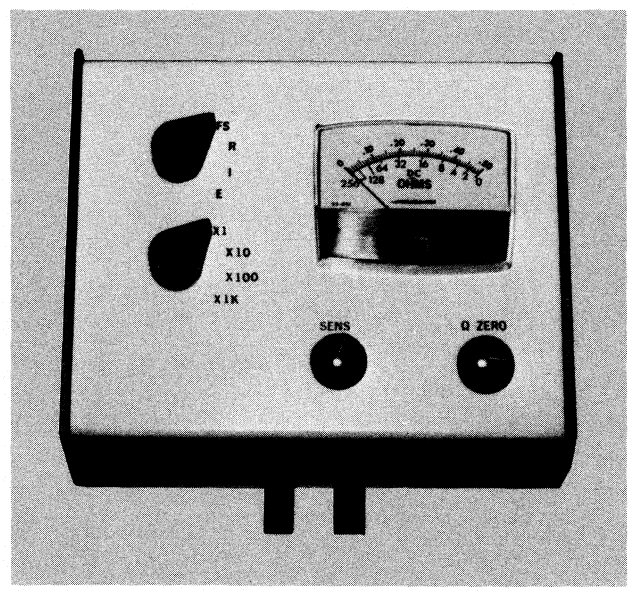
Whether you want a weekend project or would like to get your club involved, here's a project that can be used to teach theory, construction practice, and troubleshooting while providing a basic piece of test equipment. The FSVOM is a combination field strength meter and volt-ohmmeter. It evolved as part of a training program for prospective amateurs. The objective was to make learning fun by providing a balance of theory and practice. My hope was that the finished product would provide a degree of confidence and encourage home construction, rather than being something that would end up on the shelf, like the old "two-transistor super" often used as a first project. In this respect, the FSVOM has been a success.

I don't think anyone need have any qualms about building this unit; there are no exotic or hard-to-find parts. Calibration is minimal. If you don't have facilities for etching the printed circuit board, an etched and drilled board (see parts list) is available. It's an easy weekend project for the experienced builder, and an excellent learning experience for the neophyte.

the FSVOM

The unit is a basic VOM, with the added feature of a field strength meter. It has a sensitivity of 20k ohms per volt, so that most readings will agree with standard-notation schematics. Circuit loading is minimal. The controls and meter scales are designed to minimize operator error. Only one voltage/current scale and one resistance scale are on the meter face, and only one set of jacks is used, so there's very little chance of error. I find the FSVOM much easier to use than many conventional units with their many scales and ranges.

The FSVOM contains four functions: field strength, resistance, current, and voltage. A multiplier (range) switch provides four ranges, **X1, X10, X100,** and **Xlk,** for each function with the exception of the field strength function, which has an independent sensitivity control. The multiplier scheme allows the use of a single meter scale. The reading is then multiplied by the factor set into the range switch. A separate meter scale is used for the resistance function, which is nonlinear. Dc voltages are read at full-scale ranges of 0.5, 5, 50, and 500 volts. Dc current is read at full-scale ranges of 0.5, 5, 50, and 500 milliamperes. Resistance center-scale readings of 18, 180, 1800,



Front panel showing labels and parts layout.

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

and 18,000 ohms are available. Field strength is read on a relative scale of 0-50. The FSVOM is powered by a 1.5-volt cell and is fully portable.

circuit description

The FSVOM schematic looks a bit complex because of the switching functions; but when broken down into the four separate functions, as depicted in **fig. 1,** each is quite simple and each is an exercise in Ohm's law. The fundamental component is the meter movement, and each function is the application of 50 μA to the meter through series, parallel, or combinations of resistance values. You can readily see why this would make a good training project.

Fig. 1A depicts the basic voltage-measuring function. In this configuration we have the meter and voltage-calibration resistor in series to form a total circuit resistance of 10k ohms. Applying 0.5 volts across this series circuit yields a current of 50 μA and full-scale deflection of the meter. Smaller voltages will, of course, yield lower readings because of the reduced current flow in the series circuit. With the range switch in the **X10** position, additional resistors will be added to the series circuit to form a total resistance of 100k ohms.

The application of 5 volts across the 100k series circuit will again yield a current of 50 μA through the meter movement, causing full-scale deflection. This process is repeated again for the **X100** and **X1k** ranges, each time increasing the series resistance of the meter circuit by a factor of ten and increasing the voltage required for full-scale meter deflection by a factor of ten. This very simple circuit forms the voltage measuring portion of the FSVOM.

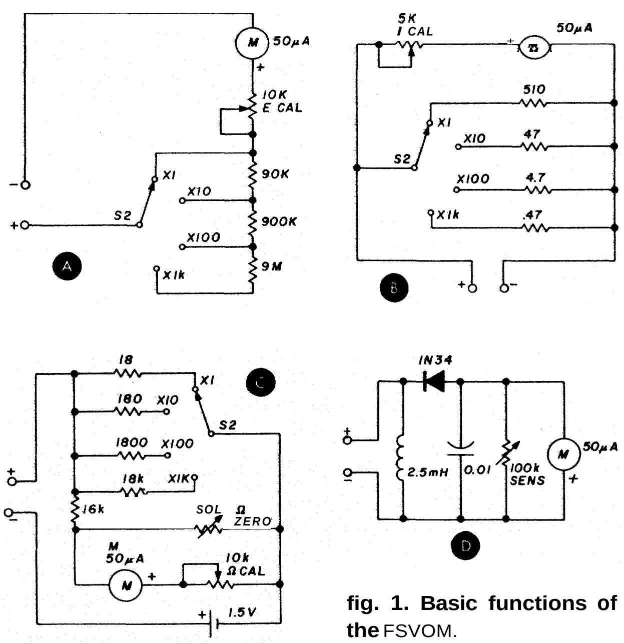


fig. 1. Basic functions of the FSVOM.

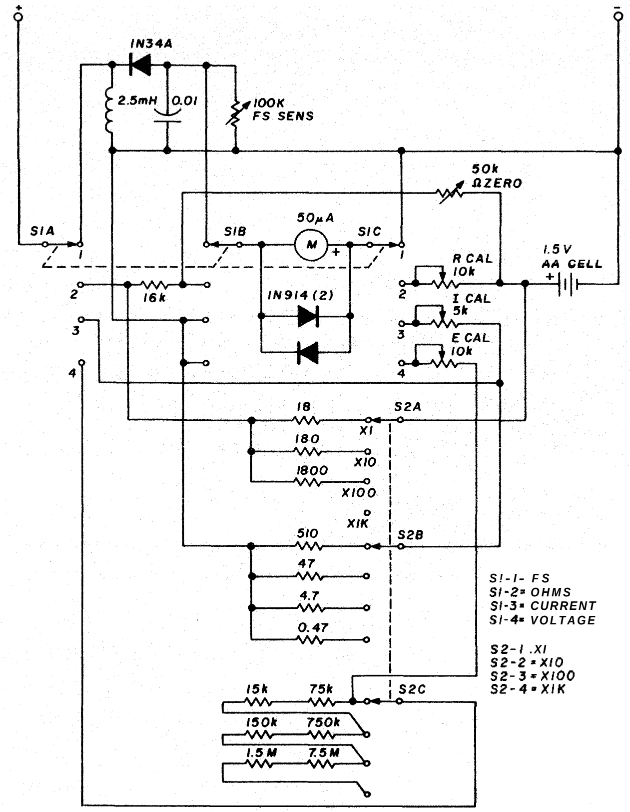


fig. 2. Complete schematic of the FSVOM.

Fig. 1B shows the basic current-measuring circuit. In this function we have the meter movement and current-calibration resistor in series to form a total circuit resistance of 5k ohms. In contrast to the voltage-measuring circuit, where we used series resistors to achieve the desired ranges, we now switch resistors in parallel with the metering circuit to change ranges. The meter movement and calibration resistor form a 5k-ohm circuit and would require 0.25 volt applied across this circuit for a full-scale meter deflection. In effect, we're using this 0.25-volt, full-scale voltmeter to read the voltage drop across each shunt resistor as current from the circuit under test passes through the shunt resistor selected by the range switch.

In the voltage-measuring function, we increased the resistance values in proportion to the voltage being measured. In the current-measuring function, we decrease the resistance values as we select high ranges. Once again, our objective is 50 μA full-scale on the meter. The same scale and multiplier factors are used in the current-measuring function as in the voltage function.

Fig. 1C depicts the basic ohmmeter circuit. As with the previous circuits it's a matter of controlling the amount of current through the meter movement, but in this function things get a bit turned around.

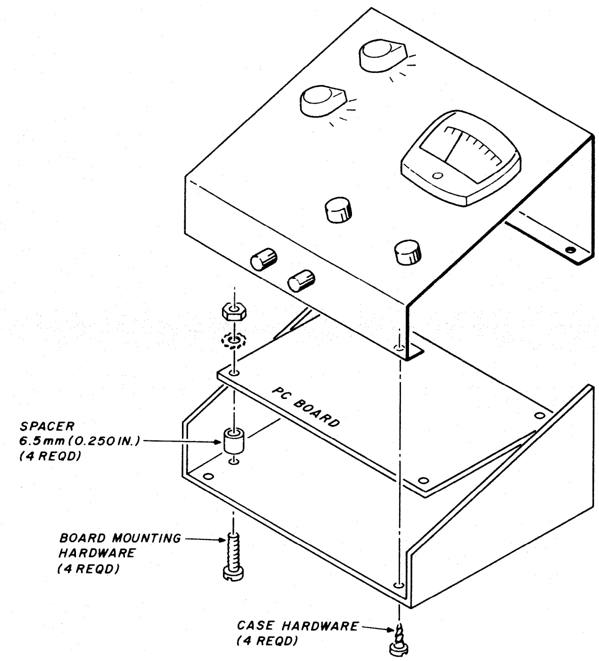


fig. 3. Mechanical details of the FSVOM enclosure.

Zero on the OHMS scale is reflected by full-scale deflection of the meter (50 μA), and for maximum resistance, or infinity, there is no deflection. This fact explains the second scale on the meter face.

In the ohmmeter function we form a voltmeter with the meter movement and the resistance calibration trimmer, just as we did in the current function. The voltage is furnished by the internal battery; it is applied to a voltage divider formed by the resistor selected by the range switch and the external resistance value being measured. When these two values are equal, the meter will read half scale.

We're comparing an internal known value against an external unknown. An OHMS ZERO adjustment control is included on the front panel for setting the meter to zero as the internal battery's output voltage changes with age and use.

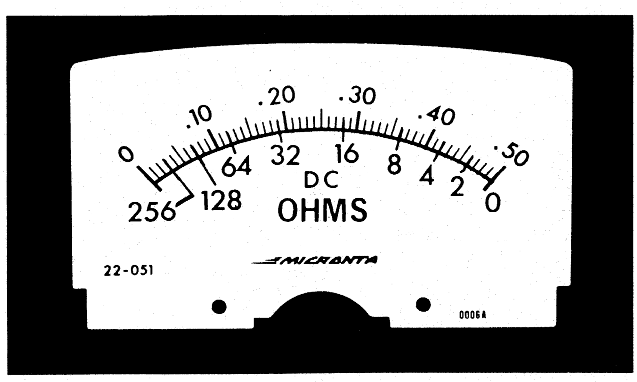


fig. 4. Modification of the meter scale using dry-transfer characters. Resistance scale is in red.

Fig. 1D is the basic circuit of the field-strength-meter function. In this function the positive input jack is the antenna input, and the rf voltage picked up by the antenna is rectified by the germanium diode. The output is applied tc the meter movement. A sensitivity control shunts the meter to allow adjustment over a wide range of field strengths. Since the field-strength function is a relative reading, no special scale is used.

The overall schematic is shown in **fig. 2.** It's a composite of the four individual function diagrams shown in **fig. 1.** The only components added are the two silicon diodes across the meter to prevent damage from overload. After looking over the individual diagrams, it's easy to understand the composite.

table 1. Parts list for the FSVOM. The parts in this listing are those used for the unit in the photos. Parts of equivalent value may be used without degrading performance.

meter	50 μA, RS 22-051
switch S1	3 pole 4 position, Centralab PA-2007
switch S2	3 pole 4 position shorting, Centralab PA-2006
control, FS Sens	100k linear taper, RS-271-092
control, Ohms Zero	50k linear taper, RS-271-1716
trimmer, I Cal	5k, RS-271-217
trimmers, R Cal & E Cal	10k, RS-271-218
jacks, input	Nylon, RS-274-662
holder, battery	1.5 V AA Cell, RS-270-1432
rf choke	2.5 mH, Miller 6302
case	178 × 152x76 mm (7 × 6 × 3 in.), Mod-U-BOX3-7-6
resistors	½W 5%

PC board, etched & drilled: J. Oswald, 1436 Gerhardt Ave., San Jose, California 95125. $4.00 prepaid. (5 or more at $3.00). Residents add sales tax.

Resistors are available from local RCA distributors in 2% at no additional cost over 5% prices (series 8300 resistors).

The RCA resistors and cases are available from Quernent Electronics, 1000 S. Bascom Ave., San Jose, California 95100.

This should be a boon if trouble with the unit is ever encountered.

construction

The first step is to drill or punch the holes in the case, using care not to mar the finish on the case (see **fig. 3**). Next, apply all the labeling to the case using Datak Letraset #K61 dry-transfer lettering. After labeling, apply two light coats of clear finish such as Datak or Krylon to protect the lettering.

While the case is drying remove the plastic front cover from the meter movement and carefully remove the meter face, which is secured by two Phillips screws. Using the periods from the K61 Letraset, change the meter scale numbers 10 through 50 to read .10 through .50, as shown in **fig. 4.** Now, using the red lines and numerals from Datak Letraset K19, add the resistance scale to the meter face. This two-

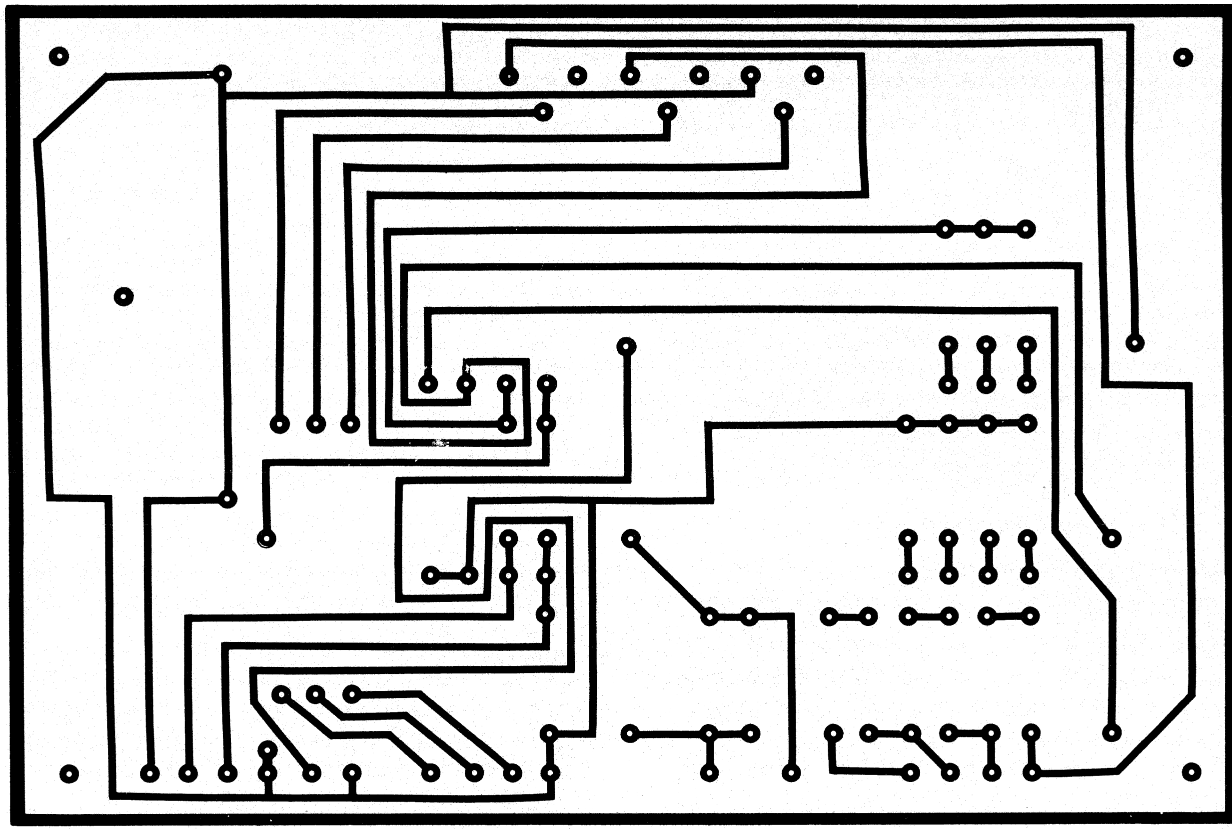


fig. 5. Foil side of the FSVOM PC board.

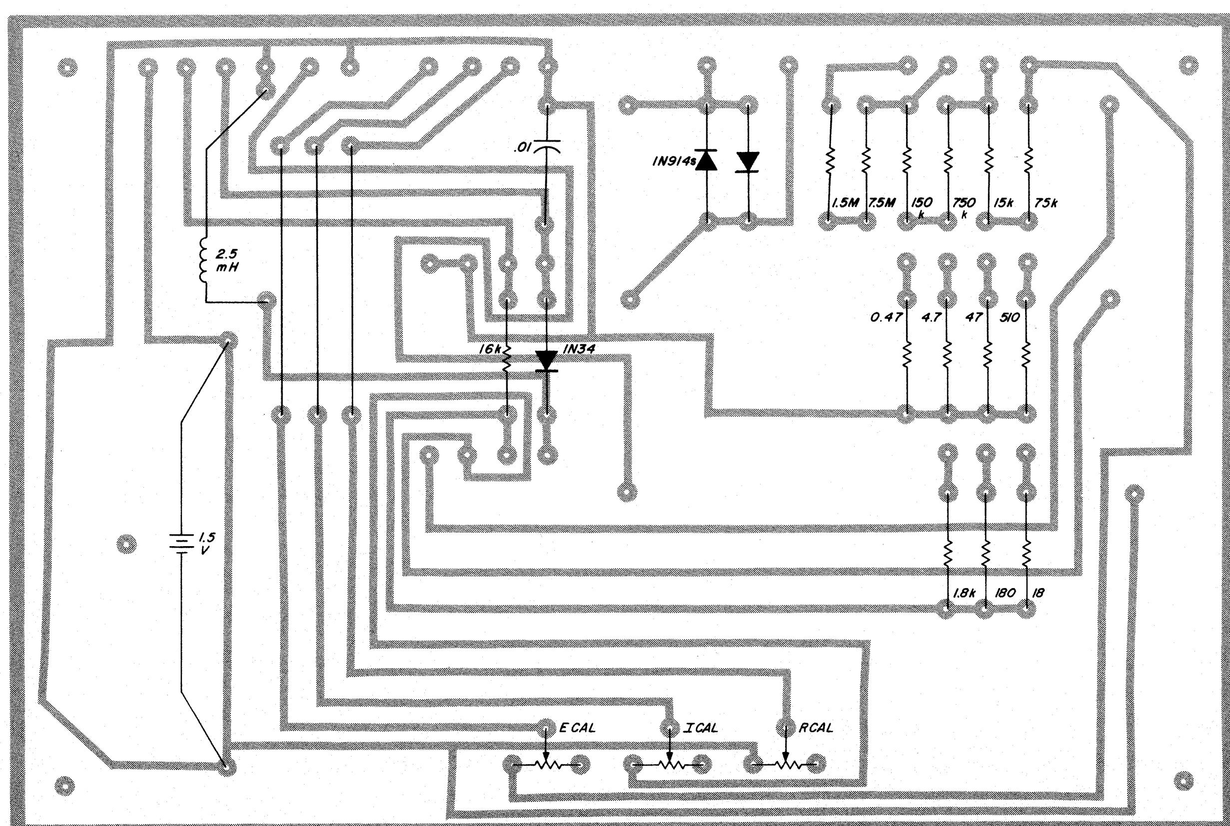
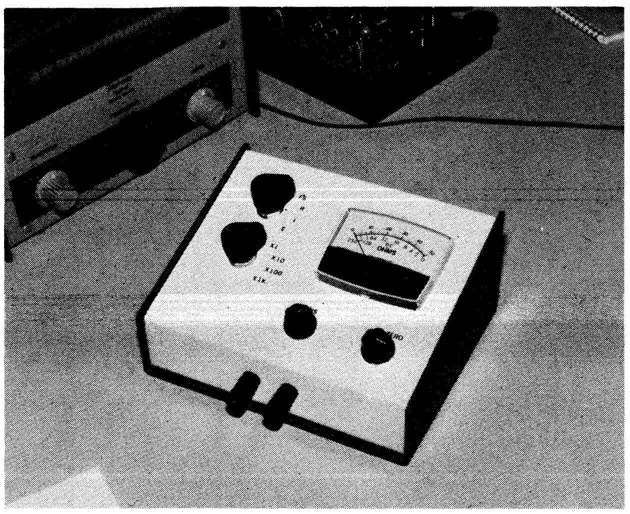


fig. 6. Component mounting.



The complete FSVOM ready for action.

color scheme will make the resistance scale easily discernible. Apply a light coat of protective spray to the meter face. Allow the meter face to dry, reassemble the meter, and mount all the front panel components to the panel half of the case.

A parts list for the FSVOM is shown in **table 1.** The PC board should be etched and drilled as in **fig. 5** and the components mounted as in **fig. 6.** Next, solder the interconnecting wires to the PC board as in **fig. 7,** leaving the wires about 305 mm (12 inches) long and in three groups: S1, S2, and panel components. Mark the far end of the wires with wire-markers or masking tape so they can be identified later.

Mount the completed PC board to the case with the wires extending toward the front. Lay the front panel face down in front of the PC board, and connect the premarked wiring to the panel components. Cut the wires to length after identifying them.

This type of prewiring saves wear and tear on the wires, as the PC doesn't have to be turned over time and again to connect the individual wires. When the wiring is done, the panel should fold back into the case and the wiring should fold neatly into place. The wiring should be laced or spot-tied for neatness. Double check to be certain that the wiring is not pinched in the edges of the case. This completes the construction. Don't button up the case yet.

calibration

The first step in calibration is to zero the meter with its zero-adjusting screw. This is best done with the meter in its normal operating position. Next, install the battery, set the function switch to OHMS

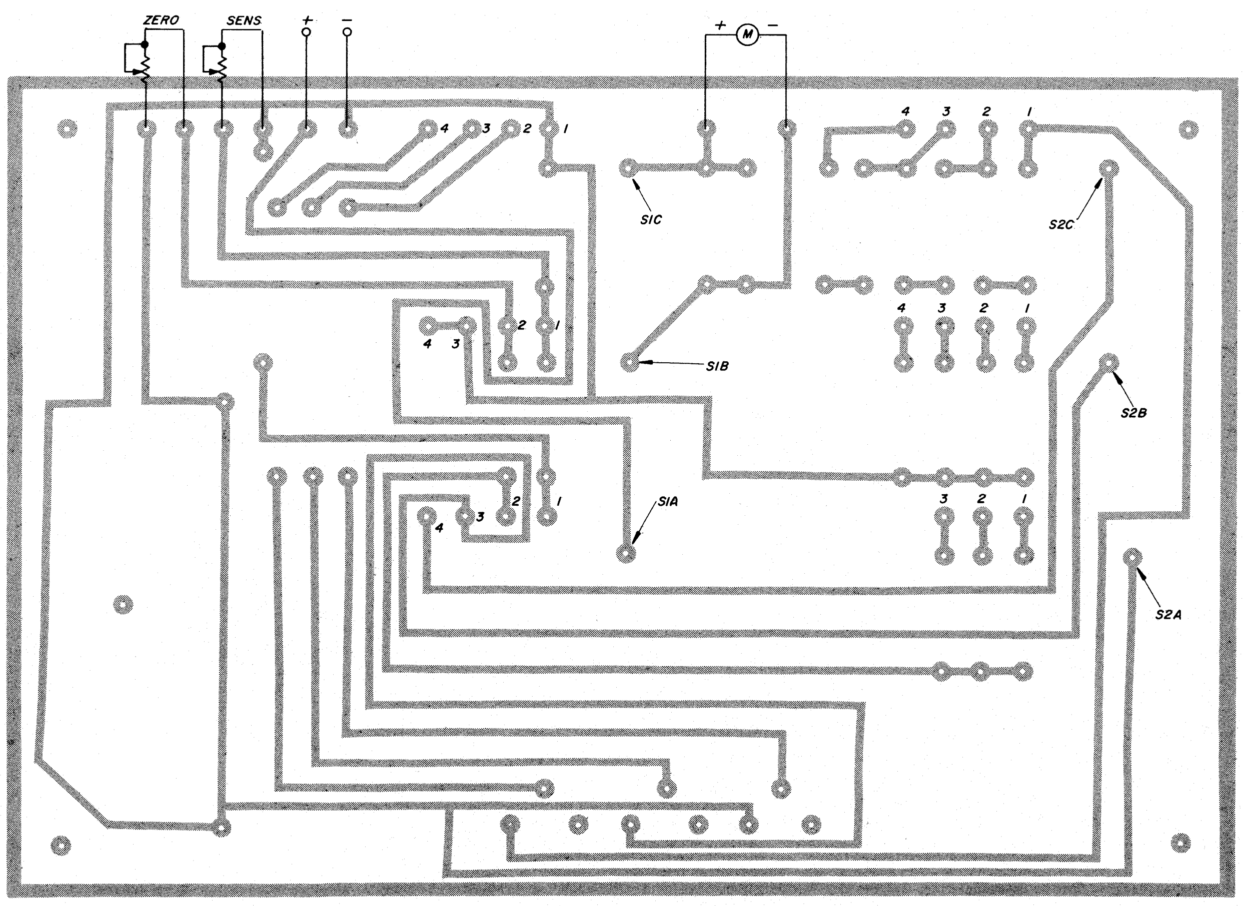
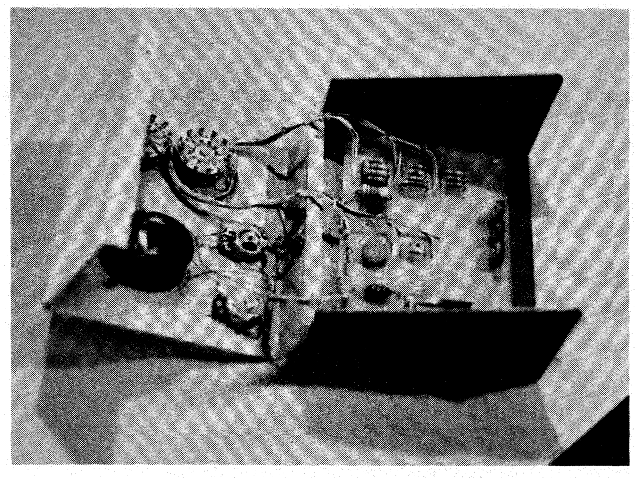


fig. 7. Board-to-panel wiring points.



Wiring details between PC board and panel.

and the range switch to X10. Leave the range switch in the X10 position for the remainder of the calibration procedure. Short the input terminals and adjust the Ω ZERO control to read 0 on the meter's resistance scale. Remove the short from the input terminals and put a 180-ohm resistor across the terminals. Adjust the R calibration trimmer for a reading of 0.25 (half scale) on the upper meter scale. This coincides with 180 ohms on the lower, or resistance, scale of the meter.

Switch the function switch to the current (I) position and place 300 ohms (two 150-ohm resistors) in series with a 1.5-volt battery across the input terminals. Adjust the I calibration trimmer for a full-scale reading of 5 mA (0.50 x 10).

Switch the function switch to the VOLTAGE position and place two 1.5-volt batteries in series across the input terminals. Adjust the E calibration trimmer for a reading of 3 volts (0.3 x 10). This completes the calibration procedure. Now you can fire up your rig and check out the field-strength function.

closing remarks

While I can't make any great claims of accuracy for the FSVOM, it's more than adequate for amateur use. As you probably know, most commercial meters use 1 per cent resistors and much more sophisticated circuitry. But the FSVOM is very easy to use and will provide a degree of satisfaction in that you know what's going on inside the little meter when you use it.

As a club project, it's fun and gets amateurs involved in areas we generally don't explore. As a useful device, it's hard to beat. When you've finished building this one you will have explored just about all phases of home-brew construction, from PC boards to rotary switches to meter movements. The experience gained on this little project should pay off well on all future building efforts.

ham radio

improving the HT-37 ssb transmitter

The Hallicrafters HT-37 is still around and has many excellent features — here are some ideas for upgrading this venerable old rig

Of all the equipment manufactured by the Hallicrafters Company, perhaps the most popular was the HT-37 transmitter-exciter. It uses the phasing system for single-sideband generation, which is a good method if the phase-shift networks are properly adjusted. Once set, however, these phase shifters rarely need adjustment. The rig is as stable as a rock and the ssb audio quality compares well with that from filter-type circuits.

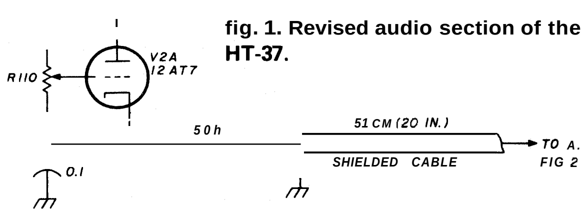


fig. 1. Revised audio section of the HT-37.

This article is presented for HT-37 users who might wish to improve its operation with some minor modifications. For the purist who wants better sideband suppression, an easy modification appears in reference 1, which shows how to install a filter-type sideband generator. If you're happy with the as-built HT-37 sideband generator but wish to make a couple of other simple modifications to improve efficiency and increase power-transformer life, you might be interested in the comments that follow.

Automatic Level Control (ALC) allows you to operate at higher audio levels without overloading the transmitter, which causes interference to nearby stations. With an ALC circuit added to the HT-37, more emphasis is given to lower-frequency speech components, and the higher dynamic range (louder) speech components won't overdrive the final-amplifier stage. Such overdriving creates splatter, "buckshot," and a broader signal.

You'll need the following parts for the ALC-circuit addition:

resistors

quantity	type
1	50k composition, 1/2 watt
1	1 meg composition, 1/2 watt
2	10k composition, 1/2 watt

capacitors

quantity	type
2	0.1 µF paper, 200V working
1	0.25 µF paper, 200V working
1	0.001 µF ceramic 200V working

miscellaneous

quantity	description
2	1N2070 or 1N2071 silicon diodes, 400V PIV
2	solder-lug terminal strips
90 cm (3 ft)	small plastic-covered shielded cable

procedure

1. Remove both top and bottom halves of cabinet.

2. Remove side rail from chassis left side.

3. Remove cover from the audio section to gain access to the audio gain control, R110.

4. Remove the direct connection between R110 and ground. Install a 0.1-µF capacitor between these points.

By Alf Wilson, W6NIF, 1068 Arden Drive, Encinitas, California 92024

5. Solder one end of a 50k resistor to the former grounded terminal of R110, then connect the center wire of a 51-cm (20-inch) length of small shielded cable to the other end of the 50k resistor. Ground the cable shield to a convenient point inside the audio compartment. The revised audio section will then be as shown in **fig. 1.**

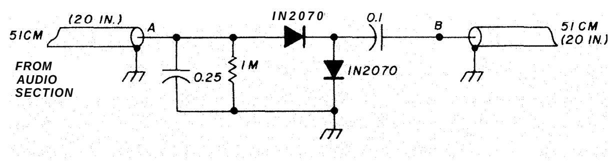


fig. 2. The ALC circuit built from instructions in the article.

6. Feed the free end of the shielded cable through a hole in the audio-section cover shield, Replace the cover shield onto the chassis.

7. Replace the side rail removed from the chassis.

8. Using a solder-lug terminal (4 or 5 lugs), build the ALC circuit shown in **fig. 2.** Be sure to use a terminal strip with a grounded mounting lug at one outer end.

9. Solder the outer end of the shielded cable coming from the audio section to point **A** in **fig. 2.** Attach another piece of shielded cable, about 51-cm (20-inches) long, to point **B** in **fig. 2.**

10. Remove the nut from one of the four machine screws securing the antenna coax fitting to the chassis rear apron. Mount the ALC assembly just constructed on the machine screw, replace the nut, and tighten securely. Of course, a grounding lug should be used to mount this assembly onto the coax connector.

11. Near the bottom of the final-amplifier 6146 tube sockets you'll find a solder-lug strip where R19 (**t**) is connected to rf choke L13, which in turn is connected to the 6146 grids.

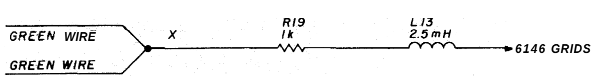


fig. 3. As-built appearance of the HT-37 (see text).

Before performing the next step, note that two green wires are connected to R19 as shown in **fig.** 3. Now proceed as follows:

12. Remove both green wires from point X in **fig.** 3. Leave them free for now.

13. Mount two 10k resistors and a 0.001-μF ceramic bypass capacitor on a 4-lug terminal strip, as shown in **fig.** 4.

14. Next mount the assembly just completed under one of the self-tapping screws that hold the lid onto the shield can next to the 6146 sockets. (This is merely a suggestion; use any mounting position that seems convenient.)

15. Connect the two green wires lifted in step **12** to point **C (fig.** 4) and run a new wire from point D **(fig. 4**) to R19 (point X in **fig. 3).**

16. Connect point **E** to the shielded cable coming from point **B** in **fig.** 2. The final-amplifier wiring will now appear as in **fig. 5.**

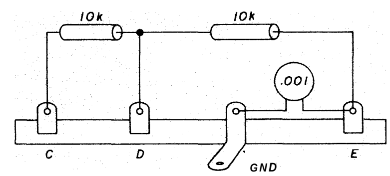
17. Dress the new shielded cables as close as possible to the chassis. Check all wiring for errors. This completes the installation of your new ALC circuit.

A schematic of the complete ALC circuit is shown in **fig. 6**. I suggest you recheck the 6146 bias voltage to ensure it hasn't changed. It should be – 49 Vdc.

power-transformer protection

The HT-37 seems to have a history of power-transformer failure. I've talked to several HT-37 owners who have had to replace the transformer because of a short circuit either in the secondary windings or between primary and secondary windings. It's pretty hard to find an exact replacement for the HT-37 power transformer today, although at least one source of help appears in the amateur ads in which a transformer rebuilding service is offered.

fig. 4. More construction details for the ALC addition to the HT-37 final-amplifier section.



In any event, it's possible to preclude catastrophic failure of the power transformer by simply adding an autotransformer, such as a Variac, in the primary voltage circuit of the HT-37.

HT-37 owners will note that, when the OPERATION switch is turned from OFF to STANDBY, a distinct *"thung"* sound will be heard if the peak of the ac primary voltage occurs at time of switch turn-on. This means that a surge of voltage is presented to the power transformer primary at the instant of switch turn-on.

Why not eliminate this surge by using an autotransformer in the transformer primary? With the OPERATION switch in the OFF position, turn the autotransformer to zero, then gradually advance the autotransformer control until the proper ac input voltage is presented to the power-transformer primary. An ac voltmeter should, of course, be connected across the power-transformer primary.

Another cause of power-transformer failure, according to many HT-37 owners, is sheer carelessness

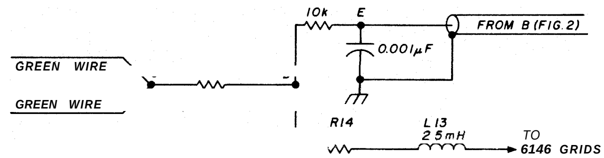
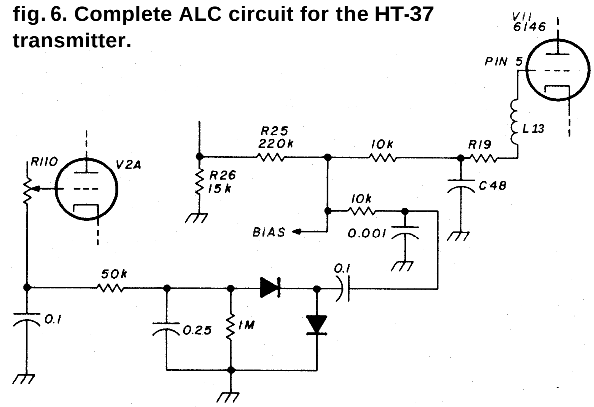


fig. 5. Schematic showing modified wiring in the HT-37 final-amplifier section.

during operation. The HT-37 instruction manual stresses that, when switching to **STANDBY** from any of the operating modes (MOX, **VOX**, or CAL), it's important to wait for a few seconds before switching the set **OFF.** There must be a message here. According to other HT-37 owners, it's a dead cinch that the powertransformerwill blow if the set is rapidly turned from one of the operating modes to **OFF** and back on! It's easy to do this with this equipment, especially during the heat of a contest. A Variac won't help in this case, of course, because during operation the Variac will be adjusted for full input primary voltage.

fig. 6. Complete ALC circuit for the HT-37 transmitter.



So if you value your HT-37 power transformer, respect the precautionary advice in the instruction manual for the **OPERATION** switch. The HT-37 is a great rig, even by today's standards. But if you can't replace a blown power transformer, you may as well try to sell the rig for junk.

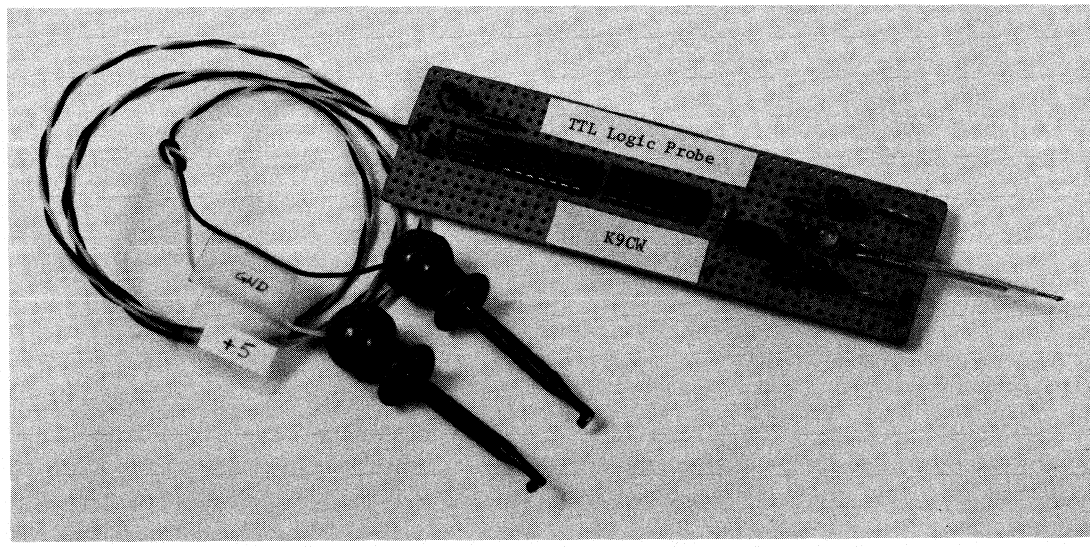
acknowledgment

The material on adding ALC to the HT-37 was taken from a paper by WØNCK and KØTYO. This paper was included with the instruction manual for my HT-37, which I purchased second hand. I built the circuit and it is an improvement over the original HT-37 design. Any credit for this improvement should go to WØNCK and KØTYO.

reference

1. Milton L. Pokress, W3CM, "Increased Sideband Suppression for the HT-37," *ham radio,* November 1969, pages 48-51.

ham radio



high-performance TTL logic probe

Design of a simple yet high-performance logic probe using three low-cost integrated circuits

Have you constructed a frequency counter or keyer that didn't work when you first turned it on? Has some piece of equipment that uses TTL integrated circuits failed recently? If you answer yes to either of these questions, you've been in a situation in which a logic probe could have been very useful. This article describes a simple TTL logic probe that can be constructed for less than **$5.** Yet its performance equals that of units costing as much as **$45.**

operation

The logic probe is to the digital designer what the VOM is to the electrician. With a single touch, the probe displays the logical state or condition of the selected circuit connection. A well-designed probe will indicate an open circuit, a good logic 1 or 0 state, and whether a pulse train is present. This tool enables the digital experimenter to follow signals through the various logic gates and circuit components until the faulty part or wiring error is isolated.

The logic probe has three LEDs of different colors

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mounted near the probe tip. After the probe's +5 volt and ground connections are made to the circuit under test, the tip can be touched to any point carrying a TTL signal. If the point is less than about 0.7 volts, the 0 (green) LED will be the only one ener-

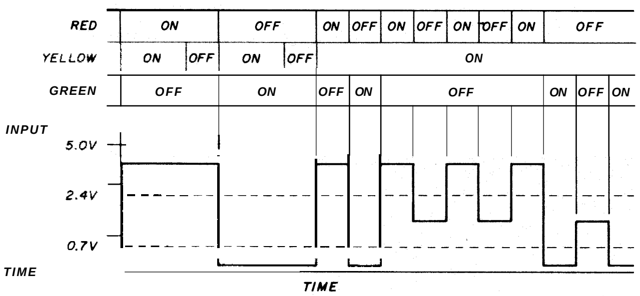


fig. 1. Graphical summary of the indicating LEDs for various input states. When connected to a floating input, none of the LEDs will be on.

gized. If the test point is 2.4 volts or greater, only the 1 (red) LED will come on. If the selected circuit is oscillating between 1 and 0, the red and green LEDs will show the relative duty cycle of the oscillation, and the PULSE (yellow) LED will pulsate or will seem to be continuously on.

The advantage of a special pulse indicator can be seen when considering a normally low or 0 signal with 200 nS high pulses every 200 mS. Since the signal is high for only a small fraction of the total period, the probe's green LED will seem to be always on and the red LED always off. However, each of the pulses is stretched to about 0.5 second and displayed on the yellow LED. Thus, the green and yellow LEDs on together indicate the presence of the narrow, high pulses.

If the line is an open circuit, has a resistance of more than about 4000 ohms to ground, or has a voltage level between 1.0 and 2.4 volts, neither the high nor low LED will be energized. An unterminated TTL input will be displayed in this fashion. A summary of the logic probe's operation, with various input waveforms, is shown in **fig. 1.**

circuit description

The schematic diagram of this logic probe, as illustrated in **fig. 2,** consists of three sections: the high-level detector and red LED driver (Q1 and Q2), the low-level detector and green LED driver (U1A and U1B), and the pulse detector and yellow LED driver (U2 and U3). Q1 functions as a threshold detector, which will turn on with an input voltage of greater than about 2.4 volts. The threshold level is determined by the three silicon diodes connected between the emitter of Q1 and ground. Each diode exhibits a forward-biased voltage drop of about 0.7 volts. When Q1 turns on, Q2 is forced on, connecting the +5 supply voltage to the anode of CR3, turning on the red LED.

The low-level threshold detector uses one section of a 74LS02 and a silicon diode. Without this additional diode voltage drop, the green LED would be energized at about 1.4 volts. With the diode, the input voltage has to be below 0.7 volts before pin 1 of U1A will go low, turning on the green LED. Notice that between the upper and lower thresholds neither CR1 nor CR3 will be energized.

The pulse detection circuit employs a dual D-type flip-flop and a monostable multivibrator to stretch the input pulse length. The detection of a valid 0 or 1 level is signaled, respectively, by U1 pin 4 or the cathode of CR3 going high. Either of these conditions causes one of the outputs of U2 to go high, forcing pin 13 of U1 low, which triggers the monostable and turns on the yellow LED, CR2, for about 0.5 seconds. As long as the output of the monostable is high, both flip-flops remain cleared. The D-type flip-flops are used to insure that U3 is properly triggered. The trigger pulse must be at least 200 nS long and cannot stay low, since that would hold CR2 on. The 74LS74 shown will respond to pulses as narrow as 25 nS.

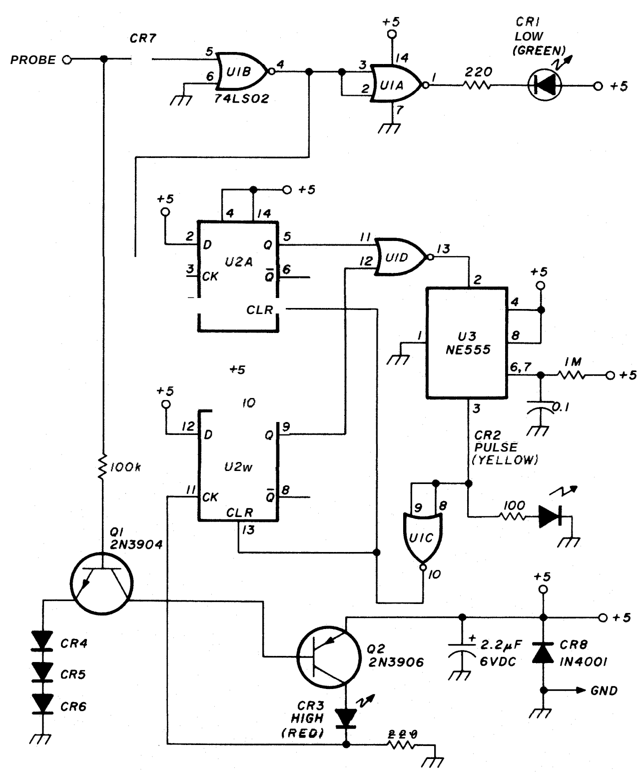


fig. 2. Schematic diagram of the logic probe. All resistors are 1/4 watt; CR4, CR5, CR6, and CR7 are 1N4148 or equivalent small-signal diodes.

A logic probe should not excessively load the circuit under test. This probe will sink only about 25 μA when the input is +5 volts. With a O-volt input, the circuit input current is about −200 μA. In the worst case, this probe would represent one LSTTL

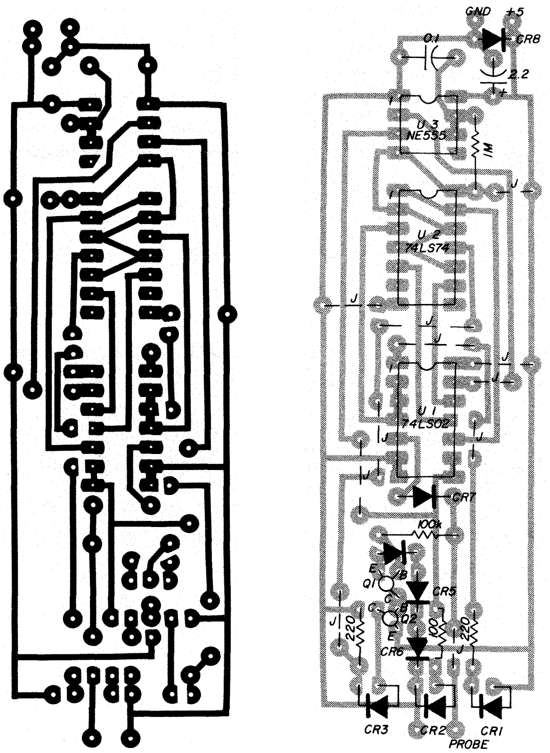


fig. 3. Suggested printed-circuit board layout and parts placement diagram for the logic probe. If desired, prepunched board and simple point-to-point wiring may be used.

load. In addition, the probe will "catch" pulses shorter than 100 nS in duration.

construction

The circuit layout for this device is not critical. **Fig.** 3 illustrates one printed circuit board layout design and parts placement for the logic probe. The overall board dimensions are 3.3 x 10.2 mm (1.3 × **4** inches). You could use point-to-point wiring on a circuit board with holes on a standard 2.5 rnm (1 inch) grid instead. The probe tip can be made from 16 AWG (1.3 mm) tinned-copper wire or a small brass nail. The +5 and ground inputs should be connected to clip leads so they may be easily connected to the circuit supply voltage source. Standard TTL chips may be substituted for the LS devices shown, but the circuit loading will be increased.

I have found a logic probe to be a valuable aid in debugging TTL projects. After you use one, you may wonder how you were able to get along without it!

ham radio

code speed counter

Do you dislike counting up the words and doing the arithmetic to check your code speed? If so, then you need a code speed counter. All it takes is an ordinary frequency counter and a couple of ten-cent ICs.

In many cases, a frequency counter can be set up to count for one second, performing the latch and reset function during the next 0.2 second. By inserting a divide-by-four counter (7493) in the clock circuit, the counter will then count for four seconds instead of one second.

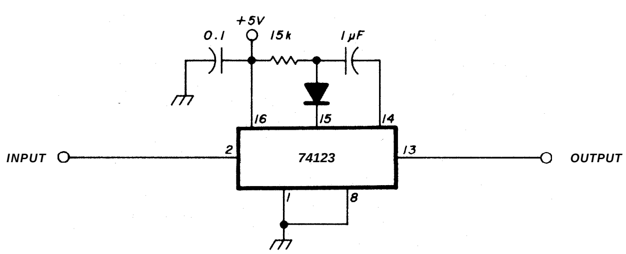


fig. 1. Wiring diagram of the one-shot multivibrator used to convert the sine wave into a pulse for the counter. Pin 2 connects to the output of the wave shaper, while pin 13 goes to the counting circuit.

The goal is to have the counter count the number of key downs during a four-second interval. This will be approximately the code speed in words per minute. A dot or dash from the speaker terminals is, in many cases, a 700 to 1200 Hz sine wave. It is necessary to generate a positive pulse from each character. To do this, insert the retriggerable monostable shown in **fig. 1** before the counter's input.

performance

Using a piece of graph paper, find out exactly what readouts the modified counter will produce. In **fig. 2,** the word ***Paris*** is represented in code. If you assume the speed is 12 wpm, then each dit is 0.1 second long. The counter reads 0 until the first four seconds have passed. It will then read out 12, since 12 key downs occurred during that time. Those key downs sent during the 0.4 second of the latch and reset cycle are ignored by the counter. The 12 will remain on display until the next four seconds have passed, then read out the number of key downs during that period, and so on.

If you plug the counter into the speaker jack and tune in the W1AW code practice, you'll find that the counter produces readouts which are very close to the announced speeds. On the faster speeds, 20 wpm or more, the counts tend to be low. This is because at faster speeds more words will occur during the four-second count period, and the counter does not make any allowance for the seven-baud spaces between words.

A very useful operating aide is provided by connecting the input of the 7490 counters to one of the decimal points in the readout. This makes it possible to see the dits and dahs which are being counted. By watching the blinking decimal point you can tell when interference is messing up the count. Using the gain control on the receiver you can tune out the interference, if the desired signal is strong enough. Just turn down the volume until only the desired signal is visually evident. You can also tell, by watching the flashes of the decimal point, which four-second intervals are representative of the speed being sent, and ignore counts of intervals containing long pauses or interference.

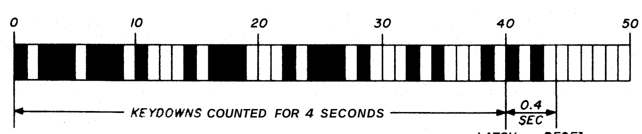


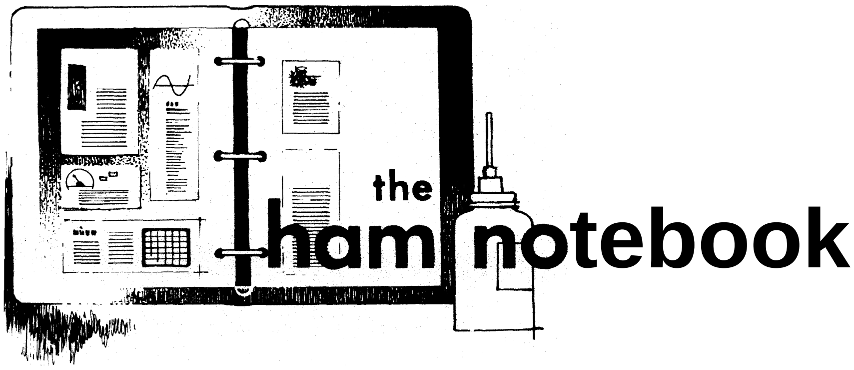
fig. 2. Diagram of the standard 50-baud word PARIS.

If you don't have a counter which you wish to modify, then you can build the code counter from scratch by adding the above modifications to any of the plans which have appeared in the literature.

The code speed counter will add another element of enjoyment to operating CW. It will also give you an easy way to check on your progress if you're trying to build up your code speed.

ham radio

By Louis C. Graue, K8TT, 624 Campbell Hill Road, Bowling Green, Ohio 43402



the ham notebook

quick connection

An effective, inexpensive way to connect wire to a pole or pipe is to use a regular automobile hose clamp. These clamps come in various sizes and are available at most service stations.

For outside installations, stainless-steel clamps should be used. The area of the pole that comes in contact with the clamp and wire should be free of dirt and grease. Clean the pole with steel wool or very fine sandpaper. Be sure to wipe off any dust from sanding.

After the clamp is secured to the pole, it should be taped with plastic electrical tape for weather protection. I've employed this method to connect a grounding wire to a ground rod of small diameter, and it proved very successful.

Jim DiSpirito, AB9Q

remote crystal switching

With the rising theft rates of mobile equipment, it is becoming increasingly attractive to mount vhf radios in the trunk. However, such an arrangement can present problems if you normally use a large number of channels. In a conventional remote-switching system, the relays or diodes, which are used to switch crystals, are selected by a simple rotary switch in the control head. A separate conductor, from the control head to the radio, is required for each channel. However, the use of a large multiconductor cable can be avoided by implementing a simple binary encoder/decoder system. By using binary coding, only four conductors are needed to select sixteen channels, or five conductors for thirty-two channels. Such a system has been incorporated in my mobile installation, which was built around a modified Genave chassis.

The control-head circuit for the first ten channels (0 through 9) is shown in **fig. 1.** Of course, this can be expanded to any number of channels. Binary encoding is accomplished through the use of 1N4148 or equivalent steering diodes. If you have trouble finding a rotary switch with enough positions, you might try using a toggle switch to control the most significant bit. In this way, an 8-position rotary switch and a toggle switch can select sixteen channels.

As for decoding circuitry, there are a number of decoder ICs on the market, with a variety of specifications. The user should select one which is best suited to his particular application. The 7445 and 74145 ICs are BCD-to-decimal decoders with open-collector outputs, suitable for driving reed-relay switching.

If diode switching is used, the 7442 BCD-to-decimal decoder or 74154 4-line to 16-line decoders will work well. A 4-line to 16-line decoder may also be built from two 7442s and an inverter as shown in **fig. 2.** This illustration shows the decoder/diode driver currently being used in my mobile installation. This configuration is suitable for use with oscillators in which the crystals are selected by grounding. The 7442s were used instead of a 74154 because they happened to be available. TTL inputs normally float to a high logic level if left unconnected. The four 150-ohm resistors pull down the inputs to approximately ground

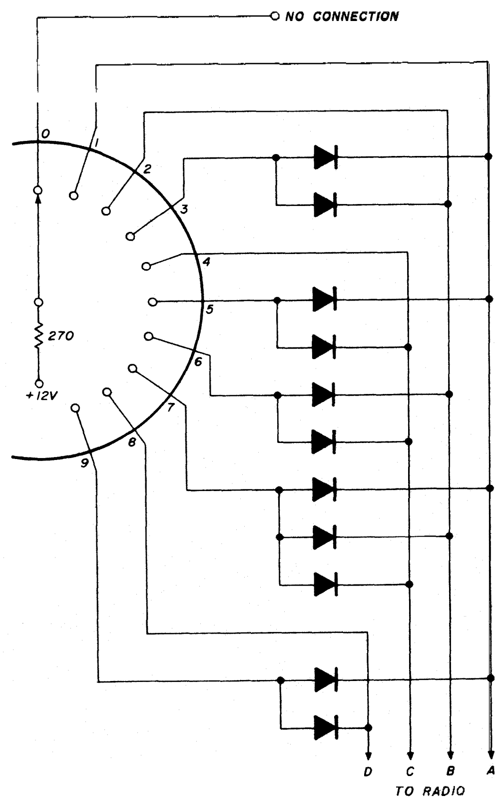


fig. 1. Schematic diagram of binary encoding circuit in the control head. The diodes provide steering such that the four lines to the radio receive the correct binary coding.

potential, unless a positive voltage from the control head is present. The reason for using this method is to provide relatively low-impedance inputs to the decoder to insure immunity to rf pickup and other extraneous signals. To be on the safe side, all four lines should also be bypassed with 0.01- or 0.005-μF capacitors at the decoder.

Inverters of the 7404 type provide a positive voltage and enough current to turn the diodes fully on. Diode current is set by the 82-ohm resistors. Larger-value resistors were initially tried, but they did not allow adequate diode current to flow. As a result, the receiver oscillator output was diminished, which in turn caused a small reduction in receiver sensitivity. A value of 75 or 82 ohms allows adequate current to flow, while remaining well within the current limitations of the diodes.

Notice that no rf chokes or bypass capacitors are required in the vicinity of the crystals or switching diodes, unlike some circuits appearing in the handbooks. A decoder could probably be designed without inverters on the outputs, but rf chokes would very likely be necessary to isolate the crystals from each other. With prices being what they are, it would be considerably cheaper to use hex-inverter ICs for isolation rather than rf chokes.

Binary channel addressing is readily adapted to more sophisticated digital control applications. For example, the radio can be converted into a scanner simply by adding a gated clock oscillator and a 7493 4-bit binary counter. The binary format is also well suited to microprocessor control.

Binary coding circuits have seen continuous service in my mobile installation since late 1974, first with DIP reed-relay crystal switching and more recently with diode switching. They have proven to be rugged and reliable, and have added considerable flexibility to the mobile system design.

J. Lee Blanton, WA8YBT

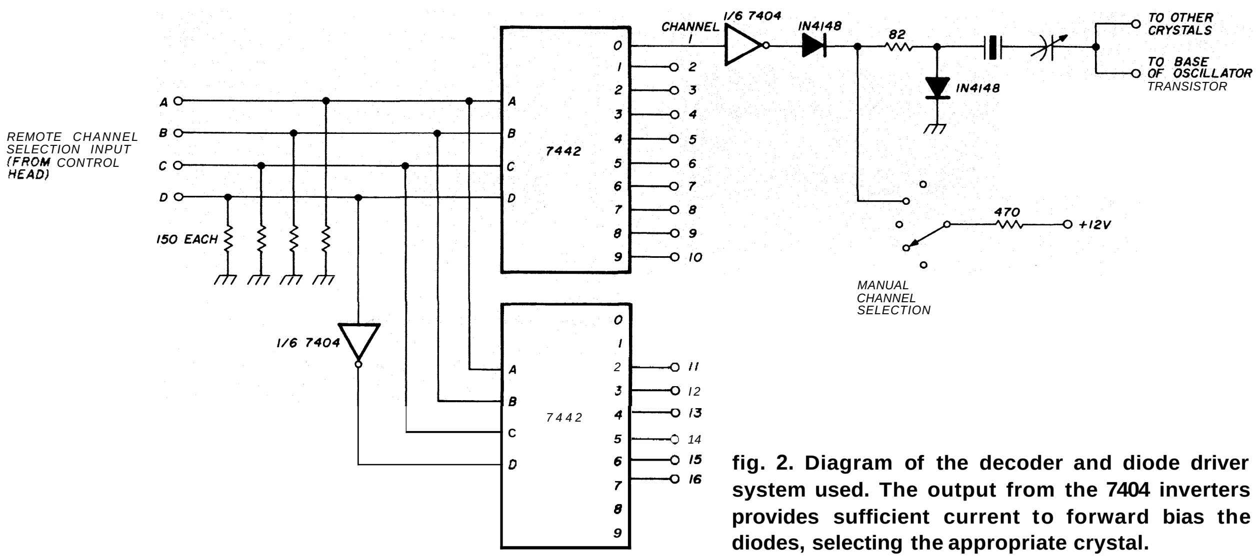


fig. 2. Diagram of the decoder and diode driver system used. The output from the 7404 inverters provides sufficient current to forward bias the diodes, selecting the appropriate crystal.

solid-state rectifiers in the Collins 516-F2

Not long after acquiring my present Collins station, I experienced a problem in the power supply which was directly traced to a cracked high-voltage rectifier tube socket. The socket becomes brittle after some years of use. This is primarily due to the large amount of heat generated by the rectifier tubes. Of prime concern though, is the power transformer which might be damaged due to such an occurrence; they are expensive to replace.

Rather than simply replacing the socket, I decided to eliminate the heat source by substituting plug-in solid-state rectifiers for the tubes. This requires a minimum amount of work since both the 5U4 and 5R4 were replaced by SS-5R4 units purchased from United Page Incorporated. Another procedure, which was suggested by the Collins Service Department, is the use of four Semtech SCH-5000 diodes (CPN 353-0425-010), two diodes replacing each rectifier tube. These are hardwired to the tube sockets underneath the chassis.

In each instance it would be a good idea to disconnect the rectifier filament leads, insulate and tie them back out of the way, although I ran my unit for some time after installing the plug-in units without doing so. The filament leads are not needed and removing them eliminates one more source of possible trouble. Also, the bias should be re-adjusted to obtain the proper resting plate current since all voltages will have increased about 10 per cent.

Paul Pagel, N1FB

short circuits

500-watt power supply

The article by WA6PEC, December, 1977, *ham radio,* page 30, contains a drafting error in fig. **1.** There should not be a connection between the collector and base of Q3.

tone-burst generator

The tone-burst generator by WA5KPG described on page 68 of September, 1977, *ham radio* contains an error in fig. **1.** The junction of the 1N914 and R2 should not be connected to the PTT line, but only to C3.

active bandpass filters

The article on active bandpass filters (December, 1977, *ham radio,* page 49) contains some errors in the equations. An errata sheet is available by sending a self-addressed, stamped envelope to either the author or *ham radio.*

admittance, impedance, and circuit analysis

In Anderson's article in the August, 1977, issue, there's a glaring typographical error in the second sentence of the right-hand column on page 76. This should read, *"if R = 10 ohms, then G = 0.1 mho."* There should be no minus sign in front of the 0.1 mho.

10-GHz broadband antenna

In the article on the broadband 10-GHz antenna which appeared in the May, 1977, issue of page 40, the chart of optimum feed beamwidth (fig. 3) is incorrectly labelled. The ordinate should be labelled as feed angle, *not* half-angle. Therefore, the simple antenna should be used with reflectors with focal length to diameter ratios, *f,* between 0.3 and 0.6. Thanks go to N6TX for spotting the error.

digital frequency counter

The counter article on page 22 in February, 1978, *ham radio* contained some drafting errors. The 9368 IC in figs. **2** and 3 should have V_{cc} connected to pin 16 instead of pin 11. Also in fig. 3, the short across the crystal should be omitted. In all cases, the collector resistor for the 2N5179 is 1000 ohms.

A circuit board layout of fig. 3 is now available from the author; a paper print is available by sending him a self-addressed, stamped envelope. A film negative is also available from the author for $5.00.

ssb phasing systems

On page 58 of the January, 1978, issue VK2ZTB states that an 88 mH toroid consists of two 44 mH coils. This is incorrect — an 88 mH toroid consists of two 22 mH coils connected in series. Thanks to N3GN for spotting the error.

30-MHz low-noise preamp

Coil winding instructions for the low-noise 30-MHz preamplifier in the October, 1978, issue, which were inadvertently left out of fig. **1,** are as follows:

L1	(0.77 pH; 17 turns no. 28 (0.3 mm) wound on Micrometals T-25-6 powdered-iron toroid
L2	(1.0 μH) 20 turns no. 28 (0.3 mm) wound on Micrometals T-25-6 powdered-iron toroid
RFC	(10 μH) 20 turns no. 28 (0.3 mm) wound on FT-230-06 ferrite bead

active filters

The letter regarding active filter design (see *Comments, ham radio,* June, 1978, page 102) contained an erroneous equation. The value for R2 is determined from the following equation:

$$R2 = \frac{1}{(2 \cdot Q - (G/Q)) \cdot C \cdot f \cdot 2\pi}$$

spectrum analyzer filters

The two ceramic 10.7 MHz filters used in the spectrum analyzer described in the June, 1977, issue of *ham radio* (FL401 and FL501) are no longer available from Vernitron. Just before the production line was shut down, however, one of the employees was able to obtain 25 matched pairs which he is offering to readers who wish to build the spectrum analyzer; the price is $6.50 per pair, plus postage. Write to William Bowen, 1939 Green Road, Cleveland, Ohio 44121. The only other source for comparable ceramic filters is the Murata Corporation in Marietta, Georgia.

semi-precision voltage calibrator

The digital voltmeter calibrator on page 68 of the July, 1978, issue of *ham radio* contains an incorrect statement. The input resistance of the precision rectifier circuit is actually 10k ohms, instead of the value of between 5 and 65 megohms specified in the article. To correct this problem, fig. **1** shows a simple voltage follower that can be inserted between S1 and the input to the precision rectifier. This will eliminate the loading effect on the reference voltage source and has negligible effect on the overall accuracy.

general purpose vhf receiver

A capacitor was deleted from the schematic diagram shown in fig. **2** of the general purpose vhf receiver published in the July, 1978, issue of *ham radio* (see page 19). A 22-pF capacitor should be inserted between the gate of the MPF102 and the rotary contact of switch S1F.

R-4C product detector

In the October, 1978, issue of *ham radio,* the schematic diagram of the new product detector for the R-4C shown on page **94** contains an incorrect component value. The 0.01-μF capacitor in series with pin 3 of the TL442 should actually be a 0.1-μF capacitor. In addition, the two inputs for the TL442 are actually taken only across one-half of the transformer's secondary.

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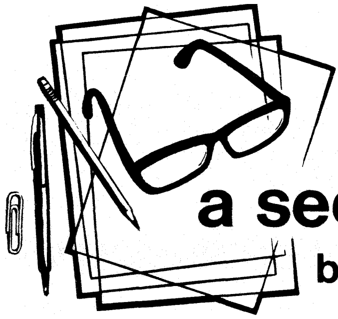


a second look

by Jim Fisk

This is the time of year when many high-school seniors are scurrying around, planning their future education, sending applications off to the college of their choice, and taking entrance exams. Seniors who are also Radio Amateurs are probably considering a career in electronics. If they're lucky, they will have a knowledgeable guidance counselor who can steer them in the right direction; if not, they'll probably pick a school with a good reputation and work from there. Sometimes this works out, and sometimes it doesn't – it depends entirely on what the student is looking for.

Electrical and electronics engineers who graduated more than ten years ago would probably not recognize the engineeering curriculum now offered by their old alma mater because, in the past few years, there have been significant changes in engineering education. During the 1960s the classical engineering educational programs tended to become more and more theoretical oriented, with less emphasis on applied engineering. The backgrounds of some electrical engineering staffs changed from being primarily applied electronics to applied mathematics, and attempts to develop practical engineering programs were not ail that successful. In recent years, however, some engineering colleges have restructured their curriculums for a better balance between the theoretical and the practical. On the other hand, some colleges have continued to stress the theoretical aspects of engineering science, so the prospective student is faced with a very important, but difficult, choice.

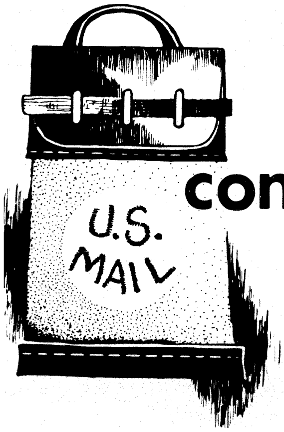
Not too long ago, the prestige of an engineering school was almost always gauged by the theoretical emphasis of its courses; each school tried to outdo the others in the theoretical sophistication of its curriculum. Unfortunately, the majority of jobs within the sphere of electronics engineering does not require such an advanced mathematical sophistication as they do a "gut" understanding of electronics. If you talk to students at a *theoretical* school, you'll find that many of them don't know how to solve a simple steady-state ac problem, although they can invert a matrix and use state-variable techniques.

The difficulty with this type of engineering education is that graduates are not adequately prepared to solve the day-to-day engineering problems they will be presented with in industry. Employers are faced with the prospect of several months of on-the-job training before the newly hired engineer becomes a fully contributing member of the staff. Obviously, a new engineer who can solve problems quickly and practically in the real world is a valuable asset.

In the 1970s several colleges introduced four-year electronic technology programs in an attempt to get back to the old practical engineering concept; the courses at these colleges emphasize electronic hardware and laboratory techniques as well as electrical theory. Although graduates of these Bachelor of Science Technology programs have been pictured as fitting into the occupational spectrum somewhere between the technician and the engineer, many professors see technology graduates as having much wider employment opportunities. In fact, technology graduates now have opportunities in many areas of electronic applications and design traditionally occupied by engineering graduates, jobs vacated because of the change in emphasis in engineering education programs.

Students who are interested in this type of engineering program should be aware that there is a wide difference in B.S. programs parading under the "Technology" banner. Some curriculums are managerially oriented, others are slanted toward applications and design, while still others are little more than two-year electronic technician training programs with added courses in the arts and humanities to fill out four years. Students who wish to enter this area should obviously choose a school carefully to be sure they get exactly what they want.

Jim Fisk, **W1HR**
editor-in-chief



comments

metric dimensions

Dear HR:

The meter is a widely used unit of length abroad. It is convenient only in that fractions and multiples are decimal. As a specific unit of length it is as arbitrary as the foot; for reference, see the NBS history book *Measures for Progress.* Navigators use the international unit of distance, the nautical mile (based on the earth's equator), rather than the result of two French surveyors' efforts in measuring the distance from Paris to Marseilles.

There are times when metric measurements are useful, other times when other standards are better. Having used so many different standards for so long, I am not really partial to all-metric or all-other. For construction and items used therein it would seem advisable to state the local measurement (depending on the author's location and supply) *first,* then the conversion. For example, the W7DI antenna in the November issue used tubing from a U.S. supplier. It comes in standard diameters in inches. The same for plumbing pipe. One is hard pressed to make a supplier understand millimeters or centimeters. It would seem reasonable to state such dimensions in inches first with cm or mm in parentheses. If DK1AG had written the article it would seem reasonable to expect the local German dimensions first, then inches. I am referring mainly to those dimensions where specific material is widely available.

A sheet of plywood in the United States is going to be 4 by 8 feet for a long while to come. Three-quarter-inch plumbing is going to remain ¾″ for a long time — in fact, any material used in the building trades can be expected to remain in established dimensions for the next ten years — perhaps longer. Many electronic items also remain in the "English" standard. Most numerous are the ICs. They remain rooted on the 0.05-inch (1.27-mm) grid for a simple reason: The United States started it and uses millions of them every year. And many are fabricated in metric countries.

There are a host of other standards. Why the 1¾″ increment on the height of rack panels? Think of how long those have persisted. Some rack panel heights have varied, but the 19-inch (42.3-cm) width is still here. Quarter-inch (6.4-mm) shafts are common. Fasteners are described in threads per inch with roughly arbitrary diameters; both U.S. and foreign types should be differentiated. Consideration should also be given to the enormous number of United States types in use and made each year.

The scientific community remains in both camps and is not certain on a few items. There is increasing use in optics of the nanometer instead of the Angstrom, for example. Seimens has yet to replace mhos. Parsecs, kilometers (why not megameters?), and light-years seem to be interchangeable in astronomical distances. At least to NASA.

We have made the transition from tubes to transistors but tubes are still here. I believe that a double standard can continue, and should do so if certain material is in common supply in the author's country; contemporary literature should reflect that fact.

It is difficult for all to go completely metric, both for editor and author. We will continue to write on 21.6 by 27.9 cm paper using 3.94 pitch, double-spaced. I am awaiting delivery of a new IBM typewriter with a 38.1-cm carriage and dual pitch (4.72 for correspondence). If it bothers me, I'll just pour a glass of milk from the 1.89-liter carton.

Leonard H. **Anderson**
Sun Valley. California

Mr. Anderson's view of metrification is reasonable and well stated. The response we received to our boxed "metrics only" editorial which appeared with W7DI's article in the November issue was both immediate and loud. Based on the letters which have crossed my desk during recent weeks, ham radio *will be using both metric and English dimensions in our magazine articles for the foreseeable future.* **W1HR**

zip-cord feedlines

Dear HR:

The article on zip-cord feedlines in the April, 1978, issue and the follow-up comments in October brought back some old memories. I first heard of lamp-cord around 1930 (twisted-pair in those days) from the Globe Wireless operators who used it for transmission line to their "noise reducing" receiving antennas. Later there were a number of articles in the Amateur magazines on how to use lamp-cord transmission lines for both receiving and transmitting dipoles; it provided a reasonable match to the feedpoint impedance. Lamp-cord may have been a little lossy, but the early forms of coax which came into use a few years later was not all that good either!

Wayne W. Cooper, AG4R
Miami Shores, Florida

high performance small beams

For antenna sizes up to that of the quad or three-element Yagi, there is no direct connection between size and gain. This article shows how to design small beams without sacrificing gain.

In view of the urgent need for small beams, the number of them in use is remarkably small. In a recent sample of 14-MHz contacts with Australian amateurs, I found that eighty per cent of the stations worked were using quads or three-element yagis; the remainder, the ones with the biggest signals, had even larger beams, beyond the resources of most of us. There are, nevertheless, some smaller beams, such as the VK2ABQ and the capacitively loaded quad,[1,2] which usually give a good account of themselves. Of particular importance, in the present context, is the fact that some of the most outstanding signals observed during my 50 years as a licensed amateur have originated from stations using driven arrays with only two elements!

The significance of this lies in the ease with which the performance of such beams can be calculated,[3] and the fact that size comes into the calculations *only when estimating the efficiency, which remains high for element sizes and spacings down to approximately 3 meters (10 feet).* Translating this fact into practice has brought to light some interesting problems, in particular the problem of overcoupling. That is so basic a problem that the absence of references to overcoupling in the literature suggests that no serious attempt has, until this time, been made to make beams as small as possible.

To achieve this reduction in size, elements *must* be capacitively end-loaded, without resorting to lossy inductances. This involves large concentrations of metal, which, besides tuning the elements, also couple them very tightly together. Following the normal behavior of coupled circuits, a large secondary (parasitic element) current and a small primary (driven element) current are generated, and there is virtually no beam action. Fortunately, I have found it quite easy to overcome this problem by neutralization similar to that used in push-pull amplifiers.

Another inevitable consequence, when small size is combined with high efficiency, is narrow bandwidth; this makes it essential to use separate feeders for each element so that fine tuning can be carried out in the shack. Also, this arrangement brings with it other important advantages. Since the beam is instantly reversible, less time is wasted in beam rotation, and, because less than 180 degree rotation is required, you can use low-loss, open-wire feed lines as well as simpler and cheaper methods of beam rotation. The direct relationship between the size and performance of large beams is well known, and some readers may find it difficult to accept that a small beam can be as good as a big one, particularly if those readers have experience with typical small beams using large loading inductances. The claim is, on the face of it, improbable, and one might think that it could be dismissed by invoking some general scientific principle, as in the case of perpetual motion. Looking for such a principle we come instead to the surprising discovery that, although the gain of *big* beams is limited, there is absolutely no limit to the *theoretical* gain from an antenna, *provided it is small enough!* You would be justified in some skepticism at this point, since it turns out that gains much in excess of 6 dB are impracticable unless the boom length is increased to half a wavelength or more. It is

By Leslie A. Moxon, **G6XN,** 1 Stoner Hill House, Froxfield, Petersfield, Hants, England

possible, though, to go down in size to about 3 meters square (10 feet square) without dropping below about 4.5 dB gain.

If such statements are found puzzling, it is probably because of the failure in most of the literature to distinguish between two completely different methods of beam formation, additive and subtractive. The subtractive method is typified by the W8JK array, and, as I have shown elsewhere,[3,4] most amateur high-frequency beams can be regarded as derived from or related to this array. The radiation patterns are calculable without using any variable other than the direction. Therefore, the gain is independent of size, provided the efficiency remains high enough to ensure that most of the power is radiated. It is this constraint, allied with the need for adequate bandwidth, which in practice limits the gain of subtractive beams to about 6 dB.

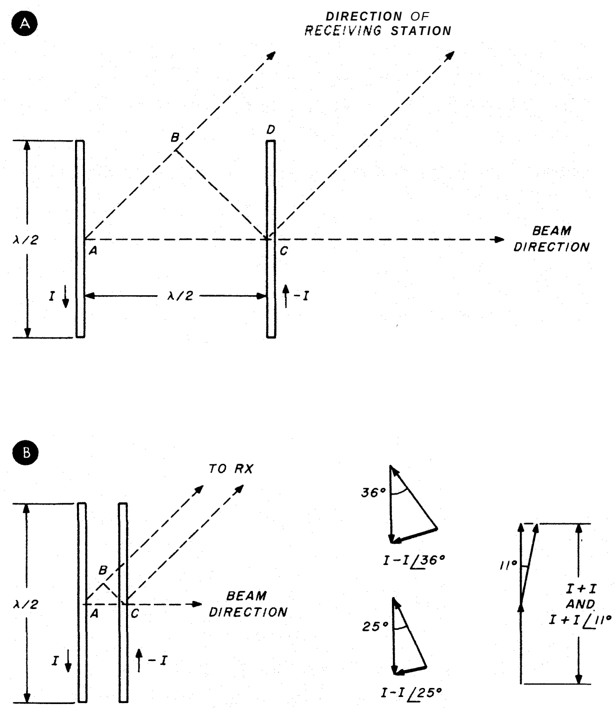


fig. 1. In (A), the elements are fed 180 degrees out of phase so that the fields add in phase along the line at right angles to the elements. A receiver located 45 degrees off the main beam sees a phase difference corresponding to AC-AB, *i.e.*, 54 degrees, causing a drop of 4 dB in signal level in addition to the 3 dB which would be expected for a single element. Thus, the use of two elements has produced a narrower beam by virtue of the wide spacing. For subtractive gain, as shown in (B), the elements are closely spaced. Radiation would be cancelled except for a small phase shift of which the maximum value corresponds to the distance AC, or 36 degrees for a spacing of $\lambda/10$. At 45 degrees to the beam, the phase shift is reduced by the ratio of AB to AC (11 degrees), which translates into a 3-dB drop in signal level. If the fields are additive in the wanted direction, the 11-degree shift is of no consequence and the radiation pattern becomes that of a single dipole.

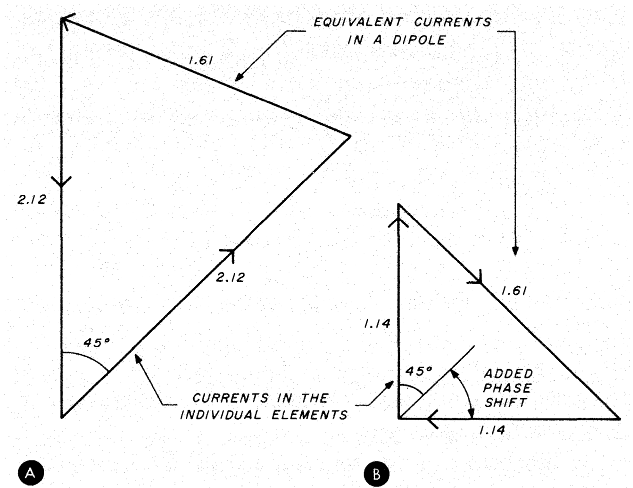


fig. 2. Method for estimating gain from the radiation resistance. Diagram (A), which is drawn to scale, shows how the W8JK element currents of 2.12 amps "add" to give the equivalent of 1.61 amps in a single element. The currents are correct for a radiated power of 72 watts. This would produce a current of 1 amp in a dipole, so that the voltage gain is 1.61, *i.e.* 4.2 dB. (B) shows how if one element is given a phase lead equal to the spacing, only about half as much current is needed to produce the same field strength. In each case, the current is obtained from $I = \sqrt{P/R}$.

The following discussion is intended to provide further insight into the problems of designing small beams, so that you can make your own choice from the available options and then design the "best-possible" beams for the space and facilities you have available. Practical details and some performance data are given for a number of designs, but these are intended as guidelines and not as blueprints. Although the feasibility of the 3 meter square (ten foot square) design has been proved, my own preference, given the space, is a 5.2 meter square (17 foot square) 3-element design, which, besides providing slightly more gain and greater bandwidth, lends itself to a particularly light form of construction and easier methods of achieving multiband operation.

It is a strange paradox that the main advantages of increased size relate to operation at the higher frequencies, to the extent that considerable extra gain can be obtained at 28 MHz by using full-size, 14-MHz elements. For no reason that I can discover, this advantage is usually thrown away by using traps or nesting to "cut the elements down to size" at the higher frequencies.

gain variation vs antenna size

Gain results from concentrating the radiation in one particular direction at the expense of other directions. This can be achieved by arranging more-effective addition or, as with the W8JK, less-effective cancellation of radiation for the wanted direction. Addition requires wide spacing, as in fig. **1A**, where the elements are arranged so that their fields add in phase for the wanted direction. Viewed at an angle of 45 degrees to the line of fire, the elements are closer together by 0.29 wavelength, which translates into a phase shift of 104 degrees and a corresponding drop of 4 dB in signal level. This drop of course, is in addition to the $\cos\theta$, or "angle to the wire" effect, which applies equally to beams and dipoles and amounts to – 3 dB at 45 degrees. But for the wide separation of the elements, there would be no phase shift, no narrowing of the radiation pattern, and hence no gain.

On the other hand, given an array of n elements sufficiently far apart for mutual interaction to be ignored, we can provide each $1/n$ of the power. Since the voltages at the receiver add in phase, there is a power gain equal to the number of elements. Because the antennas have to be separated by at least $\lambda/2$, it turns out, not surprisingly, that gain is proportional to size. During reception, such an antenna collects most of the energy contained in the volume of space which it occupies, thus giving rise to the concept of aperture, for which it is sometimes claimed "there is no substitute." It is perhaps difficult to conceive that a tiny beam can collect as much energy as a big one, but the explanation lies in the high Q of the smaller antenna. Just as tuned circuits couple together more tightly when their Q is increased, so reduction in fhe radiation resistance is accompanied by "tighter coupling" of the antenna into the surrounding space.

To understand the mechanism of small beams, the W8JK system shown in fig. **1B** is the easiest starting point. Radiation would be completely cancelled in all directions except for that caused by the phase shift, which results from the elements not being exactly the same distance from the receiver. Field strength is proportional to the apparent separation AB, which, like the radiation from a single element, varies as $\cos\theta$. In simpler language, the field strength, at a given distance and direction, is proportional to the apparent length multiplied by the apparent separation of the elements. The separation factor applies equally in a plane at right angles to the diagram, so that the radiation pattern of a horizontal W8JK array is given by $\cos^2\theta$ in the horizontal and $\cos$ 0 in the vertical plane. For a single horizontal wire, the pattern is $\cos$ 0 in the horizontal plane and an omni-directional pattern in the vertical plane. The important point to note is that none of these patterns contain any seference to the dimensions, though we are of course assuming them to be small. In fact, the conclusion that gain is independent of size turns out to be accurate within about 0.5 dB for spacing up to $\lambda/4$ and element

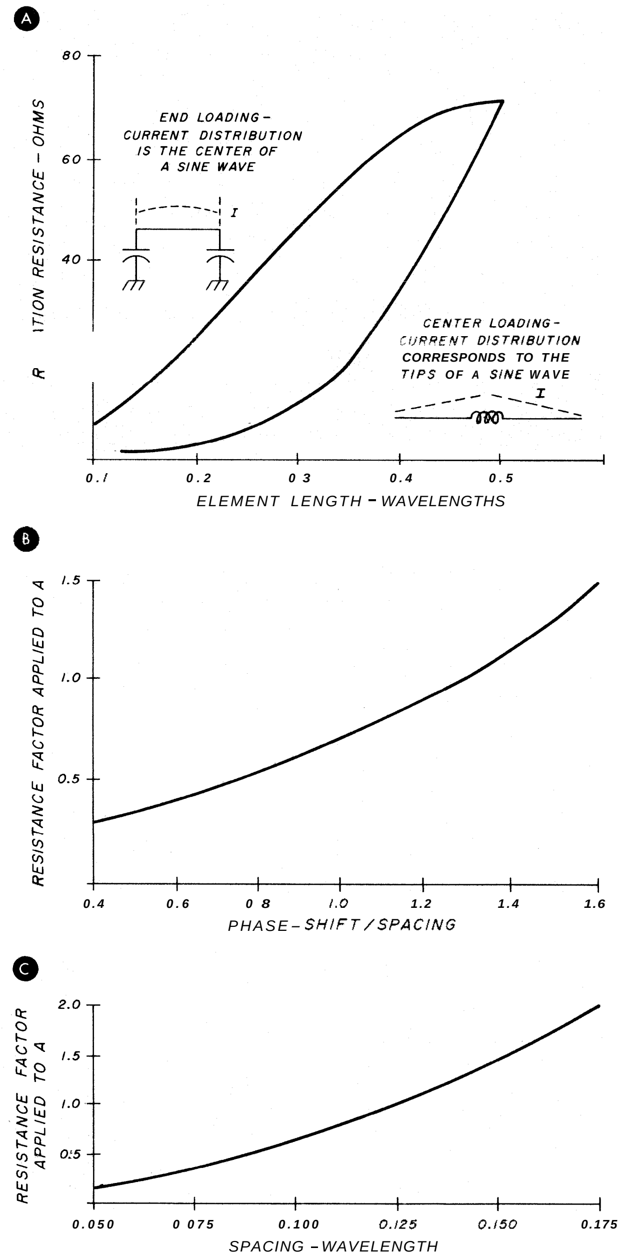


fig. 3. Graphs for determining the radiation resistance of short elements. (A) shows the variation of radiation resistance with both end and center loading. The multiplying factor [see (B)] is used to give the sum of the radiation resistance for a pair of elements with $\lambda/8$ spacing. For miniature beams, such as those shown in figs. 9 and 11, this approximates to the impedance seen by the feeder going to the front element. The value for spacing is converted to the same angular units as the phase shift, which is relative to the condition 180 degrees out of phase. The third chart, (C), is an additional factor which is used if the spacing is other than $\lambda/8$.

lengths up to just over $\lambda/2$. Up to these sizes, each element behaves in the same way, *i.e.*, as a "point source" of energy. It is only when an element is very large that appreciable extra gain, or directivity, can be expected, as for example when a 14-MHz element is used as two half-waves in phase on 28 MHz. It should be obvious from inspection that quad loops, even full-sized ones, do not meet this condition,

To prove the point, there is no harm in treating the loops as a pair of stacked dipoles. Various handbooks provide data indicating that the stacking gain for $\lambda/4$ spacing is only 1 dB. However, some of the 1 dB is lost by bending over the ends, since a half-wave dipole has only a small gain (up to 0.4 dB) when compared with shorter dipoles; a further small amount is lost due to radiation off the ends. Even more is lost when parasitic elements are added, since the stacked dipoles then become stacked yagis, and, according to the usual rules, the higher the gain of individual antennas the further apart they have to be placed in order to achieve an appreciable stacking gain.

It is unfortunate that many wild claims have been made for the quad, some of them involving professional journals and computer studies. It needs to be stressed that measurements are very difficult and computers need to be asked the right questions. The habit of accepting figures without checking them against ordinary common sense is not confined to the Novice! In fact, as I have found, better low-angle gain is obtained by omitting the lower halves of quad loops and using the upper halves as inverted V or U elements. This increases the mean height by 2.4 meters (8 feet) at 14 MHz; low-angle gain for a flat, unobstructed site being proportional to antenna height, this more than offsets any slight loss of free-space gain for heights up to about 21 meters (70feet)!

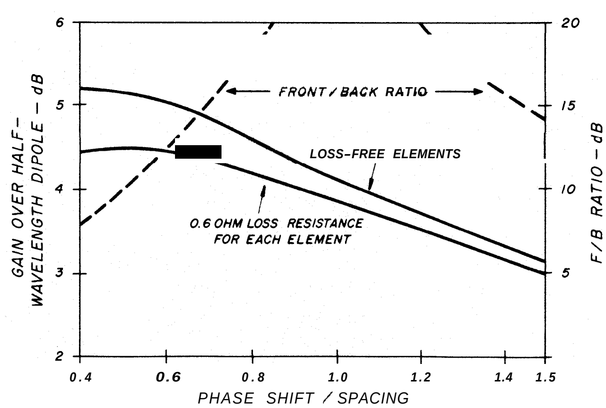


fig. 4. Gain and front-to-back ratio for the two elements, with and without losses. Loss resistance per unit of conductor length is inversely proportional to diameter and directly proportional to the square root of the frequency. The figure of 0.6 ohm is based on a half-wavelength of no. 12 AWG (2.1-mm) wire having a sinusoidal current distribution at 14 MHz. The element currents are assumed to be equal. The upper gain curve is for loss-free elements, while the lower is for 3.7-meter (12-foot) elements spaced at 2.4 meters (8 feet).

The real advantage of the quad is the large amount of extra gain (3-4 dB) obtainable by using the 14-MHz elements at 28 MHz with suitable resonators so that they become a "Bi-Square," but this is rarely exploited. Another frequently made claim, that the quad provides better DX signals by "lowering the angle of radiation", is also without foundation; the lobes of practical antennas are too broad to discriminate between direct and ground reflected waves which must always interect in the same way, resulting in a loss at low angles unless the antenna is very high or the ground sloping. To obtain the effective gain at a low angle of radiation, the free-space gain is multiplied by the ground reflection factor, which, for horizontal antennas, depends only on antenna heights and radiation angle, being the same for a quad, yagi, dipole, small rhombic, or even a minibeam!

The 4-dB gain of the W8JK can be estimated very roughly from the 3-dB widths of the radiation patterns discussed earlier, or, more accurately, from mutual-impedance data. For $\lambda/2$ dipoles spaced at $\lambda/8$, mutual impedance is 64 ohms, which has to be subtracted from the self-resistance of 72 ohms so that for each element the radiation resistance is 8 ohms. If the power available is 72 watts, you would have a current of 1 ampere in an ordinary dipole, or 2.12 amps in each of the W8JK dipoles. The phase difference due to spacing is 45 degrees, and, by "completing the triangle" as in **fig.** 2, you'll see that the distant field is the same as that which would result from 1.61 amps in the dipole, *i.e.*, there is a voltage gain of *1.61/1* or 4.2 *dB*.

Now, if the phase of the current in one element is advanced by an amount corresponding to the spacing, the total phase shift becomes zero for one direction and is doubled for the other direction. The resultant unidirectional (cardioid) pattern would require only about half the current to produce a given field strength. It happens (by coincidence) that the gain is unchanged, so that the effective radiation resistance has been multiplied by about four, proving that the beam can be made a lot smaller with the same results. **Fig.** 3, adapted from references 3-6, shows how radiation resistance varies with element length, spacing, and phase angle, from which it can be seen that *provided you shorten the dipoles by sacrificing the ends and not the middle,* the length can be reduced to 37 per cent of a half-wavelength before the radiation resistance drops to its W8JK value, giving a length of only 37 meters (12 feet) at 14 MHz. This reduction in size could be carried out by taking the W8JK beam and folding its ends to fit the available space. Whether or not the example can be improved upon, it does demonstrate that *if* the W8JK beam works properly you could expect to be equally successful at 14 MHz with a 3.7 x 2.4 meter (12 x 8 foot) miniature beam which has been adjusted to give a large front/back ratio.

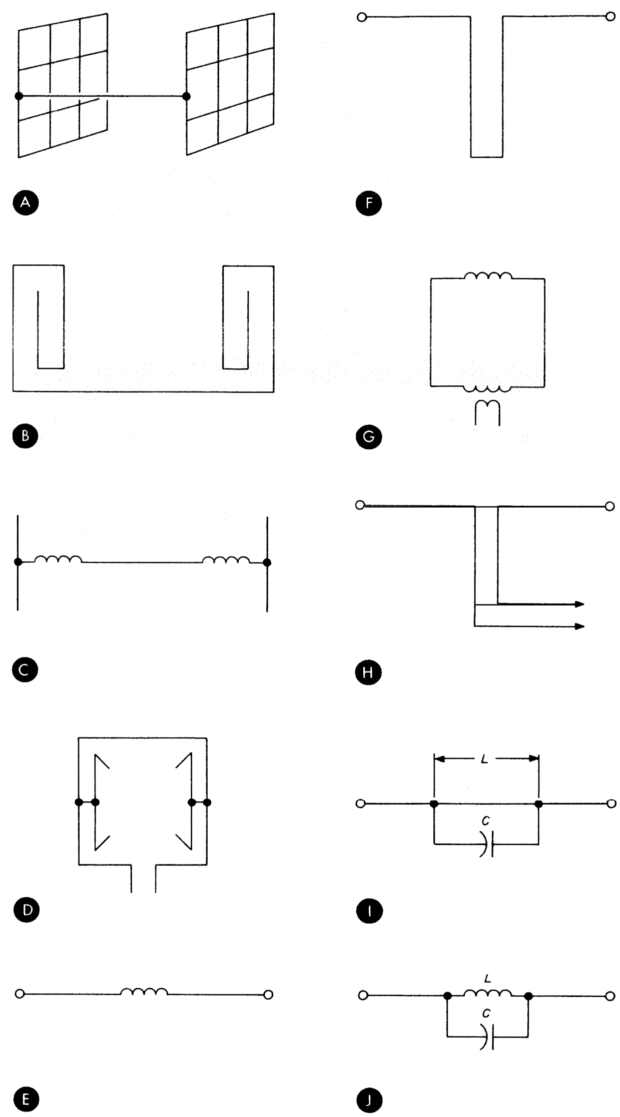
Thus, changing the size does not alter the gain, but, obviously, if the length or spacing is halved, the current must be doubled to maintain a constant signal, so that the radiation resistance has been divided by four. In the same way, you could alter the shape to just under 3 meters square (10 foot square) without affecting matters in any way, save that the diameter of turning circle is slightly decreased and it becomes easier to handle the ends.

Whatever shape is selected, radiation resistance may be obtained by multiplying the resistances plotted in fig. **3A** by the factors obtained from figs. **3B** and 3C. This is not quite the full story, since, as demonstrated by fig. 3 and **4,** an extra dB of gain results from using an intermediate value of phase shift. However, this halves the radiation resistance and doubles any power loss. From the lower gain curves in fig. **4,** it can be deduced that for a 3-meter-square (10-foot-square) 14-MHz beam, using no. 12 AWG (2.1-mm) wire elements with bent ends, a net gain of 4.5 dB should be obtained after allowing for resistance losses. There may be a slight additional loss due to the "short dipole" effect mentioned, but this is caused by a small difference in the amount of endwise radiation and is much less in the case of a beam.

To put this into perspective, the maximum gain theoretically possible for a three-element parasitic array on a quarter-wavelength boom is 7.5 dB, but in this case , the radiation resistance is only 4 ohms (even less than that of the small beam), and the maximum gain normally found in practice7 is less than 6 dB. Theory goes on to predict a possible gain equal to the square of the number of elements,[3] but before getting far along this road you're faced with huge currents and voltages, infinitesimal radiation resistances, and microscopic tolerances to the extent that the 6-dB figure is unlikely to be exceeded in practical rotary beams for 14 MHz without increasing the boom length to at least $\lambda/2$, which is sufficient to provide some "additive" gain. Reducing the size below 3 meters square (10 feet square) leads to a similar rapid increase of practical difficulties. And, as I shall point out later, there are substantial advantages, particularly for multiband operation, in an increase to about 5.2 meters square (17 feet square).

Limitations on the reductions possible in antenna size result from the following:

1. Drop in efficiency, *i.e.,* the radiation resistance becomes comparable with the losses. This is basic and imposes a well-defined, practical limit.



A Capacitance plates consisting of wire grids
B Half-wavelength elements with folded ends. The length must be increased slightly to maintain resonance.
C Small capacitance hats — the effective capacitance is enhanced by near resonance with the inductors
D A loop with capacitance hats. This is equivalent to a stacked pair of B-type elements with their ends in contact.
E Center loading with an inductor
F A half-wavelength element with a folded center. This is similar to E with a stub instead of the coil. However, the ***R*** values from **fig. 3A** are transformed by the stub to give an even lower value at the closed end of the stub.
G Loop equivalent to E
H Resonant feeders
I Version of a two-band element as used by DL1FK. The capacitor tunes the inductance of the center of the radiator to increase its effective value at the lower frequency. Series resonance of the capacitor with its connections shortens the electrical length for the higher frequency.
J Lumped circuit equivalent of **I** as used in one form of the G4ZU minibeam.

fig. 5. This figure shows ten different methods for loading short elements. The merits of each type are discussed in the text.

2. Narrow bandwidth — the acceptable lower limit depends on the skill of the designer and the skill plus patience of the operator.

Short elements have to be loaded to bring them to resonance. The aim must be to keep the radiation resistance high, which requires end loading, and the loss resistance low, which rules out the use of lossy devices such as coils or long resonant feeders. **Fig. 5** shows ten methods that have been used, with only **5A** through **5D** meeting the radiation resistance requirement. The others tend to be more convenient, but use the triangular tips instead of the center of a sine-wave current distribution, thus halving the average current and dividing the radiation resistance by four, besides requiring an inductive and therefore relatively lossy loading device in the center. Of these methods, **5F** is open to least objection, since stubs have less resistance than coils; **5H** is particularly bad, since it multiplies the losses of **5F** by the total number of half-wavelength in the resonant system. Methods **5I** and **5J** are used for multibanding,[8] which necessarily adds to the losses because of circulating currents in the resonators. Methods **5E** to **5H** are therefore applicable only for very modest reductions in antenna size.

Of the remainder, **5C** makes use of small end-loading capacitances which have their effective values greatly inflated by near resonance with the inductors. In one typical design, the total inductance is about 14 μH, so that for a Q of 200 the loss resistance would be 6 ohms and more or less comparable with the radiation resistance, the element lengths being 3.7 to 4.3 meters (12 to 14 feet). **Fig. 5A** is ideal insofar as there is negligible loss in the capacitance, but it is difficult to achieve enough loading, so that some inductance on the lines of **5C** is likely to be required; in this case, however, it could take the form of short stubs and the losses would be minimal. Arrangement **5B,** consisting of a half-wavelength dipole with its ends bent over, is very similar, but simpler and equally efficient. The bent ends are not, of course, pure capacitances and there are losses, but, as has been seen, this method allows high efficiency to be maintained down to very small beam sizes. Efficiency can be estimated with the help of **fig. 4.**

Another efficient type of small element is the capacitively loaded loop shown in **fig. 5D.** Loops with 3.2-meter (10.5-foot) sides have been used by G3YDX for a 14-MHz, two-element beam,[2] and a design patented by G3IMX achieves two-band operation by using traps to remove the loading at the higher frequencies. For a single 3.7-meter (12 foot) loop, the radiation resistance is 75 ohms. About 20 ohms could be expected for a 1.8-meter (6-foot) loop, the loss resistance (referred to the feedpoint) probably being about 2 ohms. My own attempt to compress a 14-MHz quad into a 1.8-meter (6-foot) cube was unsuccessful, but I still think it might be achieved with rigid, all-metal construction and inductive loading stubs, plus sufficient ingenuity. The problem is to dispose of enough loading in the space available. In this case, neutralization will certainly be required as the lower physical limit is approached. Neutralization will not be required with 3-meter (10-foot) loops, as sufficient loading can be achieved within the plane of the loops without bringing ends into proximity.

It is important in all cases to arrange the loading so that wires carrying appreciable current are not doubled back in a direction parallel to the driven element, in which case they subtract from the wanted radiation, bringing the radiation resistance "down with a bump!" This constitutes one of the main problems of construction.

bandwidth

The bandwidth of a dipole may be roughly estimated from the radiation resistance (R) and the characteristic impedance (Z), which depends on the ratio of length to diameter. Each per cent of detuning produces a reactance of $0.015 \cdot Z$ ohms; and the 3-dB bandwidth points (corresponding to an SWR of 2:1) may be found by equating this to R. Bandwidths are plotted in **fig. 6** for 14-MHz dipoles of various wire diameters. The figures should be valid, as an approxi-

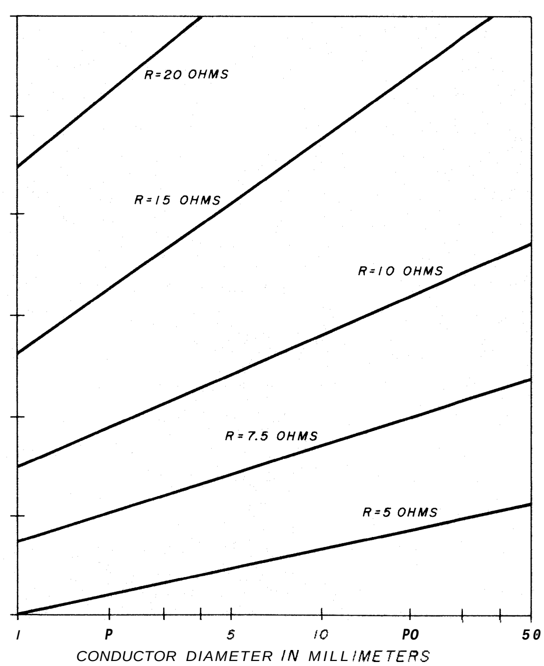


fig. 6. Bandwidth of folded $\lambda/2$ wires at 14 MHz for different values of radiation resistance. plus loss resistance. The curves are calculated for an SWR of 2:1 at the band edges, neglecting any change in Z_a caused by folding. For a coupled pair of elements, the bandwidth for a given R is increased slightly.

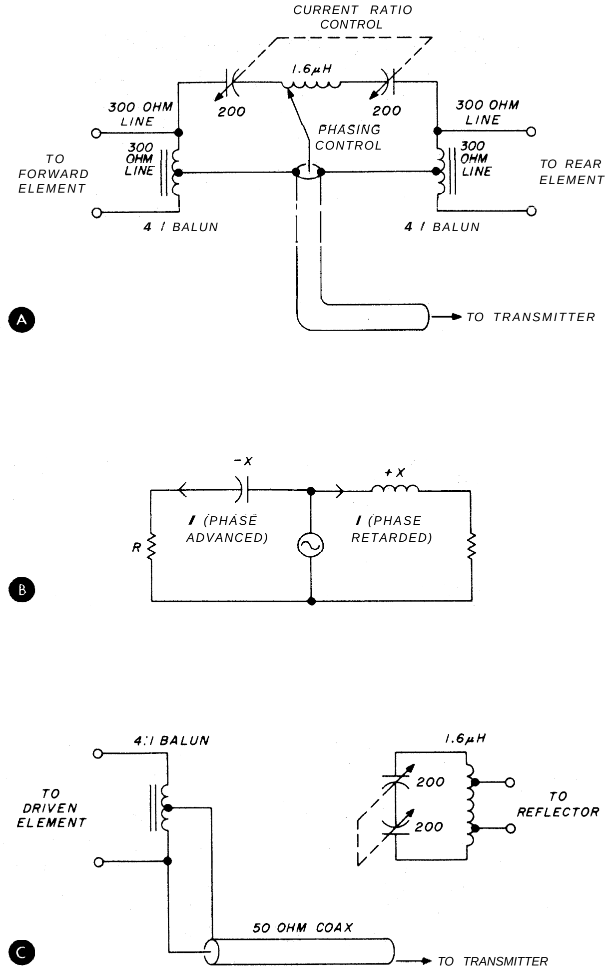


fig. 7. The top illustration shows a practical arrangement for a 14-MHz phasing control. The beam reversing switch, though not shown, interchanges the feeders. This circuit may be usable to 28 MHz, but the inductance should be reduced to about 0.8 μH; some experimentation with values may be needed. The principle of operation is demonstrated by (B); as shown the load currents are equal and the ratio of X/R determines the phase shift. If, as is more usual, the loads are unequal, current equality may be achieved by adjustment of the reactances provided in (A). The components of (A) can also be rearranged as a tuner for a parasitic type of beam [see (C)].

mation, even when the dipoles are folded as in figs. **5B** or 5F. When two elements are assembled to form a beam, the bandwidth is reduced because of the drop in R, though, for a given value of R it is increased and could be almost doubled by virtue of the "coupled circuit effect."

It is important to distinguish between three kinds of bandwidth:

1. The pattern bandwidth – the bandwidth over which the radiation pattern and gain are satisfactory

2. The bandwidth for satisfactory SWR

3. The bandwidth over which the antenna is usable subject to adjustments that can be carried out without leaving the shack.

To illustrate, one element of the first version of my small beam had an intrinsic bandwidth (as calculated) of about 450 kHz. The completed minibeam, using a parasitic reflector, had a useful bandwidth of only about 200 kHz; at one edge the front/back ratio had dropped below 8 dB and at the other edge the gain had dropped by 1 dB. in wet weather, due to an inadequately treated bamboo spider, the tuning shifted 200 kHz low, outside the desired range. In contrast, with the second version the useful bandwidth was only 130 kHz. However, it was desirable to keep within about 20 kHz of the tune-up frequency. But, by using separate feeders to each element, the beam was tunable to any frequency in the 14-MHz band either as a driven array or a driven element plus reflector. So, for practical purposes, despite the smaller intrinsic bandwidth, the useful bandwidth was greater.

With the parasitic reflector, the phase-shift is related to the ratio of reactance to the total resistance, R, of the reflector, and, in addition, must meet the requirement for reasonable gain in accordance with fig. **4.** Although the actual relationship is much more complex, a rough idea of the bandwidth for useful performance can be obtained as follows. Referring to fig. **4,** we might decide that performance is acceptable for values of phase shift from one half up to one and one half times the spacing, which means from 25 to 75 degrees in the case of a spacing of 3 meters (10 feet). At 25 degrees, the front-to-back ratio is down to 10 dB and the radiation resistance (fig. **3**) is getting very low, while at **75** degrees the gain is down to 3 dB. Phase shift is partly due to X, and partly to the mutual reactance between the elements, which, in this case, takes care (more or less) of the required mean value of 50 degrees. The maximum allowable shift of ±25 degrees occurs when X/R is roughly equal to this angle (measured in radians), *ie.,* $X = R/2$. From fig. 3, R = 15 *ohms* and for $Z = 1000$ *ohms* (a typical value) a reactance $R/2$ results from ±0.5 per cent detuning, *ie.,* the bandwidth is 1 per cent or 140 kHz, in good agreement with the observed figure of 175 kHz (*i.e.,* 14.300-14.125) from fig. **10.**

The first version of the small beam was 3.7 meters square (12 feet square), which would be expected to double the radiation resistance and bandwidth; in this case, too, the calculations are in line with the observed performance and there was good agreement in regard to the observed bandwidth for a 20-dB front-to-back ratio. From fig. **4,** the theoretical 20 dB points correspond to a change of ±18 per cent in phase angle, or 36 per cent in reactance, *i.e.,* 6 ohms, which translates into a bandwidth of 0.4 per cent, or 56 kHz in comparison with a measured figure of 55 kHz; however, this must be regarded as coincidental, since both figures are very rough.

phasing

It is essential for the currents in the elements to be equal (or nearly so), as well as correctly phased. With parasitic arrays there is usually no independent control of these quantities, though equality tends to be achieved in the case of quad reflectors. With $\lambda/2$ dipole elements the inequality, though it degrades the front/back ratio, does not upset gain to a serious extent. In contrast, the small-beam elements behave as overcoupled circuits resulting in large ratios of reflector to driven element currents.

Neutralization allows the coupling to be reduced to any desired extent, so that it can be adjusted in conjunction with the tuning of the reflector to obtain equal currents with any required value of phase shift. In this way, very deep nulls can be obtained in one or more back directions without resorting to driving both elements. This is subject to a number of assumptions: For example, the neutralization, which has to be adjusted at ground level, must remain "right" when the antenna is raised to its full height, at all frequencies, and in all types of weather.

A driven arrangement with provision for adjustment of phases and amplitudes is obviously more versatile and allows compensation for considerable errors in tuning or neutralization, but the methods usually employed for driven arrays are based on false assumptions that are particularly disastrous in the case of small beams with reactively coupled elements. It is usually assumed, for example, that $\lambda/8$ of line provides a 45 degree phase shift – which is true if the lines are perfectly matched. But often there is little attempt at matching, which will, in any case, be upset if the phasing is adjusted. A further difficulty is that for the elements to have equal impedances the mutual coupling between them must be a pure resistance, a condition that applies in practice only for straight $\lambda/2$ elements spaced just over $\lambda/8$. This, incidentally, is an important special case since it is then easy to calculate gain and radiation resistance.[3,5] Having established that gain is independent of size, the numbers obtained can be applied to other sizes on the basis of **fig.** 3.

If there is capacitive reactance coupling, as happens with wider spacing or miniature elements, the radiation resistance of the reflector may be zero or even negative, in which case its feeder returns more power to the transmitter than it receives! Matching is clearly impossible under these conditions. The method I use, which can take various forms,[6,9] is based on resonance; in **fig.** 7, the two feeders are connected through a series-resonant circuit and the complete system, minus the connection to the

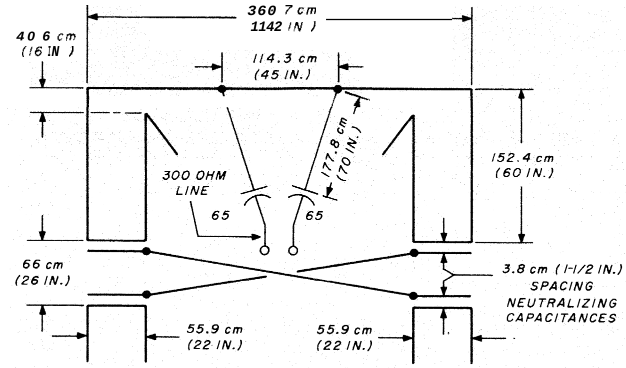


fig 8. Original version of the small beam showing one element plus the neutralizing wires. The elements are identical, with a wire length of 11.4 meters (37.3 feet). For initial adjustment, one element without the feeder may be tuned as a parasitic reflector, in which case the length is about 10.9 meters (35 feet, 10 inches). The elements and neutralizing capacitors are made from no. 14 AWG (1.6-mm) wire, but the delta match and neutralizing cross connections can be made of a lighter gauge wire. The spider projects beyond the elements, which are suspended by lengths of polyethylene cord.

transmitter, is made resonant. Phase-shift is obtained by off-center connection of the feedline from the transmitter. If necessary, element currents can be made the same by detuning the series circuit so that one element or the other is brought closer to resonance. Matching should be carried out to reduce the SWR in the line to the forward element, a high SWR in the reflector line being of no importance as there is very little power transferred along it. Crossover of the elements as with the W8JK, though normally essential with full-sized elements, is sometimes not required when coupling to highly reactive elements. Using this arrangement, it has been possible to work with different or even unequal lengths of feedline, and to compensate for quite large errors of adjustment, though never to the extent of being able to use it as a substitute for neutralizing. The beam can be reversed either by interchange of feeders or moving the coil tap. It is possible, in principle, to null out from any given direction "off the back." A curious feature of reactive coupling is that, despite the remarkable effect it has on relative impedances, the *sum* of the two radiation resistances is not affected; for **fig.** 3 to have universal application, it was therefore necessary to plot the sum and not the individual resistances. The sum is in any case a more useful figure for the present purpose since it is into this value that the feeder for the front element has to be matched (so that the antenna impedance is "repeated" in the shack). It is often possible, particularly if the feeder length is a whole number of half wavelengths, to use the phasing circuit with very little modification as a remote tuner for driven element and reflector operation (see **fig. 7**); this reduces the

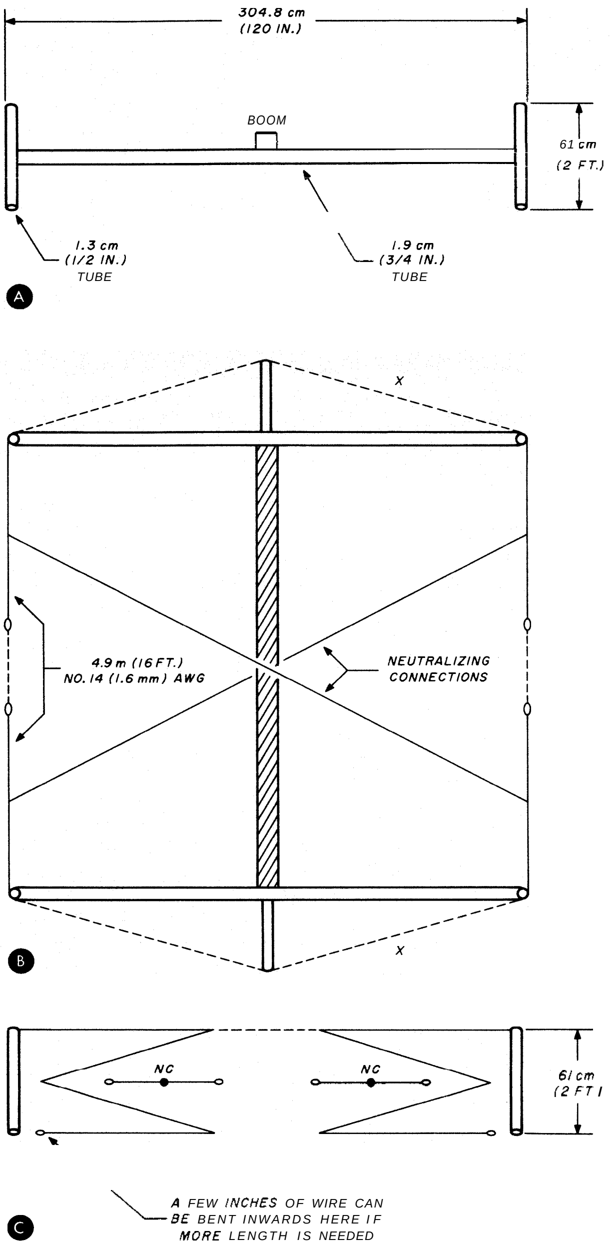


fig. 9. Construction details for a 3-meter-square (10-foot-square) beam for 14 MHz. The front, plan, and side views are shown in (A), (B), and (C), respectively. In (B), the bow-string-type arrangement is used to counter the pull of the loading wires and to allow the use of lightweight elements. Each neutralizing capacitor, as shown in (C), is approximately 0.6 meter (1.9 feet) long. If greater wire length is needed, additional wire can be added as indicated in (C).

number of knobs and has proved more convenient in practice, provided the neutralization is reasonably accurate. It has been found that, although small differences can be compensated, care in making the elements identical pays big dividends in convenience of operation, since it allows beam reversal with no change of tuning or matching. One can tune for a minimum on either the wanted station or an interfering signal from the back direction, using the reversing switch as required.

practical designs

Several construction methods have been used.[10] Fig. **8** gives the dimensions of my first smal! beam, the elements being suspended by thin polyethylene cords from a six-armed spider made from bamboo garden canes which extended just beyond the wire. The two central arms were required to counter the inward pull of the folded sections. One element was driven using 300-ohm line with a delta match having series capacitors to tune out the reactance. I tried to tune the other for a null in the back direction, but this was barely perceptible. Current in the reflector was found to be higher than that in the driven element, so the neutralizing capacitance (formed by parallel wires) was increased by reducing the spacing, retuning the reflector, and adjusting the driven element for lowest SWR at each step. Eventually current equality and a good front-to-back ratio was achieved. At a height of 16.8 meters (55 feet) the antenna was compared with a two-element quad at 13.7 meters (45 feet) and appeared to be at least as good.

Mention has already been made of bandwidth and wet weather problems. It quickly became obvious that two feeders would have to be used. To save cost, I used the transparent, plastic-type of 300-ohm line; lacking previous experience with it, I was unprepared for the rapid deterioration, which caused a lot of confusion. Later, good results were briefly obtained using a pair of 600-lines.

However, mechanical problems arose and the next step was a new design using metal construction with 3-meter (10-foot) elements made from 19-mm (0.75-inch) tubing. Short vertical rods at the ends of the elements (see **fig. 9**) act mainly as brackets for supporting wire zigzags. I found that, because of limited space, almost as much wire was needed as in the original design so that the use of tubing provided no electrical advantage apart from getting rid of "wet weather" effects caused by the bamboo. To avoid the need for heavy-gauge elements, their ends were guyed back with polyethylene cord to the boom extensions, thus countering the inward pull of the loading wires. Two open-wire feedlines were used.

Some typical measured performance figures are shown in **fig. 10.** On-the-air performance was down compared with earlier results, on average about 2 dB; this might be due to the smaller size (radiation resistance being halved) in addition to closer proximity to tree branches, the mast being slightly shorter due to breakage and splicing! Another uncertainty arose from loss of the quad as a standard of reference; the 3-element beam, though equal to the quad in performance, was so placed that it tended to screen the smaller beam. I could not be certain that rotation to an end-on position was removing all this effect. From this, I felt that the best test for the small beam would be as a direct replacement for the larger one. Now, however, I used a new multiband version, the disposition and loading of the elements for 14 MHz being as shown in fig. 11.

For such tests my yardstick for many years has been another station using a TH6 at 11.6 meters (38 feet) and with a slightly better location for VK (long path) over which most of the tests were made. With the small beam at 14.6 meters (48 feet), I was able, for the first time on record, to equal the signals of the other station on three consecutive days. Normally with a quad at 12.1 meters (40 feet) I should be down in most cases by 1/2 to 1 S-unit. Added to the initial results of the first small beams, this would provide substance for a claim that small beams, given a few feet of extra height, are actually better than big ones! Unfortunately, such a claim could not be supported by theory or common sense, but I was confirmed in my belief that the small beam can be made fully competitive.

Any type of feeder can be used, but, because of its lightness, good-quality 300-ohm line can be recommended as an alternative to open-wire line. I have also used 50- and 75-ohm coax with 4:1 baluns and delta matches identical to the one shown in fig. 8.

three-element beam

A starting point for this development was the original VK2ABQ design[1] in which a quad loop lying on its side is converted into a two-element beam by insulators in the sides. The rather wide spacing ($\lambda/4$) unfortunately means that performance is slight-

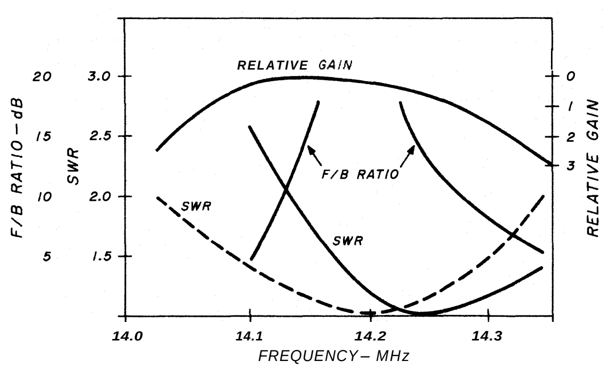


fig. 10. Typical performance of the 1-meter-square (10-foot-square) beam. All parameters were measured with the beam at a height of 2.1 meters (7 feet) and the reflector tuned for 14.2 MHz. The dotted curve shows the improvement in SWR after adding 10 cm (4 inches) to each end of both elements and retuning the reflector for best front-to-back ratio at each frequency. Note: The gain curve is listed as dB down from the maximum.

ly worse than the best that can be achieved with two elements, and it seemed logical to me to put a third element in the middle. No increase in the diameter of the turning circle dictated a maximum element length of 7.3 meters (24 feet). With some end loading by vertical rods and the use of linear resonators, I was able to design a triband element (see fig. 12) having almost as much radiation resistance as a full half-wavelength at 14 MHz, higher radiation resistance than a normal element at 21 and 28 MHz, a small amount of extra gain at 28 MHz, and no trap losses.

Tubing was used for this element, but retention of bent wire elements for directors and reflectors made engineering sense, since these elements have less current flowing in them and a lot of cost and weight could be saved. The next step was to eliminate the quad spider by using the driven element as one diagonal of a square and making it support the ends of the parasitic elements which were now V shaped instead of U shaped and filled in the sides of the square as shown in fig. 12C. The other diagonal became the "boom" and consisted of two 2.4-meter (8-foot) bamboo garden canes joined by a length of aluminum, supporting the apex of the Vs.

V-shaped elements are not suitable for multibanding by the methods used for the driven element, and ordinary traps would be unsuitable for suspension in wires supported by such a light framework, as well as being too lossy. I therefore used separate elements for each band with their ends strung out to different points on the vertical loading rods of the driven elements in the hope that this way they might not get entangled! Though somewhat haywire, this arrangement has unexpectedly survived two severe storms and, although it needs a lot of tidying up mechanically, comes fairly close to the ultimate in lightness combined with high performance. Details in fig. 12 should assist any reader wishing to experiment on similar lines. Despite the larger size, the overcoupling problem was not completely avoided and there was some slight difficulty in adjusting the 14-MHz reflector, but this was overcome by increasing the spacing slightly as indicated in fig. 12C.

The linear resonators work as follows.[9,11] The inductance L_{AB} of the central portion (AB) of the element is tuned to resonance by C1 so that it acts as insulator for 28 MHz, making the element two half-waves in phase. At 21 MHz, about half of C2 serves to eliminate the inductance by tuning it to parallel resonance, with the remainder used to series resonate the inductance (L_O) of the outer portions of the element, L_{AB} and L_O chosen to be roughly equal. At 14 MHz, the capacitors are virtually "not there," their values at 21 and 28 MHz having been inflated by series resonance with their connections, and consequently AB no longer acts as a tuning inductance

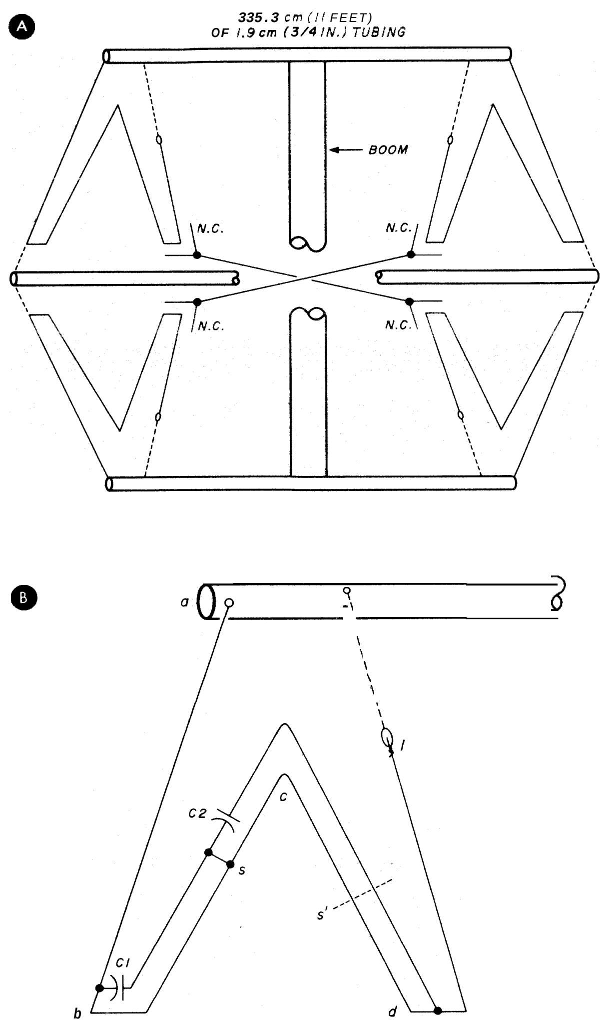


fig. 11. This is an alternative form of the beam shown in fig. 9. In this case (A), the elements have been folded in a horizontal plane to avoid any vertical projections. The metal supporting strut in the center of the boom also serves as a 21128-MHz driven element in the multiband version. The detail shown in (B) illustrates the arrangement for multibanding with linear traps. Moving the shorting bar from S to S' results in an improved trap for 28 MHz, but separate parasitic elements must then be used for 21 MHz. For monoband 14-MHz operation, the capacitors are omitted, but extra wire may be needed to obtain sufficient loading. The wire lengths, a, b, c, *d*, and e, are all equal and approximately 35.6 cm (14 inches). C1 and C2 use the wire segments bs and *sd* to form the linear traps for 28 and 21 MHz respectively.

but reverts to its normal role as the "middle portion of a dipole." This process is accelerated by mutual coupling between the capacitive and inductive branches of the resonators. Values for C1 and C2 are critical and I used selected capacitors slightly lower than the required values (about 10 pF and 20 pF respectively) making up the difference with short, open-wire, adjustable stubs.

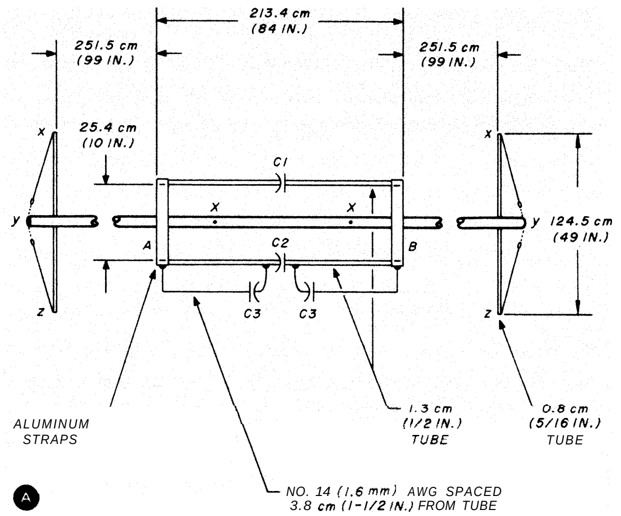
Although linear resonators have little effect at lower frequencies, the 21-MHz resonator in its simplest form has disastrous consequences at 28 MHz, where it looks like a rapidly varying inductance that results in a very narrow bandwidth. This problem is overcome by the capacitors (C3) which form additional linear resonators tuned somewhat above 28 MHz so that the effective inductance at 28 MHz is greatly increased without affecting 21 MHz. C3 is not critical. Using a grid-dip oscillator, the driven element should be tuned by adjustment of C1, C2, and the length of the loading wires to provide the required resonant frequencies, taking care to avoid spurious resonances. Using a pick-up loop with a rectifier and meter as a simple rf current indicator, currents in L and C1 at 28 MHz will be roughly equal and some four times greater than the loop current observed at point D, (see **fig. 12C**). At 21 MHz, currents in C2, L, and just outside AB can be expected to be roughly in the ratio 3:2:1. This is not critical, but any major departure could indicate a spurious resonance. It is important to ensure symmetry and this requires a balanced feed, not a gamma match.

When tuning three-element antennas, there are many performance combinations, but I tended to aim for maximum gain in the 14.1-14.2 MHz region. This yielded the following typical performance figures:

Bandwidth for greater than 9 dB front/back ratio	**200** kHz
Maximum front/back ratio	**12-15** dB
Bandwidth for less than 1 dB drop in gain	**230** kHz
SWR at band edges	2.0:1

Better frontlback ratios and bandwidths were obtainable at the expense of gain, thus emphasizing the desirability of remote tuning; in this way it should be possible to achieve effective gains between 6 and 7 dB. I tried to achieve this by using a single pair of additional feeders, attached to coupling loops placed near the centers of each set of parasitic elements, but achieved only limited success. A practical, but rather cumbersome alternative is to use separate feeders to all elements.

At 21 MHz, the SWR was 1.5-1.6 over the whole band; frontlback ratio increased with frequency from 13 to 16 dB, but was accompanied by an apparent 2-dB drop in gain. At 28 MHz, bandwidth for an SWR better than 2:1 was 350-400 kHz. Tuning for maximum gain at 28.5 MHz, there was a drop of 2 dB at 28.25 and 29 MHz. The frontlback ratio rose from 11 dB at 28.5 to 15 dB at 28.9, but was down to 6 dB at 28.25 MHz. These figures have been selected as fairly typical from a wide assortment and the variations illustrate the degree of improvement to be expected from remote tuning. The linear resonator



element lengths

frequency	reflector	director
14 MHz	11.2 meters (440 inches)	10.8 meters (426 inches)
21 MHz	6.9 meters (272 inches)	6.3 meters (259 inches)
28 MHz	5.1 meters (202 inches)	4.8 meters (188 inches)

boom length from center of driven element to apex of Vs

frequency	reflector	director
14 MHz	4 meters (156 inches)	3.2 meters (126 inches)
21 MHz	2.6 meters (101 inches)	2.6 meters (101 inches)
28 MHz	2.2 meters (86 inches)	2.2 meters (86 inches)

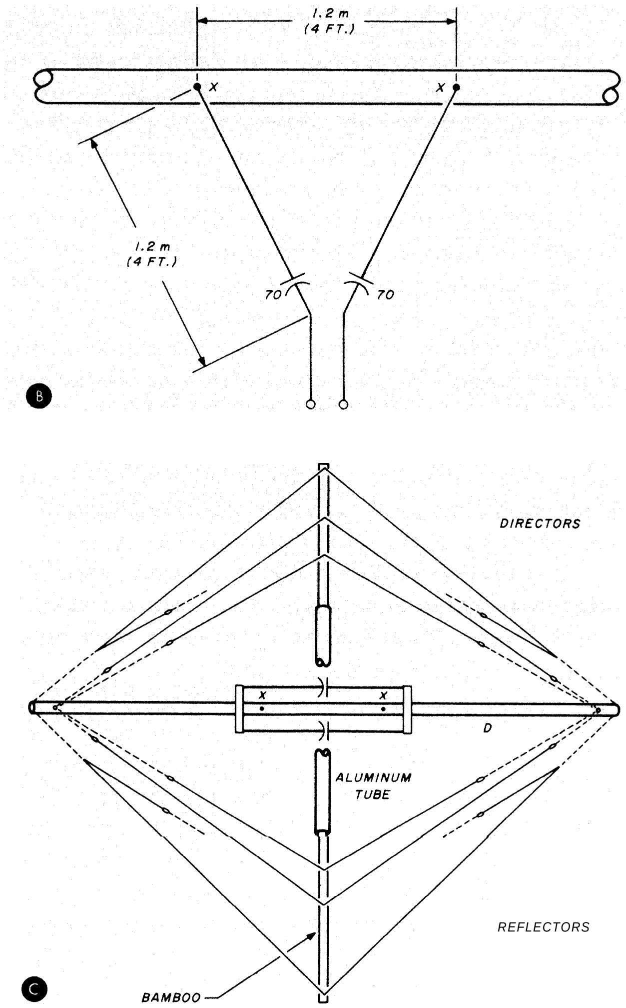


fig. 12. (A) shows the plan view of the driven element made from aluminum tubing, starting with 32 mm (1¼ inch), tapering down to 12.5 mm (0.5 inch) diameter. The feeder is connected to points *x,x* using the delta match shown in (B). Close examination of the end of the element may be necessary to fully understand the loading. Each vertical rod (*x,z*) is not only a help in providing the end loading, but is also used as an attachment point for the ends of the parasitic elements. Also attached to the vertical rods is the other portion of the end loading. short lengths [40.6 cm (16 inches)] of no. 14 AWG (1.6-mm) copper wire. These lengths are run from x and *z* to point y, *i.e.*, in the same line as the center element. The delta match detail is shown in (B). For open-wire line, the capacitors are omitted and the spacing is increased slightly. A plan view showing the parasitic elements is shown in (C). The elements are made from no. 14 AWG (1.6-mm) copper wire.

details shown in fig. 12 are applicable also to "full-size" elements, a reduction in length of about four per cent being required due to the presence of C2.

Earlier I criticized trapped beams, but the situation is quite different with small elements, including the V-shaped parasitic elements of the 'beam just described. If the folded ends are left in place, at the higher frequencies they degrade performance because of an inefficient current distribution and by radiation from the ends. The only practical way to remove the excess capacitance is by means of traps, but ordinary traps result in appreciable losses even with full-size elements. And, as the radiation resistance drops, the power loss increases in the same proportion so that large losses can be expected. This problem can be resolved with the help of the linear resonator, which, instead of degrading the performance at 14 MHz, actually has a very slight beneficial effect by increasing the total capacitance.

Another problem is caused by excessive spacing between elements at the higher frequencies. This can be avoided by using a separate, centrally placed, driven element for 21 and 28 MHz, a coax feeder with 4:1 balun, and a delta match into a linear resonator, providing two-band operation. To keep within an area 3 meters square (10 feet square), long, vertical, loading rods would have been required, but I thought it was better to increase the element lengths to 3.4 meters (11 feet) and make full use of the 4.7-meter (15.5-foot) turning circle.

Multiband operation, with good front-to-back ratios, was obtained using traps as shown in **fig. 11B**, but on 21 MHz the bandwidth was too narrow to be acceptable. The structure was braced with a lot of cord ties and I was able to use them to support separate (wire) parasitic elements for 21 MHz.

Capacitive coupling from these into the driven element was excessive, however, and the reflector had to be neutralized. Bandwidth was still rather narrow and devious means were needed to achieve even enough remote tuning for band coverage (without beam reversal) on 21 MHz. Despite satisfactory performance, the design became too complicated to be recommended in its present form. It has nevertheless proved that the problem is solvable, and guidelines have been established for further experiments. In particular, I found that proximity between the 14 and 21 MHz parasitic elements (average separation less than 0.3 meter [1 foot]) had no effect on 14-MHz performance. One obvious solution would be the use of separate parasitic elements for the higher bands, stacked about 0.6 meter (2 feet) above and below the 14-MHz elements. Linear traps remain the neatest method, and from inspection of **fig. 11B**, it is obviously possible to increase the trap length (and hence the bandwidth) by using the entire available length for 21 MHz, in which case the 28 MHz traps would have to be accommodated within the 21 MHz traps. Experiments along these lines can be recommended to anyone who feels he has the necessary skill and patience.

Unfortunately, narrow bandwidth at 21 and 28 MHz, *in comparison with a monoband antenna of the same size,* appears to be a price that has to be paid for the multibanding of a "smallest-possible" beam. (It is also part of the price usually, but mistakenly, paid for the multibanding of full-size elements which, when used without traps, have very much higher radiation resistances at the higher frequencies.)

One practical point that must be stressed is the need for accurately maintaining the shape of the linear resonators. With separate feeders to each element, small changes in resonant frequency can, of course, be corrected from the shack, but any asymmetry causes the current maximum to shift toward one end of the radiator. Probably worse, a voltage will exist in the center causing current to flow in a metal boom or dielectric losses in a wooden boom if wet! Insulating a metal element from a metal boom is not a complete cure, because there is bound to be some capacitance, but it has been found preferable to bonding.

future trends

I have tried to cover the subject in all its aspects and find it difficult to envisage a future for small beams outside the guidelines presented here, but perhaps someone will take this as a challenge and come up with something really new! There is room for plenty of ingenuity in finding ways of folding elements to fit them into small spaces, and it should be possible to improve bandwidth by the somewhat mind-boggling process of folding a folded dipole! The mini-quad also has interesting possibilities and appears to lend itself to a number of options for multibanding, of which I favor use of 20-meter loops for 10 meters, and separate, stacked elements for 15 meters.

Better construction methods are needed to improve reliability and achieve closer tolerances. A completely foolproof tune-up procedure, which can be used in all situations, has yet to be evolved. More data is needed on the range over which remote tuning from the shack can be achieved without penalties; further development is needed to find the best methods for remote tuning and reversal of small three-element beams. Perhaps someone may then get around to applying these features to the big beams — which will otherwise find themselves at a disadvantage! For multibanding of beams such as those described here, I believe the future lies in improved versions of the linear resonator trap, which would ease the problem of remote fine tuning and avoid the complications of stacking.

As an alternative to neutralizing, I have tried using the 21128-MHz element as an electrostatic screen between the 14-MHz elements, but I failed to achieve a viable system. There are difficulties, for example, in specifying exactly what is "earth," and I have found it advisable, even with neutralizing, to insulate elements from the boom and perferably to use a wooden boom.

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ham radio

high-performance voltage-tuned mosfet oscillator

One comment often heard on the amateur bands is that the art of homebrewing is a thing of the past. Nothing could be further from the truth. The amateur fraternity has acquired new vigor and growth, and more hams now than ever before, are "rolling their own." This is happening in spite of the fact that, again as a result of recent amateur growth, commercial manufacturers of ham gear are able to offer greater bargains in terms of performance per dollar. Yet there is a growing number of do-it-yourself amateurs who are dedicated to building all or some of their own equipment. The VTO described here is only one example of this trend. I am admittedly a homebrew addict.

concept

The constantly changing state of the art has provided a diversity of methods and circuits applicable to today's requirements. The VTO described in this article is more than just another of the garden variety. It was deliberately designed to perform well under the most demanding conditions. You may be interested in duplicating the design, or in using the basic concept to build one to suit your own particular needs.

This article is intended to provide sufficient specific information so that the design can be easily duplicated. If you need a VTO possessing superior performance characteristics, scrutinize the following checklist and decide if this construction project is for you.

1. Adjustable tuning range
2. Exceptionally clean spectrum – like that of a VCXO
3. Frequency essentially independent of load
4. A temperature compensation adjustment to cancel drift
5. Power output +10 dBm in a 50-ohm load
6. Remote tuning capability
7. Frequency independent of line voltage
8. Excellent short-term frequency stability
9. Excellent long-term frequency stability
10. Fast warm-up

After reading this article, you may wish to modify the design to fit your own needs. If so, there are certain things you should not do or need not do:

Don't use powdered-iron tuning slugs in the oscillator coil. It is particularly important to stay away from pot cores, bobbins, sleeves, toroids, and other ferromagnetic materials. Not only do these devices generally have relatively high temperature coefficients, but they possess a characteristic called hysteresis, which may cause the VTO frequency to wander from time to time for what appears to be no good reason. The permeability of powderded iron and ferrite materials is relatively sensitive to even weak magnetic fields.

Don't use a clamping diode on the MOSFET oscillator gate. It isn't necessary and only adds to the complexity. Note that the gate is grounded through the coil winding and therefore bias shift on gate 1 cannot occur.

The VTO described here should be operated at the fundamental, not a subharmonic. With the source bias used, the waveform on gate 1 is very nearly a pure sine wave, as is the waveform at the output. The circuit does not work well as a multiplier.

Don't operate the varicaps with too little bias. If the peak rf voltage exceeds the bias, rf rectification may occur which will generally disrupt performance. In the present design, varicap bias varies between 7 and 10 volts, although bias as low as **4** volts may be used if a wider band is to be covered.

Don't use a zener to provide voltage regulation for gate 2 or the drain of the MOSFET oscillator. (See **fig. 1**.) Use an integrated circuit regulator such as the μA723. Gate 2 voltage should be provided from a resistor divider network fed from the 723 output.

Don't mount the oscillator on a printed circuit board, since flexing or warping can cause serious stability problems. It is much better to assemble the entire oscillator on a brass or copper plate that can be mounted firmly to a shield box. Copper is preferred

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because is has higher heat conductivity, which helps to keep all circuit components at the same temperature. Use a small plate not less than 0.8 mm (0.032 inch) thick.

Don't use plastic, phenolic, or fiber coil forms. The coil forms used in this VTO are Cambion part number 1536-3-1. Only the amplifier circuit coil uses the tuning slug, which is carbonyl J (green).

Do not use ceramic trimmers in the oscillator circuit. The glass-piston trimmers specified possess a low temperature coefficient and are mechanically rugged.

Do not locate the VTO circuit near heat-generating components such as power supply regulators or transformers. Remember, since the oscillator is voltage tuned, it can be located at any convenient location because it does not require mechanical linkage with the front panel. You may want to take advantage of this feature and locate it next to the mixer where it belongs.

application

This particular VTO was designed as the local oscillator for the second mixer in a communications receiver. The design requirements were very exacting. It had to operate above 30 MHz and also drive a counter to provide a digital readout. The readout constantly monitors the frequency, which means that any drift exceeding 50 Hz would be particularly annoying, not to mention the problem associated with copying sideband signals. Furthermore, the tuning range was to be adjusted to cover exactly 100 kHz plus 100 Hz overlap at each end. The drift after warmup turned out to be considerably less than originally expected, in spite of the higher-than-normal operating frequency and the use of varicaps for tuning.

tuning mechanism

Although the circuit values can be modified so that the VTO covers other frequency ranges, this particular unit tunes 30,250 to 30,350 kHz. A five-turn, linear-taper, Bourns precision potentiometer (part number 35205-417-453) is used for frequency tuning. A 6.4-mm (2.5-inch) diameter Millen fluted knob provides approximately 20-kHz per revolution, permitting ssb signals to be tuned in with relative ease. This amounts to 18 degrees per kHz. A ten-turn potentiometer can be substituted if greater tuning resolution is desired. In any event, avoid the use of a gear train to increase resolution if backlash is to be avoided.

Originally, there was considerable concern regarding the ability of the of the potentiometer to stand up

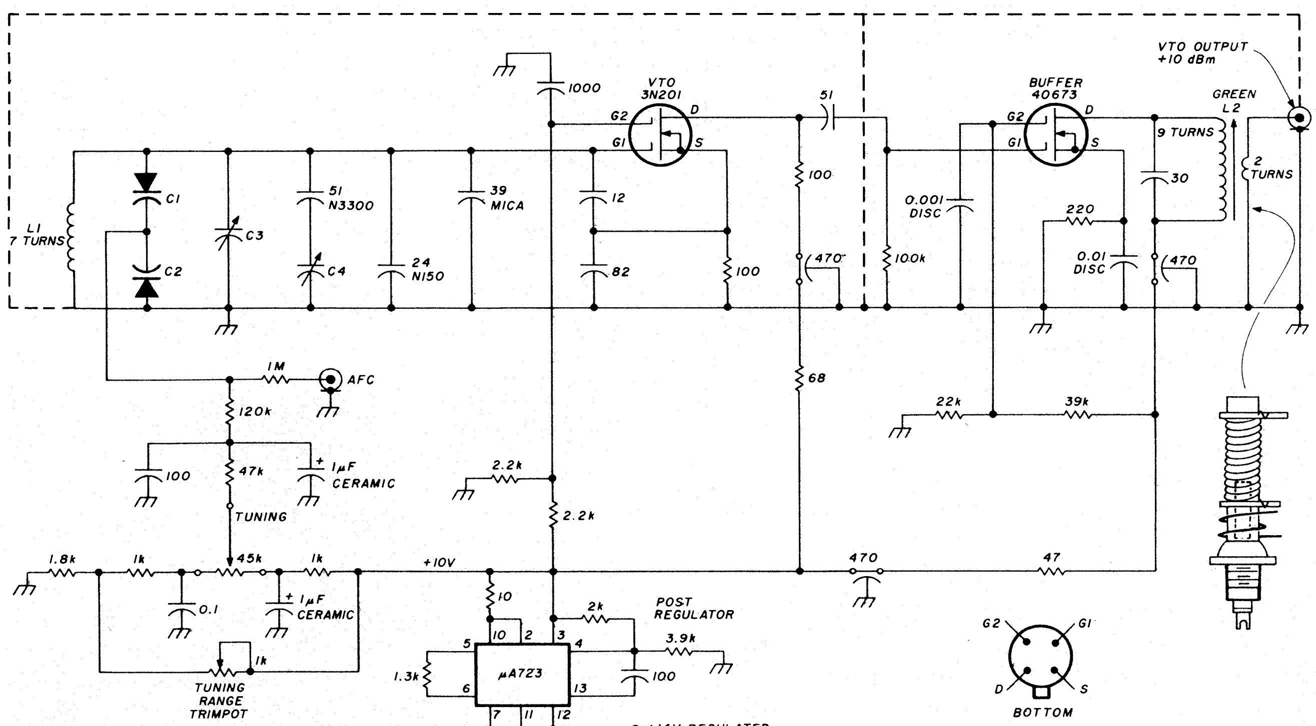


fig. 1. Schematic diagram of the voltage-tuned oscillator. All fixed resistors are 114-watt composition, and all capacitors are dipped silver mica unless otherwise indicated. The 470-pF feedthrough capacitors are from Spectrum Control (part number 54-794-002-741M). The varactors are TRW 4808s. Capacitor C3 is an Erie 0.7-12 pF piston trimmer, while C4 is a 1-18 pF Erie piston trimmer. L1 is wound on the Cambion 1536-3-1 coil form; L2 uses the same form with the carbonyl J tuning slug inserted as shown in the diagram.

to abuse day after day. However, the original unit showed no indication of wear after two years of almost daily and often grueling use.

circuit details

The VTO includes a 3N201 dual-gate MOSFET oscillator driving a low-cost 40673 MOSFET buffer amplifier. A high degree of isolation is achieved by use of the 100-ohm resistor in the oscillator drain circuit. As a result, pulling of the oscillator as the buffer coil is tuned through the oscillator frequency is less than 200 Hz. Since the buffer circuit is broadband and therefore fixed tuned, this effect has insignificant effect on drift. If the load terminals are opened or short-circuited, the change in the oscillator frequency is less than 100 Hz. Power output over the 100-kHz range is essentially constant at 10 milliwatts under load.

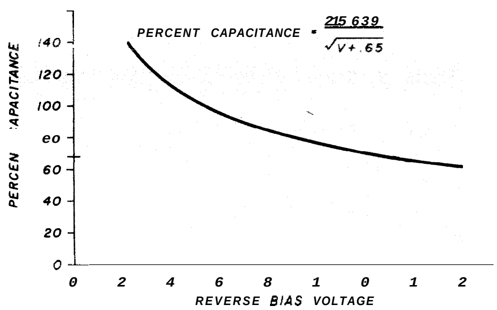


fig. 2. Manufacturer's specifications for the TRW 4808 varactor.

The coils are wound on ceramic forms with no. 26 (0.4-mm) AWG enameled copper wire and heavily doped with polystyrene Q dope. The 50-ohm output winding on the amplifier coil consists of two turns of no. 26 (0.4-mm) AWG enameled copper wire, pushed up tight against the bottom (cold) end of the tuned circuit coil. It is important that the tuning slug be positioned at the 50-ohm output end of the coil, partly overlapping both the tuned circuit and the output winding.

power supply considerations

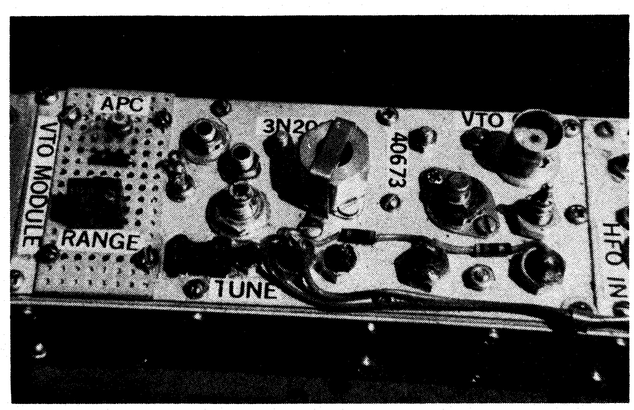
Both the oscillator and its buffer are powered from a μA723 integrated circuit regulator which accepts current from a regulated 14-volt supply. The 723 reduces the 14 volts to 10 volts. The importance of this additional regulator cannot be overemphasized. Not only does it provide the required tight voltage regulation, but by its nature the output voltage is extremely well filtered. For example, when the spectral purity of the VTO was compared while using a battery supply and then the 723 post regulator, there was no noticeable difference. By contrast, without the post regulator, there was considerable hum modulation. Small changes in supply voltage cannot be tolerated because the same supply is applied to the tuning varactors.

A pair of TRW 4808 voltage-variable, varactor diodes (varicaps) are used to control the VTO frequency. These 27-pF tuning varactors are connected in the conventional back-to-back manner with the tuning voltage fed to their junction through an RC filter. The 1000-ohm tuning range trimpot is set near its midpoint so that the voltage across the 45k-ohm tuning potentiometer is approximately 3 volts. Under these conditions, the tuning potentiometer provides the varicaps with between 7 and 10 volts. Long life and reliability is assured since the potentiometer carries only 64 microamperes while the current carried by the moving arm is essentially zero. Data on the varicap is shown in **fig. 2.**

The varicap tuning sensitivity is 30 μV/Hz. If the spectral purity of the oscillator is to be high enough for the intended application, power supply noise and ripple must be extremely low. The μA723 is specified at 20 μV rms noise typical. This could cause a random peak-to-peak fm of 1.89 Hz. Most of this noise is eliminated by means of the lowpass filter consisting of the 47k-ohm resistor and the 1.0 μF capacitor. What little noise actually remains should consist of very low-frequency components, since the RC filter has a 20 Hz cut-off frequency. In the author's application, there is no detectable reduction in receiver output noise when a battery is substituted for the μA723 regulator.

Ripple from the 723 is specified as typically 74 dB below the input ripple. Even if the input ripple were 100 mV peak-to-peak, the output ripple would only be 2.0 μV, including the filtering action of the varicap RC filter. This translates to less than a tenth of a cycle fm.

The varactor capacity as seen by the oscillator coil varies from 3.8 to 4.4 pF. Since the total circuit capa-



Since the VTO is dc tuned, it does not have to be located near the front panel. In this case, it has been incorporated into the same module as the HFO buffer.

city is approximately 90 pF, the ratio of high to low end frequencies is as follows:

$$\frac{F_H}{F_L} = \sqrt{\frac{C+\Delta C}{C}} = \sqrt{\frac{90.6}{90}} = 1.0033$$

where

F_H = high-end frequency
F_L = low-end frequency
C = total circuit capacitance

Therefore, if

$$F_L = 30{,}250\ kHz,$$
$$F_H = 1.0033 \times 30{,}250$$
$$= 30{,}350\ kHz.$$

Of course, the tuning range tripot can be adjusted to provide overlap at each end of the tuning range if that is desired.

The frequency ratio formula can also be used for different frequencies and tuning ranges. I suggest that the circuit capacity be scaled in proportion to wavelength. A VTO operating between 5.0 and 5.1 MHz, for example, could use a circuit capacity of about 500 pF. The varicaps would each be 150 pF, while C would be approximately 20.2 pF, which requires a tuning voltage range of about 6.4 to 10 volts.

tuning linearity

While the tuning potentiometer is linear, the varicap tuning characteristic is not. As a consequence, the frequency tunes slightly faster at the low end of the range than at the high (see **fig.** 3). However, the departure from linearity is small. Based on the manufacturer's data, and by measurement, the number of kilohertz per revolution at the low end is 22.4 compared with 18.2 at the high end. This nonlinearity might be objectionable if it were necessary to employ a mechanical frequency indicator, such as a planetary mechanism. In my application, the VTO drives a programmable digital counter in which case the small nonlinearity is of no consequence.

temperature compensation

Considerable effort went into developing a variable temperature compensating circuit for this VTO. While the idea is not new, it lends itself well to this particular application. The compensation circuit is simple and easy to adjust.

A piston trimmer is connected in series with a negative temperature coefficient ceramic capacitor. The coefficient was intentionally made larger than needed. The effective circuit coefficient is adjusted by tuning the piston trimmer, C4. The change in frequency which resulted is compensated for by adjusting C3. There is a particular setting of these two capacitors which will provide zero temperature coefficient, therefore zero oscillator drift after warm-up.

Based on manufacturer's data, the voltage-tuned varactors have a temperature coefficient of 250 to 300 ppm/degrees C. Of course, most of the other circuit elements exhibit positive coefficients as well, but their coefficient is generally less — in the order of 50

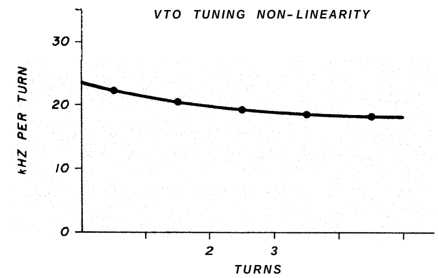


fig. 3. Chart of the voltage-tuned oscillator's linearity as the oscillator tunes from one end of its range to the other.

to 100 ppm/degrees C. Dipped silver-mica capacitors were found to be entirely satisfactory for use in the oscillator and amplifier tuned circuits.

A series of tests was performed using NTC capacitors with various temperature coefficient values. Finally, a 51-pF, N330 capacitor was selected. An additional 24-pF, NTC 150 capacitor was added in parallel to bring the circuit temperature coefficient within range of C4. C4 is a key circuit element because it allows the effective temperature coefficient to be adjusted, and therefore the amount and direction of oscillator drift. The proper setting was found after a number of drift tests were performed over a period of a few hours using different settings of C3 and C4. The results of these tests are shown in **fig. 4.**

After a period of about three months, and again a year later, the drift was checked. It was found that no change in compensation had taken place. The circuit gives you a real good opportunity to check the Other fellow's frequency stability!

After two years of drift-free use, and strictly out of curiosity, the original 3N201 oscillator MOSFET was replaced with another made by a different manufacturer. It was found that the frequency was slightly higher and needed trimming. Subsequent drift tests showed the circuit to be over-compensated. A new set of drift tests were run and the overall temperature coefficient quickly brought back to zero.

If the drift that occurs after turn-on from a cold start is annoying, or if the application requires absolutely zero warmup drift, I recommend that the drift correction circuit described by PAØKSB[1] be used. This circuit was built and applied to the AFC terminal shown in **fig. 1.** The performance was excellent.

There was absolutely no warmup drift. It was necessary, however, to include the up-down tuning push buttons recommended to periodically correct for calibration errors inherent in the design. Since this requirement was more of a nuisance than the warmup drift, the circuit was not permanently installed. From a practical standpoint, about the time the 6146s are warmed up and ready to go, the VTO has stabilized! It was also interesting to note that if the VTO design described here were to be modified for operation at 5.0 MHz, the warmup drift would be only about 100 Hz.

dryer will probably prove only that the temperature compensating circuit has a thermal lag.

Short-term drift and low-frequency shot noise effects can be determined by tuning in WWV (assuming the VTO is used as the local oscillator of the receiver). Using the BFO, set the tuning for a convenient heterodyne (500 Hz for example), and then acoustically zero beat an audio oscillator and speaker combination with the receiver output. Listen for any warbling and jumping. If you use a low-beat note, you should be able to hear and detect frequency variations as small as ± 2 Hz by reading frequency

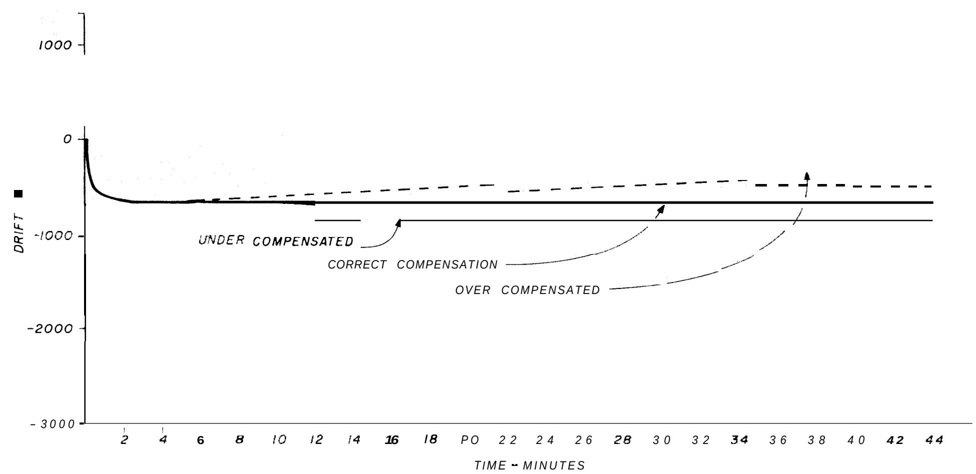


fig. 4. Typical results of the oscillator drift tests. Incorporating the drift correction circuit described by PAØKSB will reduce the frequency change to zero.

Excellent short-time stability was achieved by enclosing the VTO circuit in a metal box. Also, a brass cup heatsinked to the chassis was mounted over the 3N201 transistor and its socket. This protects the transistor from drafts to which it is particularly susceptible, and forces the ambient temperature in the vicinity of the transistor to track that of the entire circuit. This combination was found to be very effective.

One of the most frustrating problems associated with checking out a new oscillator is being able to separate out the various causes of drift. They include the effective temperature coefficient of the overall electrical circuit, the effects of voltage variations, and mechanical effects. If the oscillator varies or jumps when the oscillator chassis is tapped sharply with a screwdriver, you have mechanical problems. Don't try to temperature stabilize the circuit until it is mechanically sound. It is a good idea to run the temperature of the circuit up and down several times using a hair dryer to reduce mechanical strains before checking drift.

If you use the regulator circuit suggested here, you should be able to set the line voltage as low as 100 volts before detecting any oscillator drift. Long-term drift tests should be performed by running the circuit as it would normally be operated. The use of a hair

changes from the audio oscillator dial. If the oscillator is working properly, random variations should not exceed ± 5 Hz at 30 MHz, or ± 1 Hz at 5 MHz, and there should not be any jumping. It is interesting that the technique described here allows you to detect variations as small as one part in 10^7 or better.

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ham radio

operational characteristics of the 555 timer

Care and feeding of the multi-purpose 555 timer — a closer look at some of the reasons behind the designs

The *555* timer has become one of the most popular ICs used by both industry and hobbyists alike. Since its introduction by Signetics Corporation in 1972 (as the NE555), every major IC manufacturer has produced a 555 equivalent. Variations have been developed by Signetics and others, primarily for specific applications (NE556, NE558 quad, XR2250 programmable).

Circuits using the 555 show up in virtually every electronics-oriented magazine. The popularity of the 555 is partly due to its versatility and partly because of its cost (generally less than 50 cents). These published circuits use the timer to generate ramps and time delays, detect missing pulses, act as oscillators, and a host of other applications. The list of specifics can become quite extensive. Most of these applications now appear as "cookbook" circuits with little detail on how the design was accomplished and what variations a builder of that circuit can expect.

This article was written to help you get a better understanding of the 555 and how its eight pins react to external components. With this information, you can get a feel for how most 555 applications work. For the more creative, I'll outline the basic design rules for using the 555. (Information on the 555 also applies to the 556 dual timer.)

Fig. 1 shows the block diagram of the 555 timer. The basic components consist of an output driver, a control flip-flop, and two voltage comparators. The flip-flop drives the discharge transistor and output driver. The state of the flip-flop is controlled by the reset pin (pin 4) or one of two comparators. One comparator is controlled by the voltage on the trigger pin (pin 2) and the other controlled by the voltage on the threshold pin (pin *6).*

These comparators have separate reference points which are controlled by the three-resistor divider from V_{cc} to ground. The resistors are all of equal value (5-k ohms). The reference voltage for the threshold comparator is thus 213 of V_{cc}, and is also available on pin 5, the control voltage pin. The reference voltage into the trigger comparator is set at 1/3 V_{cc} by the divider network (Note: it is also one half of the voltage on the control voltage pin).

The output (pin 3) has the capability of providing 200 mA. Because of the totem-pole structure, this

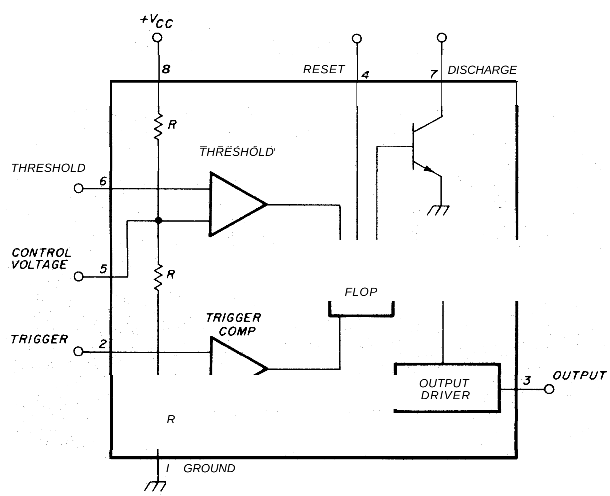

fig. 1. Block diagram of the 555 timer. The three-resistor divider network causes the voltage comparators to work on a ratio, rather than an absolute voltage level. This makes any timing functions relatively insensitive to supply voltage.

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current supply is available whether the output is high or low. Additionally, when the output is high, the discharge transistor is turned off.

monostable operation

To help understand the action of the comparators it is best to show the operation of the 555 in one of the basic timing modes. The monostable configuration of the timer requires only two external components. It is shown in **fig. 2A** (the capacitor on pin 5 is optional).

A resistor (R_A) is connected from V_{cc} to the discharge pin (pin 7) and threshold pin (pin 6). A capacitor is then connected from pins 6 and 7 to ground. The values of R_A and C will determine for how long the output remains high.

Initially, the output is low and the discharge transistor is turned on. This essentially shorts the timing capacitor *C*. Therefore, the threshold comparator input is at zero volts. When the trigger comparator receives a trigger that is less than 1/3 V_{cc}, it causes the flip-flop to switch. The output goes high and the discharge transistor turns off. When this happens, the timing capacitor starts charging to V_{cc} via R_A. As soon as the charge on the capacitor reaches the threshold voltage level (2/3 V_{cc}), the threshold comparator triggers the flip-flop. This drives the output stage low and turns on the discharge transistor, which discharges the timing capacitor, and the timer is back to its initial condition. **Fig. 2B** shows the waveforms on the timer during this time.

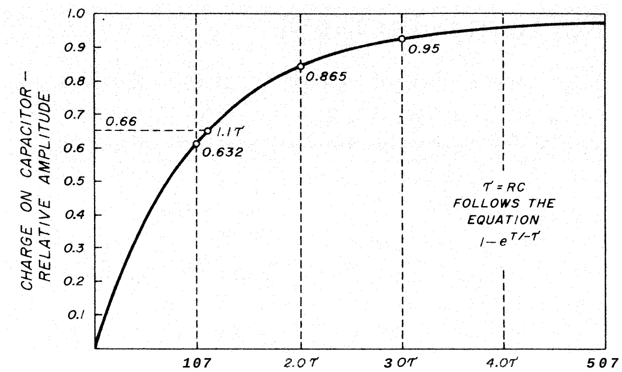


fig. 3. Diagram of the normal RC time-constant curve. When connected as a monostable, the output pulse width from the timer is determined by the time it takes the capacitor to charge between zero and 66 per cent of the full charge.

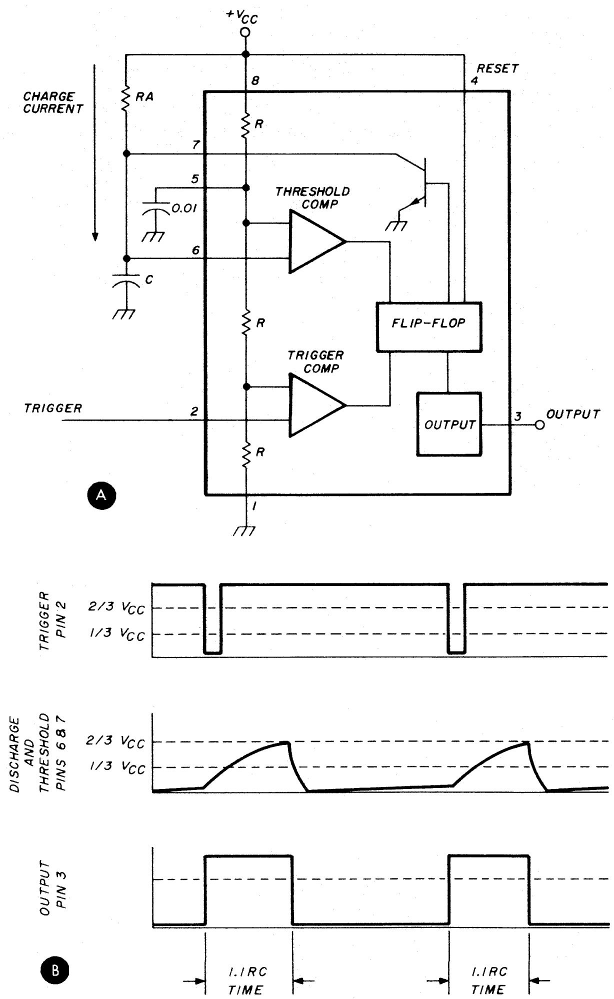


fig. 2. Configuration of the 555 timer for use as a monostable multivibrator and the associated waveforms (B). The negative-going pulse applied to the trigger input causes the internal flip-flop to change states, generating the output pulse.

triggering

Due to the nature of the trigger circuitry, the trigger comparator will trigger the flip-flop whenever the trigger voltage drops below 1/3 V_{cc}. For proper timing, the trigger level must return to a voltage level greater than 1/3 V_{cc}. This is because the trigger comparator has overriding control of the flip-flop. Should the trigger input be held low for a period longer than the timing cycle, the output will remain high, without regard to the voltage on the threshold comparator.

reset function

The 555 has a reset pin for applications that require an abort signal. That is, it may be necessary to interrupt a timing period or to inhibit a trigger. The reset control (pin 4) is normally high. If the reset is held greater than 1.0 volt, the 555 will function normally. However, if pin 4 is held below 0.4 volt the output will be held low. When the timer output is high and the reset goes low, the output will turn off immediately. If the reset pin is between 0.4 and 1.0 volt, the timer is in no man's land; some devices may reset, some may not. When using the reset function, it is important to have the reset less than 0.4 volt. This level is guaranteed to cause the timer to reset. To prevent possible noise spikes from causing an unwanted reset, pin 4 is normally connected to the V_{cc} pin when not being used.

calculating pulse width

When describing the operation of the 555 timer in the monostable operation, I mentioned that the values of R_A and C determined the time the output stayed high. The voltage on C must charge to 213 V_{CC}. Fig. 3 shows the normal RC charge curve. As shown, the voltage rises from 0 to 100 per cent in five RC time constants (τ). The 66.6 per cent (2/3) point occurs at 1.1τ, or

$$T = 1.1\,RC \tag{1}$$

where

T is in seconds,
R is in ohms
C is in farads

changing the pulse width in the monostable mode

The threshold control voltage on pin 5 is an important feature of the NE555 timer. By imposing a voltage on pin 5, the internal comparator reference levels, the timing can be varied. For example, if one-half the supply voltage were placed on pin 5 (normally held at 2/3 V_{CC}), the timing capacitor would charge only to one-half V_{CC} before the threshold trip level is reached. The time required to charge the capacitor can be found by observing how many time constants it takes for the capacitor to charge to 50 per cent using the equation $T = \ln\left(1 - \frac{V5}{V_{cc}}\right) RC$ or by using **fig.** 3. With the voltage on pin 5 equal to 0.5 V_{CC}, the new timing equation is $T = (\ln 0.5)RC = 0.693RC$.

You must remember that the trigger threshold is also affected by changes on the voltage level at pin 5. In the example above, the trigger threshold is now lowered to one-quarter V_{CC} (halfway below the voltage on pin 5). In order to trigger, the trigger pulse must now go below one-quarter V_{CC}. This feature of the timer opens a multitude of application possibilities such as pulse width modulators, voltage-controlled oscillators, and so forth.

As can be seen, any variations on pin 5 will cause a change in timing. For that reason it is recommended that a small bypass capacitor (about 0.01 μF) be used on pin 5. This will increase noise immunity of the timer to high frequency trash that could cause timing errors by modulating the threshold trip levels.

astable operation

To configure the 555 timer in the astable or oscillatory mode requires only a slight modification to the monostable configuration. **Fig. 4A** is the schematic of a 555 timer in the basic astable mode. It requires two resistors and a timing capacitor. Assume that the charge on the timing capacitor is charging toward V_{CC} through R_A and R_B. The output is high and the discharge transistor is off. When the capacitor charges to the threshold trip level (2/3 V_{CC}) the output goes low and the discharge transistor turns on, shorting pin 7 to ground. The capacitor now begins to discharge through R_B toward ground. But as soon as the capacitor discharges to 1/3 V_{cc} (the trip level of the trigger comparator), the output again goes high, the discharge transistor turns off, and the

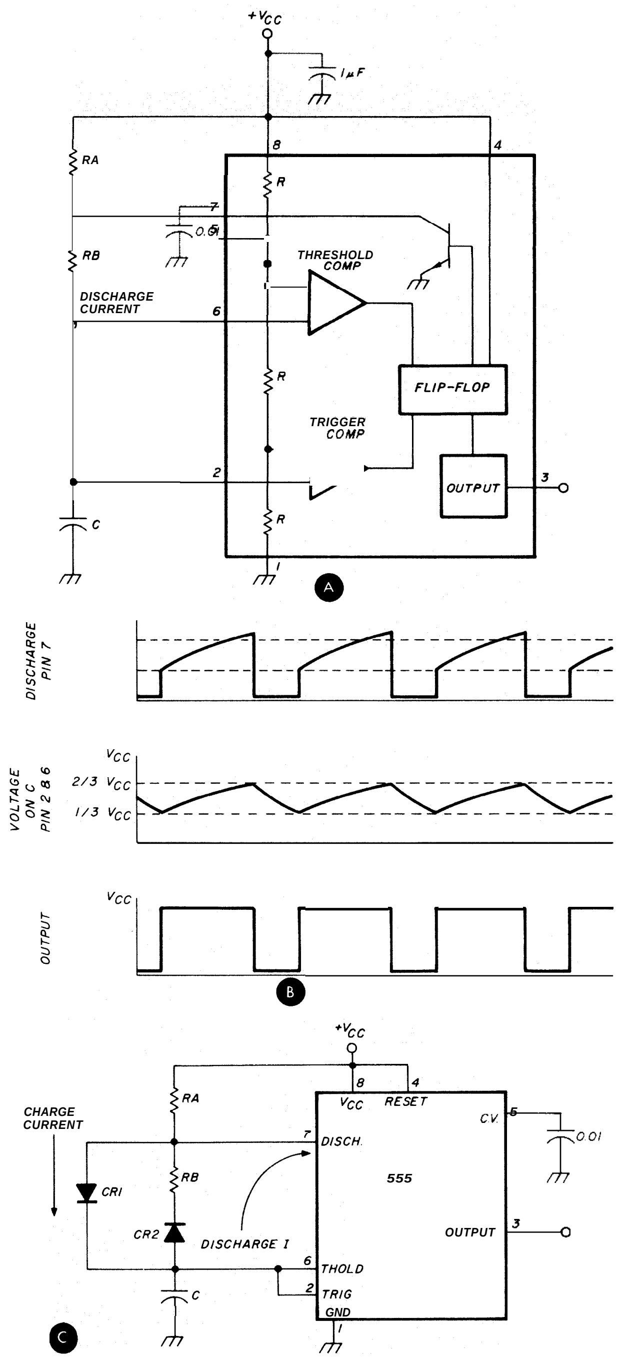


fig. 4. Schematic diagram of the 555 connected as an astable multivibrator and the waveforms at different points in the circuit (B). In applications where a 50 per cent duty cycle is required, the circuit shown in (C) can be used. In this case, CR1 effectively shorts out R_B during the capacitor charge time.

capacitor starts charging through R_A and R_B again, creating an oscillator. **Fig. 4B** shows the waveform in the astable mode.

calculating frequency

The time required to charge from 1/3 to 213 V is 0.671 RC. The charge path is through $R_A + R_B$. This is the period of time that the output is high, or

$$T_1\,(High) = (R_A + R_B)\,C \tag{2}$$

During the discharge time the output is low. The time required to discharge from 2/3 to 1/3 V_{cc} is also 0.671 RC. But the discharge path is through R_B only, so

$$T_2\,(Low) = 0.671\,(R_B)\,C \tag{3}$$

The total period of oscillation is then:

$$T_{TOT} = T_1 + T_2 = 0.671\,(R_A + 2R_B)\,C \tag{4}$$

and the frequency

$$F = \frac{1}{T_{TOT}} = \frac{1}{.671\,(R_A + 2\,R_B)C}$$
$$= \frac{1.49}{(R_A + 2\,R_B)C} \tag{5}$$

The duty cycle is given by:

$$D = \frac{R_A}{R_A + 2\,R_B} \tag{6}$$

Since the charge and discharge paths are different, the duty cycle cannot be less than 50 percent. In some applications, this may cause a problem. With a slight modification to the basic circuit, the 555 output can be made a square wave. **Fig. 4C** details the circuit configuration for a 50 per cent duty cycle when $R_A = R_B$.

In this circuit, CR1 shorts R_B and the charge time is 0.671 $R_A C$. The discharge path is still through R_B and does not change. The series diode, CR2 is optional to match the charge and discharge paths. With the diodes in the circuit the formulas for frequency and duty cycle become:

$$F = \frac{1.49}{(R_A + R_B)C} \tag{7}$$

$$Duty\ Cycle = \frac{R_B}{R_A + R_B} \tag{8}$$

With this configuration, the timer is capable of generating duty cycles from 5 to 95 per cent.

supply voltages

In selecting supply voltages, note that the 555 timer is guaranteed to operate from 4.5 volts to 16 volts. Because the threshold levels work on a ratio, the supply voltage will not change the timing calculations. The timer has a "totem pole" output structure capable of switching high levels of load current. As a result, large current spikes can develop on the supply line. This momentary loading effect can cause a degree of timing error because of changes in charging current. It can also cause noise glitches and false triggering in TTL circuits. To eliminate this phenomenon, it is necessary to bypass the supply line to ground with a capacitor. The size of the bypass required generally depends on the load to the timer. Values range from 0.1 to 10 μF, or more. The bypass capacitor should be as close to the device as possible.

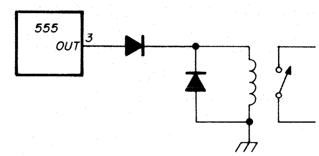


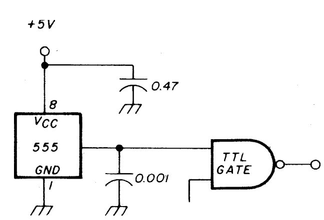
fig. 5. When driving an inductive load, a protective diode is required to prevent the inductive kick from latching up the output of the timer.

Bypassing the control voltage pin (pin 5) is generally considered good design practice. As mentioned earlier, the value is not critical, but the capacitor should be as close to the device as possible. Typically, the bypass capacitor is 0.01 μF.

capacitors

The timing capacitor size is virtually unlimited; but, the type of capacitor is important. Ceramic disk capacitors are usually unsuitable. They generally are not stable enough to operate in an RC timing circuit. Electrolitics usually have very high leakage rates and would cause drastic timing errors. The Signetics data sheet lists several acceptable types of capacitors, including silver mica, mylar, polystyrene, and tantalum. The smaller the timing capacitor used, the more apparent effect stray capacitance can have on the timing. The larger the capacitor, the more expensive the capacitor is. This is particularly true of the low-leakage types. For long delays, the capacitor has to be very large, but there are ways of getting long delays without using big capacitors.

fig. 6. The-capacitor on the output of the timer is used to prevent any glitches caused by the cross-over distortion from acting as false triggers to the TTL logic.



resistors

There are certain maximum and minimum values

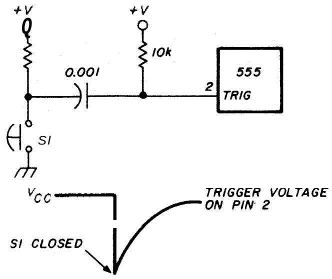


fig. 7. In this example of ac triggering, the trigger pulse must be less than the duration of the output pulse from the timer. The duration of the trigger depends upon the RC time constant of the components.

for the resistors. The threshold comparator requires 0.25 μA of current to trip the output. Considering worst case, the resistor must be able to supply 0.25 μA of current for the comparator and still charge the capacitor to 213 V_{CC}. To calculate the maximum resistance, the IR drop must not exceed 113 V_{CC} with 0.25 μA current flow, or

$$IR\ drop = V_{cc} - V_{CAP}$$
$$= V_{cc} - 2/3\ V_{cc}$$
$$= 1/3\ V_{cc}$$

The maximum resistance between V_{CC} and pin 6 is defined as

$$R_{MAX} = \frac{V_{cc} - V_{CAP}}{threshold\ current} \quad (9)$$

With a 15-volt supply

$$R_{MAX} = \frac{15V - 10V\ (2/3\ V_{cc})}{0.25 \times 10^{-6}}$$

$$R_{MAX} = \frac{5}{0.25 \times 10^{-6}}$$
$$= 20\ megohms$$

With a 5-volt supply

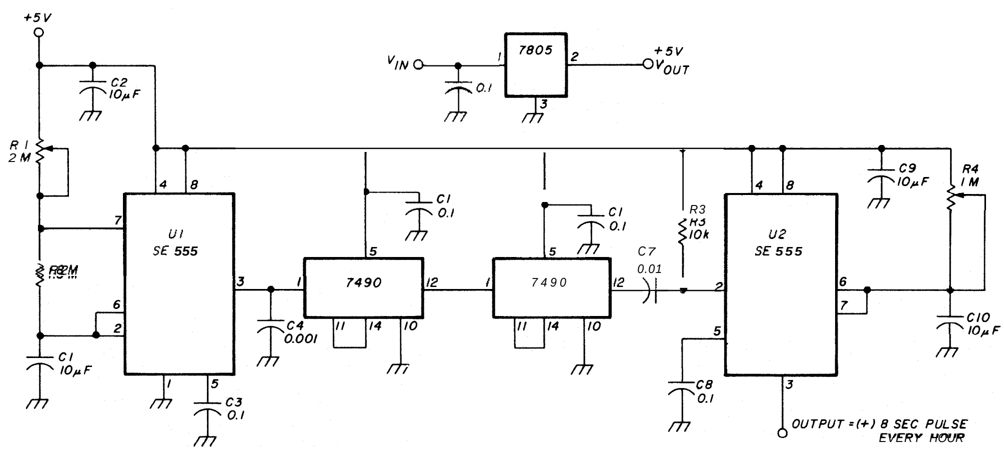
$$R_{MAX} = \frac{5V - 3.33V}{0.25 \times 10^{-6}}$$
$$= \frac{1.67}{0.25 \times 10^{-6}}$$
$$= 6.6\ megohms$$

When using large resistors, capacitor leakage can cause larger timing errors because it represents a larger percentage of the total charge current available. Excessive capacitor leakage current will also cause an IR drop and not allow the threshold comparator to reach 2/3 V_{CC}.

The minimum value of resistance is determined by the current that the discharge transistor can supply. The discharge transistor is internally current limited to about 35 mA. The discharge transistor must supply two loads. The first is the current through R_A. This should be reduced so that the second path into the capacitor (or through R_B) carries most of the load. As a general rule, R_A should not be less than 5k ohms. R_B should not be less than 3k ohms.

control voltage changes

As mentioned earlier, the control voltage pin (pin 5) is normally bypassed to ground by a 0.01-μF capacitor. By imposing a voltage on this pin, it was shown that the timing can be changed since the threshold level is changed. The voltage level on pin 5 can be lowered to about 45 per cent of V_{CC} when operating the timer in the monostable mode. The trigger level is also changed because it will equal one half the voltage on pin 5, and a certain minimum volt-



U1 — frequency $= 1/36$ Hz
$R1 = R2$, let $C = 10\ \mu F$

with $F = \frac{1.49}{(R1 + 2R2)C}$

Since $R1 = R2$, $F = \frac{1.49}{3R \cdot C}$

therefore $R = \frac{1.49 \cdot 36}{10 \cdot 10^{-6}} = 1.8\ megohms$

U2 — 8-second output pulse
let $C = 10\ \mu F$

therefore $R = \frac{1}{\cancel{1.1C}} = 0.727\ megohms$

fig. 8. Long time delays can be generated by cascading timers with other count-down devices. In this example, the basic timer, U1, runs at about 0.028 Hz, is divided down by 100, and then used to trigger another timer configured as a monostable. The output is an eight-second pulse, once every hour.

age level must be maintained on the trigger comparator reference. If the voltage drops below 45 per cent on pin 5, the timer becomes more sensitive to noise because the trigger voltage nears ground. At the other extreme, the control voltage on pin 5 should not exceed about 90 per cent of V_{cc} since the timing capacitor must charge to V_{cc}. Exceeding this voltage range in the monostable mode may cause timing error, false triggering, or other problems.

In the astable mode, the voltage level changes on the control voltage pin can be used to change the oscillating frequency about +25 per cent and still remain linear. Changes greater than this will still cause the frequency to change, but linearity decreases due to the RC timing circuit.

Before examining some specific circuits, there are several idiosyncracies of the timers that can cause problems unless you are aware of them.

Temperature. The timer exhibits a small negative temperature coefficient (50 ppm/C°). This can cause small timing changes in the monostable mode and frequency drift in the oscillator mode. In critical applications, R/C values can be selected which have positive coefficients, with the net result a lower drift. Since the astable mode relies on both trigger and threshold levels, the drift from temperature is usually higher than when in the monostable mode. An important point to remember when working with the timer (for that matter any IC) is that power dissipation of the device should never be exceeded. Power causes heat, and excessive heat will destroy the device. The 555 can handle about one-half watt of power at room temperature. This power rating is lower when the device is operated at higher temperature.

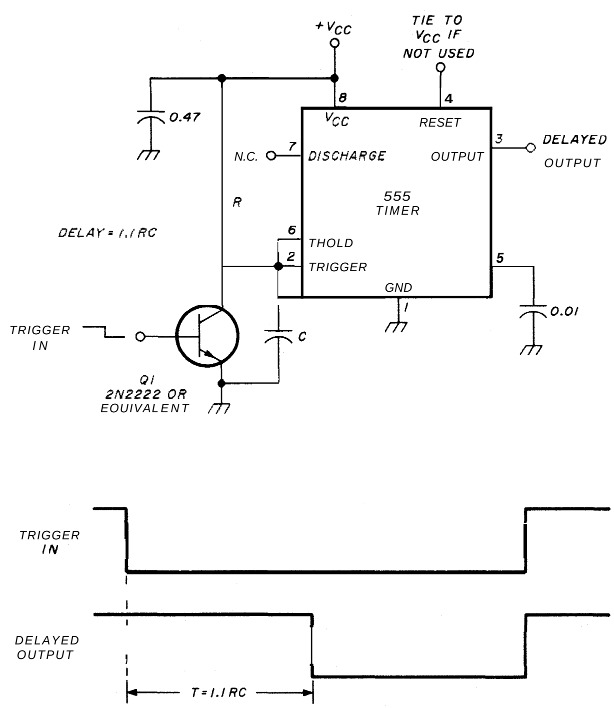


fig. 9. In this time delay circuit, the timing capacitor is effectively short circuited by the normally conducting transistor. A negative-going trigger cuts off the transistor, allowing the timer to operate and provide a pulse after the delay time.

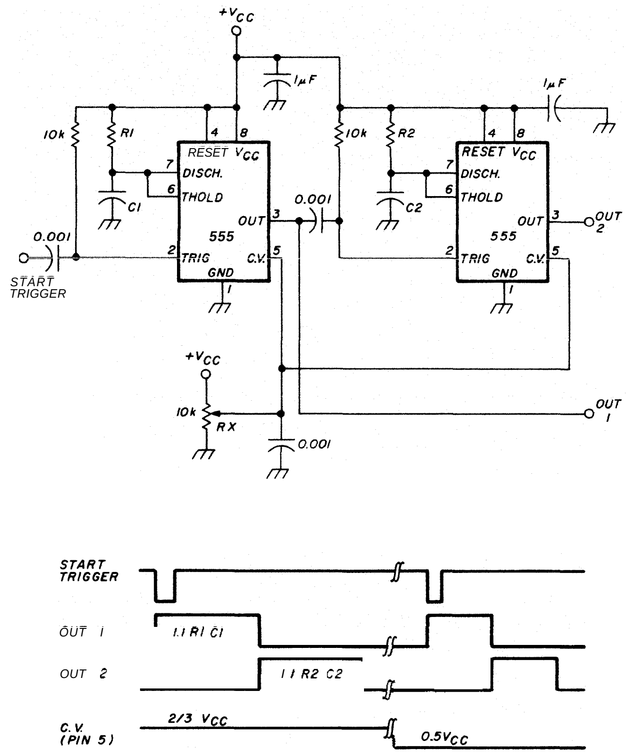


fig. 10. This version of a sequential timer uses the control voltage input to change the pulse width of the output pulses, though they do remain in the same ratio.

Output. If a negative voltage (with respect to pin 1) is applied to the output of the timer, the 555 could latch up. This can happen when the 555 is used to drive an inductive load, such as a relay. To prevent this inductive kick back from latching the timer, a diode in series with the output should be used. **Fig. 5** shows a schematic of a timer being used to drive an inductive load.

The output drive capability of the timer is 200 mA. Because of the output structure's high current capabilities, and fast rise and fall times, the timer exhibits crossover distortion. This glitch can cause false triggering of TTL circuits. By providing a capacitive load, the timer output is slowed to a point that the glitch does not occur. A capacitor of about 1000 pF from the output to ground will eliminate any false triggering of the TTL circuit (see **fig. 6**).

Triggering. I've stated that in the basic monostable mode the trigger must go below 1/3 V_{cc} (or one half

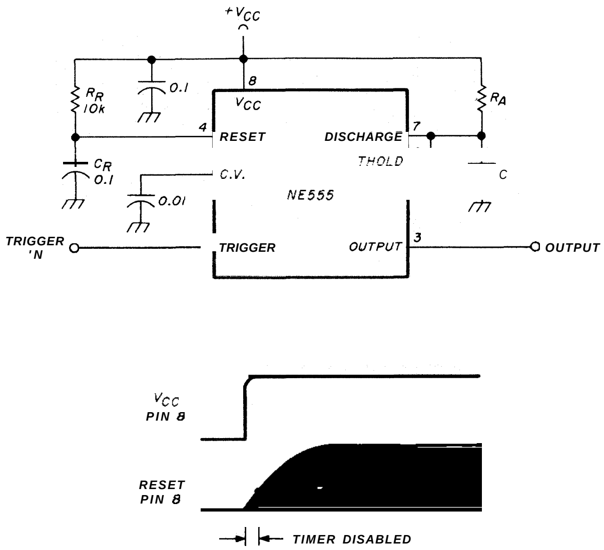


fig. 11. With the capacitor attached to the reset input, the timer cannot produce an output when power is initially applied. After C_R has charged, a trigger will produce the desired output pulse.

the voltage on pin 5) and return high before the end of the timing cycle. One way to generate such a trigger is to ac-couple the trigger. **Fig.** 7 shows such a circuit. The duration of the trigger pulse depends on how long the RC time constant is. The switch must return high again before the timer can be retriggered.

example circuit

The following circuits illustrate the use of the 555 timer (or 556 where two timers are used).

Long time delays. Because of the limitations of resistor size, long time delays (times greater than one hour) can be difficult to achieve using the basic circuit. One method of getting long delay times is shown in **fig. 8.** This particular circuit provides a positive 8-second pulse once each hour. U2 receives a trigger once each hour. The output of U2 is set for the desired pulse output by R4 and C10. R3 and C7 provide the ac-coupled trigger. The 7490 counters are set to divide by 100. To provide one cycle per hour, the input must be clocked at a frequency of 1/36 Hz; U2 is then set to oscillate at 0.028 Hz. Note that C4 provides a deglitch filter on the output of U2. Because the timer must interface with TTL counters, the timers and counters use 5-volt supplies. The entire circuit can also be built by using a single 556 dual timer and one 74390 dual-decade counter. By changing the divider network or frequency of U1, it is possible to get an almost infinite combination of pulse outputs.

Simple time delay. Fig. 9 shows a simple circuit that is a modification of the monostable operation. This circuit provides an output after some predetermined time. Initially, the trigger input to the base of Q_1 is high, causing Q_1 to conduct. Pin 2 of the timer is low and the output (pin 3) is high. When the input goes low Q_1 turns off, allowing the timing capacitor to charge. When C charges to 2/3 V_{CC}, the output of the timer goes low. The output will stay low for as long as Q_1 is turned off. The reset pin can be used to keep the output low if required; otherwise it should be tied high.

Sequential timing. Fig. 10 shows another type of delay circuit. This is a sequential timer. The output of the first timer is used to trigger the second. The control-voltage pin is used to vary the sequence time, but the ratios remain the same. The timing diagram shows how the pulse width is reduced as the control voltage pins are lowered.

Delayed triggering. When power is first applied to the circuit shown in **fig.** 2 there is a chance that the trigger voltage on pin 2 will be lower than 1/3 V_{CC} and the timer will trigger. This may not be desirable. To prevent this initial trigger when power is turned on, the reset circuit can be modified as shown in **fig. 11.**

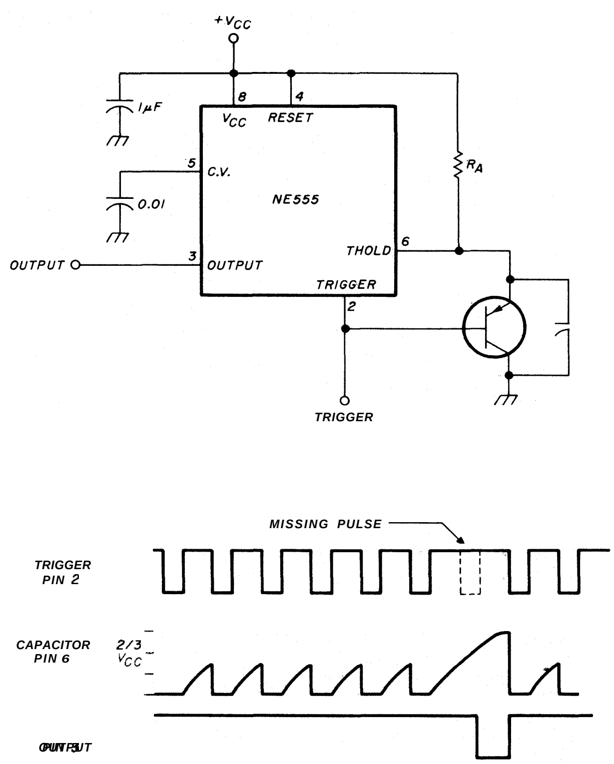


fig. 12. In the missing pulse detector, each trigger shorts the charging capacitor, preventing the output pulse from appearing. If a pulse is missing, the capacitor will charge to the required level, producing the output pulse which indicates the missing pulse.

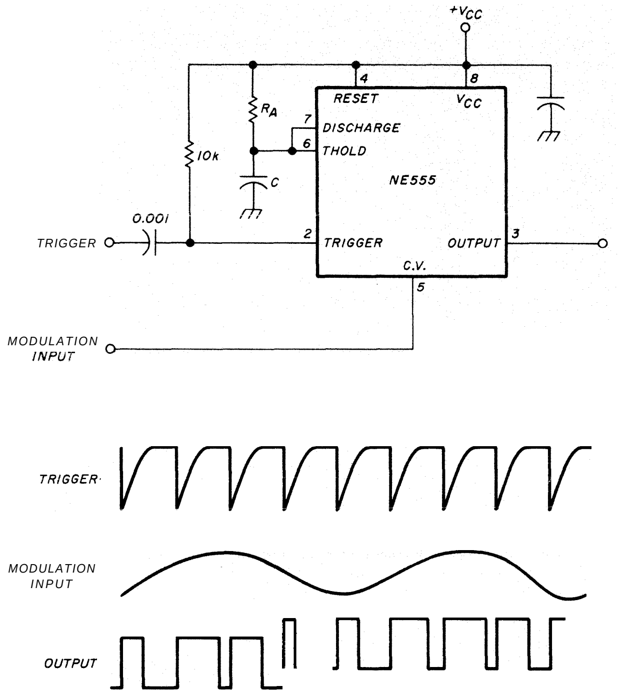


fig. 13. The modulation input is applied to the control voltage input, with the width of the output pulse varying as the amplitude of the modulation.

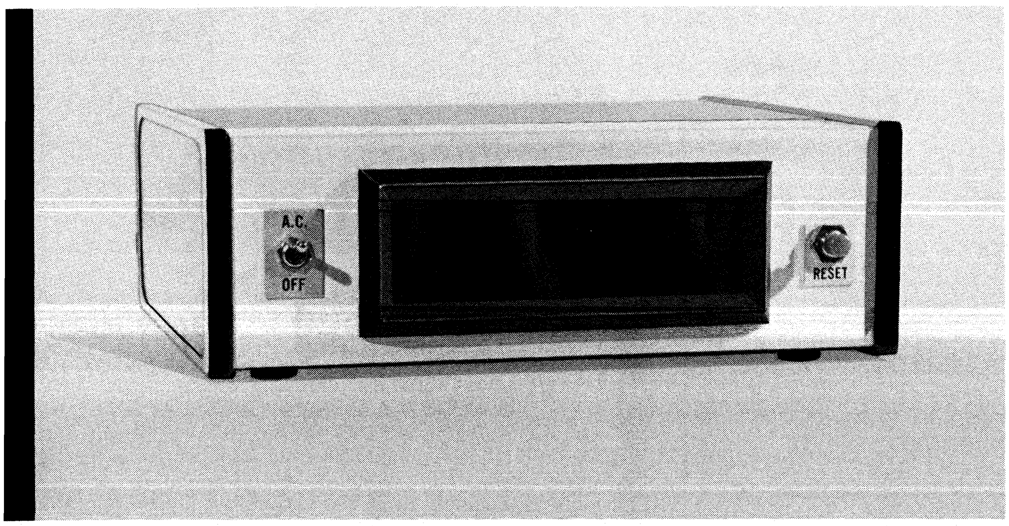
Pin 4 is held below 0.4 volt since C_R is fully discharged. When power is applied, the timer is held reset until C_R charges to above 0.4 volt.

Missing pulse detector. Fig. 12 shows an NE555 timer hooked up as a missing pulse detector. The trigger input also drives the base of a transistor. When a trigger occurs, the transistor conducts and shorts the timing capacitor before the timer can time out. If an input trigger is missing, the timer will time out and pin 3 will go low indicating a missing pulse. The values of R_A and C are set to be slightly longer than the period of incoming pulses.

Pulse width modulator. Fig. 13 shows a circuit which uses the control voltage pin to modulate the pulse width of an incoming clock signal. As the voltage on pin 5 varies, it changes the threshold level of the internal comparator and the pulse width of the output changes.

The examples covered were used to illustrate some of the possibilities of the 555 timer. The uses for this handy little IC could well take a whole book to illustrate. With an understanding of the basic operation, you can analyze timer circuits and design a circuit for your particular need. Who knows, you may come up with an original application and add your name to an evergrowing list of 555 timer circuit designers.

ham radio



receiver digital display

A digital display for your receiver featuring 100-hertz readout, single frequency input, and provisions for forward or reverse tuning VFOs

The addition of a digital frequency display to a receiver goes a long way in enhancing operator convenience. It eliminates squinting at the fine markings found on the usual mechanical dial; and, at the same time, it provides an accurate frequency readout across an entire band without any need for recalibration between band ends. A bright digital display is not only easy to read, it also adds a touch of class to a perhaps otherwise ordinary station.

The circuit described in this article provides a stable, four-digit readout that includes the 100-kHz digit through the 100-Hz digit and will accommodate forward or reverse tuning VFOs at the flick of a switch. The resolution is greater than is usually provided by the typical receiver dial and is handy for returning to a particular frequency in a crowded band. In the interest of economy and simplicity, I did not include the MHz digits; they are easily read from the receiver band switch.

The technique used here[1] requires only a single connection to the receiver's VFO, which, in many modern receivers, is usually available at a connector on the rear panel, making it unnecessary to tamper with the receiver in any way.

theory of operation

In the case of a backward-tuning VFO, the frequency to be displayed is equal to a fixed frequency minus the VFO frequency. All that needs to be done is store the fixed frequency, subtract the VFO frequency, and display the result. In the case of a forward-tuning VFO, store the complement of the fixed frequency (subtract the fixed frequency from zero), add the VFO frequency, and display the result. These additions and subtractions are easily accomplished by using up/down counters and sequentially gating the frequency to be added to the up input and the frequency to be subtracted to the down input, each for the same fixed interval. The fixed frequency is usually crystal controlled and changes very little with time. Thus, measuring and storing it once during an initial calibration is usually adequate.

The price paid for the single interconnection and the simpler circuit is the need to perform a single calibration step each time a new band is used. The operator, using the receiver's internal calibrator, tunes to the bottom edge of the band and presses the calibrate (RESET) button on the display. No further adjustments are necessary. Depressing the

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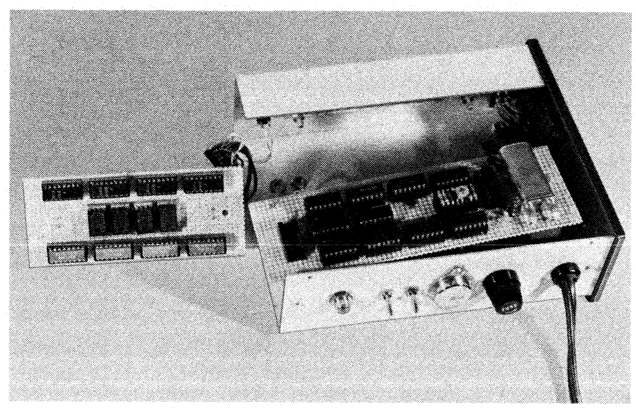
CALIBRATE button stores the VFO frequency and displays all zeros. The operator may now tune up the band in the normal manner and read the display as the frequency above the bottom edge of the band. On the twenty-meter band, for example, 14.0253 MHz would be displayed as 025.3, with the 14 read from the receiver bandswitch. This calibration requires only a few seconds and is something that should normally be done anyway.

The circuit will handle either forward- or reverse-tuning VFOs, but, for the purpose of this explanation, assume a reverse-tuning VFO. Pressing the CALIBRATE button will clear the 74192 up/down counters and cause them to count the VFO frequency in the up direction for 100 milliseconds. The result is then stored in the 7475 latches. All subsequent count cycles will start at the number stored in the latches and count down for 100 milliseconds. Suppose the calibrate button were pressed when the receiver was tuned to exactly 14.0000 MHz. As a result, 5000 would be stored in the latches because the VFO was at 5.5000 MHz. Tuning the receiver to 14.0253 MHz would cause the receiver VFO to be 25.3 kHz lower, or 5.4747 MHz. The display counter would now start at 5000 and count down for a period of 100 milliseconds. The resulting count would be 5000 minus 4747, or 0253. By permanently enabling the decimal point to the left of the least significant digit (LSD), the display would show 025.3. Notice that by counting for 100 milliseconds, 54747 cycles are fed to the counters, but, since it has only four displays, the figure 5 is lost.

For a forward-tuning VFO, the sequence switch is placed in the reverse position. Pressing the calibrate button will then cause the first count cycle to be a down count with all subsequent cycles up counts.



Inside view of the receiver digital display. The displays, up-down counters, and latches are mounted on the small board at the left. All other circuitry is mounted on the board inside the enclosure.

circuit description

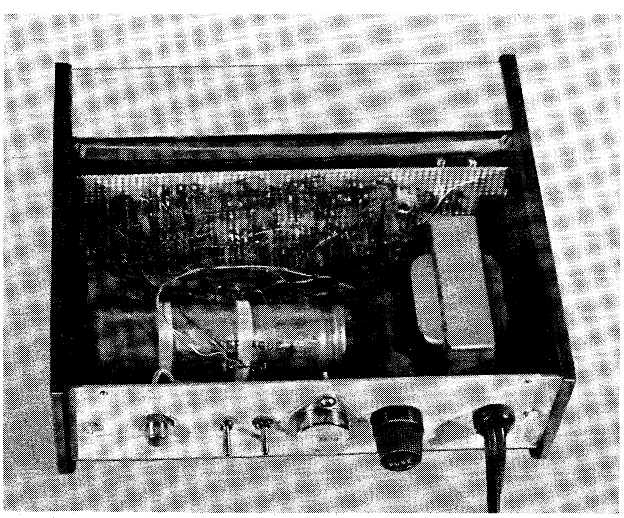
One section of U1 serves as a 100-kHz crystal-controlled oscillator (see **fig. 1**). A divider chain, consisting of U2, U3, U4, and U5, drives a 7490 decade counter, U6 with one clock pulse every 100 milliseconds. The output of U6 is decoded by U7, with output pins 1, 3, and 4 each sequentially going low for 100 milliseconds of each measurement cycle. A jumper between pins 2 and 8 of U6 causes it to have only four output states rather than ten.

Assume the sequence switch is in the normal position. Pressing the CALIBRATE button causes the sequence of the next cycle to be *clear, count up, update.* This clears the counters to zero, counts for 100 milliseconds, and stores the final count in the 7475 latches. All succeeding cycles follow the sequence *load, count down, latch.* This loads the number stored in the latches into the counters, counts down for 100 milliseconds, and latches the final count into the displays.

U11 synchronizes the beginning and end of the calibrate cycle with the state of U7. Pressing S1 causes pin 5 of U11 to go high. As a result pin 9 goes high at the end of the current cycle and causes the next cycle to follow the calibrate sequence, in addition to resetting the first half of U11. At the end of the calibrate sequence, pin 4 of U7 goes high, resetting the second half of U11. U10 is the equivalent of a double-pole, double-throw switch and determines the count sequence, either up/down or down/up.

The remaining sections of U11 amplify and square-up the VFO signal. Three 4049 CMOS inverters are used. The first is biased to serve as an amplifier and functions well to above 6 MHz. This is fine for the large number of receivers with 5 to 5.5 MHz VFOs; but if your VFO operates at frequencies above this, a

Inside view of the assembled digital display. The smaller perf board is mounted right behind the bezel.



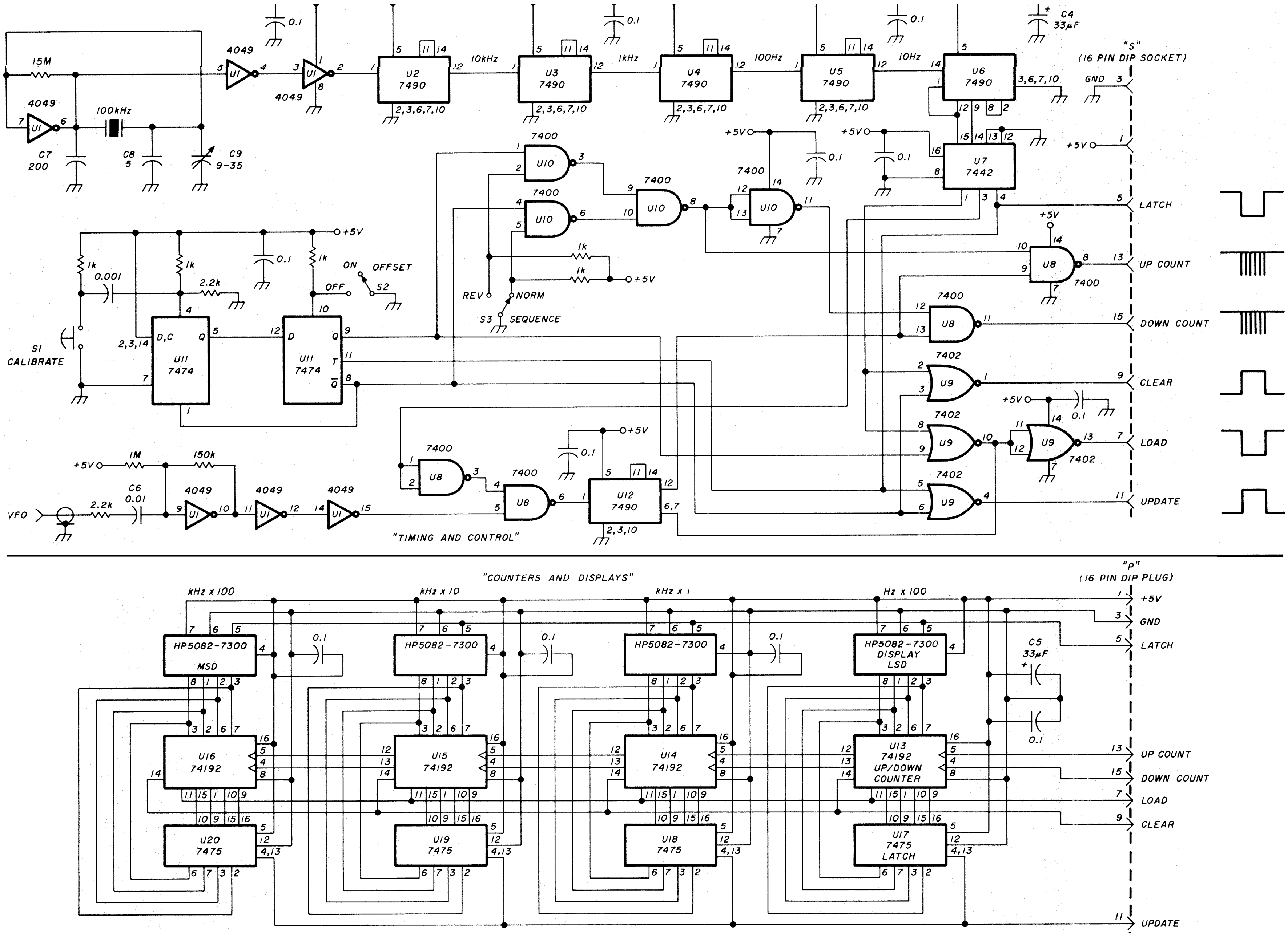
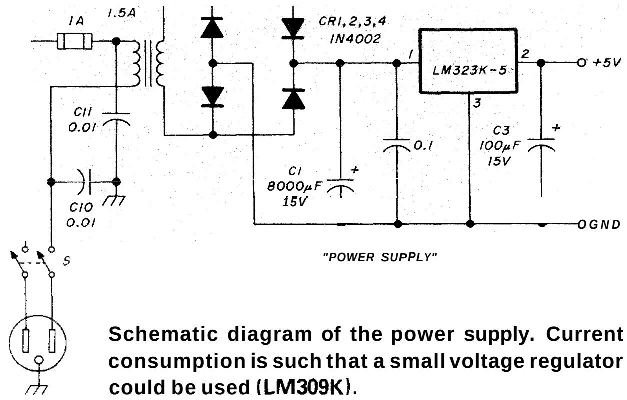


fig. 1. Schematic diagram of the complete digital display. C7 and C8 are silver-dipped micas; C9 is a small ceramic trimmer. Substitution of the LS-series ICs for the 7490s and 74192s will greatly reduce the power supply current permitting the use of a small regulator and transformer.



Schematic diagram of the power supply. Current consumption is such that a small voltage regulator could be used (LM309K).

more conventional frequency counter type of circuit using one or more discrete fets is in order.

U12 operates as a straight decade divider and provides a stable readout with no last digit flicker by acting as an undisplayed counter stage. The display may also be used as a conventional up counter or a straight down frequency counter by merely throwing the offset switch, S2, to the off position.

construction

My display was built in a 6.4 x 18 x 15 cm (2% x 7 x 6 inch) Ten-Tec enclosure using a commercial bezel assembly available from Digi-Key. No printed circuit board is used. Instead, I used two pieces of 2.5-mm (0.1-inch) grid perf board, one for the counters and displays, and the other for the timing and control circuits. All ICs are mounted in wirewrap sockets which are in turn cemented to the perf board. Most of the interconnections were made using standard number 30 AWG (0.25-mm) wire-wrap wire and an inexpensive hand-wrapping tool. The only exception is the power buses, which are number 18 AWG (2-mm) solid bare wire mounted on push-in solder terminals and run on either side of each row of IC sockets. Bypass capacitors (0.01 μF) were used liberally throughout the circuit. The voltage regulator is mounted on the rear of the cabinet using thermal compound, and several vent holes were drilled in the cabinet sides to remove internal heat.

The displays shown in the schematic are very compact and convenient, as they include decoders and latches, but they tend to be rather expensive if they must be purchased new. A functionally equivalent and much cheaper arrangement could be made using 7475 latches, 7447 decoder/drivers, and common-anode, seven-segment LED displays,[2] but a slightly larger cabinet would probably be required (see **fig. 2**).

The 100-kHz crystal I used is of unknown lineage and different crystals may require a slight adjustment of component values in the oscillator circuit. The 8000-μF power supply filter capacitor could probably be reduced to half the value without any problems and a 1.5-amp regulator substituted for the LM 323, as the current requirement at 5 volts is about 900 mA, a bit less than I originally anticipated.

calibration

Clip a piece of insulated hookup wire to pin 2 of U1 to act as an antenna and place it near a receiver tuned to WWV. Zero beat WWV with the trimmer and you're ready to go. If you have a good frequency counter, adjust the trimmers for exactly 100 kHz.

I hope that the ideas and circuits presented here prove to be informative and may perhaps encourage the construction of improved designs using hardware that will appear in the future. Counter latch combinations and dividers with a modulus greater than ten could serve to simplify construction considerably. For anyone attempting construction as shown, I would be pleased to supply advice or answer questions. Please include a stamped, self-addressed envelope with all inquiries. Now that I have had the use of this display for several months, I wish I'd built one years ago.

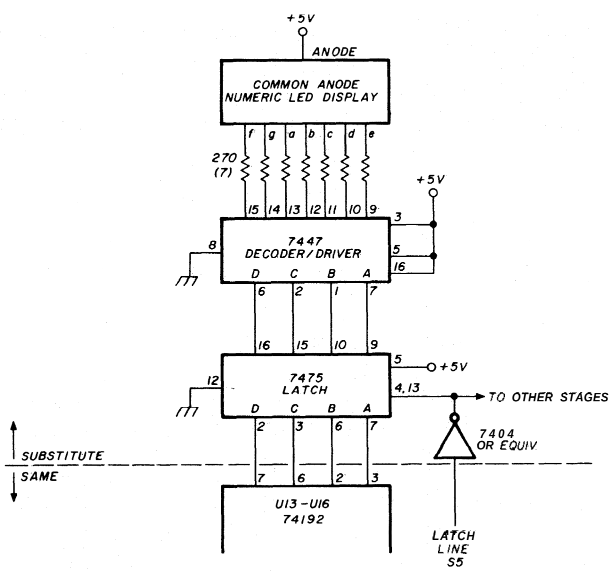


fig. 2. Diagram of a less expensive substitute for the HP 5082-7300 LED display.

references

1. Jon Hagen, W7URZ, "A Simple Frequency Counter for Receivers," QST, December, 1972, page 11.

2. Bruce McNair, WB2NYK, "A Digital Frequency Display for Amateur Communications Equipment." *ham radio*, September, 1976, page 16.

ham radio

new approach for a I-MHz oscillator

A reversion to basic oscillator concepts, this oscillator was built with optimized circuitry for each stage

Most of the oscillators built by Amateurs use a single transistor to perform several functions. However, the basic principles of operation are somewhat a mystery. The original analysis of the oscillator, done in the 1920s, showed that it could be divided into separate stages (see **fig. 1**). By actually building separate stages, you can obtain a better understanding of oscillator operation. This article presents an oscillator built using those basic concepts.

circuit description

The RCA CA3028 differential amplifier, shown in **fig. 2,** is used as an amplifier and limiter. Feedback, for the actual oscillator, is the limited sinewave from the pin 6 output. Note that this is a noninverting configuration. The amount of feedback is controlled

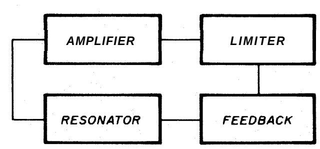


fig. 1. Diagram of the basic stages needed for an oscillator.

by the 5-kilohm pot connected to pin 6. Feedback is injected into the quartz-crystal resonator. Because the crystal acts as a bandpass filter, the output will be a sinewave, in this case at 1 MHz. However, the crystal cannot be connected to the limiter without a penalty. The input impedance of a limiting differential pair is nonlinear. To overcome this problem, the crystal must be isolated from the limiter. This is done by connecting a common-base amplifier between the resonator output and the amplifier input. The com-

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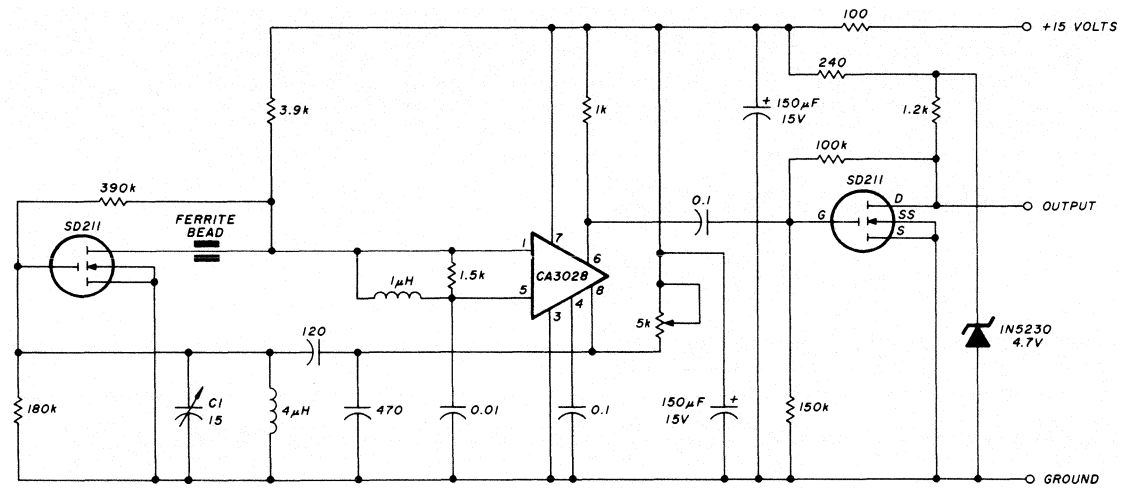

fig 2. Schematic of the 1-MHz oscillator built using separate stages for each function rather than combining functions into a single device. All resistors, except R1, can be 118 watt: R1 is a 114-watt resistor. The power requirement is 36 mA at 15 volts

mon-base amplifier provides a constant impedance load for the crystal. Because the output of the limiter is fixed, regardless of its input, and because the resonator also sees a constant impedance, the sinewave at the collector of the common-base stage has should oscillate. If it does not, increase the feedback until oscillation occurs at 8 volts. After the feedback is correctly adjusted, connect the output to a counter and adjust the crystal frequency with the small trimmer.

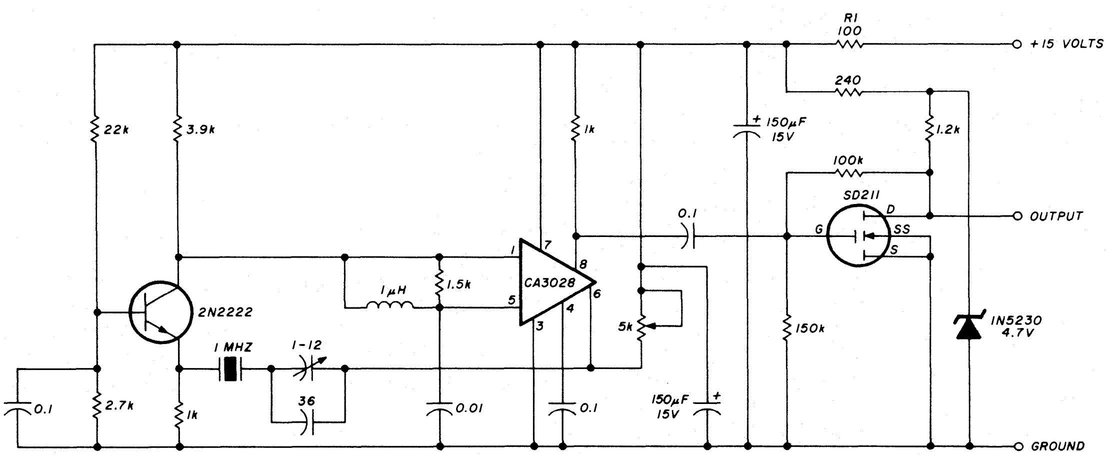

fig. 3. Diagram of a 40-meter LC-tuned VFO built using the same guidelines as the crystal oscillator.

a constant amplitude. This signal completes the loop by driving the limiter. The output is taken from pin 8, the unused inverting output. The output from the CA3028 is a limited sinewave. To make the output a clean square wave, an SD211 DMOS FET was used. The FET is driven into saturation, squaring the signal.

construction and adjustment

The oscillator was built on a piece of perforated board and housed in a 10 x 5.4 x 4.3 cm (4 x 2-1/8 x 1-5/8 inch) mini-box. Adjust the 5-kilohm variable resistor until the circuit goes into oscillation. Next, reduce the power supply voltage to zero and slowly increase the voltage. At 8 volts, the circuit

The output is a 4-volt peak signal with a rise time of 100 ns. The variation in frequency, at 1 MHz, is 0.5 Hz over a 1-second interval. This is due to the quality of the crystal and thermal effects. This variation can be reduced if a higher-quality crystal is used. Also, the environment can be temperature controlled with an oven or the use of insulation. The final frequency error was adjusted to 0.05 Hz. My present application for the oscillator is a time base in a frequency counter. However, you could modify the circuit to act as an LC-tuned VFO (see **fig. 3**). Since the SD211 FET amplifier is inverting, feedback is taken from pin 6, the inverting output.

ham radio

novel method for matching input impedance of grounded-grid power-amplifier tubes

When I was thinking about driving a tube power amplifier with my 20-watt transistorized ssb general-coverage transceiver, I noticed that uniform matching from the 50-ohm output of the transceiver into the amplifier between 1.8 and 30 MHz might present some problems. A circuit with a grounded cathode tube wasn't too attractive, because of the required neutralization and the various voltages needed for a tetrode or pentode. So I decided to use one of the modern grounded-grid tubes of the 8873-8875 series. Their input impedance is in the vicinity of 100 ohms, but since it's a dynamic input, impedance varies as drive level changes.

It's therefore highly recommended to use at least one tuned circuit between the exciter and the cathode connection of the tube to store energy and stabilize the dynamic input impedance. Since this tuned circuit requires retuning when the frequency is changed more than 10 per cent, such an arrangement would be highly impractical.

Note that this energy-storing circuit, although sometimes used in a pi configuration, does not really suppress exciter harmonics. Standard practice is to have enough harmonic suppression already in the exciter, and a 50-ohm output is provided from the exciter.

matching network

With these things in mind, and using a calculator to determine elliptical filters, I constructed a matching network that provides perfect matching and ideal energy storage between 1.8 and 30 MHz **(fig. 1).** This is a bandpass filter, which has Chebishev response in both bandpass and stop-band areas. It provides a constant impedance match over the frequency range of between 1 MHz and 60 MHz. This is important, because even high-order harmonics must be properly terminated, otherwise high transient voltages can be developed in the driver stages.

If you consider the fact that the frequency range between 1.8 and 30 MHz, on the basis of individual filters, must be split into at least seven segments

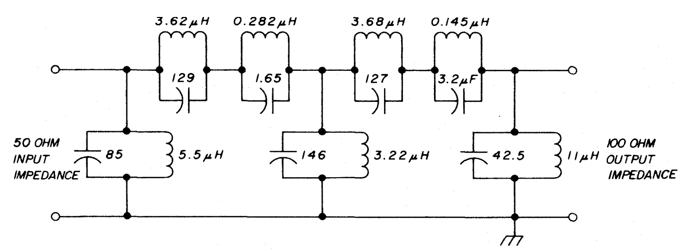

fig. 1. Simple network for constant impedance matching and ideal energy storage between a low-level driver and power amplifier using tubes in the grounded-grid configuration. System is a bandpass filter with Chebishev response in both bandpass and stop-band regions. Frequency range is 1-60 MHz. See text for recommended components.

(and assuming three components per segment), the total number of segments would be twenty-one, while here only fourteen are required.

components

All component values are included in the schematic. You must use 500-volt mica capacitors together with either air-wound coils or coils with suitable ferrite materials, such as toroids. The recommended material for the latter is either Q1 or Q2 (Indiana General), Part no. F625/9 (dimensions in inches: D1 equals 0.375, D2 equals 0.187, and H equals 0.125).

A filter of this design has been successfully used, and no contribution to intermodulation distortion has been measured at a 25-watt drive level.

ham radio

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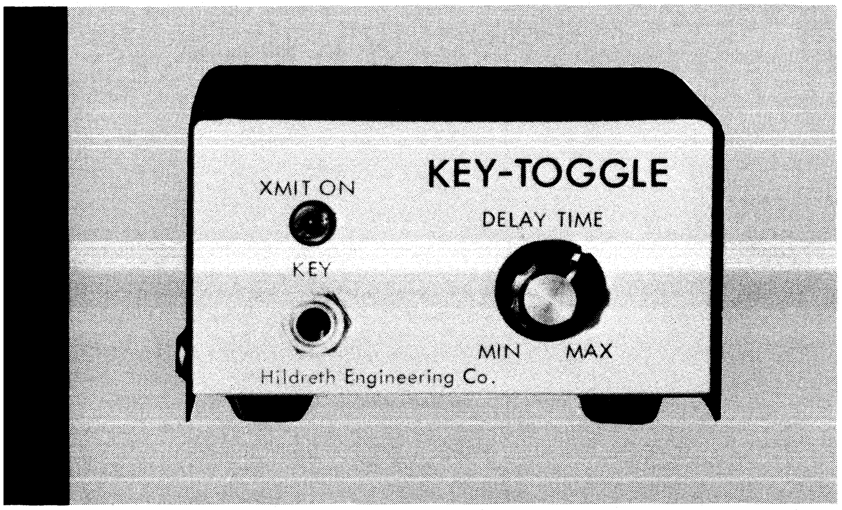

the key-toggle

Key-toggle, another form of fast break-in, featuring simple but effective hand-key control of the transmitter

I wonder whether some psychologist won't someday discover that ham radio owes much of its appeal to the fact that you can say your piece in silence and the other guy can't interrupt until you're through. Even in most VOX systems, the other person cannot force interruption. Perhaps this is one of several reasons why "full break-in" CW operation has not yet become prevalent. It is not difficult technically, but there are still those of us who choose not to use it.

For handling commercial traffic on a clear channel, full break-in is perhaps best; but for most other activities, coupled with reasonable and unselfish operating practices, I prefer my key-toggle mode. To come on the air, you just start keying. To switch to receive, you simply hold a dash a little longer than normal (time adjustable). Key-toggle can be just about as fast as full break-in, but the other guy can't break you. For example, when you are winding up a transmission, the last letter normally sent is dah-di-dah. With key-toggle, it's dah-di-dahah, click, click. The first click switches your transmitter's plate supply off and the second click, milliseconds later, switches the antenna relay and opens the receiver (optional). You are now in receive mode until you start keying again. When you hit the key, once again: click, click. But this time the antenna relay is switched and the receiver muted first; then, milliseconds later, the transmitter's plate supply comes on and you are ON again. No combination of letters or long pauses will change the relays, switching you back to the receive mode. Only a long dash will toggle your station back to receive.

circuit description

Major components, as seen in fig. **1,** include a 741 operational amplifier configured as a dual time constant integrator, a 711 comparator, a 7474 TTL flip-flop, a few transistors, and three relays. The KEY UP output of U1 is a nominal +2 volts. Q1 is conducting, which effectively limits the voltage across the 1-μF timing capacitor to near zero volts. With the +1 volt on pin 2, and 3 volts on pin 3, the output of U3 is also low. This places K2 and K3 in the de-energized (receive) state.

When the key is pressed, Q1 is cut off, which lets

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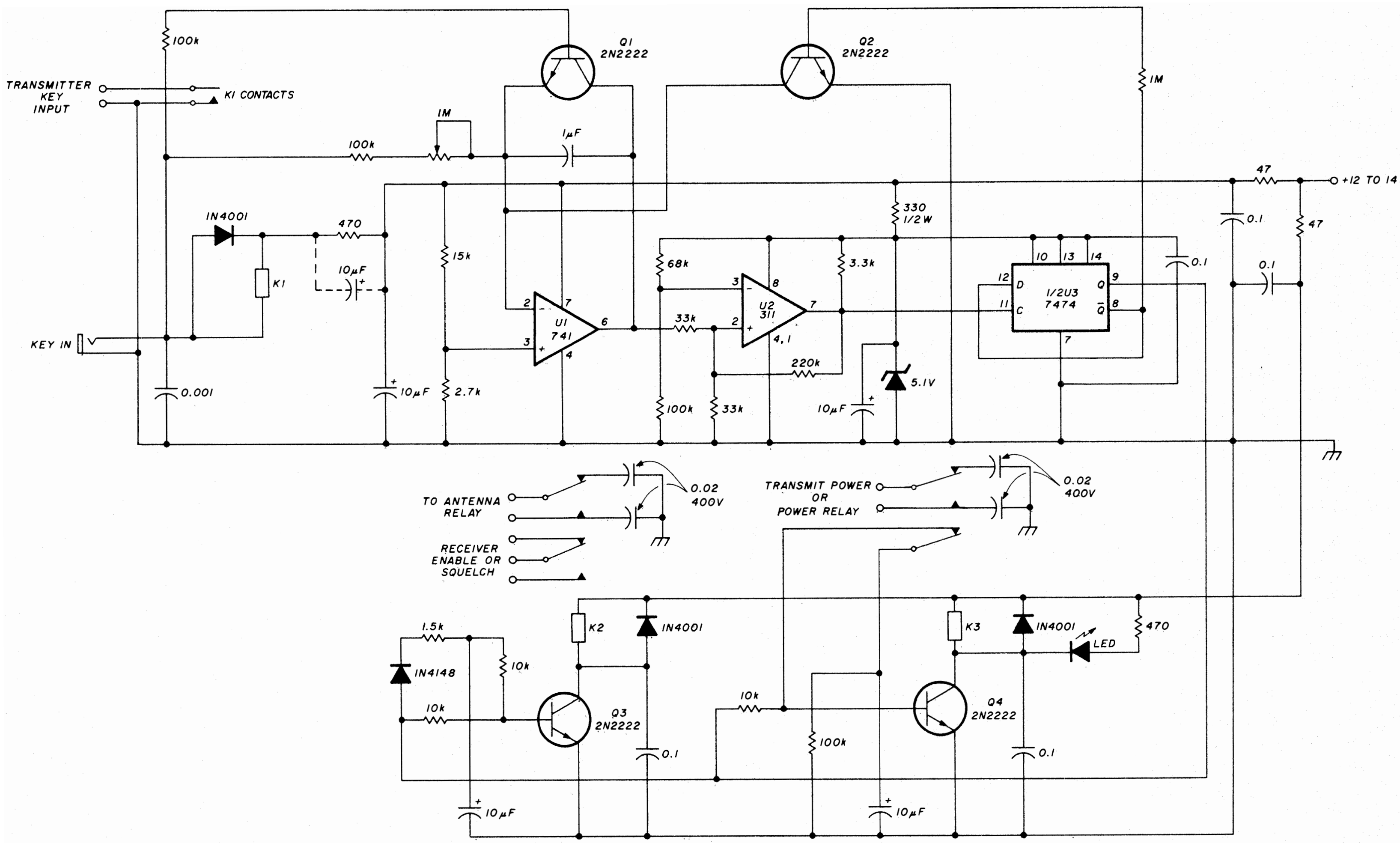

fig. 1. Schematic diagram of the key-toggle system. K1 is a Radio Shack relay (275-004); K2 and K3 are relays 275-206. Ensure that the contact ratings of the relays are not exceeded by the controlled equipment.

the op-amp circuit start ramping positive because of the +2 volts and the timing components connected to pin 2 of U1. When the level at pin 2 of U2 just exceeds the voltage at its negative input, the comparator snaps high (with the help of some positive feedback). This toggles the U3, making Q high (pin 9) and $\overline{Q}$ low. With pin 9 of U3 high, a turn-on volt-

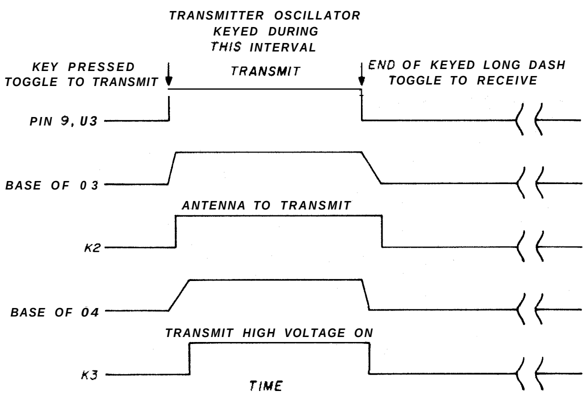

fig. 2. Relay keying sequence when switching from receive to transmit.

age is fed to Q3 and Q4 through timing control networks. The base network for Q3 comes on first and causes it to turn on very fast and turn off slowly. Q3 operates K2 for antenna control and receiver muting. The RC network connection through K3 to the base of Q4 causes this transistor to turn on after Q3. After K3 is activated, the RC network is removed, allowing Q4 to turn off before Q3. **Fig. 2** shows the timing sequence for K2 and K3. K2 is the basic keying relay and must be able to follow keying speeds. For some keying circuits, K1 may be replaced with a 1-kilohm resistor for direct keying.

By using the $\overline{Q}$ output of U3 and *Q2* as a switch, the system provides a short time constant for switching from receiver to transmit and the longer, variable-time constant when switching from transmit to receive.

why key-toggle

Back in May, 1950, Hiele showed a system[1] which, with variations, has been much used. In Hiele's system, the switch-over from receive to transmit is done quickly, just as in key-toggle, but the transmit mode switches to receive after your key is up a selectable time, sort of a "dead man's switch." This system is used in many modern transceivers through an application of the VOX system to CW operation. The relay timing system is simpler with this method because your key must be up in order to switch, but it has its own drawbacks in operation. If you adjust the time delay to allow for reasonable pauses when sending, the time seems like forever when switching over. And, if you adjust the timing for a quick turnover, your receiver blasts you when you pause. With key-toggle, you are in command. In fairness, however, it is recognized that key-toggle is not applicable to any automatic type of keyer.

power supply and construction

I used an inexpensive wall transformer supply system, although the 12-volt units usually supply 14 volts when lightly loaded, this poses no problem to the relays or op-amp. A zener diode is'used to obtain +5 volts for the comparator and flip-flop. Maximum current is approximately 150 mA. Since this device is always used where strong rf fields are generated, a metal enclosure is probably mandatory.

connections

Outputs from ***Key-Toggle*** are contact closures intended to operate existing antenna and transmitter relays. Capacitors have been chosen to prevent switching transients from burning contacts and from false-triggering U3. For low voltage operation, as is the case in solid state equipment, the closures may be applied directly to the equipment.

operating with key-toggle

Assuming you are in receive status, you switch to transmit at the initial touch of your key (the switching process will shave a small piece off of your first dot or dash). Your transmitter will remain active until you hold a dash slightly longer than normal, the exact time determined by the TIME DELAY setting, which may be set between approximately 0.2 and 2.0 seconds. Adjust the time delay pot such that the time required to toggle your system is slightly longer than the time length of the dashes at the rate you intend to send.

One note, however: If you simply press your key and hold it when switching from receive to transmit, your transmitter will remain ON as long as the key is held down. After switching from receive to transmit, your key must be released at least once before a dash can toggle the relays.

This system takes a little time to get used to, but once you get the hang of it you can actually toggle for a quick listen between words, even between letters when you are not sending too fast. And switching will only occur when ***you*** choose.

reference

1. M. E. Hiele, W2SO, "An Automatic Transmitter Turner-Onner," QST, May, 1950, page 56.

ham radio

the weekender

i-f transformers — problems and cures

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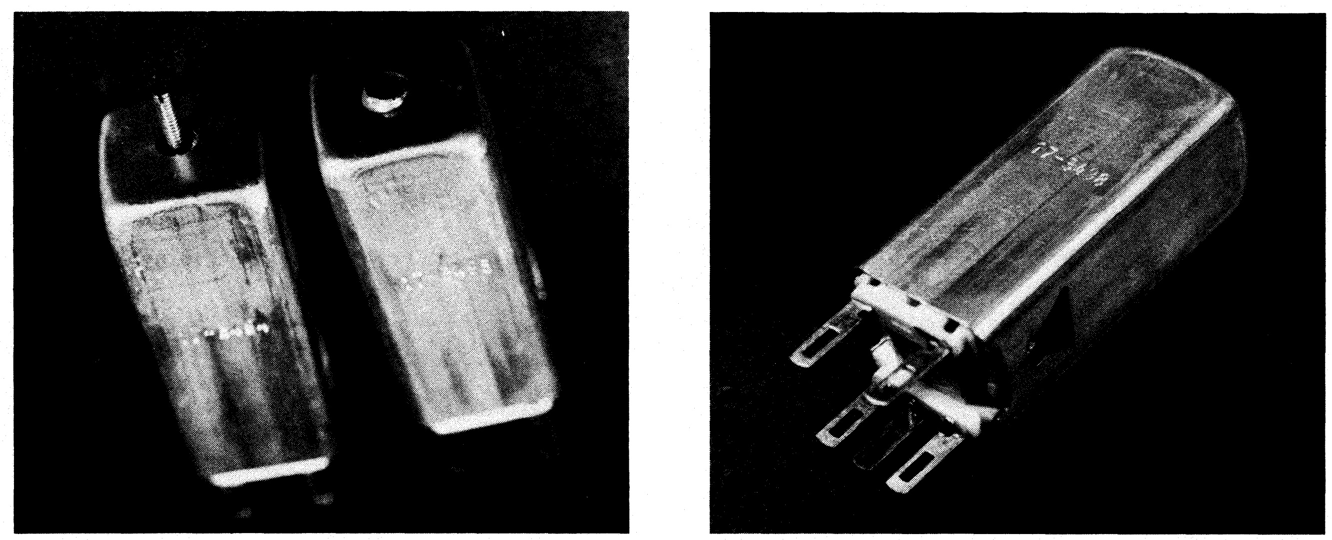
All **serious communications** receivers, transceivers, and most transmitters use the heterodyne principle for frequency generation. For receivers, various frequencies are converted to a single frequency called the *intermediate frequency* (i-f), while transmitters, on the other hand, will create the ssb signal at an i-f and then heterodyne it to the ham bands.

An i-f amplifier is a tuned radio-frequency amplifier that operates on a single frequency (*e.g.*, 455 kHz, *3350* kHz, or 9 MHz). In most cases, the tuned circuits in the i-f amplifier are tuned transformers (the primary and secondary coils are resonated by a capacitor) in a shielded can.

Fig. 1A shows two i-f transformers. Both of these are in the now-standard 1.9-cm (0.75-inch) shielded can. One of the transformers uses a threaded rod attached to the inductor core for tuning, while the other uses just an access hole to admit a tuning tool. In some cases, the inductor core will have a screwdriver slot to permit adjustment, while in others a hexagonal hole is cut through the center of each core. In the latter case, a special insulated hex alignment tool is used. The hex type offers the advantage of permitting adjustment of both primary and secondary tank circuits from the same side of the chassis.

Fig. 1B shows the pins at the bottom of the transformer. For an i-f transformer, at least four pins are required and some may have five or six pins. Some older models used wire leads instead of pins, and, of course, transformers intended for printed-circuit mounting will have solder tails instead of pins.

Fig. 2 shows several of the literally dozens of i-f transformers used in Amateur Radio gear. The transformer in **fig.** 2A uses two tuned coils. In a few early receivers, the tuning of the tanks was accomplished by the adjusting capacitors, the adjustment screws being accessible through holes cut into the shielded can. Most modern transformers, however, use a slug-tuned coil to tune the tank. The version shown in **fig.** 2A is commonly found in vacuum-tube equipment. **Figs.** 2B and **2C** show transformers designed for use in solid-state equipment. The version in **fig.**



A B

fig. 1. Two versions of the standard 1.9-cm (0.75-inch) i-f transformer, showing the different methods of tuning the inductor. (B) shows the connection pins at the base of the transformer.

2B uses taps on the primary and secondary windings to match the low impedances normally found in transistor circuits. In **fig.** 2C the secondary is an untuned link. The transformer in **fig. 2D** is very similar to the one in **Fig. 2A,** except that a built-in bypass capacitor is provided to decouple the cold side of the secondary. These i-f transformer circuits are not unusual; they represent those most commonly found in receivers.

Fig. 2E shows a pinout configuration used by many manufacturers; note the odd sequence for numbering the pins. A color dot, usually green, identifies pin number one, with pins two through four occupying the four corners. These pins are usually connected to the tank circuit in the manner shown in **fig. 2A.** Pins 5 and 6 are reserved for such things at taps **(fig. 2B)** or bypass capacitors **(fig. 2D)**.

Older units, using wire ieads instead of pins often follow the specific color code:

blue – plate
red – B+
green – grid or detector diode
black – grid or diode return

transformer failures

Fortunately, i-f transformers are relatively simple devices, with failures limited to a few problems, such as open windings, shorted or open capacitors, windings shorted together, and the inability to resonate.

Open windings rarely occur anyplace except right at the transformer pins. **Fig.** 3 shows a close-up view of the wire from one winding that has broken at the pin. Once the signal tracing gets you to the correct stage, this type of problem can be diagnosed with an ohmmeter.

fig. 3. Example of a broken transformer wire at the connection pin.

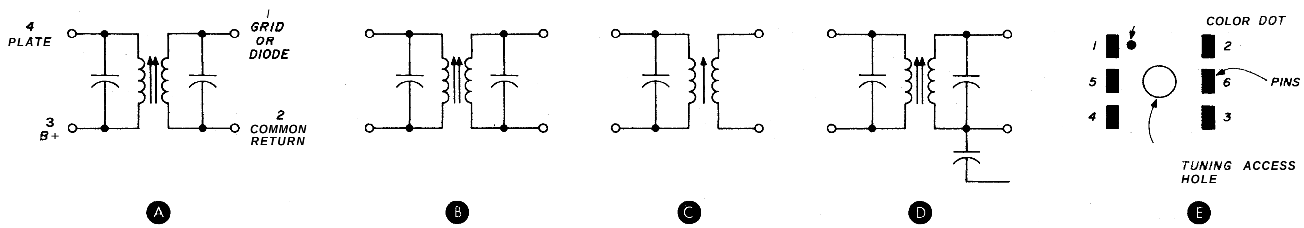


fig. 2. Versions (A) through (D) show different configurations for the i-f transformers. The pin numbering system is shown in (E) (bottom view).

The open winding is also the usual cause of the intermittent i-f. Interestingly enough, I have seen many cases over the years where the wire had never been soldered. The transformer worked nicely for many years, until either corrosion built up or a mechanical jarring knocked the wire loose.

Repairing the open-winding problem is easy in many cases – if you are gentle. Use a small screwdriver to pry open the metal shield. Most of the transformers in use have a coil form mounted on a plastic base, with the entire assembly slid into the shielded can. Tabs on the can are then bent over to form retainers which keep the coil form and base in place. It is a simple matter to pry these tabs loose and then gently pull the base from the metal shell.

Use small long-nose pliers when working on the transformer. In fact, it is best if you use tweezers instead of pliers. Avoid pulling on the wire; it *will* break, and right at the coil form! It is then almost impossible to repair the coil without rewinding it.

The wire should be resoldered to the pin using a small, pencil-type soldering iron, not a gun. There will usually be enough pretinned wire left to allow resoldering to the pin. But, if not, do not try stripping the wire or you will break it. Insulation may be removed by gently scraping the wire with a razor blade or sharp knife. Keep in mind that even this is at the risk of breaking the wire. The best method is to melt the insulation with the tip of the pencil iron.

Ordinarily, there will be enough slack to make the new connection without stretching the wire. If not, bend the pin toward the wire – do not try to stretch the wire. If this is not possible, then solder a short

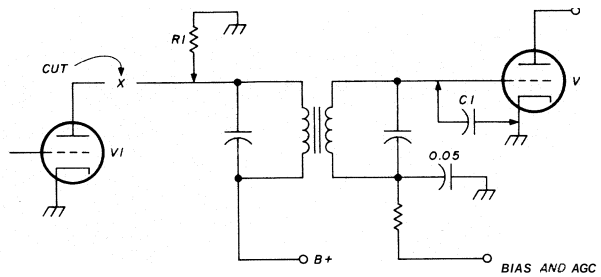


fig. 4. Technique for isolating a noisy i-f transformer. R should be equal to the plate resistance of the tube, with C being 0.01 to 0.1 μF.

piece of wire to the pin for use as a jumper. A single wire removed from a piece of stranded hookup seems about right.

Because of the low coil resistance, it is often impossible to tell if an i-f transformer capacitor is bad. Signal tracing will usually isolate the problem at the correct stage, and then dc measurements and a tube or transistor check will leave you with the fact that there is nothing left except the transformer.

Only one type of problem, the noisy i-f transformer, is easily identified as a bad transformer. This problem produces a crashing, crackling, static-type noise. To isolate the problem, remove the tube from its socket or break the plate lead between the transformer and the plate pin on the tube socket. Temporarily solder a resistor (see **fig. 4**) between the plate and cathode pins (or ground). The value of the resistor should be approximately the plate resistance of the tube. If noise persists, then suspect the transformer. As a quick extra confirmation, shunt a 0.01- to 0.1-μF capacitor across the grid and cathode of the following stage. If the noise disappears, then the i-f transformer is bad.

The i-f transformer is bad if the noise persists when the tube is removed and the resistor is connected to the circuit, or if the noise disappears or is reduced significantly when the shunt capacitor is connected. In a multi-stage, cascade i-f amplifier, each stage may have to be checked separately in succession, beginning with the stage that is closest to the detector.

In many cases, the noise is caused by a bad tuning capacitor inside the transformer can. Some transformers use individual ceramic or silver mica capacitors (which rarely go bad), but in most cases the two capacitors will be as shown in **fig. 5.** In **fig. 5A,** the capacitor is formed from a piece of mica dielectric with the silver plates deposited on both sides. A pair of contact springs connect each plate to its respective pin. This arrangement can cause noise by dc breakdown of the mica or by noisy contacts. In the latter case, cleaning and retensioning often cures the problem. The second arrangement, shown in **fig.** 5B, buries the fixed mica compression capacitors inside of a molded plastic base.

A study by the service department of a major automobile radio manufacturer revealed that, in humid regions of the country, there are roughly two to three times as many trimmer capacitor and i-f transformer problems than in dry areas. The best remedy for this is to replace the transformer. But, if that is impossible or would take too long, try replacing the bad capacitor with a disc ceramic or silver mica capacitor.

If the original capacitor is like the one in **fig. 5A,** it is a simple matter to use side cutters to clip off the top spring clip contact. Then solder a capacitor of the proper value across the terminals (see **fig 6A**) The proper value can be determined experimentally (check first to see if the manufacturer gives the value in the schematic) using a GDO or signal generator to locate a standard value that will allow the coil to resonate at the correct frequency.

If the bad capacitor is not easily removed or disconnected, there is still a possible cure. Disconnect the coil wire going to one end of the capacitor, and then connect it and one end of the replacement capacitor to an unused pin. Or add a pin in one of the blank pin slots **(fig.** 6B) by forcing a piece of hook-up wire through the slot.

The problem of primary to secondary short circuits comes about because the wires from the upper coil pass by, and may touch, the winding of the lower coil. The problem almost always results in B+ leaking through to the grid of the next stage or the plate of the detector diode.

The cure here may be as simple as moving the wires apart and then patching up the damaged spot using Q-dope, high voltage corona dope, or just glue. Replacement of the transformer, however, may be required.

You'll sometimes find a transformer that is not causing static, passes all dc checks, but signals will

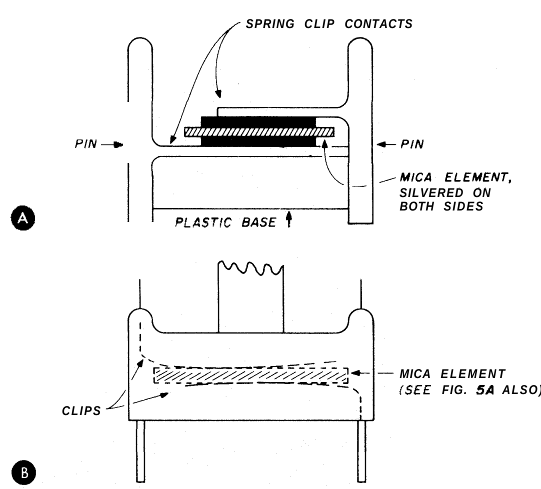
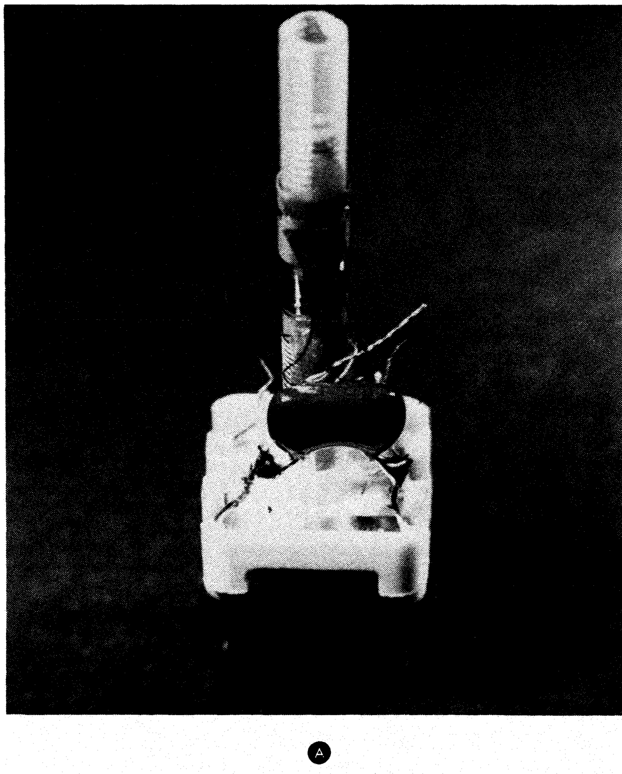


fig. 5. The compression trimmer can be one of two types, either open or enclosed in the plastic base. Cleaning and retensioning the version shown in (A) can cure the problem, though replacement is required for the molded version (B).



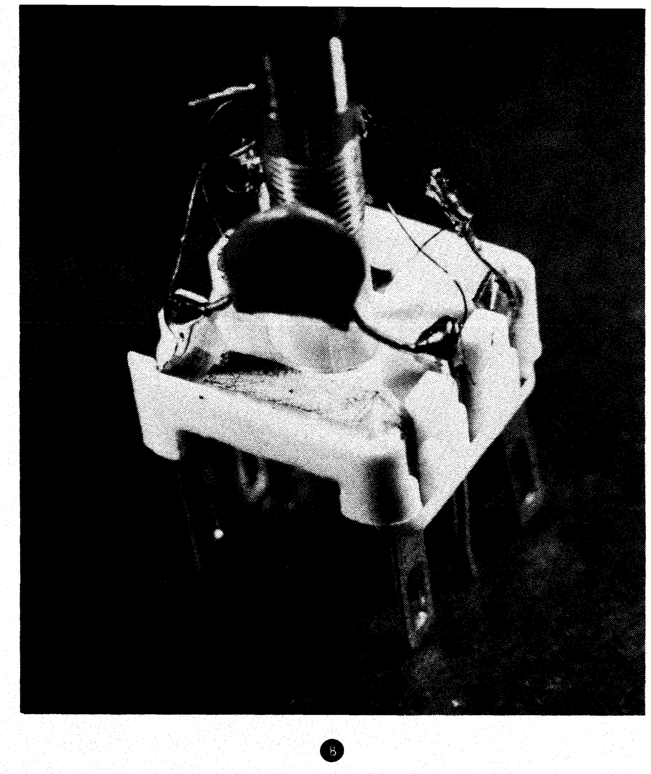


fig. 6. (A) shows a replacement capacitor soldered directly across the pins at the base. To replace a molded capacitor, a new pin is inserted in the base of the transformer (see photograph B).

not go through or are severely reduced in strength. Barring such rare problems as shorted windings and open capacitors, it is often the case that one of the tanks is off resonance due to a cracked or broken ferrite tuning slug. To replace the slug, break it into smaller pieces (it rarely will come out by its threads!) and then remove them by shaking. A replacement slug can then be installed. Find an i-f transformer of the same frequency range and with the same type of slug. Salvage the slug and install it in the bad transformer. Some electronic parts suppliers stock small assortments of tuning slugs for just this purpose. Note that 9-MHz ssb transceiver slugs can often be replaced with a slug from a 10.7-MHz fm broadcast receiver.

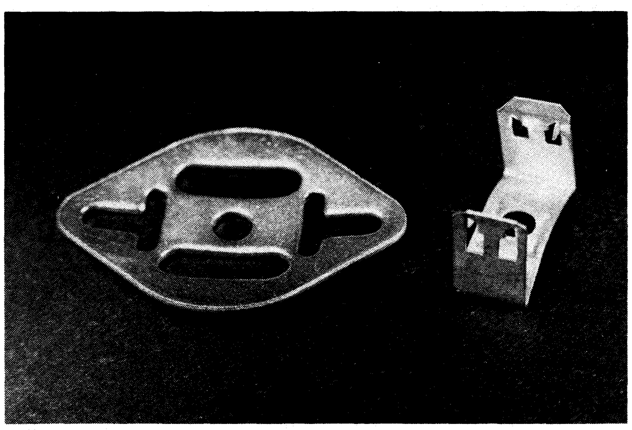


fig. 7. Mounting hardware for i-f transformers, which can be used to replace older 2.5-cm (1-inch) transformers.

Regardless of the problem, except perhaps the open coil repair, the best solution is to replace the bad i-f transformer with a new part purchased from the equipment manufacturer. However, this is not always possible, especially in older gear. Even if the company is still in business, the set may be so old that they no longer support the product. A few years after a model is discontinued, support usually vanishes.

It is possible to buy new i-f transformers. Companies such as J.W. Miller and others still manufacture both direct replacements for many types, and so-called "universal" i-f transformers. Their catalogues should reveal at least one or two currently made models which are good candidates for replacement of the bad unit in your rig. It's unlikely that a cross-reference listing for a piece of ham gear is available, but a careful reading of the specifications and inspection of the circuit diagram will lead to a replacement.

Older 2.5-cm (1-inch) or larger i-f transformers are especially difficult to replace. However, if you are willing to make a few mechanical modifications, the standard 1.9-cm (0.75-inch) transformers can be pressed into service using the adapter plates and mounting clips (fig. 7) provided with replacement transformer.

ham radio

updating the Heath HW-2036

to the HW-2036A

Modifications to the Heath HW-2036 to provide complete coverage of the new two-meter repeater subband

Now that the two-meter band has repeater allocations in the lower half, people with HW-2036 transceivers are feeling left out in the cold. No longer! With a small parts kit available from Heath, a few additional parts obtained (through Heath or purchased on your own), and these simple instructions, you can update your HW-2036 to cover the entire band. In fact, the transceiver will operate over a six-MHz range, from 143 MHz to beyond 149 MHz, although with reduced output at the extreme edges.

updating early HW-2036s:

Before actually beginning conversion of the HW-2036, check to see that your HW-2036 has all the updates that Heath has added since this model was introduced. There were addendum sheets included with the manual, in addition to several changes in the manual itself to reflect parts values different from those printed on the circuit boards. On the receiver board, R206 and R209 should be 10 kilohms, and C214, C215, and C216 should be 4.7-pF NPO disc ceramics. Also, Q204 was changed from a MPF-105 to an EL-131. If you have to replace these six parts, you will need to realign the receiver i-f. Wait, however, until the rest of the changes have been made.

On the synthesizer board, R419 should be 10-kilohms, C442 a 20-pF NPO disc ceramic, C403 a 100-pF NPO disc ceramic, and R445 220 ohms. These parts changes reduce drive to the synthesizer loop mixer. And finally, on the transmitter board, R141 should be 180 kilohms, ½ watt, which increases the zener bias in the 11-volt regulator. Only very early HW-2036s might need any of these changes. Your HW-2036 is now up to date. **Table 1** lists all the parts needed to convert the HW-2036.

conversion to the HW-2036A

There are changes to be made on each board, and we found it easier to do one board at a time and then do the alignment board-by-board when everything was bolted back together. It is necessary to have some method of removing solder from the double-sided boards before attempting to remove the parts.

Power Amplifier Board. Remove the transceiver top and bottom covers and begin with the power amp board by removing the heatsinks from the back panel. Unsolder and disconnect the ground wires connected to the corners of the transmitter and receiver boards. Carefully pull the molex connector apart, remove the two nuts that hold the power amplifier assembly to the back of the chassis, and swing the board down on its leads. A 12-pF dipped-mica capacitor (C235 in the HW-2036A manual) is added to the output connector on the inside of the board.

Prepare this capacitor by cutting the leads 22 mm (7/8 inch) long and slipping 19 mm (3/4 inch) of spaghetti or insulation over each lead. Bend the exposed 2-mm (1/8-inch) lead at right angles away from the body of the capacitor. Lay the capacitor on the foil of the board and solder one lead to the antenna jack center pin and the other to the foil where the ground wire pokes through the board. This forms a filter across the output. The lead length is the inductance part of the filter. Reassemble the board to the rear apron, but leave the ground wires to the transmitter and receiver boards loose for now. Replace the molex connector.

VCO Board. Heath now supplies a parts kit for modifying the VCO; it consists of two mica capaci-

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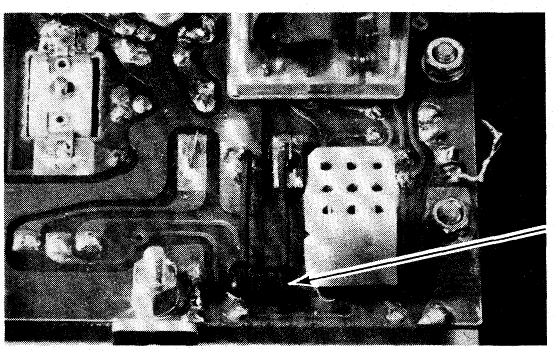
tors, a coil, and a new varicap diode (parts kit number 830-29, $3.95). Begin by removing the VCO cover screws and unsoldering the two lugs on the side (a soldering gun may be necessary for this step). Break loose the epoxy bead between the coil and the cover. Remove the nuts and washers and pull the board up on its leads to gain access to the bottom. Cut the epoxy bond between C503 and L501. If C503 is broken it will be necessary to replace it (Heath 21-192 $1.35).

Unsolder and remove L501, C513, C509, and VD502. Replace each component with the parts supplied in the Heath package. Reglue L501 to the board, C503 to the side of L501, and C513 and C509 to each other. Now, replace the board on its bolts and turn on the power. Set the frequency switches to 146.000 MHz and run the slug of the new coil up and down, checking that the synthesizer lock light on the front panel goes out at some point, indicating the VCO is operational. Install the cover, gluing the coil to the top and soldering the lugs on the side. These changes increase the tuning range of the VCO by increasing the oscillator L/C ratio and making the varicap diode a larger part of the total capacitance. A full alignment of the VCO will be completed later.

Synthesizer Board. First, remove the black wire attached to point C on the power amp board to allow room for the synthesizer board to swing up. Remove the thirteen wires that attach the thumbwheel switches to the front end of the board and the four nuts and washers. Now, swing the board up on its remaining leads and remove R427 and R428, changing the values so that R427 becomes 5.6 kilohms and R428 10 kilohms. Page 61 of the manual can be used to locate these two resistors.

Replace the board on its bolts, attach the black wire to point C on the power amp board, and refer to pictorials 4-10 and 4-11 for color code and wire placement when attaching the thirteen wires to the frequency switches. Again turn on the power and check to see that the synthesizer lock light goes out.



Location of C235 which, in conjunction with its lead length, forms a filter on the output of the transmitter.

table 1. Parts list of components needed to update the HW-2036 to an HW-2036A. Unless otherwise noted, all resistors are 114 watt, 5 per cent tolerance.

Designation	value/description	Heath number	Heath price
update parts			
C214, C215, C216	4.7-pF NPO disc ceramic	21-168	$0.29
C442	20-pF NPO disc ceramic	21-51	0.25
Q204	EL-131 (replaces MPF-105)	417-241	2.10
R141	180 ohms, ½ watt	1-112	0.22
R206, R209	10 kilohms 10 per cent	1-9-12	0.24
R419	10 kilohms	6-103-12	0.25
R445	2.2 kilohms	1-4-12	0.24
C403*	100-pF disc ceramic	21-75	0.25
conversion parts			
C503	0.1-μF monolithic ceramic	21-192	$1.35
C509*	22-pF 5 per cent dipped mica	20-99	0.40
C513*	125-pF 5 per cent dipped mica	20-117	0.50
C201*	5-pF NPO dipped ceramic	21-78	0.25
C235*	12-pF 5 per cent dipped mica	21-130	0.45
L501*	0.25 μH	40-1855	0.95
VD502*	Motorola MV2110	56-640	1.65
R103*	100 ohms	6-101-12	0.25
R233*	15 kilohms	6-153-12	0.25
R427*	5.6 kilohms	6-562-12	0.25
R428*	10 kilohms	6-103-12	0.25

"Note These parts can be obtained as a kit with instructions (Heath part number 830-29) for $3.95

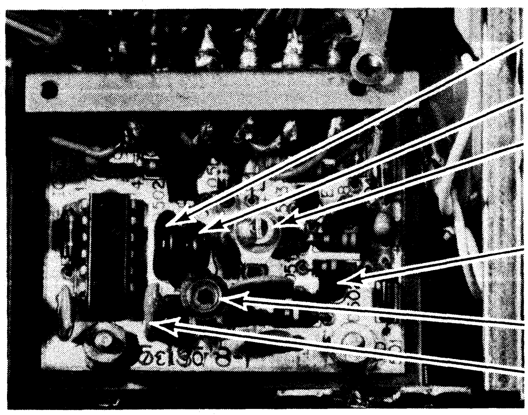
Receiver Board. Remove the four nuts and washers, remove the coax from points A and B, and tilt the board up on its remaining leads. Referring to pages 45-52 in the manual, change the value of R233 to 15 kilohms. This increases the drive to the receiver doubler.

Next, remove C201 and the shield alongside of L201 and L202. Remove 3 mm (1/8 inch) of the paper insulation from the vertical edge nearest L213. Cut a piece of 0.2-mm (0.01-inch) brass shim stock or copper flashing to 22 x 16 mm (718 x 518 inch), bending 3 mm (1/8 inch) of the long side at right angles to form a tab. Clean the shim stock and shield with steel wool and solder the tab to the existing shield where the insulation was removed. Alligator clips will hold the pieces together during the soldering.

Remove L202 and install the modified shield on its original pins. Solder the shield pin nearest L204 to hold it in place and run a bead of solder along the shield extension, joining it to the board foil behind C203 and L202. Also, run a bead of solder along the back of the shield next to C201. Replace L202, noting the polarity, and change C201 to a 5-pF NPO disc ceramic; solder both components in place, as well as the second shield pin.

Bolt the board down again and resolder the ground lead from the power amplifier board to the corner of the receiver board. Connect the power, turn on the transceiver, and tune in a local repeater to see that everything works properly. Full alignment will be done last.

Transmitter Board. Remove the phono plugs from J101 and J102, and remove the five nuts and washers holding down the board. Disconnect the molex connector to the power amplifier hoard and also the



This photograph shows the locations of components that were changed on the VCO board.

four large power connections labeled G, P, R, and V. Pull off the three-pin connector on regulator IC1 and unsolder the lead of the 0.001-μF capacitor from ground lug DA (refer to page 106). Carefully tilt up the board on its remaining leads.

Referring to page 74, remove R103 and replace it with a 100-ohm resistor. Remove the connectors on pins B, L, and N and solder the base of these pins to the board foil. These pins are the shield connections for interconnecting coax leads and should be rigidly grounded to both sides of the circuit-board ground plane!

Pull the red wire from pin S. Clip off and discard the 0.001-μF capacitor from the end and remove the two ferrite beads. Strip 3 mm (118 inch) of insulation from the end of the red wire and add approximately 7.5 mm (3 inches) of wire of a similar size, covering the soldered splice with heat shrink tubing or spaghetti. Now strip 16 mm (5/8 inch) of insulation from the extended wire and slip on the two ferrite beads just removed. There should be about 3 mm (1/8 inch) of wire sticking out. Solder this end to pin OUT, alongside the yellow wire from regulator IC1. Next, take a 2.5-cm (1-inch) piece of solid wire, bend a 3-mm (118-inch) tab on one end, and solder this tab to the circuit board foil next to pin N.

Install the board on its bolts again. Be sure to position the long coaxial cable along the top of the front of the board, as shown on page 106 of the manual. Replace all board connectors, phone plugs, and the connector on IC1. Don't forget the shield connector to pin B, hidden under the wiring harness. Now, solder the other end of the 2.5-cm (I-inch) wire from the board foil near pin N to solder lug DA on the rear apron. Make this lead as short as possible and trim off the excess. Reconnect the ground wires from the receiver and transmitter boards to the power amplifier board.

Install two ground lugs at stud X on the power amplifier board. Solder one lug directly to the chassis. Remove the coax shield from point B on the power amplifier board and solder it to the second lug on stud X.

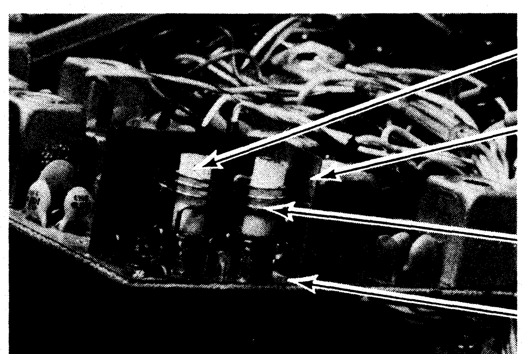
alignment

VCO Board. Set C511 on the VCO board to mid capacity. Attach the dc probe of a VTVM to TP401 on the synthesizer board; set the frequency switches to 146.000 MHz; and set the mode switch on the front panel to SIMPLEX. Turn the slug of coil L501 up until it shows at the top of the coil. Key the transmitter with the dummy load attached and rotate the slug down until the VTVM reads 2.2 volts. The SYNTHESIZER LOCK light should be out at this point.

With the transceiver *not* keyed, adjust C501 until the VTVM again reads 2.2 volts. This completes the VCO adjustments.

Transmitter Board. Since the transmitter strip is already aligned to a center frequency of 147.000 MHz, an abbreviated procedure is used to center the transmitter at 146.000 MHz, providing full output over the entire frequency range. First, install the dummy load. Set the thumbwheel switches to 146.000 MHz and connect the rf probe to TP101. The kit-supplied probe is fine and can be used with the dc scale of a standard VTVM for better sensitivity. Adjust L101 and L102 for maximum output with the transmitter keyed. Several adjustments may be necessary.

Move the rf probe to TP 102, key the transmitter,



The addition to the shield can be seen in this photograph of the receiver board. Part of the paper insulation was removed to allow the addition to be soldered to the original shield. When it's installed, the bottom of the new piece is soldered to the circuit board.

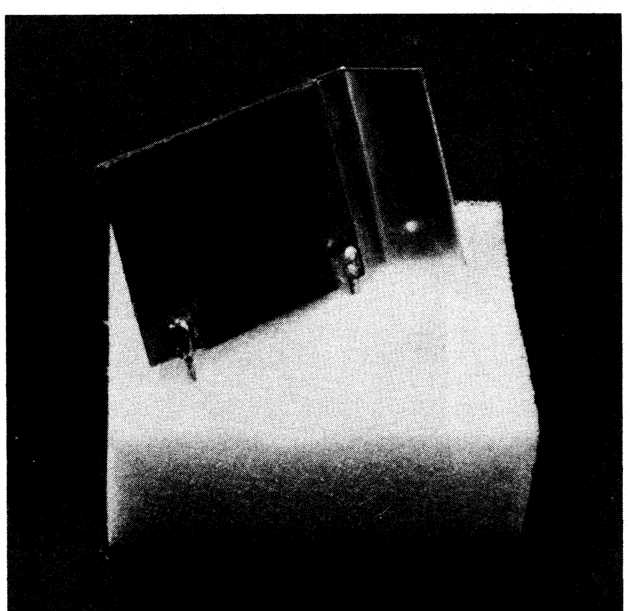
and adjust L103 and L104 for maximum. Remove the probe, and, with the transmitter keyed, adjust L105, L106, and L107 for maximum output as indicated on the front panel's relative power meter. The peak on L106 will be very broad.

Now, set the frequency switches to 144.500 MHz and adjust L101 for maximum output at this lower band edge. Then, with the frequency set to 147.500 MHz, adjust L102 for maximum. Repeat these two steps, which are designed to stagger the tuning over the full band. Finally, set the frequency to 144.000 MHz, key the transmitter, and check the relative power meter to be sure the transmitter has full output at the extreme band edge. Repeat this step at 147.990 MHz. There may be a slight difference at each end, but not much.

Power Amplifier Board. Set the frequency switches to 145.000 MHz, and, with the transmitter keyed into the dummy ioad, adjust the power amplifier trimmers in this order: A, B, C, D, E, D, C, A, B. Do not adjust trimmer capacitor A any further clockwise than is necessary for maximum output. This single adjustment series should provide nearly equal output across the entire band. Remember to replace the heatsinks!

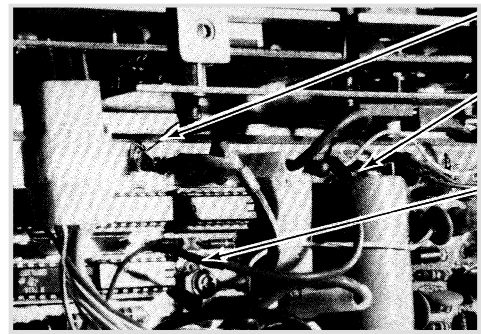
Receiver Board. Set the frequency to 146.000 MHz, connect the rf probe to gate 2 of 0202 on the receiver board, and adjust L402 and L403 on the synthesizer board and L212 and L213 on the receiver board for maximum level. Perform these adjustments in the order listed and repeat at least once. This provides maximum local-oscillator injection to the first mixer.

Next, remove the antenna coax from points A and B on the receiver board and connect the kit-supplied



Close-up of the receiver shield with the addition soldered to the right edge.

jumper from points A and B to points TP106 and TP107 on the transmitter board. Now, adjust L201, L202, L203, and L204 for maximum S-meter reading with the frequency set to 146.000 MHz. It may be necessary to adjust the frequency up or down slightly to pick up the harmonics of the reference oscillator being used as the alignment signal source. Also, since there is a high signal level, C147 on the transmitter board may be adjusted to maintain the meter



Changes to the transmitter board can be seen in this photograph. Note that the wire extension is now soldered to pin OUT instead of pin S.

reading at half scale or less. If this does not provide enough signal reduction, slightly detune i-f coil L207 on the receiver board. Once L201 through L204 are adjusted properly, L207 can be repeaked by removing the jumper from point A and holding it a small distance from the pin. This will provide enough signal to peak L207 without overloading the i-f strip.

At this point, if you replaced the parts in the i-f strip, refer now to the manual instructions for aligning the i-f stages. Few will need this however, and, if not disturbed, the rest of the i-f will not need attention. This completes the conversion.

remarks and notes

We found that the birdies on 146.000 MHz and 146.520 MHz were slightly decreased in signal strength but not eliminated. The reference oscillator and offset oscillator should not need attention if care was taken not to disturb their trimmers during the modification.

Reducing power consumption. Several ICs can successfully be replaced with low-power Schottky versions (LS series). IC104, IC105, and IC106 (7490s) may be replaced with 74LS90s. IC107 (7492) may be changed to a 74LS92. IC405 (7400) may be changed to a 74LS00, and IC404 (7473) may be changed to 74LS73. IC103 (7400) may *not* be replaced! These changes will reduce current drain about 20 per cent. It may be possible to change these devices to the CMOS series for further power reduction, but we haven't yet tried that substitution.

ham radio

the dasher

Update your Vibroplex bug by using the Dasher — a simple circuit that provides a continuous series of dashes

Back about **1946,** during my days as a commercial CW operator, I used to spend an eight-hour shift sending traffic in strings of 40 or more messages. Many of them were weather messages which had groups of five digits and many, many groups per message. I was getting a glass arm doing this at 40 words per minute or better every working day.

I saw a circuit in an Amateur magazine which used two tubes in a multi-vibrator circuit that could be mounted on top of a regular Vibroplex bug and would automatically make the dashes while you still get the dots from the mechanical side of your bug. I built and used one for years, saving my arm a thousand times over. It had the advantage that you could make the dashes much longer than the traditional 3 to 1 ratio. Therefore, you could retain your original sending style, yet send perfect code without fatigue.

Since then I haven't been too active — until recently when I got back into both MARS and CW work. That's when I remembered that silly thing I once built into a cake pan and mounted on top of my old Vibroplex.

Today, however, with all solid-state devices and small parts, one can be built in a small box, using a 9-volt battery for power. I used the 555 IC as a timer and a small 6-volt, dc-sensitive relay.

circuit description

Fig. 1 shows the circuit I used for the *Dasher.* In some circuits, pin 7 on the 555 was keyed. I tried this configuration but it caused a slight "hangover" on the last dash of a train, making it very difficult to handle. I decided to try keying the 9-volt supply to the entire circuit, but this caused the first dash of a series to be stretched. Then, I tried leaving the 9 Vdc connected to all parts of the circuit and keying the lead to pin 8 This got rid of the problems, resulting in perfect dashes.

The two pots permit adjustment of both speed and weight. The only change needed on a Vibroplex bug is to disconnect the lead under the base from the dash contact post. Then, use a two-wire, shielded mike cable to provide three connections from the bug to the input of the *Dasher.* The shield forms one lead from the frame post of the bug, and the other two leads connect to the dot and dash contacts. Inside the *Dasher,* the lead from the dot contact is connected to the output lead, with the dash contact used to control the 555.

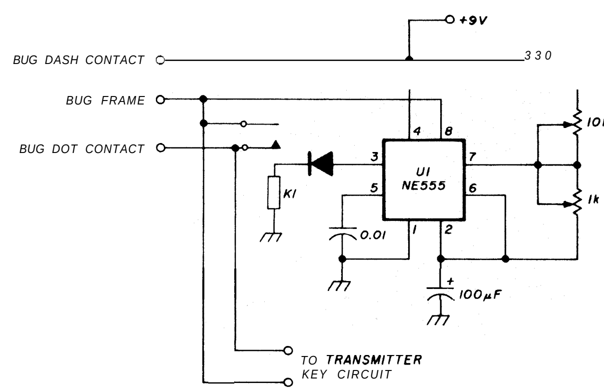


fig. 1. Circuit diagram of the *Dasher.* The relay is a small 6-volt unit available from Radio Shack. The diode can be any small signal switching diode.

Two things I might add for information. I used one of the rechargeable, 9-volt, ni-cad batteries and brought leads from the battery to screw terminals on the side of the box. This way I can connect a charger without opening the box. Also, a note of caution. Do not ground any of the 9-volt points to a ground that is common to the relay output or the bug common lead. When I tried the *Dasher* with my Collins KWM2A barefoot, no problems were encountered. However, when I ran the linear with a full kilowatt, rf was induced into the leads, causing erratic dash length. Putting ferrite beads on the input and output leads cured the problem. So have fun and don't end up with a glass arm; use the *Dasher.*

ham radio

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formation of passive lumped constant 90-degree phase-difference networks

Construction details for designing equal amplitude, 90-degree, phase-difference networks

A variety of **networks,** both active and passive, have been used over the years for splitting a signal into two equal amplitude signals while maintaining a quadrature phase relationship between the two components. Roger Harrison's article[1] presented a number of such circuits and rekindled my interest in these devices. I first became interested in these circuits while designing image outphasing systems for sweeping microwave receivers about 20 years ago.[2,3] At UHF and microwave frequencies, 90-degree phasing networks are readily achieved with coaxial or stripline hybrids which can be designed for bandwidths of several octaves. These distributed element hybrids become large and difficult to build for the lower frequencies. Sometimes, it is also desirable to have circuits capable of greater bandwidth.

I have prepared a table (see **table 1**) of prototype values from which passive, lumped-constant, first-order lattice networks (fig. **1**) may be designed. This table is an adaptation of early work by Darlington[2] and later work by Bedrosian. Darlington's work is basic and requires background in network theory plus a comprehensive table of elliptic functions to arrive at a practical design. Bedrosian's[3] paper included a number of tables of prototype values which can be used to calculate inductances, but you are stuck with the arbitrary bandwidths he chose unless you have access to a table of elliptic functions. Even then, there are several calculations to be done before arriving at a design. I have chosen a number of bandwidths which I believe would be most useful for Amateur Radio users, 2 to 1 through 20 to 1.

network description

The phase-splitting networks consist of two separate lattice networks, a P-network and a Q-network. For a given bandwidth, the amount of phase ripple (deviation from 90-degree phase difference) depends upon the number of lattice sections, the more networks the smaller the deviations. N is the total number of lattice sections in both branches.

Although the first-order networks are simple, they may not be the most practical. This sometimes becomes apparent when you try to produce a pair of networks operating over a relatively broad range.

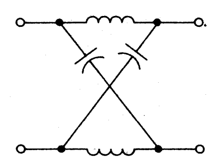

fig. 1. Diagram of a first-order lattice network.

You may discover that the networks, which are operating over the low-frequency end of the band as "all pass" networks, are also behaving at the higher frequencies like lowpass filters with a cut-off frequency within the design bandwidth. The reason for this is the self-capacitance of the inductors, which makes it difficult to place the self-resonance of the coils in the low-frequency lattices outside the design band.

Orchard[4] presented a transformation whereby two first-order networks may be combined in a single second-order network as shown in fig. 2.

description and use of the table

Once you have determined the bandwidth required (ratio of upper to lower frequency) and the phase ripple that can be tolerated, you are ready to select the appropriate prototype design from the table. Assume a 3 to 30 MHz 90-degree network is needed, requiring

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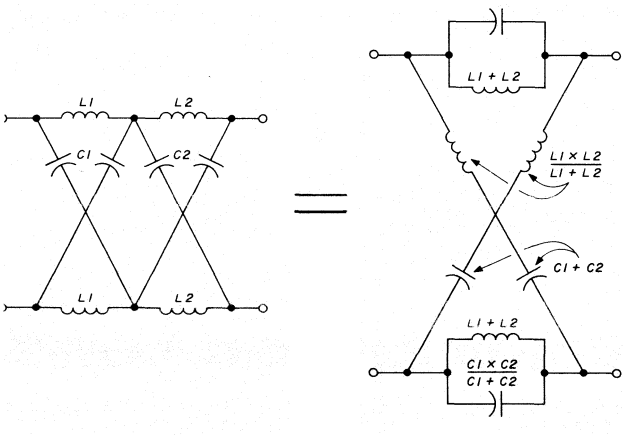

fig. 2. Transformation of a second-order to a first-order lattice network.

a phase deviation of less than two degrees. Table **1** shows that a network with four lattice sections will yield a 10 to 1 bandwidth with a phase ripple of 1.08 degrees. Table **1** also has normalized prototype values which are used to arrive at the actual designs.

denormalization and design

The lattice networks are denormalized for the geometric center frequency, in this case $\sqrt{3 \times 30} = 9.487\ MHz$. The inductances for the **P** network are calculated by:

$$L = \frac{P\,value\,(Z_o)}{2\pi f_o} \qquad \textbf{(1)}$$

where P = value from table **1**

$Z,$ = characteristic impedance of network

f_o = geometric center frequency

table 1. Prototype values used for prototype designs.

BW ratio	N	phase tolerance	P1	P2	P3	Q1	Q2	Q3
2	2	1.687	0.405616			2.4654		
3	2	4.12254	0.3933			2.543		
3	4	0.0742	0.1859	1.539		0.65	5.38	
4	4	0.178	0.179	1.564		0.64	5.6	
4	5	0.03	0.142	1	7.04	0.476	2.1	
5	4	0.308	0.173	1.586		0.63	5.787	
5	5	0.06	0.137	1	7.3	0.466	2.15	
6	4	0.455	0.167	1.61		0.622	5.97	
6	5	0.1	0.133	1	7.54	0.456	2.19	
6	6	0.02	0.11	0.731	2.75	0.363	1.37	9.1
8	4	0.77	0.1585	1.64		0.6084	6.31	
8	5	0.184	0.1254	1	7.97	0.441	2.267	
8	6	0.044	0.1038	0.72	2.87	0.348	1.388	9.63
10	4	1.08	0.151	1.675		0.597	6.61	
10	5	0.28	0.1196	1	8.36	0.428	2.33	
10	6	0.07	0.0989	0.711	2.97	0.336	1.41	10.1
15	4	1.91	0.137	1.74		0.573	7.28	
15	5	0.577	0.108	1	9.25	0.403	2.48	
20	4	2.55	0.1285	1.79		0.559	7.78	
20	5	0.829	0.101	1	9.9	0.387	2.59	

You are now ready to complete the design using the P and Q values from table 1. For the P network assume $Z_o = 50$.

$$L_1 = \frac{P_1 \bullet Z_o}{2\pi f_o} = \frac{0.151 \times 50}{2\pi \times 9.487 \times 10^6} = 0.1267\,\mu H$$

$$L_2 = \frac{1.675 \times 50}{59.6 \times 10^6} = 1.405\,\mu H$$

$$C_1 = \frac{P_1}{Z_o \bullet 2\pi f_o} = \frac{0.151}{50 \times 59.6 \times 10^6} = 50.7\,pF$$

$$C_2 = \frac{1.675}{50X\ 59.6 \times 10^6} = 562\,pF$$

For the Q network:

$$L_1 = \frac{Q_1 \bullet Z_o}{2\pi f_o} = \frac{0.597 \times 50}{2\pi \times 9.487 \times 10^6} = 0.5\,\mu H$$

$$L_2 = \frac{6.61 \times 50}{59.6 \times 10^6} = 5.54\,\mu H$$

$$C_1 = \frac{Q_1}{Z_o \times 2\pi f_o} = \frac{0.597}{50 \times 59.6 \times 10^6} = 200\,pF$$

$$C_2 = \frac{6.61}{50x\ 59.6 \times 10^6} = 2218\,pF$$

Now the complete circuit can be sketched (see **fig. 3**). An example of an $N = 5$ network for the same bandwidth is shown in **fig. 4.**

The first-order lattice is a balanced circuit whose unbalanced equivalent requires the use of a transformer. Several types of balanced-to-unbalanced transformers, which are suitable for adapting the lattice to an unbalanced circuit, are described in reference handbooks. You have the option of choosing the impedance level of the lattices to effect the desired impedance match at the common junction of the networks.

theoretical and experimental results

The values for table **1** were obtained from a BASIC computer program (PHASE90) which is capable of designing a wide range of first-order networks with bandwidths up to 100 to 1. Several designs from this program were analyzed using a BASIC (GNET) network analysis program and were found to perform as predicted.

Finally, to prove the usefulness of the theoretical design, a four-section ($N = 4$) network consisting of first-order lattices was built for the 20 to 150 MHz range. The phase deviations from 90 degrees were about 5 degrees, which is greater than the theoretical tolerance of 0.7 degree, but adequate for most requirements. This experimental data shows that the first-order lattices can be used for an 8 to 1 band-

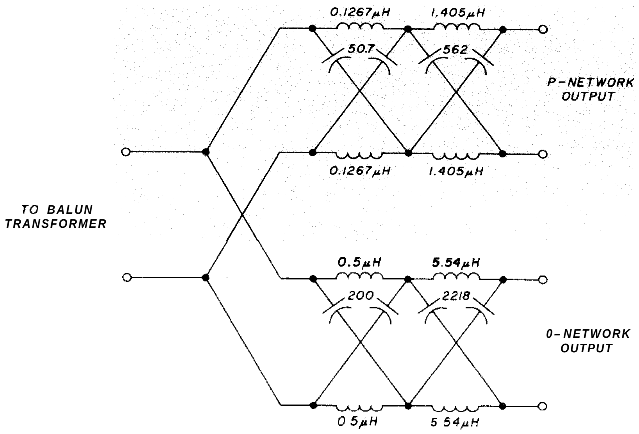

fig. 3. Component values for the 3 to 30 MHz, four-section network.

width. For greater bandwidths you may have to resort to the second-order lattice configuration.

acknowledgments

I am indebted to R. E. Booth, WB6SXV, who wrote the GNET network analysis program, and whose assistance in writing the PHASE90 program is gratefully acknowledged.

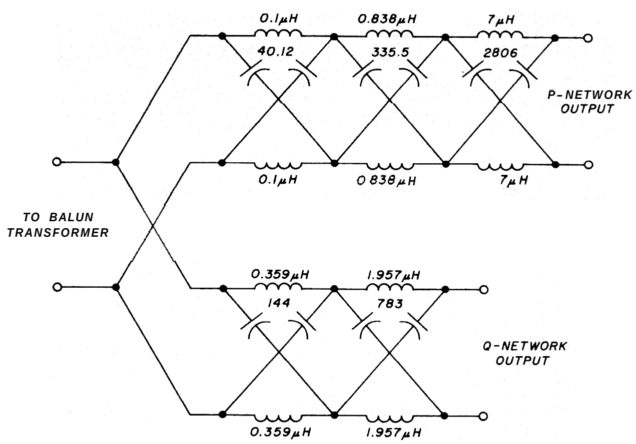

fig. 4. A five-section equivalent of the network shown in fig. 3.

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ham radio

cabinet construction techniques

A new construction technique which allows the builder to build a cabinet to fit his equipment — rather than fitting his equipment into the cabinet

All builders of electronic devices eventually come up against a continuing problem – what type of cabinet to put their creation in. The magnitude of the problem increases in direct proportion to the dimensions desired. Commercially manufactured answers to the problem are few. If a cabinet is inexpensive, it is usually too small or flimsy; if it is rugged enough it will usually be too expensive. Trying to find a cabinet with the desired size and configuration often seems next to impossible. It has been my experience that few local dealers ever carry a large percentage of a cabinet manufacturer's total line. Therefore, mail order must be resorted to, which takes time.

Many times, the builder must choose "the best of the worst" from stock on hand and then tailor construction of the device to fit the cabinet. This practice usually results in unnecessary work and design trade-offs. Another problem that occasionally occurs is the need to increase or decrease the depth of a cabinet, something not easily accomplished. All of this adds up to what I call the "cabinet frustration syndrome."

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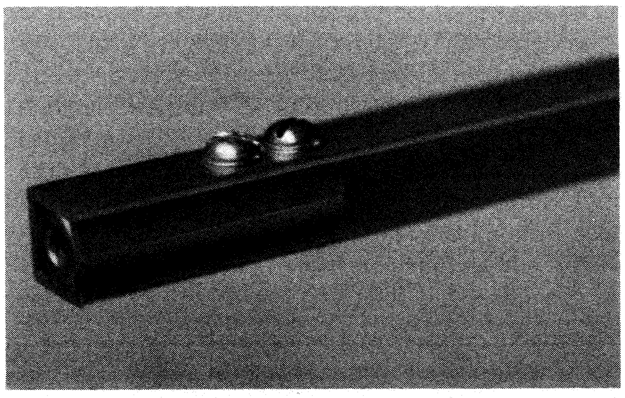
What follows is an idea that evolved from attempts to solve these problems. The result is a rugged, inexpensive cabinet which is easy to construct in any needed size. Fabrication materials are universal in that they can be used for most any size cabinet. Most of the pieces from scrapped cabinets can be used again. With this method, the depth of the cabinet can be easily extended or reduced.

fabrication

The basic cabinet is made of only two items, the end plates and the end-plate support brackets. The end plates can be the front and rear panels of your unit or they can be subpanels, using a method to be described later. The two end plates are simply cut from aluminum sheet stock to the desired front-panel dimensions. A mounting hole for each bracket is drilled in each corner. The end-plate support brackets are the key to this whole method of construction. Bolted between the end plates, they provide cabinet rigidity, while at the same time allowing construction flexibility.

The brackets are made of aluminum U-channel, cut to the desired depth. Aluminum U-channel was chosen because it is light, relatively inexpensive, and readily available at most hardware stores. My first attempt, for which I used 12.5-mm (112 inch) U-channel, was acceptable but resulted in a more rugged (and heavier) cabinet than was necessary for that application. Cabinets built of 9.5-mm (318-inch) channel proved to be best for most of my uses.

Once the U-channel size is decided upon, the next step is to locate or fabricate something to lay in the channel that will accept the end-plate mounting screws. My junk box yielded some threaded metal standoffs about 5 cm (2 inches) in length. These fit snugly in the channel. The standoffs had a 10-32



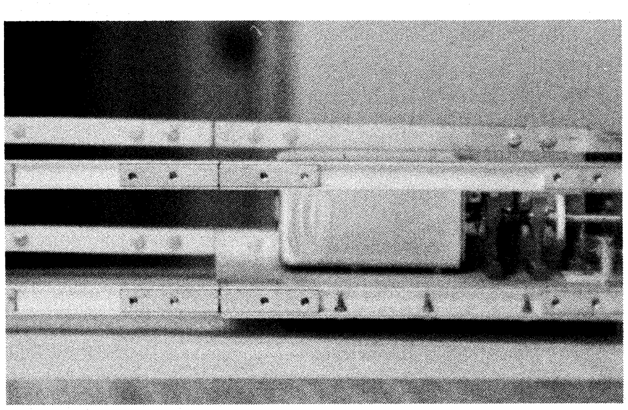
As seen in this photograph, the standoff was drilled and tapped for two additional holes, providing a means of securing the standoff in the channel stock.

(M5) threaded hole through their length. Two other mounting holes were drilled perpendicular to the length and tapped for a 10-32 (M5) thread. These holes are for bolting the standoffs to the U-channel.

If you don't have any standoffs, square aluminum stock is usually sold along with the U-channel. This can be cut to length, drilled, and tapped as necessary. Of course, the square stock can be cut to a length equal to the cabinet depth and used as the entire mounting bracket. In that case, only one hole must be drilled and tapped in each end. However, a solid bracket will have a few disadvantages when compared with the U-channel approach; it will be heavier, and it will offer less flexibility in the mounting of parts to the brackets.

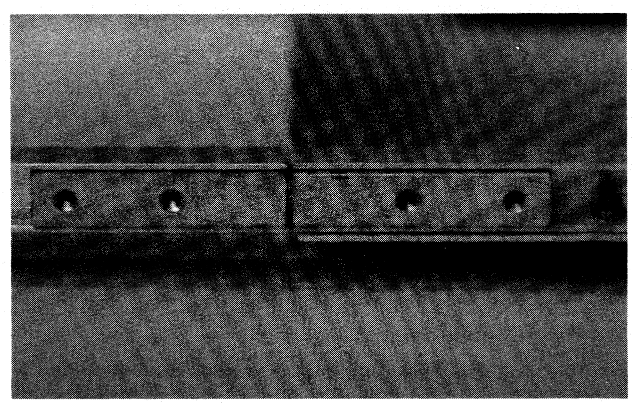
Once each bracket has a standoff mounted at each end, the four brackets are bolted to the end plates, and the basic cabinet is finished! Component mounting plates can be above or below the brackets. If a center mounting plate is needed, this can easily be accomplished with L-brackets. Incidentally, a good source of L-brackets for any project is L-shaped aluminum stock, usually sold in the same place as the other materials. Merely cut the stock to the width of L-bracket desired. The L-bracket will be much stronger than one made by bending a piece of flat stock.

Example of a chassis that has been lengthened by using additional side bracket pieces.



The basic cabinet is essentially a frame that can be covered by cutting and bending lightweight sheet aluminum. The design is such that the cover is not needed to provide the basic structural integrity and strength. This is a big problem with all but the most expensive commercial cabinets. Here again, the lightweight aluminum cover material is usually sold where all the other materials are.

Rather than making a cover by bending, cutting individual pieces of stock and attaching them to their respective sides with small screws is also a possibility. Where a heavier gauge metal will be used for the cover, it is possible to entirely eliminate the brackets.



Close-up of the connection between the two sections of a side bracket.

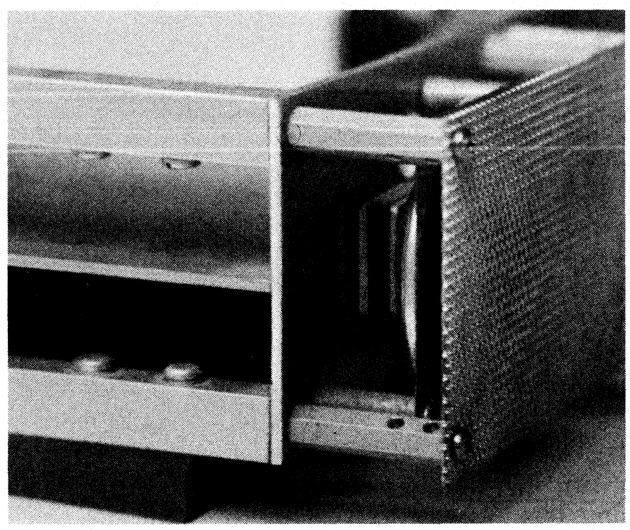
In this case, a central chassis plate is fastened to the end plates with L-brackets. Unless the cabinet is very small, the aluminum sheeting used for the end plates and chassis plate should be heavy enough to provide sufficient rigidity. Most of the aluminum that I use for chassis and end plates is 3 mm (1/8 inch) thick, although I've also used material 1.5 mm (1/16 inch) thick with success.

Aluminum sheeting is used because of its widespread availability and low cost. In the past, I have used some magnesium alloy sheet which provided excellent rigidity and light weight. The problem is, of course, that it is both expensive and difficult to come by.

As mentioned earlier, cabinets made with the U-channel technique can be easily lengthened or shortened. To shorten a cabinet, merely remove the standoffs from one end of each bracket, cut the brackets to the desired length, and remount the standoffs in the ends.

The process of lengthening a cabinet is just a little more involved. If materials are available, just make longer brackets. However, if you don't have a piece of U-channel long enough to make four new brackets you can use four small pieces and add to the length of the existing bracket. This approach will require more work, but could be the most expedient and most economical under certain circumstances.

First. cut and fashion four extension brackets for the desired added depth. Remove one of the end



This photograph illustrates a false panel speaker grille using standoffs mounted on an end plate.

plates. Insert adequate lengths of threaded screw stock (I favor 10-32 [M5]), into the ends of the brackets to be lengthened. Leave about 12.5 mm (1/2 inch) protruding. Screw the extension brackets onto the exposed screw stock. Now comes the only tricky part: when the two brackets are snugged up tight against each other, they should be aligned. If they do not align, insert a piece of metal shim stock or a thin washer between the two pieces and try again until alignment is achieved. Mount the other end plate on the extended brackets. If desired, one end plate can be used as a subpanel for false front panels or speaker grilles. This is accomplished with metal standoffs.

The only item which might prove difficult to procure is the aluminum sheet used for the chassis and end plates. Sheet metal dealers will usually sell remnants at scrap value. Scrap metal dealers sometimes have acceptable pieces. Your best bet is to get out and browse around to see what is available.

It's my hope that this article has provided you with a basic method (plus some variations) for relieving "cabinet frustration syndrome."

ham radio

GITAL TECHNIQUES

digital circuits — propagation delay and flip-flops

So far, the basic gate has been examined under dc or static conditions. This part will discuss dynamic conditions of switching, propagation delay, and the most simple flip-flop, the latch.

All digital devices have delay from internal charge storage. Input and output may switch rapidly, but the delay between input transition and output transition is usually longer than the transition itself. Fig. **1** is an example of dynamic conditions in a simple inverter.

The term t_p or t_d is associated with propagation delay and will have a number of different subscript letters to distinguish different types of delay. A table in the front of the Texas Instruments *TTL Data Book* lists most of them. The term *propagation* is used, since the input-to-output path within a device may go through several stages.

The inverter shows only two delays. t_{pLH} is the input transition-to-low delay to output transition-to-high state; t_{pHL} is the opposite. They may occur at different times. Measurement is made at half amplitude between maximum low-level voltage and minimum high-level voltage.

Output-state transition time is also listed, almost always at a specified load of maximum capacity. Load capacity affects output transition delay strongly but has only a slight effect on propagation delay.

Multi-function devices have a number of different delays. Data sheets must be carefully studied to understand the effect of delays on the overall circuit. Lack of understanding of delays can waste a lot of debugging time (most times are too short to measure on inexpensive scopes) and also can cause a circuit to fail entirely. Delay time can't be underestimated!

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the RS latch flip-flop

Fig. **2** shows the block symbol, waveforms, and an equivalent circuit constructed of NAND gates. The *RS* term comes from reset/set. A flip-flop is a bistable circuit (stable in two conditions) and is either set or reset by an external input. Because it's bistable, it's said to be latched in one state or the other.

The circuit of fig. **2** has active low set and reset inputs, indicated by inversion bubbles on the symbols. Outputs are not shown with bubbles, since the flip-flop is a bistable device.

Flip-flop terminology sometimes refers to the Q output as true, $\overline{Q}$ as false, or NOT. Output-to-input cross-connection would appear to create an oscillator. This would be true with linear devices, but it's the key to holding a bistable condition with digital devices.

Assume initial conditions of both inputs and Q high, $\overline{Q}$ low. NAND gate G1 is held high by the low from **G2. G2** is held low since both inputs are high (the *NAND RULE).* The initial state is stable. Either output could have been high at power turn-on, depending on which gate was the fastest.

The first $\overline{S}$ low input does nothing; G1 is already held high. The first $\overline{R}$ low will force **G2** high. G1 then

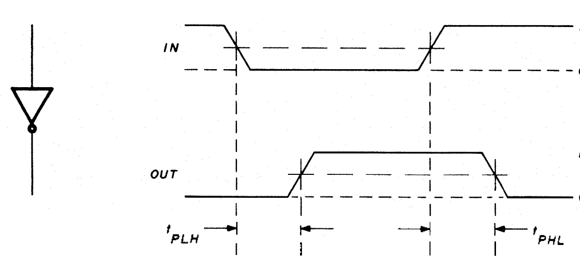


fig. 1. Propagation delay of an inverter.

goes low, since all inputs are high and the latch has flipped to a reset condition (Q low). This is the other stable condition.

Note the exaggerated time delays and sequence of changes. Each input must hold at a low state long enough for both gates to change state. Flip-flop spec sheets will give this parameter of time as t_{hold}.

A following $\overline{R}$ low will do nothing: the latch is already reset. A second $\overline{S}$ low will force G1 high. G2 will go low since all inputs are high; the latch has flopped into a set condition.

What happens if both inputs are low? Both outputs will be forced high as long as that condition persists. The unknown condition occurs when both inputs return high. Either stable condition could occur and, as in power-on, it depends on which gate is the fastest. This is the indeterminate state and should be avoided.

a switch debouncer

All switches and relays, except for mercury wetted types, have contact bounce. Contact bounce is the opening and closing of contacts because of mechanical vibration. It lasts between 0.2 and 50 milliseconds, depending on construction, and can raise havoc with certain digital circuit inputs; a cure is found in **fig.** 3.

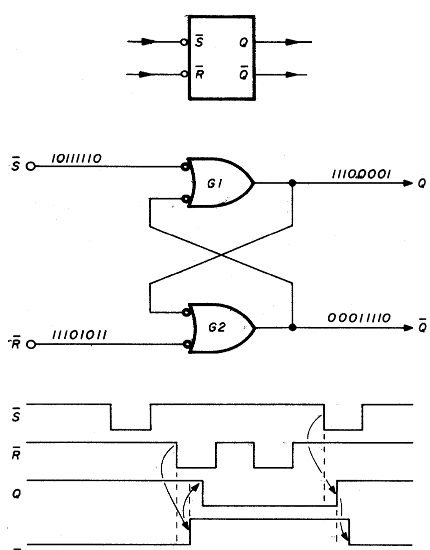


fig. 2. Simple RS flip-flop latch symbol, circuit, and waveform.

The RS latch is an interface between the switch and controlled circuit. The grounded switch arm provides the active low latch input signal. Pull-up resistors provide a high state for open contacts. The 4.7k pull-up is for TTL; CMOS values can range from 56k to 270k.

Contact bounce still exists, but the controlled circuit does not see it. The first contact closure will flip the latch. Subsequent bounce will not affect the latch as long as the switch is break-before-make. Compare this circuit with that of **fig. 2** for full understanding.

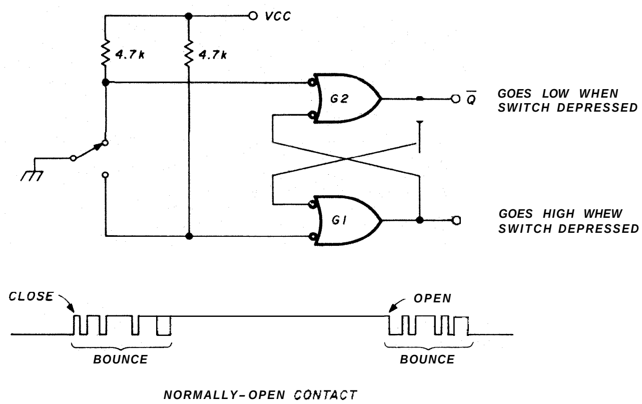


fig. 3. Switch debouncer.

Fig. 3 could be made with CMOS NOR gates, switch arm to V_{cc} and pull-down resistors. Try this on scratch paper. A hint: Use the gate truth tables in **Part 1.1** The same thing won't work well with TTL. Why?

The difference lies in the high-TTL logic-0 current. Pull-down resistors would have to be 180 ohms or less for 0.4 volt maximum at 1.6 mA. It's fine if you can afford to waste 28 mA when the pull-down is connected to V. A 4.7k pull-up resistor wastes only 1.1 mA.

Can a single-throw switch be substituted? The thought comes to mind that an inverter could be used for the other latch input. Sorry, this won't work; the same bounce is repeated in the latch. A single-throw switch can be applied only to static switching inputs or those unaffected by bounce.

clocked flip-flops

The term *clock* refers to a timing signal and came from early computer work, when the clock synchronized all functions. Applied to flip-flops, the clock input is really a trigger to initiate a change of state.

Fig. 4 shows the most common clocked flip-flops, the JK and D, in symbolic form with truth tables. Two things should be noted here: These truth tables are time-dependent, and the clock input may or may not have an inversion bubble. The clock edge is the trigger. A positive-going (low-to-high) edge is shown direct as in the D flip-flop. A negative-going (high-to-low) edge is shown with the bubble.

J, K, or D inputs are control or data inputs. Each truth table shows the effect of these inputs on the Q output after receipt of a clock edge. If both J and K are held low, Q of the JK does not change. This is indicated by Q Holding one low and one high will force the 0 and 1 condition. With both high, Q will flip to the state opposite what it was before the clock edge arrived. This is what $\overline{Q_n}$ means.

The D flip-flop is easier to understand. Its Q state takes the D input state after receiving a clock edge.

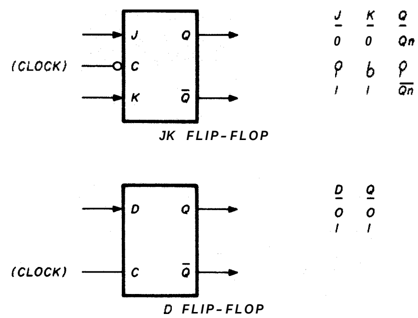


fig. 4. Symbols and truth tables for the basic clocked flip-flops.

Both flip-flops will change Q based on the control input state before a clock edge occurs. Control inputs do not directly affect output.*

Clocked flip-flops require an additional time parameter: setup. J, K, and D states must exist a specific minimum number of nanoseconds before a desired clock edge occurs. Too short a setup time will result in skipping a clock period for a Q state change.

An output state change is sometimes called toggling. A JK with both inputs high will toggle on each clock. Output period will be twice clock period, and the flip-flop becomes a divide-by-two device. Connecting $\overline{Q}$ to D in a D flip-flop will accomplish the same function.

Both flip-flops may have direct set and reset inputs. These override any clock input and may occur at any time. As such they're sometimes called asynchronous. Terms vary and ***set*** is sometimes ***preset; reset*** is sometimes called ***clear.***

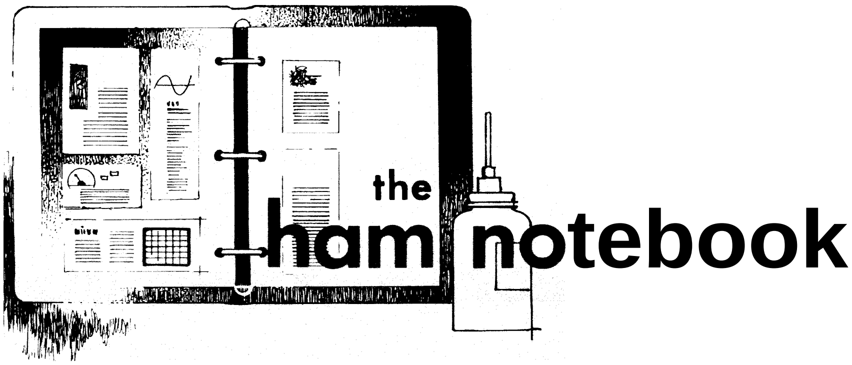
The next part of this series will look inside clocked flip-flops, present NAND gate circuits, and discuss timing.

"Some old JK types did flip when both J and K changed together with the same state.

reference

1. Leonard H. Anderson, "Digital Techniques, Basic Rules and Gates," *ham radio,* January, 1979, page 76.

ham radio



integrated circuit tone generator

The MK5085/86 as shown in the February, 1977, issue of *ham radio,* can be used to make a very simple, inexpensive, and reliable tone encoder. In fact, with the exception of the addition of the transmitter keying relay and driver transistor, the circuits shown have been taken directly from the Mostek data sheet.

A couple of additional comments are in order, however. First, it is usually inadvisable to feed the encoder's output into the microphone line as indicated in the article. The pre-emphasis in the audio amplification circuits will distort the level differential between the high and low group tones, which should be on the order of 3 dB. Thus, the levels actually transmitted, *i.e.,* presented to the decoders at the receiver, will not be correct and decoding difficulty may result. In order to use the microphone input, some kind of rolloff circuitry is necessary, and this adds to the complexity of the overall encoder.

A second problem with using the microphone input is that of background noise; the tones receive as much amplification as the ambient noise. Since the output of the MK5085/86 is on the order of 700 millivolts for the low group and 1100 millivolts for the high group, these high-level signals do not need high amplification.

The chip was primarily designed for telephone application, and to feed a load of approximately 600 ohms. In the telephone application, a convenient output load is the telephone line itself. However, the line is also the power source. With this background, it is easy to see that the chip really prefers to deliver its output right into the power line. Using an independent power source, this characteristic can be ignored. Powered from the radio, however, the effect will be to modulate the entire primary power. In short, the tones will wind up in all kinds of places where you don't want them.

A simple solution is shown in **fig. 1.** A large capacitor will swamp the tones, provided the power source impedance is low enough; take power where the 12-volt line enters the rig, and not directly from the rig's power supply. The temptation to do the latter may be great in view of the less-than-10 volts allowed by the MK5085/86. The diode in **fig. 1** is not really necessary if power is taken directly from the car battery, but it should be noted that the MK5085/86 is (mostly) CMOS and can generate appreciable spikes.

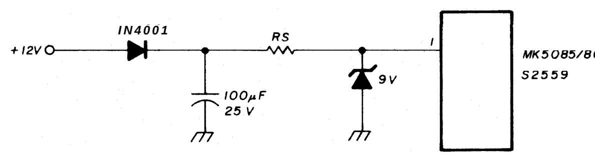


fig. 1. Diagram of the method used to eliminate tones from the power line.

I prefer the AMI S2559 over the Mostek 5085186 primarily because of its ability to stand power of up to 15 volts, eliminating the need for the zener and dropping resistor shown in **fig. 1.** The S2559 is pin-compatible with the Mostek chips and can direct-

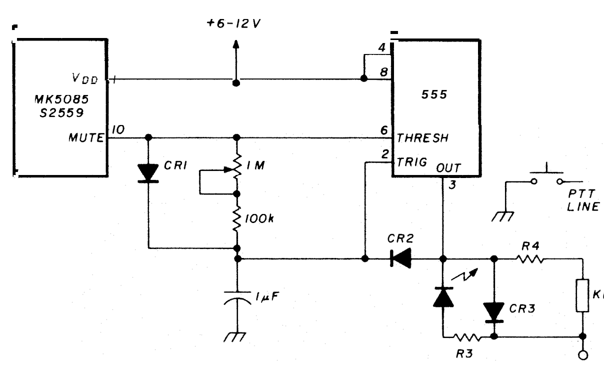


fig. 2. Schematic of a more satisfactory method for generating a variable dropout time. CR1 through CR3 are all 1N914 diodes. R3 is an appropriate current limiting resistor for the LED used. The maximum operating current permissible for the relay is 150 mA. Varying the 1-megohm resistor will change the dropout time from approximately 0.1 to 1 second.

ly replace them. It is also driven from the standard 3.58-MHz color-burst crystal. Aside from the wider power tolerance (down to 3 volts or so), there is yet another advantage: The S2559 can be used with either a Class A or 2-of-8 (or 2-of-7) DTMF keyboard. The choice of which Mostek chip to use is dictated by the keyboard key configuration.

A final observation concerns the transmitter keying arrangement shown in the article. It will work, and is simple and inexpensive, but it does have a potential drawback. The two-second time delay is too long (easily changed by choice of capacitance), but it is not very constant or reliable. This could cause problems when the circuit is used with an autodialer or repertory dialer.

In my opinion, a better (and only slightly more expensive) arrangement can be made using the workhorse 555 timer. The circuit, as shown in **fig.** 2, gives an easily adjustable hang time of from 0.1 second, to about 1.1 seconds, by adjusting the 1 megohm pot in series with the 100k fixed resistor. Once set, the hang time is extremely consistent.

The circuit works as follows. With no key pressed, the MUTE output at pin 10 of the MK5085/86/AMIS2559 is low, holding the threshold voltage at pin 6 of the 555 low. The trigger voltage at pin 2 of the 555 is also low, the capacitor having discharged through the timing resistors. The 555 output, pin 3, is high (because trigger went low), charging the capacitor through CR2. This does not change the timer state, even though trigger is now high, because threshold stays low. The 555 is stable in this state: threshold low, trigger high, output high, and capacitor charged. K1 is not energized.

When a key is pressed, the MUTE output from the encoder goes high, tripping the threshold of the 555 and driving the output low. This energizes K1. Full charge is maintained on the capacitor through CR1. When the key is released, the capacitor begins to dischage through the timing resistors. If no key is pressed before the capacitor discharges to 1/3 Vcc (approximately one time constant) the 555 triggers, returning the 555 to its previously stable state. The capacitor recharges virtually instantaneously through CR2. If another key is pressed before the capacitor discharges to 1/3 of Vcc, it immediately recharges to full voltage through CR1 and remains there until the key is released.

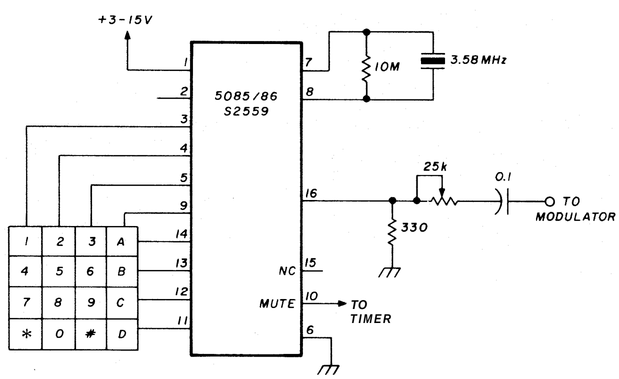


fig. 3. Circuit for a complete tone decoder. The AMI S2559 is a pin-for-pin replacement, featuring the capability of a higher input voltage, 15 vs 10 volts. Pin 15 is grounded for dual tones only. When left open, a single tone is generated when two keys in the same row or column are generated. For all practical purposes, the 330-ohm load resistor on pin 16 can remain the same value when Vdd is changed, though the optimum value is 270 ohms for a 12-volt supply and 628 ohms for a 5-volt supply. If an audible side tone is desired, a 4- to 8-ohm speaker can be inserted in the supply line.

CR3 is transient suppression for K1. I found it useful to use a relatively low-voltage relay with series resistance because (**a**) I got a real buy on 5-volt reed relays and (**b**) by adjusting the resistor the timing circuit will operate down to the lower functional limit of the 555 and the S2559. The LED and its associated series resistor are optional.

If this timing circuit is used with the MK5085/86, an 82k resistor should be used in series with CR1. The MUTE output current is capable of providing 10 μA. (The S2559 will provide a couple of mils.) However, since the capacitor never discharges below 1.3 Vcc and because the majority of charging current comes from the 555, the 82k resistor suggested isn't really required. However, it is cheap insurance to protect an $8.20 IC.

It is preferable to use a pot in series with the encoder output (see **fig. 3**). This provides both level adjustment and isolation concurrently. Taking the output from a pot in the output emitter line (as shown in the article) is not recommended, as it more greatly affects the impedance the chip is working into. I have found a 25K pot to be about the right value, using it in place of the 20k isolation resistor previously mentioned.

One final thought: A useful addition would be to short out the microphone (or open the microphone audio line) when the encoder is active, eliminating background noise. This can be done in a variety of ways, the most obvious being to use another pole on the keying relay connected to the 555 timer.

Concerning the relay, the 10 mA operating current does *not* include the relay current. Care must be taken if a 9-volt battery is used; they're not made to drive current-hungry devices like relays. Also, the 10 mA figure is with Vdd equal to 6 volts, but the circuit shows a 9-volt source. Under these conditions the current will be around 20-25 mA.

Frank Bates, W6IPB

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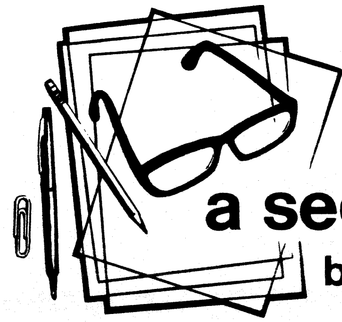


a second look

by Jim Fisk

It's spring and hamfest time again. It seems that no matter where you look, you'll see an announcement for yet another convention, radio auction, hamfest, or fm talk-in, and the Dayton Hamvention in Dayton, Ohio, which is billed as the ***original*** hamvention, is one of the biggest of the year. Scheduled for the last weekend of April, the Dayton Hamvention draws upon the large amateur population of the Midwest and has provided the model for many successful Amateur Radio conventions around the country. Several years ago, in fact, a contingent of Japanese amateurs came over to take a look at the Hamvention and its management, then returned to Japan where they staged a very popular and successful ham convention in the shadow of Mount Fujiyama.

Growing by leaps and bounds in recent years, more than 19,000 hams were in attendance last year, and even more are expected in 1979! But even with this huge influx of Amateur Radio enthusiasts, everything runs smoothly, and this year, as always, the Dayton Hamvention Committee has gone all out to ensure a lively, interesting weekend for all. Bright and early Friday, April 27th, the Amateur Radio manufacturers and distributors will start setting up their exhibits. At high noon the exhibition doors will be opened to the public — the exhibit hall will remain open until 8:30 PM, so you'll have plenty of time to browse through the many commercial exhibits. And, if this year's exhibit area is anything like those in the past, you can bet some major new Amateur Radio products will be unveiled by the manufacturers.

When the lights go out in the exhibit hall on Friday evening, they'll come on in other parts of town with banquets or gatherings scheduled for the QCWA and Old Old Timers as well as groups interested in DX, slow-scan TV, and fm repeaters (the well-known FM Bash). By early Saturday morning things will really be humming around Hara Arena, weekend home of the Hamvention, as vendors and traders from miles around start setting up shop for the famous Dayton Flea Market. Like other parts of the Hamvention, the flea market grows every year, and now takes up more than 10 acres of real estate behind the main convention hall. Whether you're interested in good used ham gear, replacement parts, vacuum tubes, new components, or antique radios, chances are good that the item you want will be for sale — your problem is to locate it among the thousands of items on display.

At 9:00 AM Saturday morning the first of many forums will be kicked off with sessions devoted to antennas, microwave techniques, space communications, microprocessors, and code proficiency. Between 1130 and 1300 there will be meetings for the DXers and Oscar enthusiasts. After lunch the emphasis will be on contests, QRP rigs, ATV, fm repeaters, vhf/uhf, and moonbounce. The traditional Saturday-night cocktail hour and banquet begins at 7:00 PM.

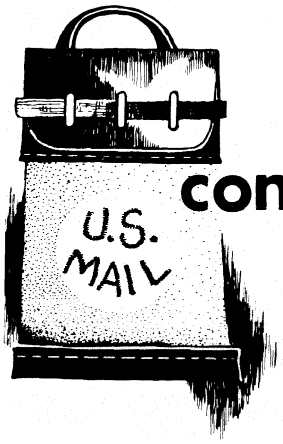
Early Sunday morning, just about sunrise, the Flea Market will open for another day's business; a few hours later the exhibition hall will open once again and the ARRL Forum will get under way. If you've never been to a major ham convention, the ARRL Forum offers you a chance to ask the ARRL's officers and directors questions about League affairs. The FCC Forum scheduled for Sunday afternoon gives you a similar opportunity to ask questions of FCC staff members from Washington. Both the ARRL and FCC Forums are among the most popular get-togethers at any convention, so be sure you arrive early to get a good seat.

In addition to the various sessions and forums, there will be technical and group meetings for ARPSC, OSSB, and MARS. Other special groups attending the Hamvention are the Buckeye Belles, Mid-Cars, Ten-Ten, Firebirds, Young Ladies Radio League, and others. If past performance and the 1979 schedule are any indication, this year's Dayton Hamvention will be another great show.

For amateurs who arrive in trailers and campers, parking will be permitted only in specially designated areas (no campers or travel trailers will be permitted to park in the Arena lot, including the flea market area). For those who stay in the downtown hotels, free bus service will be provided out to the Hamvention. A large allotment of rooms has been set aside for the Hamvention by the local hotels and motels; all room requests should be directed to the Accommodations Committee so that rooms can be alloted within the available supply. For more information, and a Hamvention brochure, write to the Dayton Hamvention, Post Office Box 44, Dayton, Ohio 45401.

If you've never been to the Dayton Hamvention, but have considered it, this is the year to go. If you've been before, you already know what I'm talking about. See you there!

Jim Fisk, **W1HR**
editor-in-chief



comments

Dear HR:

I've just finished reading the article on phase-locked loop demodulators in the September issue, and noticed a couple of minor errors which I'd like to comment on. In the discussion of capture and lock range, the PLL is referred to as a positive feedback system. That is not correct, as a PLL uses negative feedback (like any other closed-loop system) to reduce the error to a minimum. In this case, it is the phase error between the reference (the frequency you want to track) and the controlled variable (the VCO frequency) which is reduced by developing a signal proportional to their difference.

Another gremlin showed up in the caption for **Fig. 4** and the accompanying text, where the acquisition beat note is described as sinusoidal. It is not sinusoidal, as an examination of photographs of that signal will show. It can't be a sinusoid, because a sinusoid doesn't have a dc component. As the author points out, the output of the phase detector and low-pass filter has to have a dc component which will drive the VCO toward the reference. The most common phase detector in PLL systems is a multiplier which multiplies the input signal by a squarewave from the VCO. The multiplier produces a positive halfwave rectified output when the inputs are 90 degrees out of phase, and a negative half wave rectified output when the inputs are 180 degrees out of phase. The acquisition beat note consists of a series of these rectified sine waves which increase in period as the difference frequency decreases. The low-pass filter acts as an integrator which reduces the ripple and provides an average dc voltage which drives the VCO toward the reference. The PLL will lock when the signals are the same frequency and differ in phase by 90 degrees, which is the point where the phase detector and low-pass filter has zero dc output.

Academic nitpicking aside, I was pleased to see a nonmathematical discussion of amateur applications for the PLL. The PLL offers the possibility of some improvement in threshold sensitivity over conventional discriminators, due to its lower noise bandwidth. The noise bandwidth of the PLL is determined by the lowpass filter (approximately 3-5 kHz for an fm speech demodulator). A conventional discriminator must contend with the noise from the full i-f bandwidth. Since the i-f bandwidth for amateur fm is usually 14 kHz (and the noise bandwidth is even higher), the threshold sensitivity of a PLL demodulator should be several dB better than a conventional discriminator. I plan to run some experiments in the near future to determine the actual improvement in practice, and hope to see more discussion of this subject in *ham radio.*

John J. Murphy, K6JLF
Post Office Box 1875
Ridgecrest, California 93555

pi network design

Dear HR:

A small short circuit in my article on "Pi Network Design" (March, 1978 issue) has surfaced courtesy of correspondence with Robert F. White, W6PY. Three values at the top left-hand column on page 37 should be:

$$X_c' = -608.64 \qquad C_1' = 36.572\ pF$$

In the second paragraph following the tabulation, total $C_m = 222.59$ pF. Total mid-point capacitance, due to improper calculation of the first (prime-valued) section, is off by exactly 5 pF.

Upon checking my original notes, I found there were two calculations for the dual-section pi network; I inadvertently used the wrong set. The equations are correct; only the calculated values are in error. W6PY found the error with his TI-59 calculator program.

Leonard Anderson
Sun Valley, California

data package for the programmable hf receiver

Dear HR:

Because of the large response I received to my programmable high-frequency receiver, which appeared in the October, 1978, issue of *ham radio,* I have put together a limited number of data packages as a help to prospective builders. These data packages include over 40 pages of circuit and wiring diagrams, mechanical diagrams, and other data; the cost is $9.75 per set, to cover costs of printing and mailing. The data package is intended as an aid in duplicating the receiver or using the same basic scheme to build a receiver to suit your readers' own needs. Printed-circuit layouts are not included, but the mechanical and wiring diagrams will be helpful in circuit-board development.

Norman J. Foot, WA9HUV
293 East Madison Avenue
Elmhurst, Illinois 60126

40-meter receiver

for home construction

A project for those wishing to build their own high-performance superhet — it features double conversion, all-variable-diode tuning, and digital frequency readout

This article describes an easily duplicated 40-meter receiver project. The design minimizes the use of "bells and whistles" and emphasizes performance. A variety of add-on features can be incorporated in the design.

Most Amateurs have at one time or another expressed interest in building their own receiver; however, most receiver projects fail. The reasons are usually ascribed to lack of parts, time, interest, or know-how. Many receiver projects fail simply because the builder tries for too much accessory performance before the basic receiver system is operating. That is, many hours may be spent attempting to incorporate certain special-purpose features such as a fast attack agc in a detector long before the i-f and oscillator sections are operating. The builder usually loses enthusiasm for the project by being engrossed too long in the details of getting some basic function to operate in some superhuman fashion. My recommendation is to concentrate on getting the basic system operating at least to the point where signals are heard; then real improvements and add-on features will keep the inertia of the project moving along to completion.

By M. A. Chapman, K6SDX, 935 Elmview Drive, Encinitas, California 92024

Some receiver construction projects can fail because the builder wants long-imagined, ultra-high performance and operating flexibility. To have "everything" in a receiver is impractical. A more rational approach is to think "reasonabiy." Compromise the first time, learn to get the basic functional biocks operating correctly, then add on the bells and incorporate your own whistles.

some ideas

Fig. **1** illustrates the functional blocks found in a superheterodyne ssb and CW receiver. One can argue the need for an rf stage; however, to reach out and snag the really tough signals and minimize spurious mixer products, the rf stage is preferable.

The biggest single problem with the superhet are image signals also available to the i-f stage. The use of moderate Q values in the rf and mixer tuned circuits can minimize these image effects. In some operating areas, where the image signal strength is many orders of magnitude greater than the principal tuned signal, tuned traps in the antenna feedline can reduce the image-signal amplitude to acceptable levels.

The conversion of the desired signal frequency to an intermediate frequency is the responsibility of the first mixer. Mixers are nonlinear devices, which approach a power-law transfer function; because of this nonlinearity they also send along other spurious products available from the antenna to the i-f amplifiers. These products are usually second- and third-order signals but represent other noise and image sources with which your receiver must contend.

The audio stage. Audio detection normally occurs in the second mixer. The beat oscillator provides an injected carrier for the amplified i-f signal to beat or heterodyne, resulting in a detected audio signal. The signal at the second mixer output contains, in addition to the desired audio, a wide spectrum of mixer-signal components. Most of these are in the rf range and are easily bypassed to dc ground. Some audio harmonic products exist and usually represent

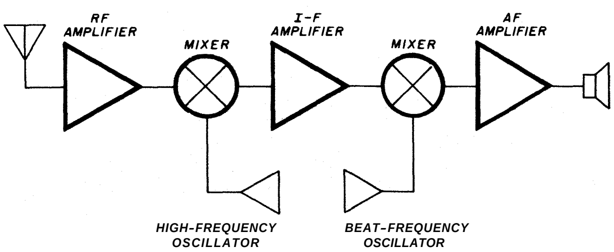


fig. 1. Block diagram of an ssb superhet receiver.

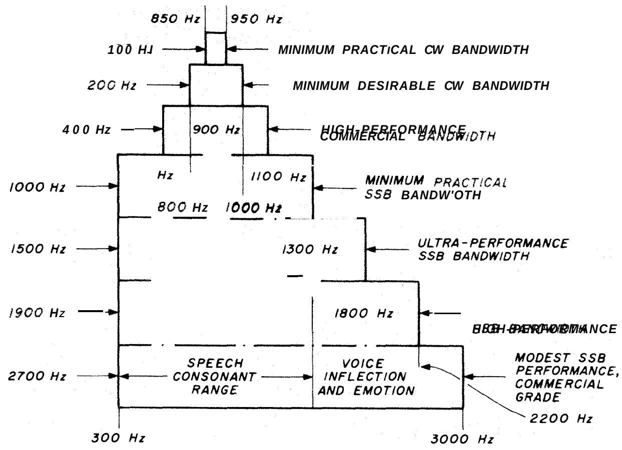


fig. 2. Summary of receiver audio bandwidth requirements for Amateur ssb and CW communications.

receiver background noise. Other noise sources are white noise from the antenna, which has an instantaneous frequency component equal to or near the audio signals of interest, and, active-device thermal noise from the receiver front end. When we speak of signal-to-noise ratio, we are usually referring to the amplitude ratio between a detected signal of say, 1 μV, with the receiver input grounded or shorted. Many commercial manufacturers also perform this test in an rf-free screened room, where discrete sources can't penetrate the chassis and only the receiver internal noise is actually measured.

Restricting bandwidth. We may improve our system signal-to-noise ratio by restricting the bandwidth of detection to the smallest possible interval that will support communications. In the reception of a CW signal we're accustomed to a tone between 800 and 1000 Hz; one could visualize a system that has only a 200-Hz bandpass. Many superior receiving systems have a CW bandpass on the order of 100 Hz, centered near 900 Hz. The signal-to-noise ratios are extremely good, since only a random number of noise products fall within this bandwidth. For ssb communications, we need only a bandwidth wide enough to pass the voice consonants. Amateur receiving systems exist that employ filters limiting the audio bandwidth to approximately 1000 Hz. Various audio communications tests have shown that the average person requires audible communications bandwidths of 1500-2000 Hz. These ideas are illustrated in fig. **2.**

If the useful audio bandwidth requirements for ssb communications are limited to 2 kHz the i-f bandpass need not exceed this interval; however, using LC-tuned circuits in the i-f or rf sections to achieve a sharp-skirted 2-kHz bandpass would require almost superhuman effort. The value of bandpass can be satisfied using crystal or mechanical filters that have extremely sharp passbands, so that infringing adjacent signals don't spill into the interval of interest.

High-frequency thermal noise in the receiver front-end active devices is normally outside the filter cutoff; however, noise originating in the detector and audio frequency sections should be considered. Most RC cutoff techniques are suitable for a-m and fm broadcast reception but don't provide sufficient attenuation of undesirable audio components for ssb and CW reception. A variety of active low, high, and bandpass circuits exist for audio processing, and most readers are at least tentatively aware of their implementation in a receiver. The ARRL *Handbook* illustrates several audio-processing designs that will provide good post-detector audio processing for optimizing ssb and CW reception.

design

We'll start with **fig.** 3, a block diagram of a 40-meter superheterodyne with all-variable-diode tuning and digital-frequency readout. There's no real difference in the diagrams of **figs. 1** and **3** — only some minor digital equipment, which allows 100-Hz tuned-frequency determination. The i-f has two separate frequencies, hence dual conversion — a desirable feature for high gain and narrow-bandpass tuning.

by a summation of signals from each of the principal oscillator sections and displayed using a five-decade LED arrangement.

Referring to **fig. 4,** signal preselection and initial receiver gain are achieved in the rf and first mixer stages, a major roie in performance. We should have a good understanding of these circuits. The rf tuned circuit consisting of L1, C1, and VVCI are parallel connected to Q1 gate 1. This is a parallel-tuned circuit. The 0.001 μF capacitor in parallel with the 620k resistor puts the bottom end of L1 at ac ground, as does the 0.05-μF capacitor on the VVCI cathode. With a little arithmetic it's easy to see that L1 is also series resonant near 1.5 MHz, the i-f.

Parallel-tuned circuit L1, C1, VVCI provides a high impedance to incoming signals at 7 MHz and a low impedance to signals outside the effective bandwidth.

The parallel tuned circuit is frequency selective by providing maximum signal voltage to Q1 gate at the frequency of interest and a minimum voltage at frequencies outside the bandwidth. A similar tuned circuit L1, C2, VVC2, is on the input of Q2 gate 1. Variable-capacitance diodes VVCI and VVC2 are ganged for peaking the incoming rf signal.

The series-tuned circuits in L1 and L2 act as image shunts for signals originating at the input. The shunting action occurs because of the low impedance to ground offered by the 1.5-MHz series-tuned circuit.

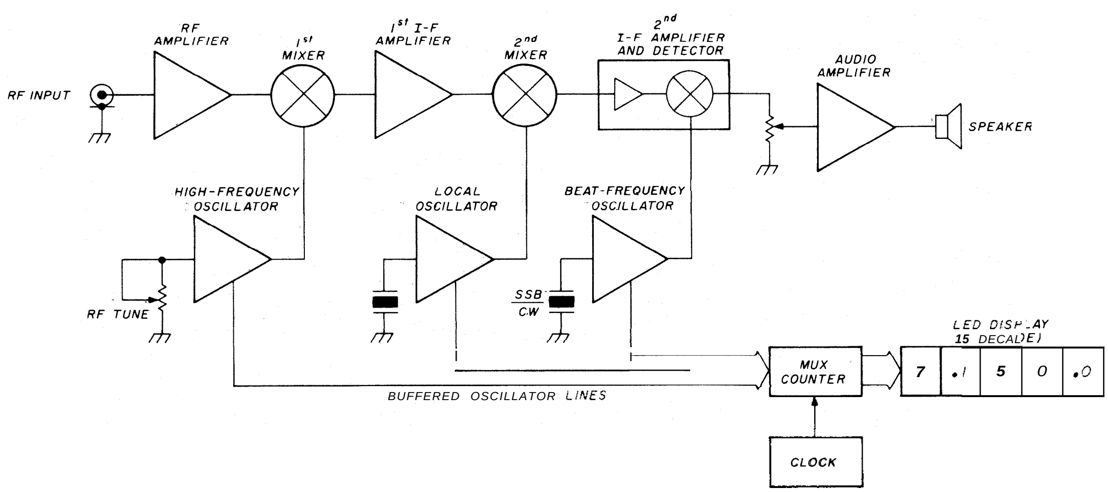


fig. 3. Block diagram of the 40-meter receiver.

Front end. The rf and first mixer tuned circuits are ganged independently from the HFO control voltage for improved signal peaking. Heterodyne oscillators are crystal controlled for stability. Over-all system bandpass shaping is achieved with a Collins mechanical filter in the second i-f. Tuning accuracy is derived

Q1 drain circuit is a simple inductive load ac coupled to Q2 input. Q2 drain load is the primary of the first i-f transformer, T1. In both cases, the drain circuits represent reasonably high ac impedances for good voltage gain. Care in the selection of the Q1 drain inductar is important, because stray winding capacitance can have the opposite effect, *i.e.*, shunting the output signal to ac ground.

Fig. 4 illustrates the Q1 and Q1 mosfet gate bias voltages, which are set for near optimum gain even though Q2 is acting as a mixer amplifier. Because Q1 and Q2 gate currents are small, we can use high volt-

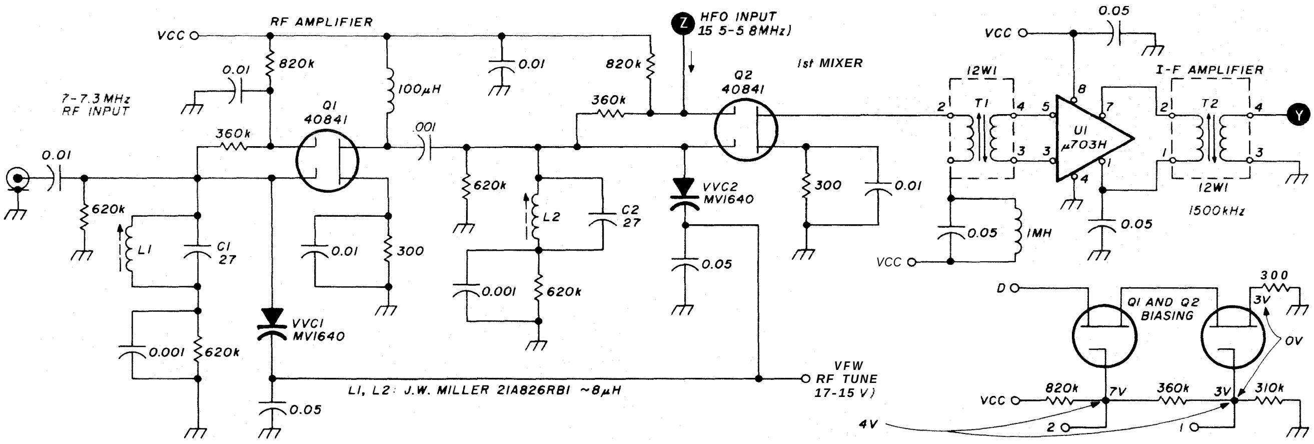


fig. 4. Rf amplifier and first-mixer schematic.

age-dividing resistance values for these circuits and minimize tuned circuit loading.

It may appear strange at first that the rf and first mixer are gate biased at the same point. Readers familiar with the principle of active device power-law mixing, where device biasing is set near the optimum nonlinear portion of the transfer curve, will realize the dual-gate mosfet acting as a cascode arrangement performs a modulation of the drain-circuit current for mixing, rather than a true nonlinear power-law mix. The mosfet can be biased for nonlinear mixing as well, and this is accomplished in the familiar circuit for Q6, the second mixer (see **fig. 6**).

Second i-f, detector, and BFO. Fig. 5 illustrates the i-f scheme for the second frequency conversion. The first i-f amplifier is a high-gain IC, U1, which provides about 28 dB signal gain. I-f transformers on the input and output preselect the desired signal and attenuate undesired heterodyne first mixer components. Q6, the second mixer, provides little actual system gain; the drain circuit load, T3, is the second i-f transformer. T3 includes a simple ceramic filter in the secondary for initial band shaping. This filter is inadequate for any serious ssb or CW use. The principal i-f amplification and detection occurs in U2, shown in **fig.** 6.

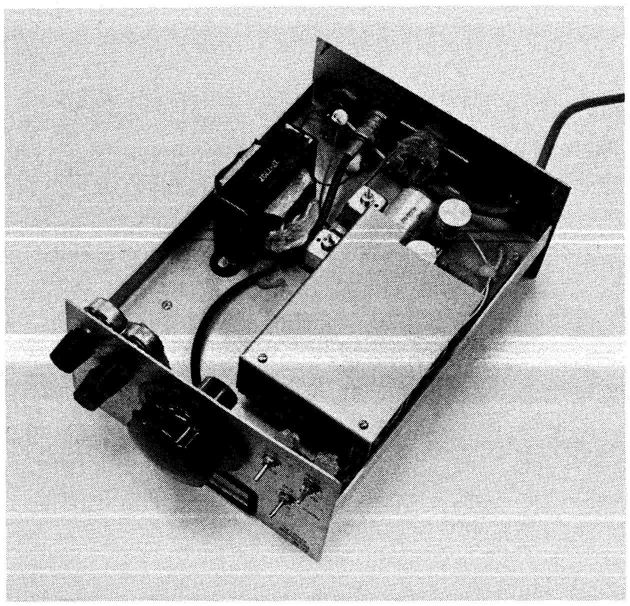
Preselected 455-kHz signals are ac coupled from T2 and internally amplified before presentation to a highly selective Collins mechanical filter for i-f bandpass shaping. I have used a Collins 1.9-kHz filter here; however, a variety of similar filters can be used, and adequate system gain is available for all but a few very exotic types whose attenuation characteristics exceed 30 dB. Nominal attenuation for the Collins unit is ≈10 dB. The selection of 1.9 kHz for bandpass shaping represents a compromise between CW and ssb signal reception. Superior ssb reception is accomplished with this filter; however, only good CW reception is possible.

Following the bandpass amplification section of U2 is a product detector agc feedback stage. Detected audio is RC filtered on the output to decouple rf components before audio amplification. The agc feedback circuit is not connected in this design but may be incorporated as an option.

Oscillator circuits. The oscillators are shown in **figs. 5, 6,** and **7.** Each of those oscillator circuits is a simple Colpitts arrangement followed by a fet bipolar isolation and shaper stage. The fet buffer minimizes oscillator loading and the effects of frequency distortion from the CE shaper stage. The high frequency oscillator (HFO) is tuned using a variable-capacitance diode, VVC3. The tuning voltage (V_{FV}) provides the range of reverse bias necessary for tuning the HFO for proper heterodyne-frequency conversion in the first mixer stage.

Audio amplifier and power supply. Detected audio amplification and dc power generation are shown in **figs. 8** and **9. Fig. 9** also includes the scheme for deriving the V_{FV} voltage and a method of band spreading the HFO frequency for ≈20 kHz/revolution of the front panel tuning control potentiometer.

Digital counter. Fig. 10 is the schematic of the digital counter. The clock from which the entire accuracy of our system depends is derived from the 10-MHz crystal oscillator, Q1, and multi-decade divided



Underchassis view of the 40-meter receiver.

by U1, U2, and U3. The combination of division in U1, U2, and U3 is 106, which provides a 10-Hz clock pulse. The QB and QC strobes are used for gate and transfer timing. U4 may be thought of as a ring counter, which further divides our 10-Hz clock into four equal time periods shown on the gate timing diagram, **fig. 11.** U5 is an AND gate to control the count periods. U6 is a large, multiple three-input AND device used to select a discrete period. The oscillator input counts and guides the signal in serial sequence through to the LED display.

We can best visualize how the digital counter works by remembering that the actual tuned frequency is a summation of the various rf stage oscillator outputs. That is, if we take a typical case, the osciilators can be summed to provide our actual tuned frequency, as in the following example:

HFO:	5.650.0 MHz
LO:	1.045.0 MHz
BFO:	0.456.0 MHz
SUM:	7.151.0 MHz (display)

Our counter works exactly as shown above. It takes each oscillator pulse train and looks at it for 100 ms, transferring the gated puise train through the final gate, U7, in sequence. If we examine, for instance, the high-frequency oscillator at some nominal case, say 5.65 MHz. we see that there are 5,650,000 counts in one second. But since we have a gate window for the HFO of only 100 ms, then only 565,000 counts are seen by U7 in any one count cycle period. The same is true of the local oscillator and BFO counts. The total number of counts is proportional to the period established by the clock timing interval, and gated as shown in the timing diagram of **fig. 11.** The count cycle periods shown in **fig. 11** are derived from the ring counter AND circuits of U4 and U5. Each complete count cycle is 250 ms wide and has four separate parts:

1. Part 1 is count oscillator 1
2. Part 2 is count oscillator 2
3. Part 3 is count oscillator 3
4. Part **4** is transfer and display

Each of these periods is sequential so that only a serial stream of counts is seen and passed by U7. U8 serves only to delete the unwanted decade before passing the train into the LED display. U8 may be deleted for six-decade presentation if the LED selected has sufficient speed.

The transfer and clear signals are derived during the fourth time periods of our count cycle and are strobed for delayed timing by the QB and QC pulses so that there is adequate internal ripple and settling time before updating the display.

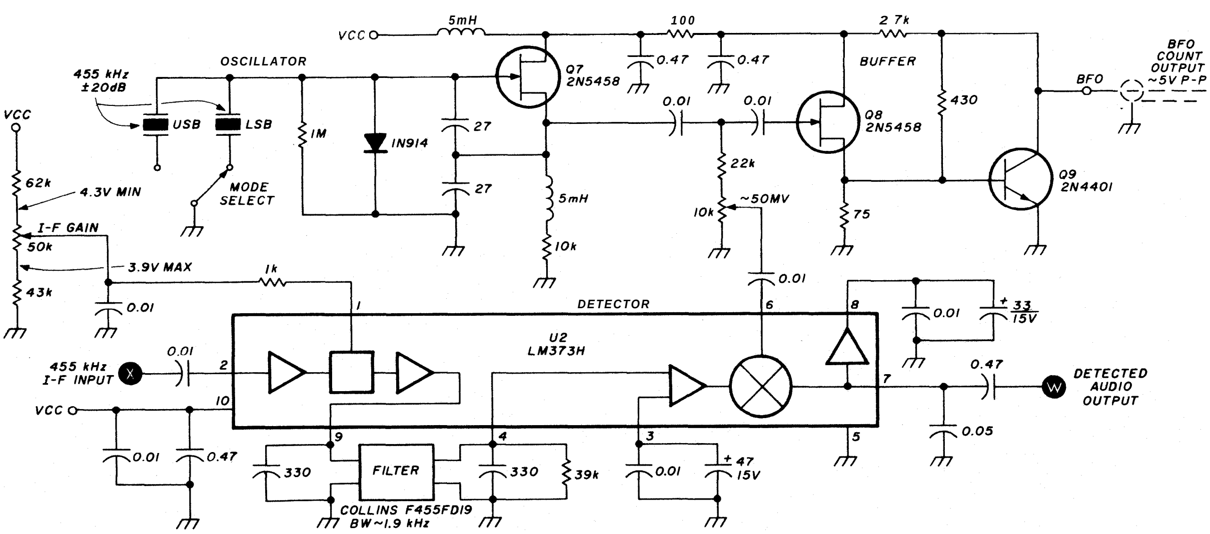


fig. 5. I-f detector and BFO schematic.

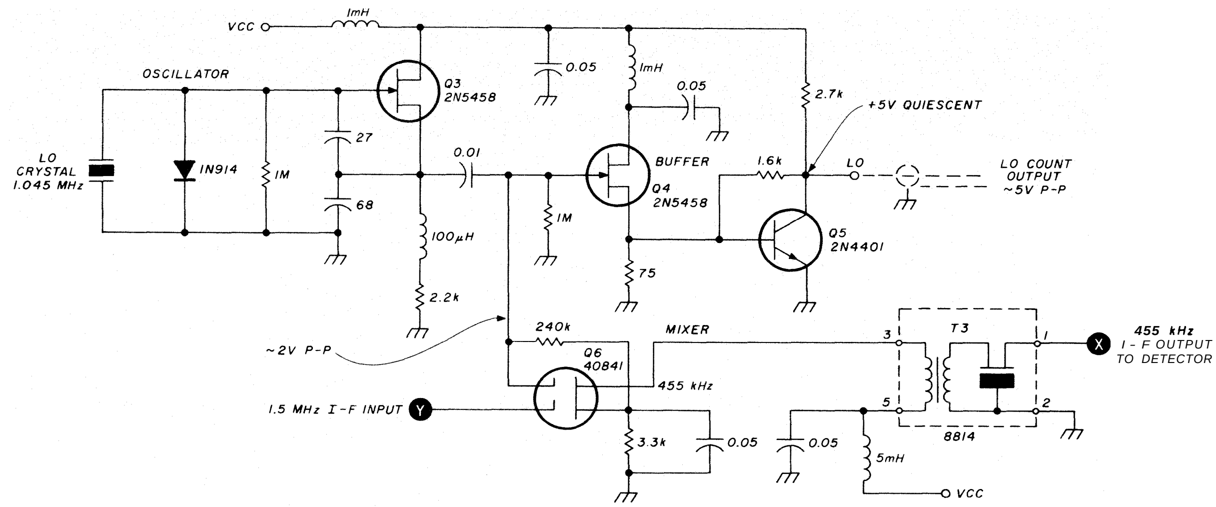


fig. 6. Local oscillator and second-mixer schematic.

U9 through U13 serve as a five-decade count, latch (transfer), and LED* display. Notice that a standby switch controls the 5 volt bus to the LED display. The purpose here is to provide power dissipation limit control but not to interrupt the continued operation of the oscillators. This will avoid the thermal shock that receiver on-off cycling would effect and minimize oscillator drift.

construction and device selection

Construction of the rf section is on one PC-board assembly." The digital circuits are divided between two boards, the gate segments and the LED display. There's probably an ideal assembly sequence; however, being practical, one never has all the parts available at the initial starting point and we must compromise by building up various sections of the system. Any of the subfunctional blocks on a PC board can be built and tested, or the entire board assembly can be constructed and initially aligned. t

The chassis assembly shown in the photographs should in no way represent an optimum or the only way to do it; however, your chassis approach should include effective shielding if birdie and image rejection is to be achieved. An rf shield around the gate and low-frequency clock circuitry is necessary to isolate the large 10-Hz pulse from being coupled to the audio or causing harmonic interference to the rf stages.

There's no magic in device selection. With a little common sense, each of the discrete devices can be replaced with an alternative that is from a similar family type.

Consider Q1, Q2, and 03; these are RCA 40841 dual-gate mosfet devices, and there are similar devices within the RCA family and other manufacturers' lines. The key items when considering substitution here is to become familiar with the transconductance characteristics with respect to gate 1 and gate 2 to source biasing and quiescent drain current. Most of the RCA family devices have very similar biasing. Alternatives such as the 40671 and 3N187 are almost identical in operation. Care in the substitution of units such as the Motorola MFE3006, 007, and 008 types is required, because the gate 2-to-source transconductance curve differs. There are a number of mosfet enhancement types available. You should avoid these in this arrangement because of the difficulty in obtaining sufficient reverse bias to properly operate the tuned circuit VVCs.

The 2N4401 device was standardized as the workhorse NPN. This is a general-purpose and switching unit having moderate beta at low collector currents. Any unit having an $F_b \left(\frac{f_T}{\beta}\right)$ greater than 6.25 MHz, or with similar ratings, can be used here. One needs only to put a small 5k pot in place of the present collector base feedback resistor and adjust the pot until collector saturation occurs, with the output amplitude and frequency as shown in the diagrams. For determining the alternative device feedback resistance value, remove the pot, measure its value, and use the closest standard resistor available.

The N-channel fet, 2N5458, is a general-purpose, medium-frequency device. Any reasonably similar type may be used by placing a pot in place of the source-feedback resistance and adjusting it until the output amplitudes coincide with those shown in the

*PC boards for the receiver are available at nominal cost from the author. Send a self-addressed, stamped envelope for prices.

†A copy of the printed circuit board layouts and parts placement diagrams can be obtained by sending a self-addressed, stamped envelope to *ham radio,* Greenville, New Hampshire 03048.

diagrams. Remove the pot and install a fixed resistor with a value close to that measured on the pot. The output signal should also be examined for obvious harmonic distortion if substitutes are used, because poor mixer gain and increased receiver noise could result.

There are many VVC units that can be used provided they have similar +4 volt ratings and reasonable tuning ratios greater than two. Substitute VVCs should not be self-resonant near these operating frequencies. Alternative ICs to those shown exist in manufacturers' data books by various part numbers; however, the TI74490 is proprietary at the time of this writing. It is my understanding that other manufacturers expect to have equivalent devices on the market soon (*i.e.*, Motorola, National, and Fairchild). Type 74L, 74LS, and 74H units may be used for all or any of the gate circuit elements with slight increases or decreases in current requirements for this assembly. There are CMOS replacements for the low-frequency portions of the gate circuitry; however, their use should be examined carefully to avoid frequency degeneration, since many CMOS equivalent units are inoperable above 5 MHz in the +5 volt V_{DD} condition. Type LM374 may be used in place of the LM373 by including a 1k resistor between pin 9 and V_{CC} on U2. The LM703LH may be used to replace the LM703H provided the ac decoupling cap on pin 5 is moved to pin 7.

alignment and test

Before alignment or assembly of the rf section to the chassis, power should be applied and a verification made of the dc voltage levels and oscillator levels in the quiescent state. Incorrect dc-voltage levels are a sure indication of probable malfunction! The dc values for all significant voltage points are shown in the schematics. These values were obtained using a high-impedance VOM and a nominal +15 volt V_{CC} voltage. Type 7815 regulators will provide outputs of ±5 per cent of nominal and may have a mild effect on the expected dc levels; but any circuit having dc values ±10 per cent from those indicated should be closely scrutinized.

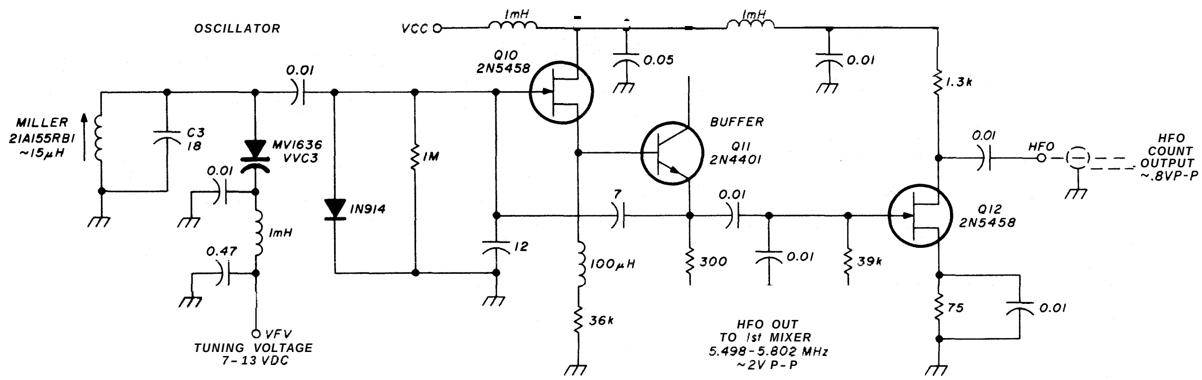


fig. 7. High-frequency-oscillator schematic.

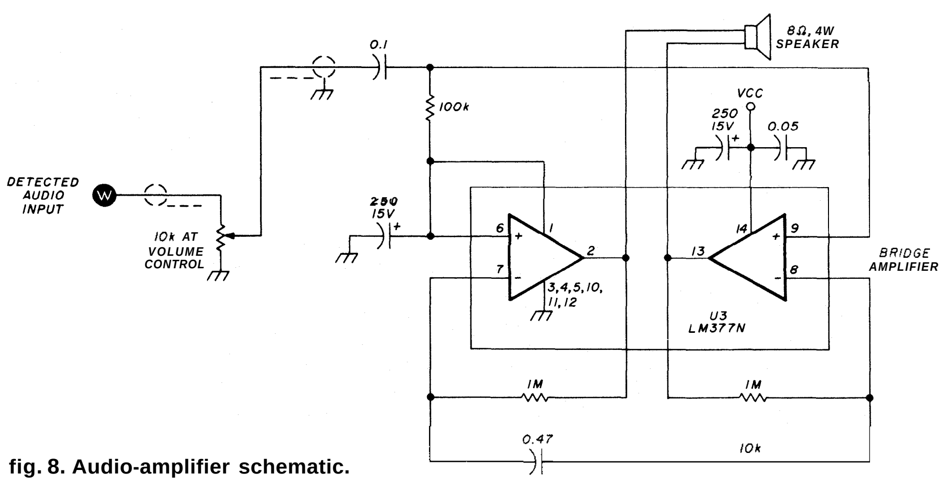


fig. 8. Audio-amplifier schematic.

test sequence

1. AC couple a 50-mV peak-to-peak 1-kHz signal through a 0.01-μF capacitor to U3 pin 7. With power applied, a 2-volt P-P signal should be apparent on pin 12 with a 10-ohm resistive load substituted for a speaker load. (All signals are referenced to ground.)

2. Adjust the BFO level on U2 pin 6 for −50 mV P-P. Adjust the i-f gain control for 4.0 volts on U2 pin 1. Ac couple a 10-mV, 455-kHz rf signal (unmodulated) through a 0.01-μF to U2 pin 2. Adjust the 455-kHz rf signal for 1 kHz detected audio on U2 pin 7. The amplitude of the detected audio should be 60 mV P-P; detuned noise should be ≈5 mV P-P. A verification of i-f gain control should be made by varying the voltage on U2 pin 1 from ≈3.9 to 4.5 volts. Maximum gain should occur at ≈4.0 volts.

3. Ac couple a 455-kHz rf signal of 10 mV into T3 pin 3. Peak T3 for maximum detected audio at U2 pin 7.

4. Ac couple a 1.5-MHz rf signal of 100 μV into T1 pin 2 and peak T1, T2, and T3 for maximum. Detected audio signal on U2 pin 7 should be approximately 200 mV P-P. The signal may also be monitored on U2 input where a 455-kHz signal appears and T1, T2, and T3 adjusted for maximum.

5. The HFO V_{FV} fine band-edge voltage levels should be set as shown in the schematic. With the tuning potentiometer set at the indicated levels, or using an external supply, apply ≈1[1] volts to VVC3 cathode. With V_{CC} applied, adjust L3 slug until a 5.65-MHz signal is available on Q2 gate 2. Adjust the source swamping resistor (30k) for the output amplitude shown in **fig. 7.** Traverse the tuning range and alternate the bandset switch so that the output signal frequency is 5.498 MHz at the low end and 5.802 MHz at the high end of the HFO. Monitoring should occur at Q2 gate 2.

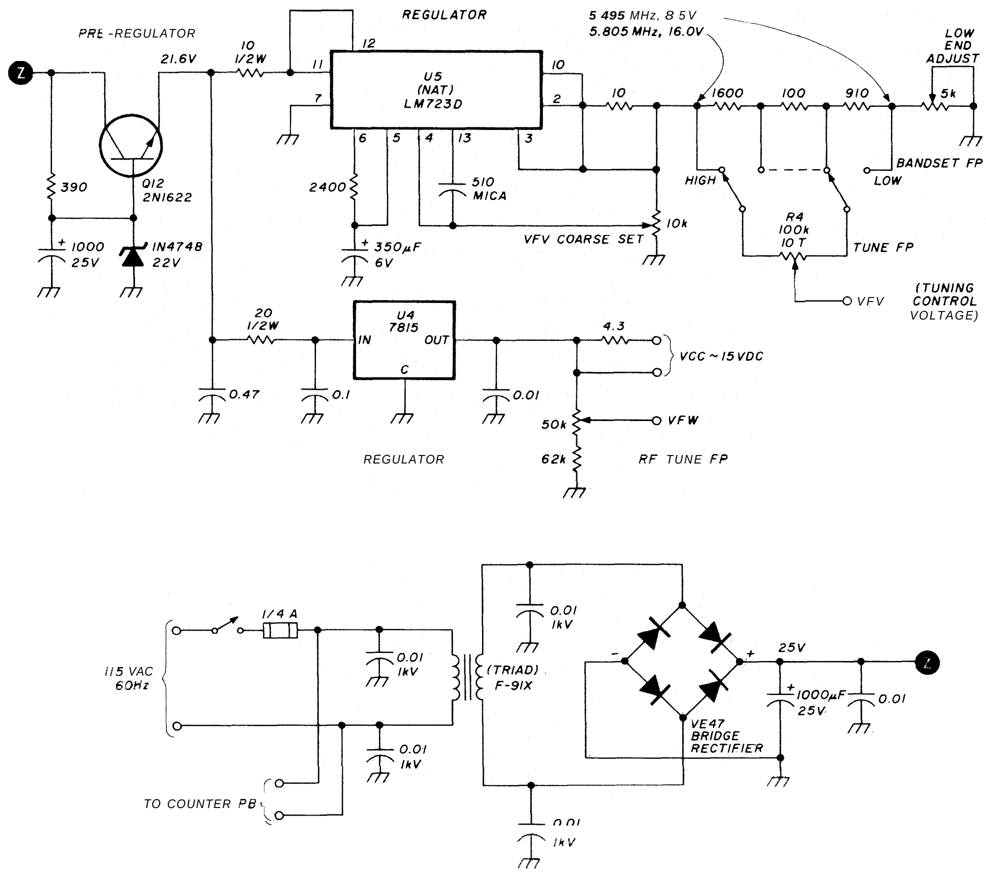


fig. 9. Power-supply schematic.

6. Ac couple a 7.15-MHz signal of 100 μV into Q2 gate 1 and monitor the audio output while adjusting the HFO for detector conversion. After obtaining detected audio from U2 or U3, with the HFO set for conversion of 7.15-MHz signals, repeak T1, T2, and T3 for maximum. Monitoring the rf envelope on U2 pin 2 may provide an easier point for verification of optimum alignment.

7. Ac couple a 30-μV signal of 7.15 MHz into Q1 gate 1. Adjust the HFO frequency for a maximum 455-kHz envelope. Adjust the rf tuning pot for maximum-detected audio or rf envelope. Inject a 5-μV signal into the antenna input, and peak L1 and L2 for maximum detected signal; then go back through all tuning elements and peak for maximum. The peak detected rf envelope at U2, for a 1-μV input of 7.15 MHz, should be 60 mV P-P (≈95 dB!.

as shown in **fig.** 11; however, scope monitoring of the logic pins should indicate that a pulse exists.

2. Inject a 1-MHz rf signal of 100 mV into oscillator no. 3 input, and monitor U7 pin 12 and U8 for oscillator clock train output occurring at regular 250-ms intervals. If the display board is available, temporarily

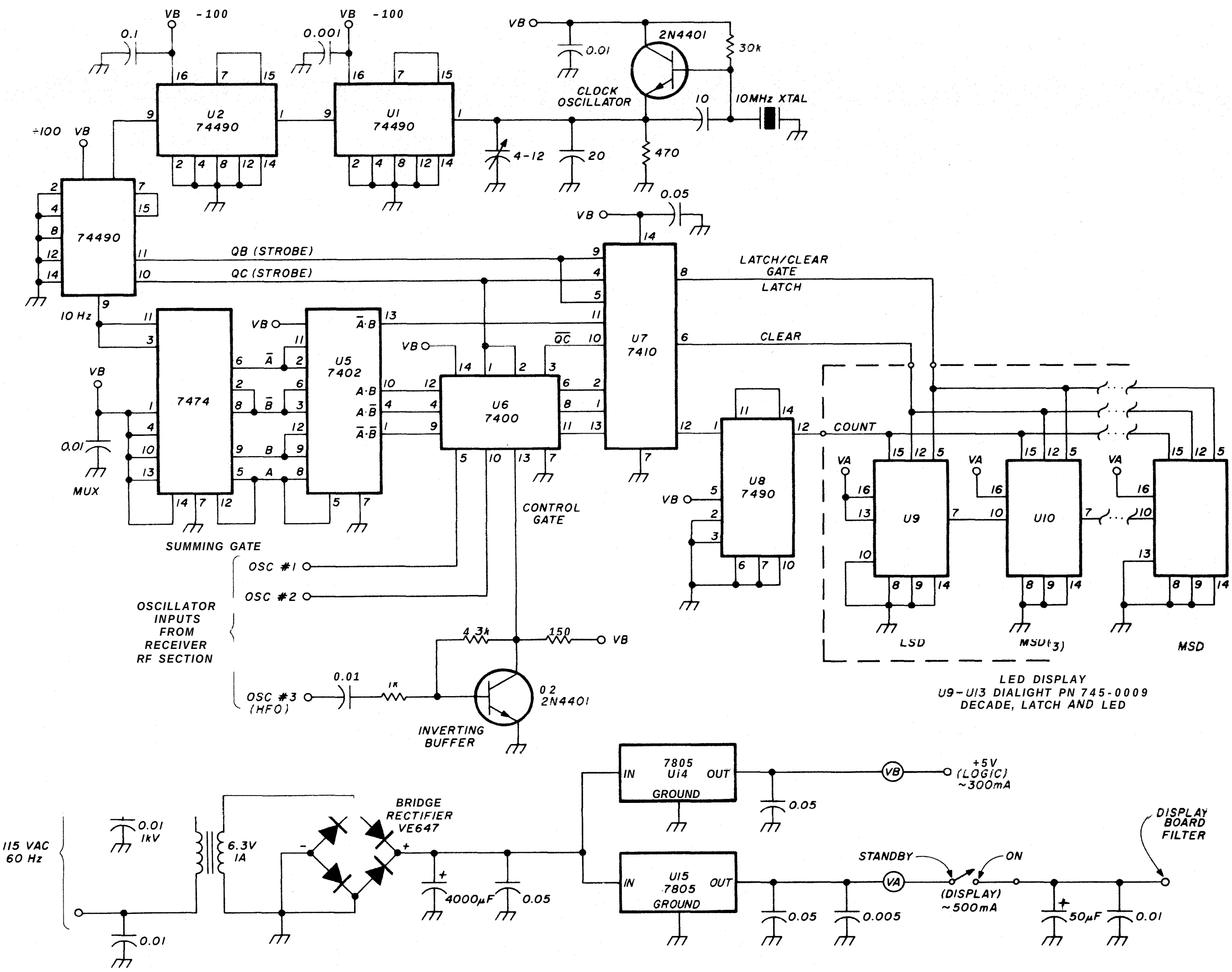


fig. 10. Digital-counter and display schematic.

8. Final alignment should occur after the rf and digital boards have been installed on the chassis, because the effects of stray capacitance will shift all of the oscillators and rf-tuned circuits slightly. This final alignment should occur with rf input signals in the order of 3 μV.

digital-circuit testing

1. Apply power to the gate control board and monitor the frequency of the quiescent states of the logic circuits, as shown in **fig.** 10. It may not be possible to verify the various timing pulses relative to each other

jumper the latch and clear lines. A display of 1.000.0 should result, Inverted latch and clear lines, or unconnected inputs to the display, will result in LED garbage, or an output of 0.000.0. On occasion, when power is first applied to the display, a residual numerical value will appear. In normal operation the next count cycle coming up (250 ms later) should clear the erroneous display.

3. The 10-MHz clock should be trimmed for the closest possible accuracy by zero beating against WWV at 10 MHz, using a separate receiver and a simple link coupling around the clock oscillator.

hints and kinks

Accuracy in a digital system is defined as time base accuracy ± 1 digit. This accuracy must include aging and temperature stability. The digital-counter accuracy will then be as accurate as the basic 10-MHz clock oscillator. All crystal oscillators have both long- and short-term drifts due either to aging, temperature, or circuit-component changes. The display accuracy then will drift with the crystal-oscillator drift.

Without temperature compensation you can expect to have drifts in the order of 5 ppm/°C plus additional aging drifts of perhaps 10 ppm/year. The net result is that faith in the absolute accuracy of the LSD of this counter is wasted. It might impress visitors, but without constant recalibration of the 10-MHz oscillators against WWV or other prime standard at regular intervals, the LSD is only a guess. The actual consistent counter display accuracy is to the closest kilohertz over any significant period of time. This could be improved to ± 200 Hz ± 1 digit by using TCO (temperature-compensated oscillator) techniques.

Selection of the BFO crystals can be made by measuring the bandpass mechanical filter for the upper and lower 20-dB points. The BFO crystal frequency should coincide with these measured values within -50 Hz. Most suppliers will furnish the filter with this information, marked either on the filter or with the shipping data package, for a small fee. Significant errors in the BFO frequency will degrade ssb detection.

In the local oscillator and BFO circuits, the amplitude of the output signals from the isolation and CE shaper will vary with different crystal types. Excessive CE saturation will cause a 10-Hz clock pulse to be coupled into the V_{CC} line and will result in pinging of the detected audio. Conversely, low outputs will not stimulate the gate levels required for proper counting.

The suggested cure is to temporarily install a 5k

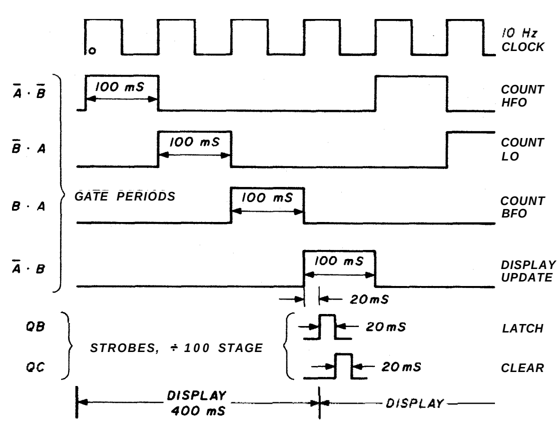


fig. 11. Digital-counter timing diagram.

pot in the 2N4401 collector to V_{CC} load position. Adjust each of these until the amplitude of the output signal is just high enough to stimulate the gate input but will not keep the collector circuit in saturation for a period that allows the 10-Hz signal to load the V_{CC} line. The schematic illustrates the peak-to-peak signal levels I found achieved these goals.

A 100-kHz clock oscillating crystal may be used as a substitute for the 10-MHz unit in fig. **11** by deleting the second divide-by-100 (74490) stage. Some adjustment in the oscillator bias values may be required, and counter accuracy will be limited by your ability to trim the oscillator frequency properly.

improvement suggestions

The following are areas of individual user improvement:

1. RF gain. Wide dynamic range is necessary to meet the demands of the Amateur-band receiving environment. Signal levels vary from parts of a microvolt to perhaps a millivolt — a three-decade range. The front end of this receiver is optimized for maximum transconductance (gain). Strong signals $>100\ \mu V$ will cause channel saturation in the rf- and first-mixer stages. Two easily implemented methods of controlled rf gain are available to the builder:

a. source current limiting: use a 2k pot in the source lead instead of a 300-ohm resistor.

b. gate-2 control: lift the gate from the board and add a bypassed flying lead to a 0-5 Vdc voltage.

2. Agc. Positive agc voltage is available on U2 pin 8, proportional to detected audio. The quiescent value, -4.0 Vdc, is very close to the i-f gain maximum value. As the audio level from the U2 audio mixer stage increases, this agc voltage may be fed through a 1-k resistor to U2 pin 1.

Agc voltage may be fed through a 1-k resistor to U2 pin 1. The i-f gain circuit must be open or disconnected for the agc feedback system to operate.

3. Variable bandwidths. The addition of a diode switch and other Collins-type filters will allow optimization for both CW and ssb reception.

4. HFO stiffness. Thermal — physical isolation of U5 will improve V_{FV} regulation and **drift/°C** rate; transient — increased beta characteristics for Q12 will aid in improved ripple rejection for both the HFO and general receiver operation.

5. Images. A simple series-tuned 1.5-MHz shunt in the antenna input will improve the image rejection by 20 dB. Provisions are on the PC board for the installation of the components.

ham radio

CW operator's PAL

An accessory for CW operators that reduces difficulties with interference, noise, and fading

Described in this article is a multifunction magic box, called a *Pal,* which I designed to relieve some of the operating problems faced by CW operators. Some of those problems are outlined below. The *Pal* is based upon old, well-known techniques; it is novel primarily because it merges — and modernizes — past solutions to these problems:

1. Many Amateurs endure listening to CW signals but enjoy listening to telegraph sounders. Most Amateurs own or can obtain sounders, which are now obsolete, and old-timers usually retain a nostalgic interest in telegraphy.

2. Received signals vary in pitch, due to frequency drift at the transmitter or receiver, which requires frequent readjustment of the receiver's BFO. Chirps due to poor transmitter power-supply regulation, or jumps due to sudden line voltage changes, can also cause problems for the CW operator.

3. Received signals vary in volume due to fading and due to variable propagation conditions; sometimes gain changes at the transmitter or receiver can occur during warm-up periods. Changing audio volume forces the receiving operator to frequently readjust the gain control to maintain normal volume.

4. Received signals often include annoying interference from adjacent frequencies, even with very selective receivers.

5. Received signals incur variable background noise due to a combination of atmospheric, static, and manmade electrical noise; noise limiters help but don't eliminate all noise.

6. Transmitter operators are said to have poor "fists" when they send imperfect code characters with straight or semi-automatic keys. Poor sending can often be attributed to inability to monitor a transmitter not equipped with sidetone output.

7. Many Amateurs don't own a code-practice set. Such a set is more than a means of learning and teaching code; it permits high-speed, off-the-air sending practice and proper adjustment of a semi-automatic key.

circuit description

In the *Pal* (fig. **1**), a rectifier-filter circuit is used to actuate a telegraph sounder; a tone oscillator-amplifier is used to provide a code-practice set. Both circuits are used to improve CW reception and transmission. The input voltage to drive the *Pal* is derived from either the receiver (J1) or a voltage picked off the transmitter (J2).

The external power supply may be a 12-volt storage battery, eight dry cells in series, a regulated 12-volt power supply (such as Radio Shack's no. 22-124), or a homemade unregulated power supply which delivers 12-15 volts at about 1 ampere. An internal 9-volt transistor radio battery of moderate size will suffice for code practice.

An ac-connected power supply should have a switch and an indicator lamp — either a small 120-volt type across the switched line or a 6.3-volt, 0.15-ampere pilot lamp in series with a 40-50 ohm, 2-watt resistor across the 12-volt output. Lamp types 40 (screw base) and 47 (bayonet base) are suitable.

construction

Nothing in this circuit is critical. The circuits may be breadboarded and simply placed in a box. A 5 x 5 x 7 inch (12 x 12 x 18 cm) metal box is recommended because it will easily house all the parts except for the sounder (on top) and the key.

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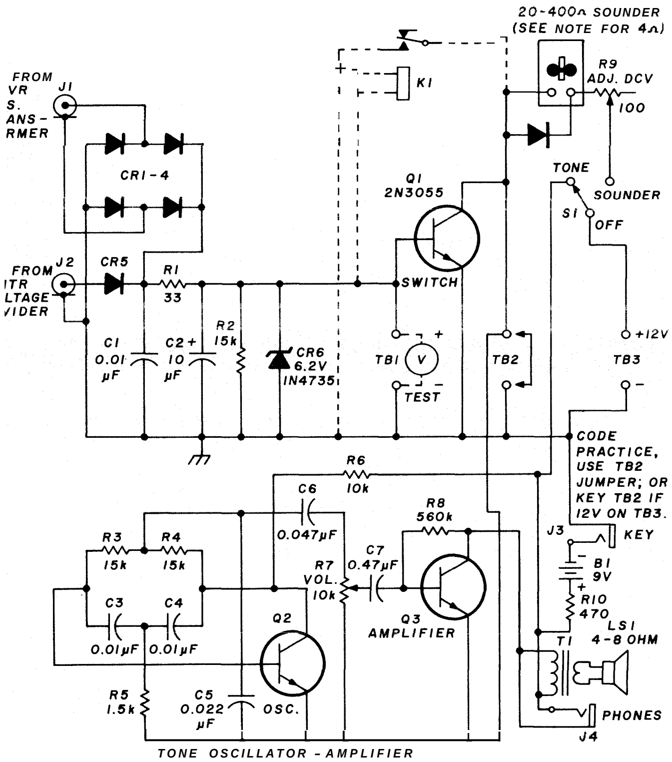

NOTE. ALL DIODES 100PIV, 1A SILICON

fig. 1. Schematic diagram of the CW operator's Pal. Q1 is preferred and relay K1 (Radio Shack 275-004) is optional because Q1 is more sensitive; it needs less input and perhaps no step-up transformer if fed directly from a high-impedance phone jack instead of a voice-coil input. Relay K1 can be used to operate a 4-ohm sounder (1.5 Volts, 375 mA) by replacing the latter in the given circuit, setting R9 to 500 ohms (or using a 560-ohm, 1-watt fixed resistor), and putting K1 contacts in series with the sounder and a large dry cell or a 27-ohm 10-watt resistor in the 12-volt line. Transistor Q2 is an NPN low-power device.

sounder operation

Any telegraph sounder with 20 to 400 ohms resistance is usable, but the power supply voltage must be dropped with a series resistance. For example, if a 20-ohm sounder operates at 3 volts and 0.15 ampere, the necessary 9-volt drop requires 60 ohms resistance at 1.35 watt, so a 56-ohm, 2-watt resistor will do. Obviously, if various sounders are to be used, an adjustable resistor or rheostat must be used; a common 12%-watt, 100- or 500-ohm rheostat is a good choice.

A communications receiver usually delivers only a fraction of a volt to a 4-8 ohm loudspeaker. The CW-dc converter must have a higher ac input voltage to deliver about 0.8 volt to a switching transistor or 5 volts to a sensitive relay. The dc output voltage of a full-wave bridge rectifier is, theoretically, 90 per cent of the input ac voltage. Under these conditions, it is necessary to use a step-up transformer unless the output from a high-impedance phone jack is sufficient. The step-up transformer can be made from a filament or output transformer used backwards and shunted across the voice coil of the speaker.

A switch in series with the voice coil will disable the loudspeaker while the ***Pal*** is in use, thus avoiding confusion between CW tones and sounder clicks. Before disabling the loudspeaker, advance the receiver gain control to a point somewhat beyond the triggering level. A lower level invites drop-out during fading, and a higher level invites atmospheric static and manmade noise, which seldom affects the sounder under normal operating conditions.

The pitch and volume of the original CW note are of little importance during sounder operation; in fact, the note may be much higher or lower than desired during normal CW operation. Since pitch is now unimportant, you may zero-beat or drop the frequency of an undesired signal so that only the desired signal passes efficiently through the receiver-to-converter transformer. If a phone signal happens to be received, the sounder will faithfully click on voice peaks; no harm is done, and the effect is rather amusing.

tone-oscillator operation

A center-off switch (to avoid needless energy loss) is used to replace the sounder with a tone oscillator, amplifier, and loudspeaker. The converter circuit is unchanged and the operating procedure is much the same. The pitch of the new CW tone is fixed by the circuit constants; the volume is fixed after a preliminary setting of the gain control. Pitch and volume are now unrelated to receiver output; they should never need readjustment regardless of fading, interference, or noise.

During tuneup, there are two tones because there are two loudspeakers with different inputs. Thereafter, confusion between the fixed and variable tones may be avoided by opening the receiver loudspeaker disabling switch.

transmitter operation

The sidetone for transmitter monitoring is produced in much the same way as in the CW-to-CW conversion just described. A separate half-wave rectifier is connected to a voltage divider shunted across the transmitter output. The half-wave rectifier and the bridge rectifier share a common output and their back-to-back polarity connection avoids interaction unless both inputs are active. Fortunately, transmitter and receiver functions do not occur simultaneously, hence the ***Pal*** can serve both functions at the same time without changing connections. Thus, both plugs can remain in both jacks (phono and coaxial) regardless of the choice or number of functions desired.

It is easy to obtain the rf voltage needed for the half-wave rectifier input; a typical transmitter which delivers 200 watts to a 50-ohm load also delivers 2 amperes at 100 volts. Even a %-watt transmitter can deliver 50 volts, of which only a small fraction is required. The proper voltage divider step-down ratio is, therefore, rather large. It is determined easily by experiment but not by calculation. It is useless to know that a half-wave rectifier, in the absence of voltage drop, has a dc output voltage which is 45 per cent of the ac input voltage. Variables include transmitter power, voltage divider resistance, rectifier load, rectifier voltage drop, and rectifier circuitry.

Under these conditions, you can connect the transmitter to a dummy antenna and gradually raise the tap on the voltage divider until the desired dc voltage appears across the relay coil or the switching transistor base-to-emitter input. The dc voltage should not exceed 6 volts or 1 volt, respectively[1].

The voltage divider should be located in or near the transmitter, if possible, to minimize any effect on the VSWR. The divider might be housed in a can, like a filter, in series with a line from the transmitter to the ***Pal.*** In any case, a short jumper made from RG-8IU or RG-58/U coaxial cable may be provided with PL-259 plugs for attachment to SO-239 jacks at both ends. The transmitter antenna jack may be provided with a T-adapter which has one male and two female outlets (amphenol M-358), one of which will accommodate the jumper.

code-practice operation

There are at least three possible ways of keying the ***Pal*** as a code-practice oscillator. The key may be connected across the normally open relay contacts or the collector-to-emitter leads of the switching transistor. If you wish, a small battery may be placed in series with the key to the phone jack used for normal receiver connection, making sure that the plug tip is positive.

A third way of keying the practice oscillator replaces the external 12-volt power supply with an internal 9-volt transistor battery in series with a key jack, thus providing greater portability.

conclusion

There you have it, a multifunction magic CW box, which is a ***Pal*** to me; other amateurs should be able to obtain similar results. I wish to thank W2SSJ and his correspondents, W9KSR and W9YZE, for their suggestions on CW-to-telegraph converters.

reference

1. *Lew McCoy, "An RF Actuated CW Monitor," QST, November, 1968, page 39, or The Radio Amateur's Handbook, 1971, page* 183.

ham radio

calculator-aided propagation predictions

An automated approach to propagation predictions by using a programmable calculator and published ionospheric predictions

Those people who use high-frequency propagation predictions know them to be accurate. Commercial interests use the predictions as part of the normal course of business. DX contest operators use them, among other things, to determine band openings into high-density population areas to establish a high contact per hour ratio before skip lengthens into the rest of their country. DXCC operators use them to determine at what time and on what band a specific long-haul station will be coming through so as not to waste time when no path is possible.

Propagation predictions became practical after World War II through the work of the Central Radio Propagation Laboratory of the National Bureau of Standards. In 1947, Newell Atwood, W3KTR, published an article[1] describing a method whereby CRPL propagation prediction overlay charts were moved over a basic world map, showing what areas of the world were open from a home station for any band and time of day. That information was issued by CRPL three months in advance and proved very useful, as well as accurate.

Much has been written over the years on the subject. At least one magazine[2] devoted a complete issue to the subject, while another[3] devoted a monthly article for many months to prediction techniques and their results. Finally the ultimate was reached in 1971 when *Telecommunications Research and Engineering Report 13* was issued by the United States Department of Commerce.[4] The report consists of four volumes on ionospheric predictions covering all frequencies, time of day, and day of the year, and by interpolation, for any solar activity with a Zurich sunspot number from 10 to 160.

The use of these predictions is explained in Volume 1, but a detailed description was published by Jerry Hall, K1PLP.[5] Every Amateur interested in DX should have a copy of these four volumes in his library. The use of these predictions is tedious and time consuming, but well worth the effort. First, let me sell you on their importance by illustrating how you can use the prediction techniques to get that long-haul DXpedition when others can't. Then I will explain how I've streamlined the use of them, and made it more fun by use of an SR52 computing calculator with the PC100A printer.

The programs discussed in this article are directly related to the TI-58/59, newer models of the SR52/SR56; by use of the equations and flow charts, they are applicable to the HP-25 and similar calculators. The PC100A printer is a desirable luxury, but is not necessary to the functioning of these programs.

Back in December, 1975, the Northern California DX Foundation sponsored a DXpedition to CR9AK in Macao, a country I needed. Macao is a long way from New Jersey; since I work from 8:15 AM to 5 PM, I had to know when I could expect to work him – to increase the possibility of being there when his signals were coming through.

The sunspot cycle had not reached bottom in 1975, but it was pretty low, around a Zurich sunspot number of 30. Under those conditions, when could CR9AK be workable in New Jersey, and on what band? A propagation prediction using *Report 13* was performed (see **fig. 1**) and, much to my dismay, there was no opening above 13.8 MHz via the short path. Happily, though, there were two openings via long path, one at 2400 GMT (2400Z) with an MUF of

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14.4 MHz and a second one from 1100Z to almost 16002 with an MUF peaking to 21.7 MHz.

Unfortunately, that particular week of the DXpedition found the ionosphere very unsettled and nothing was heard on either opening the first three days of the DXpedition; the beam was pointed long path, while I listened on 20-meter CW. Little was heard from anyplace that week. On the fourth day, a very weak pileup was heard around 14025 kHz; it had to be CR9AK. When you are desperate you have to try, so a short call was given after the pileup stood by; he came back to me with a "surprised to hear East Coast." We exchanged signal reports, I was 459, and back to the faint (in New Jersey) pileup he went. Although CR9AK made amost 4000 contacts in 71 countries, few of them were with the East Coast. My working him was part luck, but also planning, as finding him would have been hit and miss without the prediction.

In April, 1976, Bill Rindone, WB7ABK, in cooperation with the NCDX Foundation, visited Christmas Island in the Indian Ocean, operating as VK9XX. Christmas Island is 3700 km farther from New Jersey than Macao. In addition, the Zurich sunspot number was lower than it had been in December — so low that I used 10 to allow me to use *Report* 13, Volume 2 directly. Again, when should I listen for him and on what band? A propagation prediction was made, (see **fig. 2**) and I had several choices, both short and long path on 20 meters. I'll take the short path anytime I can, as the distance is about 18,300 vs 26,000 km (9900 vs 14,000 miles) for the short and long paths respectively to Christmas Island. With a TH6DXX antenna at 15.2 meters (50 feet), the signal makes quite a few hops at even 18,300 km (9900 miles).

From **fig. 2,** you can see that the opening at 24002 is sharply peaked. I have not been very successful with that type of opening and concentrated on the

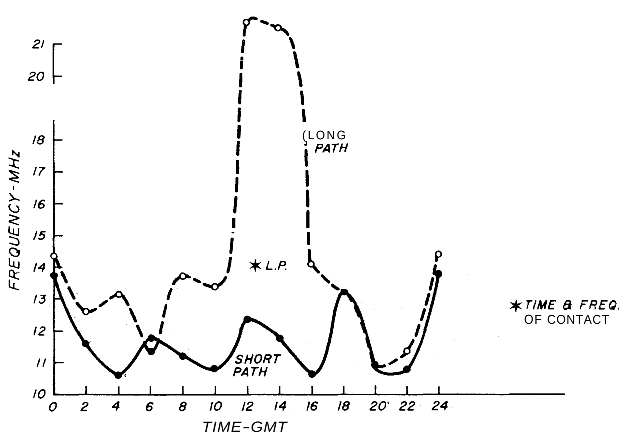

fig. 1. Propagation prediction for the long and short paths to Macao. The prediction showed that for December, 1975, there was no 20-meter short-path opening. However, the long path was open, even as high as 15 meters.

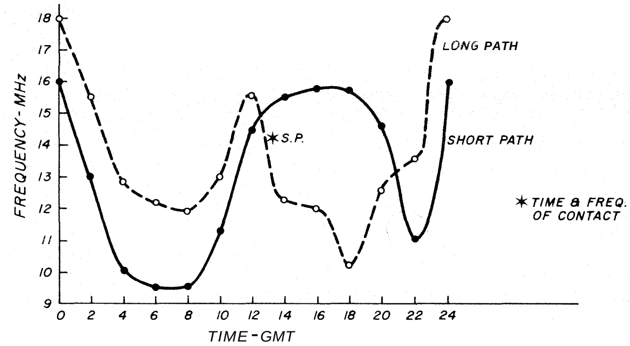

fig. 2. A similar propagation prediction for the path from New Jersey to Christmas Island in April of 1976.

opening at 1200Z. This worked out quite well — I got VK9XX shortly after the band opened on 20-meter ssb before the pack became aware of him. I left for work only 20 minutes after my normal departure time.

I know that many top DXers rely only on their experience, patience, and operating skill to snag elusive DX. But, when it comes to a DXpedition, you know he's going to be continually working. So, why not take advantage of all the miracles of modern science to get him!

Propagation predictions using *Report* 13 and reference 5 are easy to do, but they take a long time. You must first draw on transparent paper a great circle route between your home and the desired DX location and then establish an ionospheric control area distance at both ends of the great circle. The great circle is then over-layed on the MUF (Zero) F_2 map for the solar activity level, month and time desired, and the MUFs at the two recorded locations. The process is repeated using the MUF (4000) F2.

The F_2 MUF is then determined by means of a chart relating the (zero) and (4000) MUFs, which tends to decrease the MUF for all distances greater than 4000 kilometers. Next, the process is repeated for the MUF (2000) E, as some propagation may be by the E layer as well as the F_2 layer. Actually, the MUF (2000) E exercise is mainly for distances less than 2500 kilometers. By cranking the numbers and filling in the recommended work sheets, you arrive at an MUF for every two hours during the day. A plot such as **fig. 1** or **2** can then be drawn and analyzed. However, if the Zurich sunspot number is not 10, 110, or 160, which are Volumes 2, 3, and 4, the whole thing must be repeated and an interpolation made for the desired sunspot number. It's easy, it's all explained, but it is tedious. I'm sure the dogwork inhibits many from using the technique. Let me repeat: it's well worth it to give you a competitive advantage against the big guns.

Hang in there, though, if you're lazy like me. Pare the work to the bone. Being hams, we are highly op-

LATITUDE	22.0	PRT
LONGITUDE	-115.0	PRT
NAUTICAL MILES	7219.2	PRT
BEARING	343.1	PRT
KILOMETERS	13369.9	PRT

fig, 3. Example of the printout for the great-circle distance (both in nautical miles and kilometers1 and bearing to the desired DX station.

timistic when it comes to band openings. Why not ignore the MUF (Zero) F_2, which tends to decrease the MUF anyhow, forget the MUF (2000) E, unless you're looking for short haul, and record only the lower MUF of the two control areas, since that's the one you are going to use anyhow? That saves a lot of work. Unfortunately, you still have to go through the interpolation procedure untii the Zurich sunspot number hits 110, when you can use Volume 3 directly. That won't last long though, as the number will go either up or down.

making the prediction

The first thing you need for the prediction is the great circle distance (in kilometers) between your home station and DX location. Report 13 shows you how to do it graphically, but why not use the calculator? Other sources of theory include the articles by Marquart[6] and Hall[7]. If you have a Hewlett-Packard HP-55, Chester Brent, WB4GVE, has the program all worked out for you*.

The program for the SR52 provides the beam heading and distance in nautical and statute miles, as well as kilometers. (The kilometer value is automatically stored in memory 98 of the SR-52, a storage location not cleared except by removal of power.) The program is contained on a magnetic card, requiring only the insertion of the DX latitude and longitude, which are also printed out on the PCIOOA for reference purposes. The home location is part of the program, although a second location may be manually inserted. The equations for D, distance in nautical miles, and H, heading in degrees, are:

$$D = 60 \cos^{-1}[\sin L_1 \sin L_2 + \cos L_1 \cos L_2 \cos(\lambda_2 - \lambda_1)]$$

$$H = \cos^{-1} \frac{\sin L_2 - \sin L_1 \cos(D/60)}{\sin(D/60) \cos L_1}$$

where

L_1 and λ_1 are your respective latitude and longitude

and

L_2 and λ_2 are the respective latitude and longitude of the other station

When the problem is solved, bearing from true north, distance in nautical miles, and distance in kilometers are displayed and automatically printed on the PC100A tape. In addition, by pushing a button, statute miles may be displayed. **Fig.** 3 shows the printout obtained.

Since 200 of the 222 storage iocations were used for the bearing and distance program," another program is required for the prediction. The prediction program, also on a magnetic card, is inserted in the SR-52 without de-energizing the unit; it is necessary to retain the kilometer value stored in memory location 98. The program is then actuated and the PCIOOA prints out the number of hops required for the path, the reference hop length, and the ionosphere control area distance (see **fig. 4**). These equations are very simple:

$$\textit{Number of hops} = \textit{Great circle distance}/4000,$$

but increased to the next higher integer; *ie.,* if the answer is 3.3 hops, increase it to 4 hops.

The flow chart shown in **fig. 5** indicates the steps taken by the calculator to arrive at a whole integer for the number of hops.

$$\textit{Reference hop length} = \textit{Great circle distance/number of hops}$$

$$\textit{Ionospheric control area distance} = \textit{Reference hop length}/2$$

At this point the smoothed sunspot number is placed into the computer, because from now until 1990, we are going to have to interpolate. There are two buttons to be pushed on the SR-52 while extracting the data from prediction volumes. One will be called Label A for MUFs at the home location, and one Label B for MUFs at the DX location. Open two volumes to the same month, 2400Z and MUF (4000) F_2. The two volumes cover the sunspot numbers which are to be used for interpolation to the desired sunspot number.

The great-circle overlay is placed on the lower sunspot number chart first, carefully aligning the equator and the vertical line reference. I use a Greenwich meridien vertical line as well as a vertical line on the right border of the charts. The MUF under the home control area point is determined, inserted in the key-

NO. HOPS	4.0	PRT
HOP LENGTH	3342.5	PRT
CONT. AREA DIST.	1671.2	PRT

fig. 4. Printout of the hop length, number of hops, and control area distance as computed by the calculator.

"User instructions and complete coding forms may be obtained from *ham radio* by sending a self-addressed, stamped envelope.

board, and Label A pushed. The MUF under the DX control area point is next determined, inserted into the keyboard, and Label B pushed. The printer will print out the lesser of the two numbers, since the lower MUF is the controlling one for the path. A little time may be saved during this procedure. If the MUF at the DX location is higher than that of the home location, it will be discarded by your calculator. Therefore, rather than inserting that data into the keyboard, just push Label B and it will use Label A data which shows on the computer display.

Next, the great circle overlay is placed on the higher sunspot number chart and the above procedure repeated. The printer tape will now display MUFs for the lower and higher sunspot numbers. I have found it helpful to print these two numbers to verify that the interpolated number which follows appears correct. If a printer is not used, the numbers can be recorded from the display.

The program for this calculation is quite simple and may be performed on any hand calculator that has "If Pos" and "If Flag" features. I wanted to use the same labels for the minimum solar activity and the average solar activity (or the maximum activity). Label A is used for the home location MUF, and Label B is used for the distant location MUF. For any given solar activity we want to save for interpolation use the lower MUF for the high and low sunspot values.

Fig. 6 is the flow chart for the calculation. Check it out with actual numbers. From Volume2 (SSN of 10) the MUF is 18 MHz at the home control point and goes into the calculator as MUF A, from the same volume, 25 MHz at the distant control point, and goes into the calculator as MUF B. The difference is $18-25 = -7$, a negative number. Since it is not "If Pos," it goes to a flag check. Since the test for flag is

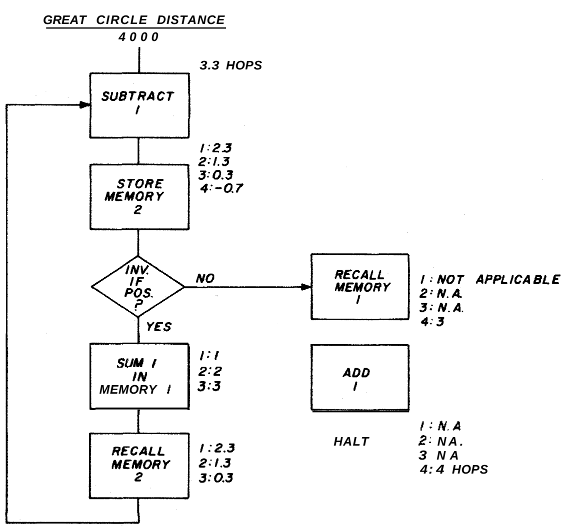

fig. 5. Flow chart for use with the TI-series calculators to determine the number of hops. Hewlett-Packard calculators with the INT function can perform this entire program in five keystrokes.

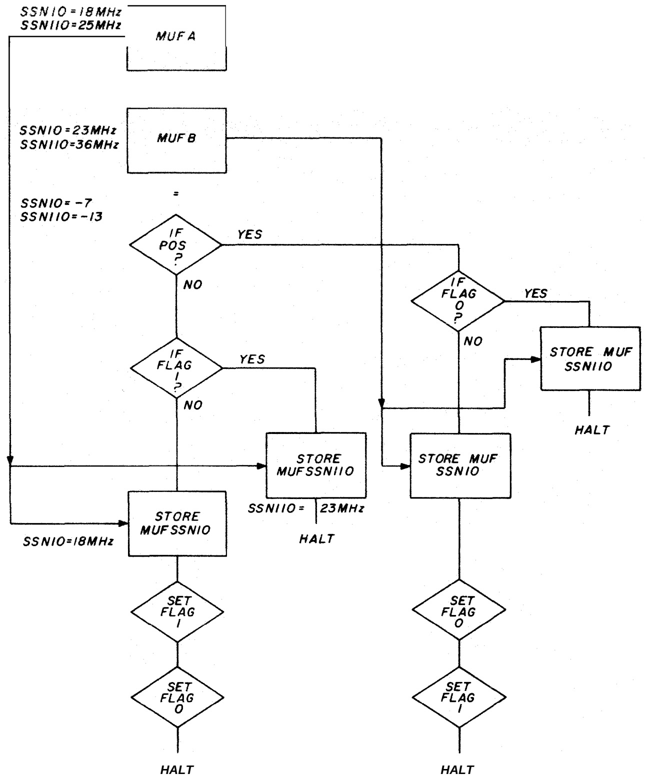

fig. 6. Flow chart to determine and store the lesser MUF value at the home or DX location. This information is needed to interpolate the MUF for the current sunspot number.

the first time through, none exists, and the proper MUF for SSN 10 is 18 MHz. The value is displayed and stored in memory. At the same time, two flags are set to prevent using this path for the MUF determination at SSN of 110.

Transferring to the SSN 110 chart, the MUF at the home location is 23 MHz and 36 MHz at the distant location. Again the "If Pos" check reveals a negative number and the calculator drops to "If Flag 1." Since that flag has been set, the value of 23 MHz is displayed and stored for interpolation purposes. If MUF A had been higher than MUF B, the calculator would have branched to "If Flag 0" with similar results.

To get the final interpolated answer for the hour used, press the Label D button. When that button is pushed, the calculator solves the equation:

$$MUF = MUF_{10} + 0.01(MUF_{110} - MUF_{10})(R_{12} - 10)$$

where

R_{12} is the current sunspot number

MUF_{10} and MUF_{110} are the MUFs at the low and high sunspot numbers.

When the sunspot numbers exceed 110, Volumes 3 and **4** will have to be used, and the program changed to use:

$$MUF = MUF_{110} + 10.02(MUF_{160} - MUF_{110})(R_{12} - 110)$$

Either of the interpolation equations can be programmed in your calculator or done by hand. With the programmable calculator, it is necessary only to recall MUF_{10} and MUF_{110} from memory and wait for the answers. It is at this point that the two flags have to be resei so thst the program can start over for the next time increment.

if a printer is riot used, the information is recorded from the calculator's register. When finished, you will have twelve points to plot for a 24-hour period, and the printout will be as shown in fig. 7. Plot them on graph paper using a suitable scale for the frequency to give a presentation such as shown in **fig.** 1 or **2.** You will now have a 24-hour short-path MUF presentation to your desired DX location. Altogether the procedure has taken 10 to 15 minutes.

Don't stop at this point, however; also compute the long paths. Some of them are rather exciting. Don't be confused by the referral in Volume 1 to short and long paths. When they say short path, they mean a short distance, like Washington, DC, to Ottawa, Canada. A long path to them is Washington, DC, to Berne, Switzerland.

The long-path computation is treated just like the short path computation, but with different numbers and control area points. Since you already know the short-path distance, subtract it from 40,000 km to determine the long path distance. For our purposes, 40,000 km is sufficient for the world circumference.

Now when using the program, the calculator will give long path information to establish the ionospheric control area distance. The long path great circle route and the new control area must be put on the transparent overlay. Before repeating the prediction procedure, it will be necessary to reinsert the sunspot number as that information was cleared from memory when the program was activated. The "clear memory" operation is automatically accomplished when the prediction program is actuated to eliminate stored information from being carried over from the bearing/distant program.

Repeating the prediction exercise will give twelve new points to plot on your graph for the long path.

SUNPOT NUMBER			SUNPOT NUMBER		
30.0		PRT	30.0		PRT
MUF	**TIME**		**MUF**	**TIME**	
12.0	**0**	PRT	12.0	**12**	PRT
21.0		PRT	14.0		PRT
13.8		PRT	12.4		PRT
10.3	**2**	PRT	11.2	**14**	PRT
17.0		PRT	14.0		PRT
11.6		PRT	11.8		PRT
10.0	**4**	PRT	10.0	**16**	PRT
13.0		PRT	13.2		PRT
10.6		PRT	10.6		PRT
11.0	**6**	PRT	13.0	**18**	PRT
15.0		PRT	13.8		PRT
11.8		PRT	13.2		PRT
10.5	**8**	PRT	10.0	**20**	PRT
14.2		PRT	13.5		PRT
11.2		PRT	10.7		PRT
10.1	**10**	PRT	10.0	**22**	PRT
13.5		PRT	13.8		PRT
10.8		PRT	10.8		PRT

fig. 7. Calculator printout for the MUFs during two-hour time increments.

Now, you have a complete picture and can establish your strategy and working hours to snag that elusive DX. Sure, you're still going to fight the pileups at times, but you will be there waiting for him with the beam pointed in the correct direction, and get him before many of the other guys wake up to his presence. Or, you'll be listening at a time and on a band which is most favorable to you.

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Volume 1, The Estimation of Maximum Usable Frequencies from World Maps of MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E, stock number 03000318, price $0.30.

Volume 2. Maximum Usable Frequencies MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E for a Period of Minimum Solar Activity, R_{12} = 10, stock number 03000319, $3.00.

Volume 3, Maximum Usable Frequencies MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E for a Period of Minimum Solar Activity, R_{12} = 110, stock number 03000320, $3.00.

Volume 4, Maximum Usable Frequencies MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E, for a Period of Minimum Solar Activity, R_{12} = 160, stock number 03000321, $3.00.

Write to the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402.

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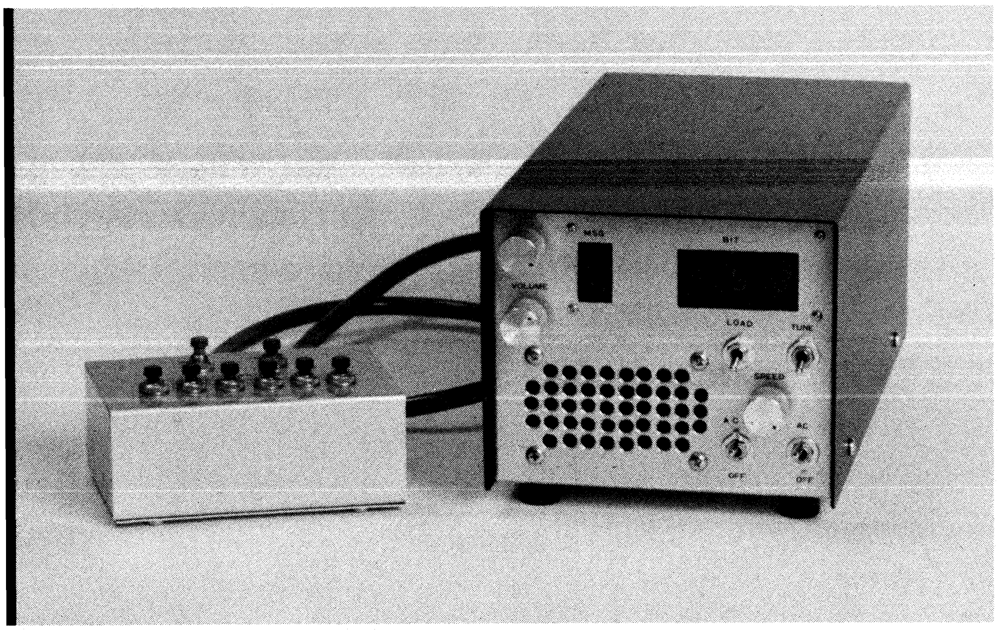


deluxe memory keyer

with 3072-bit capacity

A revision of the famous WB4VVF Accu-Memory, featuring improved automatic character spacing, six-message memory, and remote control

A memory keyer with the right characteristics offers many advantages in contests, net-control operations, and DX chasing. It can relieve an operator of much of the routine repetitive sending, generally reducing fatigue and providing time for efficient log keeping, checking for duplicate contacts, and occasionally taking a sip of coffee, even during rapid-fire contest exchanges.

The Accu-Keyer,[1] as originally described in QST, was a nonmemory keyer with many desirable features. These included self-completing dots and dashes, next-dot and next-dash memories, iambic operation, and optional automatic character spacing (ACS). Later, WB4VVF and W4YUU described the companion Accu-Memory,[2] which, when used along with the Accu-Keyer, permitted storage of up to 2048 bits (about 200 CW characters) keyed in from the keyer. These could be read out as four individually selectable messages of up to 512 bits each, or as one or more longer continuous messages.*

*The term bit is used throughout as the element of Information contained in a dot or a space in a CW character. A bit interval is the time duration of a single dot or space. A dash is three bits long, the standard space between characters is three bits, and a normal word space interval is seven bits long.

By Robert C. Cheek, W3VT, c/o Ebara-Infilco Co., Ltd., 1-1-1 Hitotsubashi, Chiyoda-ku, Tokyo 100, Japan

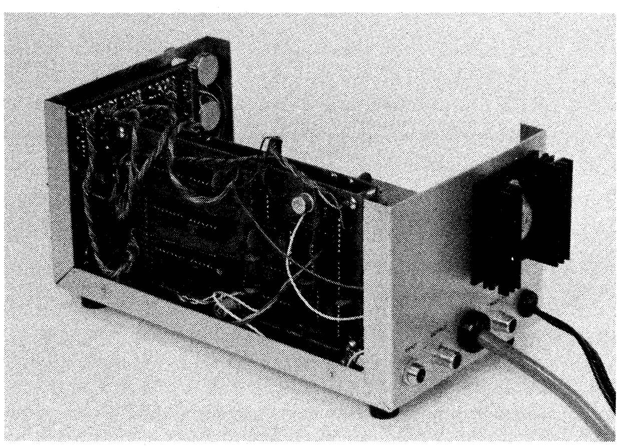
I built a modified version of the Accu-Keyer with the Accu-Memory, incorporating those modifications I thought desirable. My own experience with it, and the comments and questions of others, indicated that further improvements and enhancements could be made. I have now gone through the redesigning and building process five times, with the memory keyer described in this article the final result.

The memory capacity of this keyer is 3072 bits, in six selectable, 512-bit segments or in longer segments up to the full capacity of the memory. The entire system, except for the power supply and the display LEDs, is contained on two circuit boards measuring 80 by 150 mm (approximately 3-1/8 by 6 inches). ICs are used throughout, except for the output, which employs a keying transistor. When the memory is being loaded, the IC clock is synchronized to manipulator movement. Readout from memory may be automatically interrupted at any time by manual keying. A choice of keyer output circuitry to accommodate either grid-block keying or positive-line keying is provided for in the circuit-board layout. A cable-connected, remote-control unit is used for manual start, stop, and message selection functions. It can be conveniently placed on the operating table to be played with the left hand like the buttons on an accordion while the right hand makes log entries and operates the keyer paddle for station calls, nonstandard message insertions, and so on.

system description

A functional block diagram of the complete system is shown in **fig. 1.** Those who wish to follow the explanations in detail should also refer to the schematics of the two main boards, **figs.** 2 and **3** for the keyer-driver-monitor and memory boards, respectively. This explanation assumes a knowledge of the specific IC characteristics and truth tables given in IC data manuals.

Rear view of the six-message memory keyer showing the driver-monitor board; an adequate heatsink must be used on the 5-volt regulator.



Keyer. The keyer proper (included in **fig. 2**) is basically the well-known Accu-Keyer, with all of the original features retained. Keyer operation was explained in the original article[1] and will not be repeated here, except for those portions of the circuit that have been changed.

The original clock pulse generator, made up of discrete transistors, has been replaced by an NE555 IC clock located on the memory board. As will be explained later, the clock output approximates a square wave rather than sharp pulses. This characteristic is used to greatly reduce the error rate in manual keying when the ACS feature is in use.

The ACS circuitry inserts precision, three-bit spaces between characters, correcting premature timing of one character following another in the manipulation of the keyer lever. Without ACS, this shows up as a shortening of the proper three-bit character interval. For example, DE may sound almost like B if the E is keyed too quickly after the D.

However, the original ACS feature is difficult to use at higher speeds without errors, which show up as unintentional extra spaces. This happens because lever manipulation must not only be fast, but must actually slightly lead, or at least never lag behind, in the formation of characters in which a dot is followed by a dash, such as an A. Those of us who grew up with the bug or semi-automatic key tend not to lead while keying, but to insert our own approximately correct one-bit interval between the dot and the dash in the formation of such characters. But, if the dash lever is tapped just a fraction of a bit space too late, an A becomes ET in the original keyer when ACS is in use. This occurs because the outpin at pin 6 of U5A (see **fig. 2**) is always low during the one-bit space following a completed dot or dash, and then always goes high for the next one-bit interval. It is this high transition that transfers a waiting dash or dot from the next-dash or next-dot memory, U1A, U1B, or U2C, U2D, as seen through the iambic gates, to the present dash or dot memory, U3A or U3B, starting the output of the dash or dot. If the lever is tapped even very slightly late, this transition is missed and two additional bit spaces are inserted.

U8 in the keyer has been added to provide some

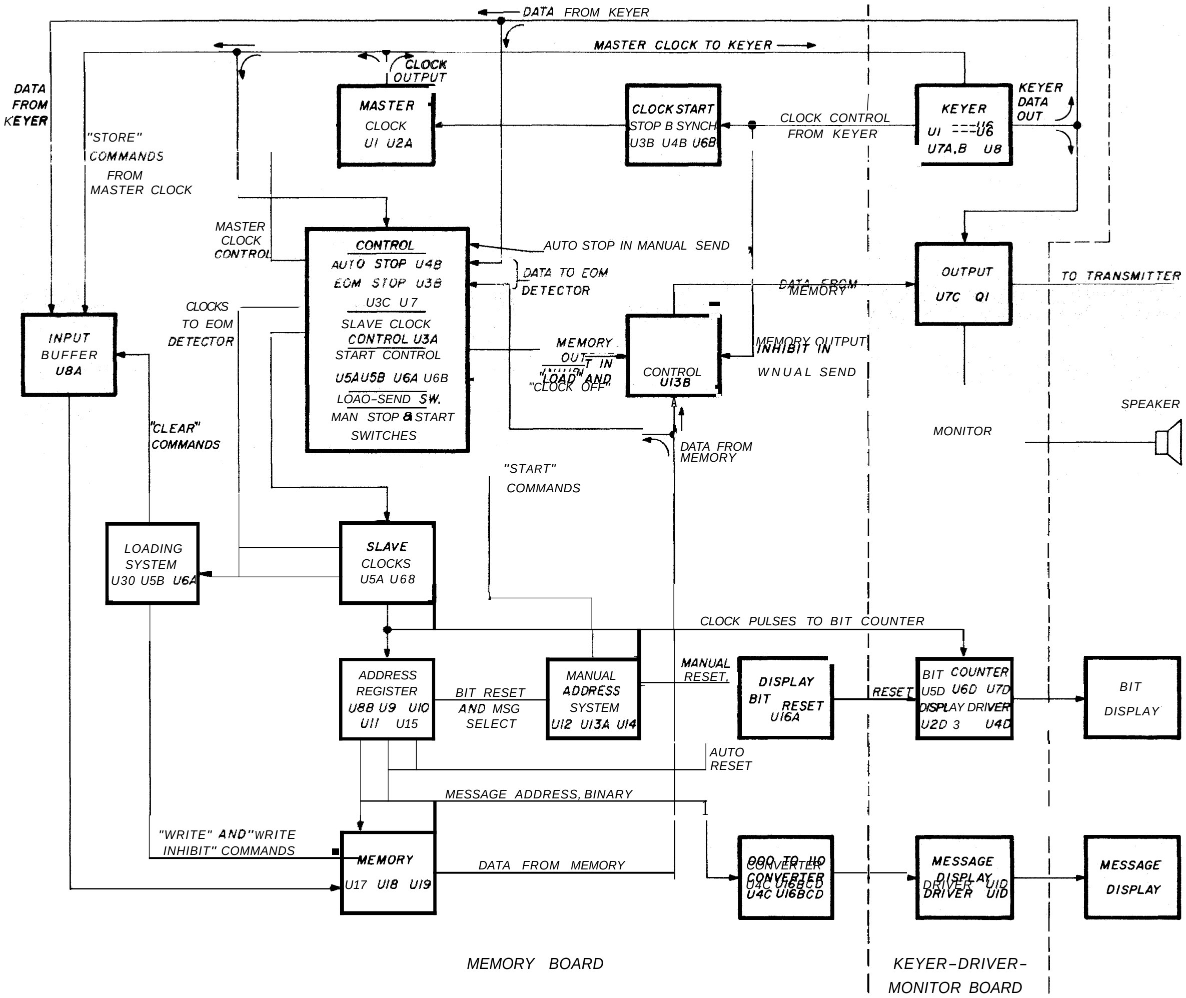


fig. 1. Functional block diagram of the memory-keyer system. Details of the keyer block are not shown; these are given in reference 1.

tolerance for slight manual timing delays in the keying of dashes following dots as described above. The approximate square wave from the clock is high at pin 1 of U8 for the first half of the bit interval after pin 6 of U5 goes high. If neither a dash nor a dot has already been called for, pin 8 of U6 will also be high. If the dash paddle is tapped during this half-bit period, U1D, pin 11, will go low. This low, inverted by U8B, will cause U8A, pin 6, to go low because all four inputs are high. The low will set U3A, pin 6, high and start the dash, although a little late. However, even though the interval between the dot and the dash may be lengthened by as much as half a bit and the dash correspondingly shortened by this amount, the result is almost undetectable to the ear and is far better than the three-bit interval that otherwise occurs, causing a completely erroneous character to be formed. Also, since data as it exists at the *end* of the bit interval is what is recorded into memory in the loading process, the correct one-bit interval between the dot and the dash is restored when the message is sent from memory.

The keying of a dash following a dot *after* the half-bit tolerance interval provided by U8 results in the insertion of the normal three-bit character interval, retaining the desirable features of automatic character spacing. At the beginning of the third-bit interval during a character space, pin 6 of U5A again goes low when the clock pulse starts. The RC filter at pin 1 of U8 delays the clock pulse slightly so that pin 6 of U5A has time to go low before the clock, as seen by U8A, goes high. This prevents a shortened dash from prematurely appearing if the dash lever has by that time been tapped for the next character.

By adding another IC equivalent to U8, it would be possible to provide the same tolerance for manual keying of a dot or dots following a dash. However, nearly all the tendency for errors at high speed is the result of the very short and critical interval for the keying of a dash following a dot. Since the available interval is inherently much longer going the other way, the extra complication would not be worthwhile.

Clock. The clock, as shown in **fig. 4,** is the heart of the keyer and the memory. Both require a clock signal which is at zero level in the quiescent state, remains at zero for the duration of the first bit interval after the clock is called for, and then provides a steeply rising positive pulse or square wave at the beginning of each bit interval thereafter. With the control scheme adopted, the combination of the NE555 timer and an inverter provides the necessary characteristics, including the square wave required for the ACS error-reducing circuit in the keyer.

Pin 4 of U1, which is normally used as the control input in NE555 applications, is maintained high at all times, keeping pin 3 normally high, with the clock output being low. When pin 11 of U3B is low, calling for the clock to be off, the timing capacitor is essentially discharged through CR1. Pin 3 of U1 remains high and the output of U2 stays low. When the clock is called for by either the keyer or the memory, pin 11 of U3B goes high and the timing capacitor begins to charge at close to a linear rate. The output from the inverter remains low during this process. When the charge reaches 2/3 of V_{cc} (the supply voltage), pin 3 of U1 goes low, forcing the inverter high, marking the end of the first bit interval. Thereafter, the charge on the timing capacitor oscillates between 1/3 and 2/3 of V_{cc}, and the output at U2 goes high at the end of each subsequent bit interval until pin 11 of U3 again goes low, stopping the clock.

CR1 and CR4 permit independent discharge of the timing capacitor for clock synchronizing purposes, as will be explained later.

Memory system (load mode). The basic memory loading system of the Accu-Memory2 is essentially retained. In the load mode, with a message address manually selected, the clock control line from the keyer goes low at the moment keying begins and pin 11 of U3B (see **fig. 3**) goes high, starting the master clock. At the same time U6B, pin 1, goes low, causing a pulse at pin 13 of U6. This pulse is passed along by U5B and then appears at pin 5 of U6A, from where it is applied to pin 3 of U7, the end-of-message detector. This pulse at pin 3 performs an AND function with the data at pin 2 of U7, setting the output at U7 pins 1 and 11 to zero, whereupon pin 8 of U3C will go high and pin 12 of U2B low, keeping pin 11 of U3B high. The high at U3C pin 8 also enables slave clock control U3A. Pins 9 and 10 of U3C are both high only when U7 is at a count of nine. Thus, even though the clock control line from the keyer may go high, U7 will keep the clock running until it counts nine consecutive pulses at its pin 14 from slave clock U5A, pin 5, without being reset by the pulses at pin 3 coincident with data at pin 2.

At the end of the first bit interval, the leading edge of the master clock pulse transfers the data at the D input of the 7474, U8A, pin 12 to U8A pin 9. Halfway through the second bit interval, the trailing edge of the first master clock pulse, having been inverted by U3A and applied to pin 10 of U5A and U6A, causes positive pulses at both Q outputs. The write inhibit, provided by gate U3D, is removed by the pulse from pin 5 of U6A, and the first-bit data is recorded into the memory. The pulse from U5A is somewhat longer in duration. Its negative-going trailing edge, occurring later, advances the address register by one, and at the same time the positive-going $\overline{Q}$ output at pin 12 of U5A causes pin 4 of U5B to pulse low, clearing the buffer, U8A, to zero. At the same time, pin 13 of U5B pulses high, and its subsequent trailing edge causes another positive pulse at pin 5 of U6A, again removing the write inhibit and recording the previously mentioned zero from U8A in the second memory address. All of this takes place during the latter part of the second-bit interval.

At the end of the second-bit interval, the loading process is repeated for the second data bit, and the zero previously stored in the second address is changed to whatever the contents of the data buffer may be for the second bit. If there is a pause in keying, the end-of-message counter U7 will stop the master clock when it counts nine consecutive bits without being reset by pulses from U6A along with the data. The ninth bit will be changed to a one if keying is resumed, and the pause, no matter how long, becomes a word space (slightly lengthened to eight bits) when the message is read our in the send mode. However, if the message has really ended and no additional data is loaded, the zero remains in the ninth bit position, and in the send mode U7 will stop the clock at that point.

The master clock free runs during the word intervals in the load mode, and it is necessary to synchronize the clock to the paddle movements during loading so that shortened dots and dashes do not appear, as they would if the paddle were tapped to start a new word just before a clock pulse was about to occur. Synchronization is accomplished by U6B, U4A, and U2C. When the paddle is tapped during loading, the clock control line from the keyer goes low, calling for the clock (which normally is already running). This low, applied through CR5 to pin 1 of U6B, causes the $\overline{Q}$ output (pin 4) of U6 to pulse low. This pulse forces the output of U4A high and U2C low. The timing capacitor is momentarily discharged through CR2 by this pulse and then allowed to charge normally, ensuring full bit intervals for the

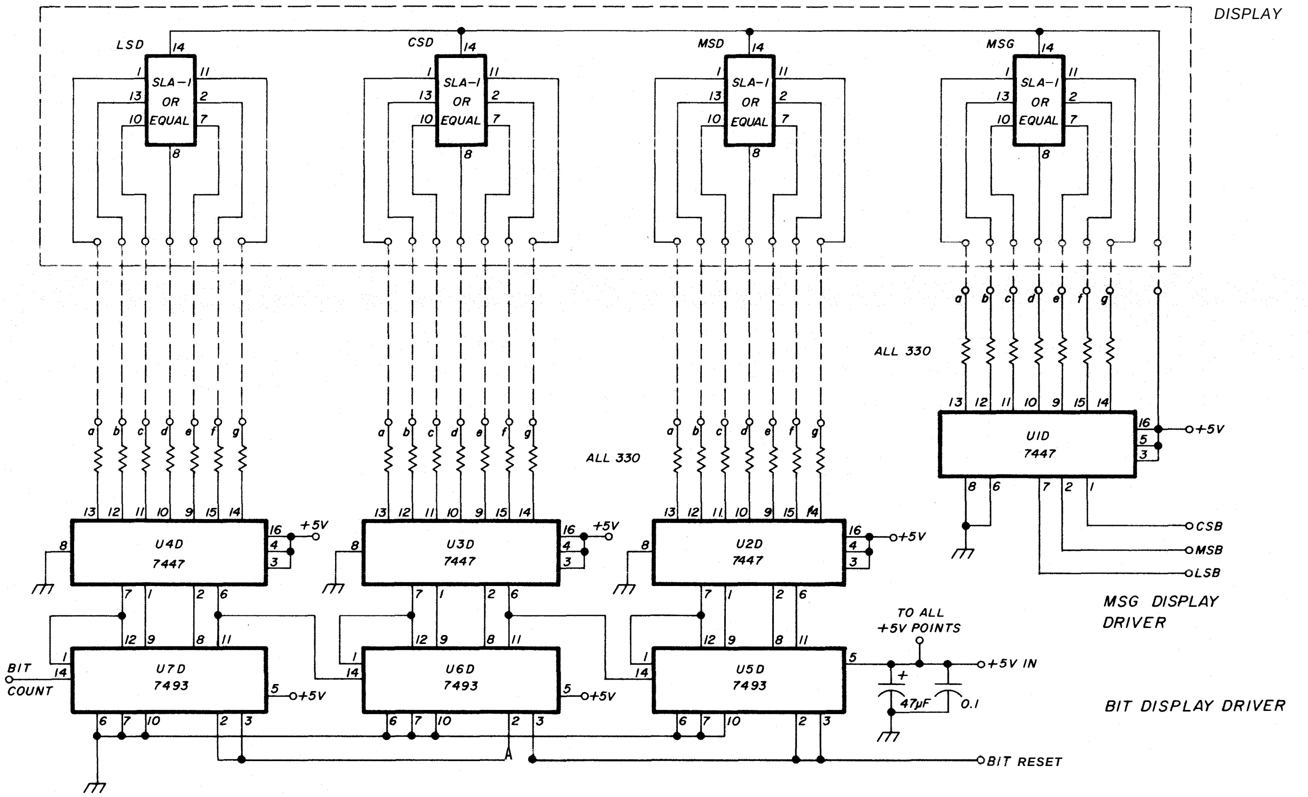


fig. 2. Schematic of keyer-driver-monitor board and display. All resistors are ¼-watt. Capacitors with polarity indication are upright electrolytic, 15-volt minimum rating; all others are disc ceramic, 25-volt minimum rating unless otherwise specified. The diodes are small signal silicon, switching diodes, 1N4148 or equivalent. Dotted lines indicate connections to external switches and controls. Small circles with labels, except +5V to ICs, indicate similarly labeled points on the board or power supply. In the output circuit for negative grid-block keying, Q1 may be a 2N4888 or any high-voltage audio, switching, or rf silicon pnp transistor with a V_{CEO} rating adequate for the open-circuit voltage to be keyed. The unit pictured uses a Japanese type 2SA510 (RCA SK3025 is also a recommended equivalent). For positive-line keying, an equivalent npn transistor is used at Q1.

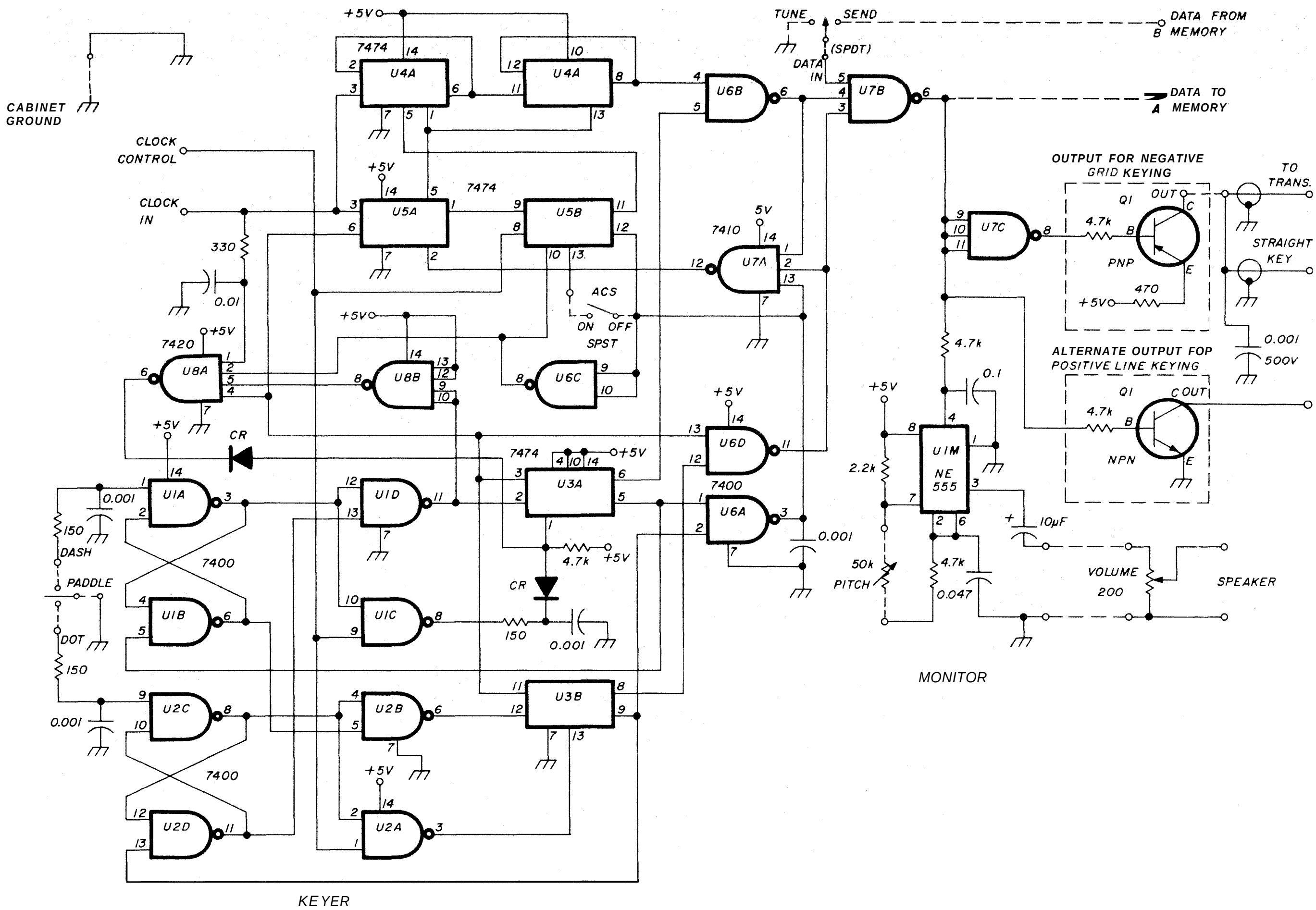
first bit and subsequent initial bits of manual sending.

In the load mode, the load/send switch grounds pin 12 of U13, preventing any output of random or previously stored data in the memory from reaching the output data line. While loading is progressing, only the new data from the keyer can key the transmitter.

Address register and manual addressing system. The address register consists of U9, U10, and U11A for the lowest nine binary address bits (512 of the 1024 positions on each memory chip). The tenth address bit, determined by U11B, becomes the least significant message address bit. Thus, U11B divides each 1024-bit memory chip into two 512-bit message segments, selected by making the output from pin 9 either a zero or a one. U8B, U15A, and U15B constitute a shift register or ring counter which activates one memory chip at a time for loading or readout by placing pin 13 of the selected chip low, while maintaining pins 13 on the other two chips high.

The entire address register is controlled by the manual addressing system, consisting of U12, U13A, and U14. When a message button is pushed (assuming the clock is not running) pin 6 of U13 goes high, setting U9, U10, and U11A to zero. Pin 9 of U11 is set either low or high, depending upon the message selection, and the Q output of one of the three flip-flops in the ring counter is set low to address the desired memory chip. The other two Q outputs in the ring counter are simultaneously set high. Whenever a manual message selection is made, the display bit counter is simultaneously reset to zero by U16A.

Since memory-chip selection is made by setting one of the Q outputs in the ring counter low, the $\overline{Q}$ outputs of U15A and U15B are used as the central significant bit and the most significant bit, respectively, of the message address. When both these Q outputs are high ($\overline{Q}$ outputs both low), the memory chip that is activated by a low from the Q output of U8B is in use, and the message address is either 000 or 001 binary (decimal 0 or 1), depending upon the



Q output at U11B. The other binary message addresses are 010, 011, 100, and 101 (decimal 2, 3, 4, and 5). Since it is better in operation to have the six messages numbered 1 to 6 rather than 0 to 5, U4C and U16B, C, D were configured to convert 000 binary to 110 binary, leaving the other binary indications unchanged. Thus, message 0 becomes message 6 for all selection, display, and operating purposes. The entire addressing system is a continuous counter, so that a message as long as the entire 3072-bit memory capacity can be started at any selected beginning address and loaded continuously in memory, with automatic transition through the consecutive message segments back to the starting address.

Send Mode. Before a selector button is pushed, the clock is stopped by a low at pin 11 of U3. This low also inhibits U13B, blocking the continuous output that would otherwise appear if the address register happens to be stopped at a memory address where a high exists. When a selector button is pushed, the high at U13, pin 6, which resets the sit address register to zero, is inverted by U2D, simultaneously causing a low at pin 1 of U6B. The resulting pulse at U6, Pin 13, triggers U5B which in turn causes a pulse at pin 5 of U6A.

A high always exists at memory bit address 000 when a message segment is loaded. This high is applied to pin 2 of U7, and the pulse from U6A is applied to pin 3. The AND function at these two points resets U7 to zero, unblocking the memory output and starting the clock. The high at pin 11 of U3B also causes U13A to return to low, even if the message selector switch is kept closed, removing the reset on the address register. Slave clocks U5A and U6A are enabled through U3A when U7 resets. U5A advances the address as described for the load mode. During each bit interval, a pulse from U6A is applied to U7, resetting U7 whenever data appears at pin 2 of U7. Unless the stop button is pushed, sending from memory continues until U7 counts the nine consecutive bits with zero data which mark the end of the message.

Automatic Stop. A message readout in progress

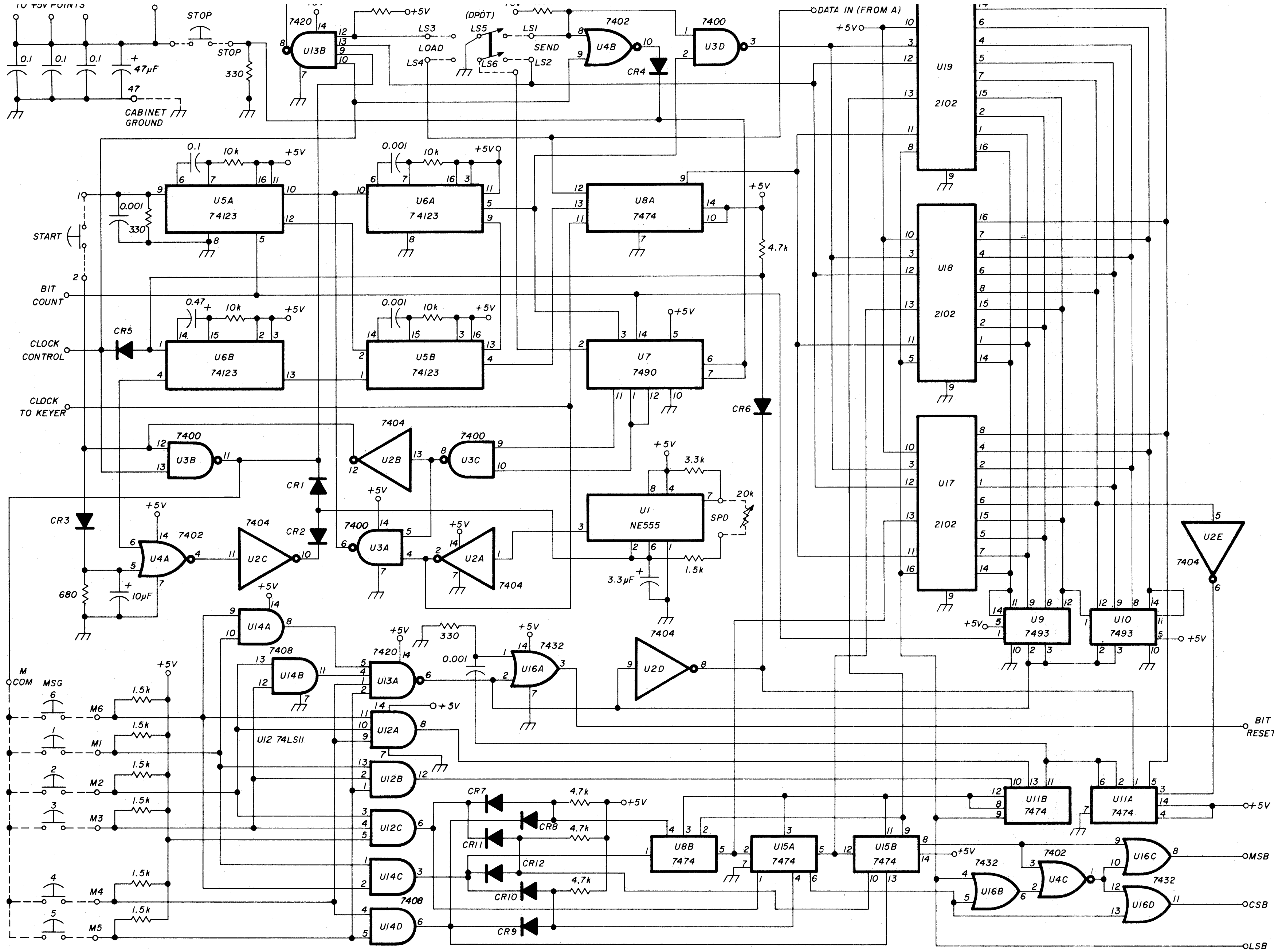


fig. 3. Schematic of memory board. See fig. 2 for explanation of component designations and external connection symbols. All diodes are small signal silicon switching diodes, 1N4148 or equivalent. START, STOP, and message selector switches are all momentary push-button type.

can be interrupted by manual sending. At the first tap of the paddle, the clock control line from the keyer will go low, forcing U4B pin 10 high. This high, applied through CR4 to U7, forces a nine count, instantly turning control of the master clock over to the keyer and stopping the input to pin 10 of slave clock U5A. Meanwhile, the low on the keyer clock control line causes U6B, pin 4, to pulse low. This pulse causes U4A to pulse high (provided pin 5 is low) and U2C to pulse low, synchronizing the master clock to the first bit of manual sending. The forced nine count at U7 causes U3C to go low and U2B to go high, but the 10-μF capacitor at pin 5 of U4 keeps that pin momentarily low, allowing the synchronizing pulse to pass. Thereafter, the capacitor charges, U4A stays low and U2C high, and the clock is under normal control of the keyer.

Power Supply. The system requires 700 to 800 milliamperes at 5 volts. A schematic of an adequate 5-volt, 1-ampere supply using an LM309K regulator is shown in **fig. 5.** Based upon its specifications, the TTL 5-volt, 1-ampere regulated supply kit offered by Jameco should be equally suitable.

construction

Assembly of the circuit boards* **(figs. 6, 7,** and **8)** offers no unusual problems, but the usual cautions apply — use a low-wattage iron and small diameter 60-40 solder; be sparing of solder, but cover the eyelets completely; above all, make sure there are no unintentional solder bridges between IC socket terminals or between adjacent separate conductors in the vicinity of component terminals.

The use of IC sockets or Molex pins is recommended in preference to soldering the ICs directly to the boards. There is a big difference, however, among IC sockets, so be sure to use the compact, slender type rather than any of the more bulky types, as the latter may cover up a few of the component or jumper eyelet holes adjacent to the ICs. The digital display LEDs are intended to be soldered directly to the display board. If you prefer to use sockets, the unused socket pins must be clipped off to permit their mounting in the holes provided.

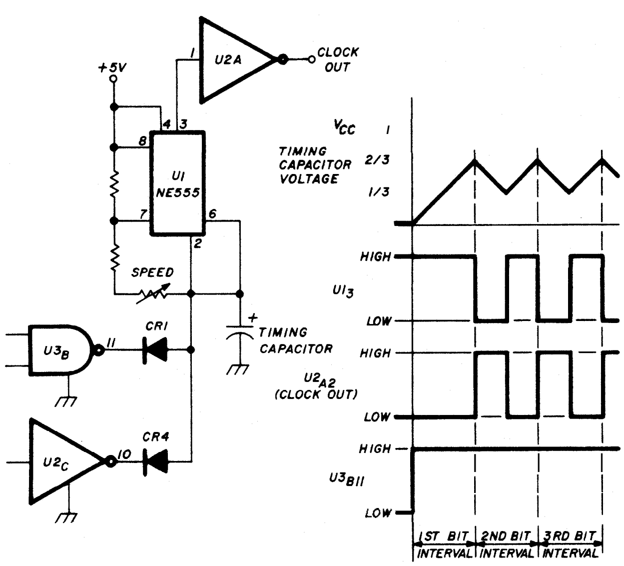
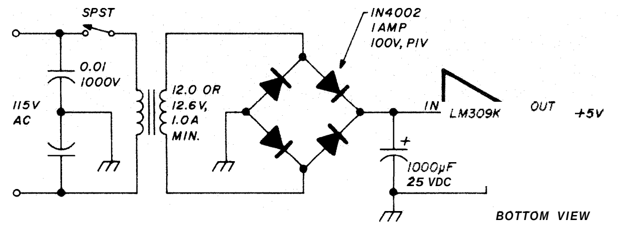


fig. 4. Diagram of the basic clock circuit and clock control logic.



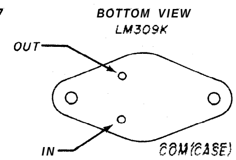


fig. 5. Power-supply schematic. The TTL 5-volt, 1-ampere supply kit offered by Jameco should also be suitable.

Some of the wire jumpers on the memory board must be routed around the IC sockets. Mount the sockets first, and then install the jumpers before mounting the other components. All jumpers are installed on the component sides of the boards. The short, isolated, straight jumpers can be made of bare wire to save space, but obviously the longer ones and those which cross over uninsulated jumpers or come close to other component leads should be of insulated wire.

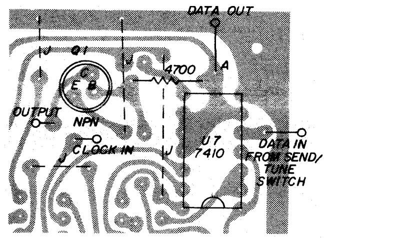
For interconnections between the boards and leads to be connected to external switches and controls, use small-gauge flexible (stranded) wire to prevent strain on the circuit board eyelet terminals. In making the main board-to-board interconnections, lay the two boards side by side, foil sides down, with the top edges adjacent to each other about **4** cm (1-1/2 inches) apart. Recheck this spacing each time you cut a new lead to be sure its length is sufficient to permit the leads to come over the top edges of the boards when they are finally installed.

All connections to the main boards are made with leads inserted from the component side. In the case of the display board, however, all connections are inserted into the eyelets and soldered from the *foil* side, with the excess wire clipped off flush on the component side. In this way, the leads do not interfere with the mounting of the LEDs against the transparent, colored plastic strip that covers the inside of

*A complete set of circuit boards can be obtained from G. R. Whitehouse. For more information on the circuit boards and other components, write to G. R. Whitehouse and Company, Newbury Drive, Amherst, New Hampshire 03031.

the front-panel cut-outs. The use of color-coded wire will permit the seven leads for each seven-segment LED and the one +5 volt lead to be twisted together before final soldering, providing a much neater job than the mess that will result if each of the twenty-eight wires is run separately between the keyer-driver board and the display.

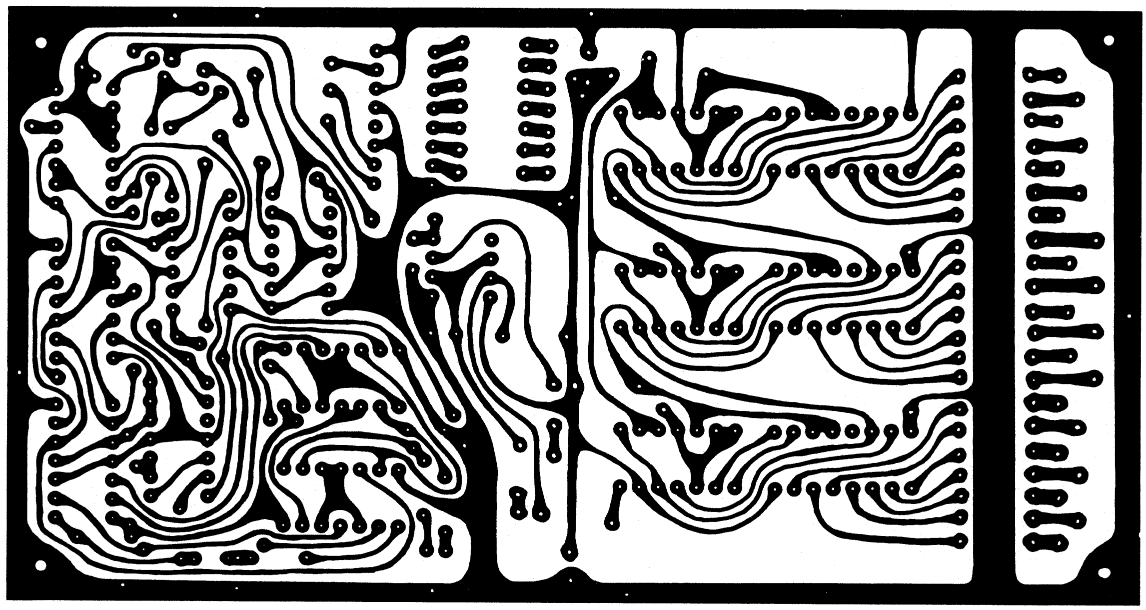
After assembly and interconnection, the two main



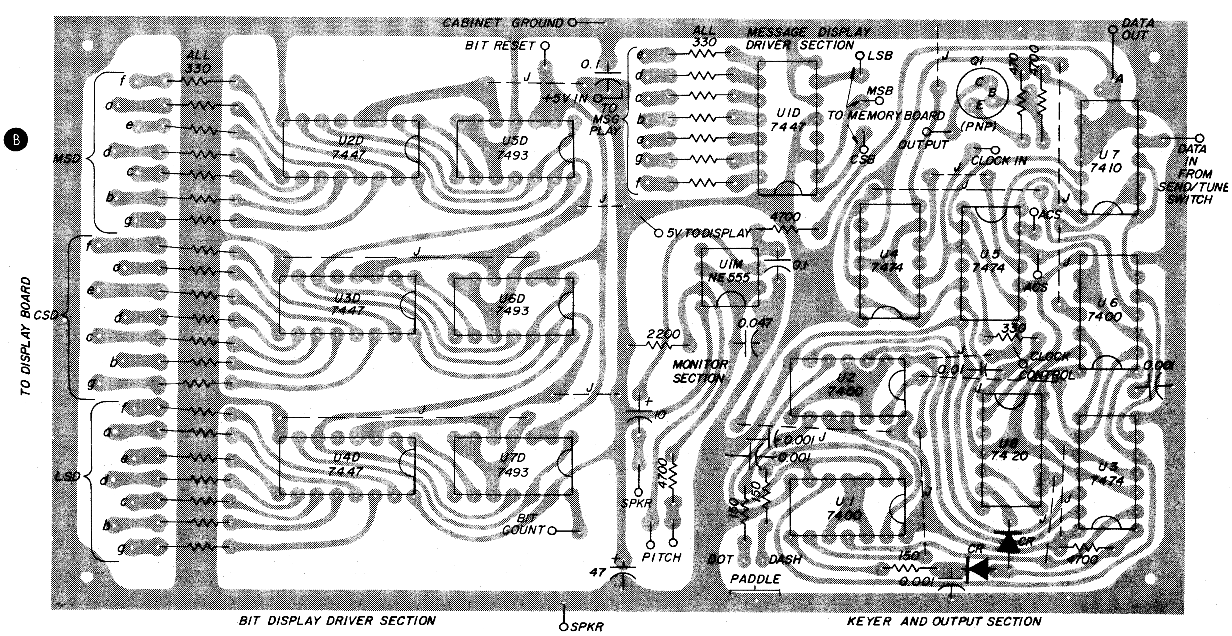
C

fig. 6. Circuit pattern of keyer-driver-monitor board (A), and component placement for negative grid-block keying (B). Fig. 6 (C) shows alternative component placement for positive-line keying.

A



B



circuit boards are mounted to each other with foil sides facing, separated by 9-mm (318-inch) cylindrical spacers at each corner. This assembly is mounted to the bottom of the cabinet with four L brackets, one on each side of the bottom corners of the board assembly. This makes a very solid unit, easily accessible for replacement of ICs. The board assembly is mounted on the right-hand side of the cabinet to make room for the power supply at the left rear.

The only precaution worth mentioning in connection with the power supply is to use an adequate heatsink with the LM309K regulator, which is mounted on the rear panel outside the enclosure. The regulator output terminal, projecting inside, will serve as the +5 volt source terminal for the leads to the main boards and the +5 volt lead to the stop switch. Note that both sides of the ac line cord should be bypassed to ground near the point at which it enters the cabinet.

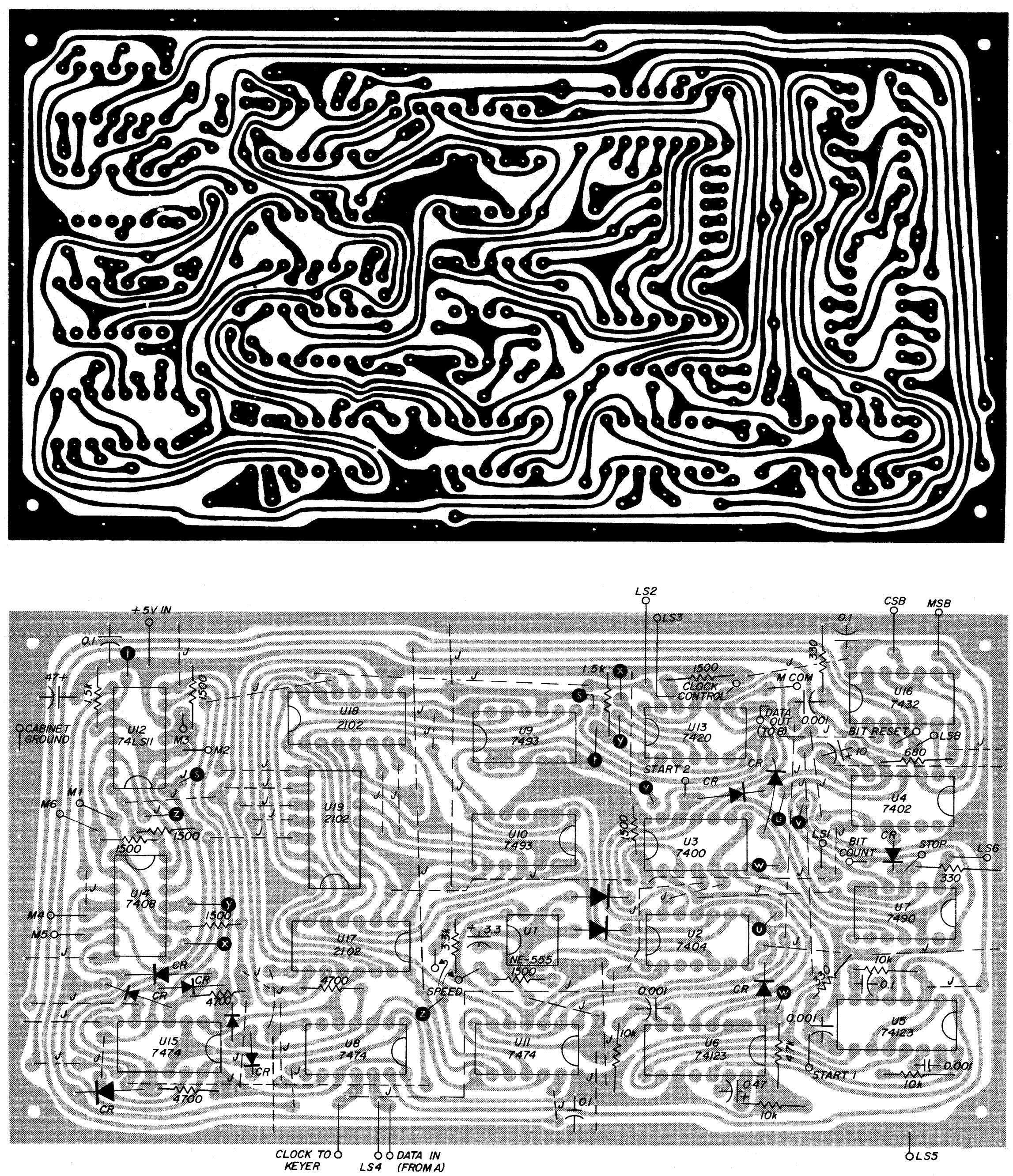


fig. 7. Memory-board circuit pattern (above) and component placement (below). For clarity, several jumpers were omitted from the parts placement diagram. These can be added to the board by connecting the jumper between the lettered designators.

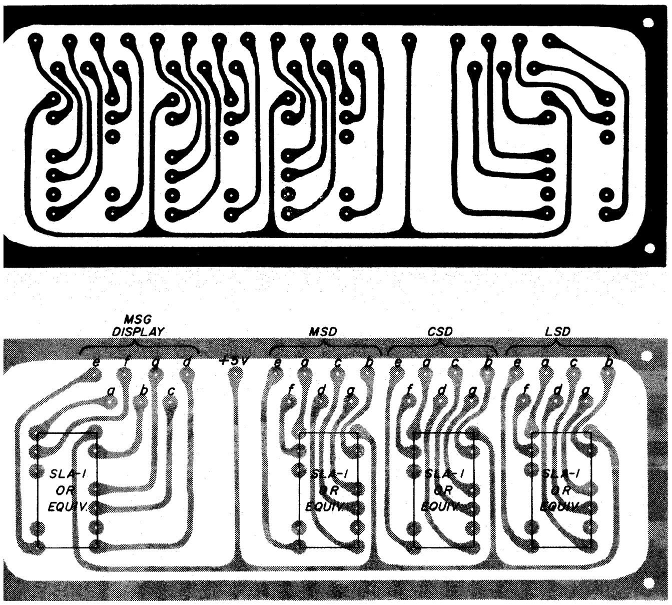


fig. 8. Display-board pattern (above) and LED placement (below). The display LEDs may be SLA-1 or any of several equivalent 8-mm (0.3-inch) common-anode displays with 14-pin DIP pin spacing and the same pin-out configuration for the seven segments. Decimal points are not used and their pin-out locations are not important.

RF shielding of all external leads to the main boards and the use of an all-metal cabinet are absolutely necessary. The leads to the remote-control box (which also must be a metal box) should consist of a shielded cable (11 conductors plus shield are needed) with the shield grounded both in the box and in the main cabinet. The keyer-paddle leads should also be shielded. A two-conductor shielded cable, with the shield used as the ground lead, will be satisfactory.

One last precaution — be sure to insert the correct ICs in the correct locations with the positioning notches in the proper direction, and then recheck each one before applying power. This is very obvious, but I've ruined a few ICs myself by being a bit careless on this point.

initial tests

After assembly and wiring are completed and checked, set the tune/send switch and the mode switch to send, all controls at about two-thirds scale, and apply power by turning on the ac switch. If some gibberish comes out of the monitor, it is a good sign. Pushing the STOP button should stop the output. Check that the output terminal of the regulator is +5 volts. Each display LED should show some digit at this point, although the actual indication is not important.

Next, check the clock, keyer, and monitor for proper operation as follows: Alternately ground the dot and dash inputs to the keyer. Continuous dots and then continuous dashes should be heard from the monitor. Grounding both inputs simultaneousiy should produce alternating dots and dashes. Switching the tune switch to TUNE should produce a continuous tone. Return it to normal. Now set the mode switch to LOAD and press each message selector button in turn. Bit count indication should be 000 and message numbers 1 through 6 should be displayed in sequence as the associated buttons are pushed. Pushing the START button should cause the bit address to advance by ten bits and then stop.*

Again selecting message 1, ground the dot input to the keyer, causing it to send continuous dots. The bit indication should advance as these dots are loaded into memory and after bit 511 the message address should change to 2 000. Continue loading dots part way into message 6, stop the keyer by ungrounding the dot input, set the mode switch to SEND and again select message 1. The entire series of dots should play back on the monitor without errors or omissions, and it should be possible to stop the playback at any point by pushing the STOP button or by momentarily grounding the dot or dash input of the keyer. Next, in the LOAD mode, again select message 1. While holding the STOP button down, push the START button down once and release it to get a bit count of 001. (Because of contact bounce it may be necessary to try several times to get just a one-bit advance when the button is pushed.) Load continuous dots from 1 001, again part way into message 6. In the SEND mode, select message 1 and this time push the START button to start playback. Again, the whole series of dots should play back smoothly. These two tests will indicate whether all memory cells in this address range of the memory ICs are okay. Perform similar tests by loading dots beginning at addresses 6 000 and 6 001, continuing into message 1 in each case, to check the remaining range of memory addresses.

using the keyer-memory

The memory may be loaded with a message starting at any selected message address. Simply switch to the LOAD mode, select the message number, and start keying. When the message is completed, another may be begun at a different message address until all beginning addresses are used. (If a

*This may seem contrary to the previous statements that U7 stops the clock after counting nine bits with zero data. Actually the START button causes the address register to advance by one bit at the start of the first bit interval as the count at U7 goes from 9 to 0. U7 then counts nine additional bits and stops the clock message is too long to fit into a 512-bit memory space, it will continue into the next message address.) In the SEND mode, pushing the message selector button for a particular message will instantly start transmission of that message from memory. In contest work, a CQ SS or CQ DX can be loaded as message 1, with several different replies (different reports, etc.) as the next few messages, and *R TU QRZ . . .* etc, as message 6. For contests in which the exchange requires a variable element such as a QSO serial number, load the standard information up to the point where the variable element must be inserted. Then, holding the stop button down, advance the memory two or three bits by pushing the start button. Continue loading the subsequent standard information. In the SEND mode the memory will stop at the point where variable information is required. After this is keyed manually, the rest of the standard message will continue when the START button is pushed. This technique can also be used to load a request for a repeat after a standard number transmission, both in the same message address, to be used in case you miss the number sent back to you.

If you are going to call a station on a schedule, you may wish to send his call sign more times than provided in the usual 3 *x* 3 calling format. To do this from the memory, load his call just once at the message starting address and pause until the bit indication stops advancing. Then continue with DE, your own call several times and *AR.* In the SEND mode, hold the message selector button down continuously instead of pressing and releasing it. The called station's call sign will be repeated as long as you hold the button down. When you release it, the rest of the message will continue to conclusion. (You can send long CQs this way too, but they're not recommended!) If you want to reduce the number of times your own call is signed, simply manually send *AR* or K at the appropriate point to stop the memory.

To call the same station often, as in a DX pileup, switch to the LOAD mode, select a message segment, and proceed to make your first call. Then, in the SEND mode, press the selector button and the memory will do all the subsequent repetitive calling for you. The above are only a few of the possibilities. Once you have gained some familiarity with the unit's capabilities, many other operating applications will suggest themselves.

references

1. James M. Garrett, WB4VVF, "The WB4VVF Accu-Keyer," QST, August, 1973, page 19.
2. James M. Garrett, WB4VVF, and D. A. Contini, W4YUU, "The Accu-Memory," *QST*, August, 1975, page 11.

ham radio

active bandpass filter for RTTY

Construction details for an active bandpass filter which will help eliminate interference and provide improved RTTY copy

This active bandpass filter was designed to work ahead of the NS-1A PLL Demodulator,[1] but it will provide improved copy when used ahead of any RTTY terminal unit. The filter is connected between the audio output of a receiver and the input of the terminal unit.

A bandpass filter is just what the name implies. It will pass only a specified band of frequencies, attenuating all frequencies above and below the specified limits. The objective is to make this attenuation as high as possible, passing nothing other than the desired signals. Without complex circuits, this is not always possible. Simpler filters, like the one described here, will always pass some unwanted frequencies, attenuated enough, however, that they are of no consequence.

For RTTY, we want to pass the standard narrow-shift tones, 2125 and 2295 Hz. For this 170-Hz shift the bandpass should be about 200 Hz wide, centered on 2210 Hz.

circuit description

As can be seen from **fig. 1,** the filter consists of two separate, two-section filters using cascade-connected operational amplifiers. It uses only one IC, the LM3900, which contains four so-called "Norton" amplifiers. These op-amps will do almost anything the standard op-amp will do. The principal advantage is that the LM3900 requires only a single power supply. Also, the LM3900 differences the input currents, whereas the conventional-type op-amp differences the voltages. The noninverting input function has been made possible by using what is called a current mirror circuit.

Several formulas are used to determine the required values of the resistors and capacitors. The

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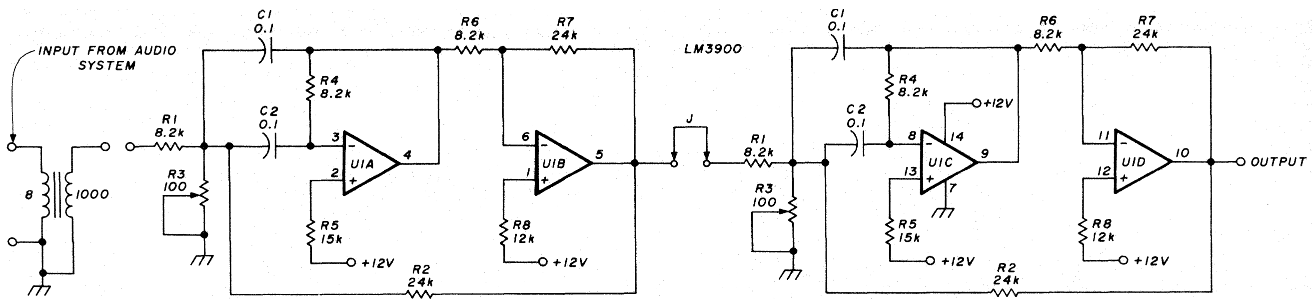

fig. 1. Schematic diagram of the active bandpass filter. The center frequency is 2210 Hz, with a − 3dB bandwidth of approximately 160 Hz. A single LM3900 Norton op-amp is used for the filter.

nearest standard values of 5 per cent resistors are shown in the diagram. The capacitors can be 10 per cent tolerance. R2 sets the Q of the circuit (and also the gain) and works out to be 12.5 kilohms for a Q of 11, or a −3 dB bandpass of 200 Hz centered on 2210 Hz. When both filters are connected together, however, the increased Q makes the bandpass too narrow. The value of 24 kilohms was chosen by trial and error to give the best passband.

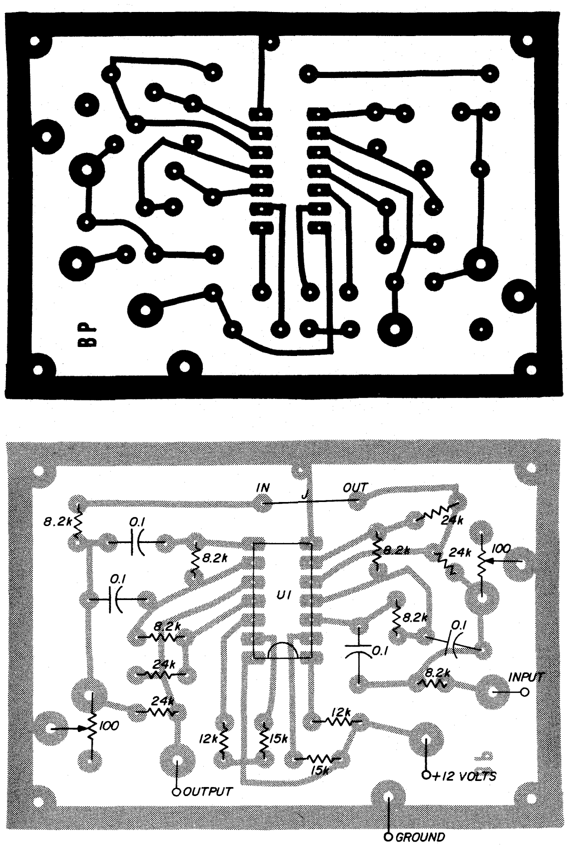

fig. 2. A circuit board foil pattern for the active bandpass filter is shown above. The filter can also be built on a small piece of perforated board using 5 per cent resistors and Mylar capacitors.* The bottom illustration is the parts placement diagram for the circuit board.

The complete filter, with standard resistor values, actually gives a −3 dB passband of 160 Hz, or a Q of 14. This may seem a little narrow for 170-Hz shift, but on-the-air tests have shown this not to be the case. The bandpass can be easily widened by changing R2 or by stagger-tuning one filter 50 Hz below and the other 50 Hz above the center frequency. The filter turns out to be about 800 Hz wide at the −20 dB points. There is no insertion loss; in fact, there is a small gain.

alignment and connection

Before connecting. the jumper between the two filters, separately tune each filter to 2210 Hz. This can be done by injecting a 2210-Hz signal and adjusting R3 for peak output as shown on a VTVM or high-impedance ac voltmeter. The jumper is then connected between the sections. Another method of tuning the filters, though not as accurate, is to tune in a strong RTTY signal with the filter in the line and adjust R3 on each filter for peak output or best copy.

The transformer is used to step up the input from the receiver speaker output to more closely match the input impedance of the filter. If the output irnpedance of your receiver is at least 500 ohms, the transformer will not be needed.

A good way to check the performance of the complete filter is to have some means of switching it in and out of the circuit. You will find that the filter will sometimes be the difference between copy and no-copy. High-impedance headphones (2000 ohms) connected across the output will enable you to hear the difference.

"A complete kit of parts and wired/tested units are available from the author. See Flea Market ads.

reference

1. Nat Stinnette, "Update of the Phase-Locked Loop RTTY Demodulator," *ham radio*, August, 1976, page 16.

ham radio

new audio amplifier for the Drake R-4C

A new audio amplifier for the Drake R-4C, suitable for direct substitution in all R-4C versions

Improvements in the Drake R-4C receiver, up to now, have been confined mainly to the i-f and detector systems.[1,2,3] One remaining area which needs improvement is the audio strip, which suffers from buzz and higher-than-desirable distortion; it also dissipates 7 to 10 watts of heat near the PTO. The audio amplifier, diagramed in **fig. 1,** eliminates these problems. While intended as an R-4C retrofit, this circuit performs so well that we also recommend it for other communications uses.

Our circuit is designed around National Semiconductor's LM383T, which, with the R-4C low-voltage supply, can deliver in excess of 2 watts into a 4-ohm load. The LM383 and associated components* can be mounted on a copper-clad board 3.8 cm (1½ inches) square, or another appropriate small heatsink (for a V_{cc} of 16 volts or less). It should be installed just behind the front-panel phone jack, between the passband-tuning capacitor and long i-f shield on which the Sherwood CF-60016 may be mounted. This location provides access to the speaker lead and detected signal at the audio gain pot. It also keeps the circuit away from power transformer hum fields in the chassis.

circuit precautions

The secret of making the LM383 an unconditionally stable audio amplifier (suitable for field installation in various layout configurations) is our output stabilization network. Proper stabilization is accomplished by connecting a 1.0-μF monolithic ceramic capacitor (such as Sprague 5CZ5U105X0050C5) with 19-mm (3/4-inch) leads directly between pins three and four of the LM383. Use of a lower-value capacitor with significantly longer or shorter leads will virtually guarantee oscillation problems. Tantaium or aiuminum electrolytics *cannot* be substituted for the monolithic capacitor.

Other circuit values have been chosen to tailor the audio response for greatest communications intelligibility. As in the original R-4C circuit, low frequencies are rolled off at one end of the needed spectrum; high-frequency shaping is similar to that of our suggested modification.[1] The feedback network has been chosen to provide nearly 40 dB of power-supply ripple rejection, minimizing the need for abnormal amounts of filtering. Gain at 1 kHz is 40 dB.

component selection

As with any high-gain amplifier, feedback and hum loops between the input and output should be avoided. Return all signal and power leads to pin 3, except for V_{cc} bypass, which should be returned to the IC tab with a solder lug.

To reduce component size, the 0.22-μF and 10-μF capacitors can be 16-volt (or greater) tantalums. The 200-μF electrolytic at pin 2 can have a 3-volt rating. The 300-μF output capacitor should have a minimum rating equal to V_{cc} (20 volts maximum). Sixteen volts is adequate for the R-4C. As mentioned above, a small heatsink is used for a V_{cc} less than 16 volts; above 16 volts a large heatsink. Never exceed a V_{cc} of 20 volts.

installation

To disable the existing amplifier, lift the output

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*A parts kit will be available from G. R. Whitehouse, Newbury Drive, Amherst, New Hampshire 03031.

transistor's collector lead at its solder lug. Also, remove its base or emitter wire, and/or disconnect one end of the driver's 100-ohm collector resistor. Connect the new amplifier's output to the phone jack terminal *with* the sleeved wire from the audio output transformer, and bypass with a 0.01-μF capacitor. This secondary is still used to provide the needed step-up for anti-vox system.

The only ground return should be a short, thick, insulated wire, run from pin 3 directly to the cable-braid terminal of the audio gain pot (rear section of

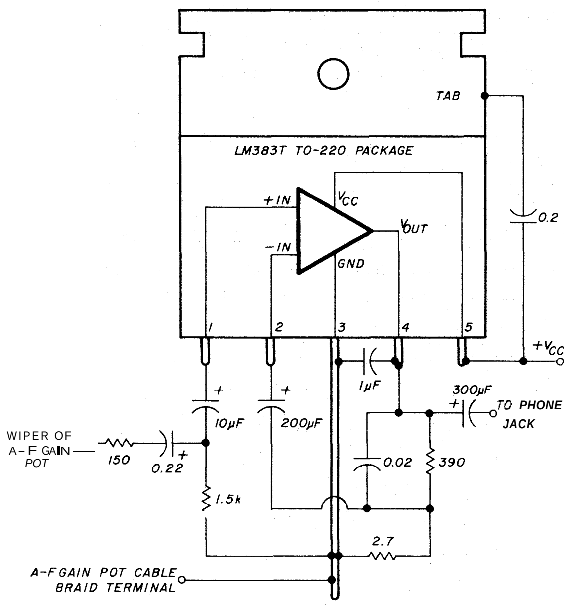

fig. 1. Schematic diagram of the new audio amplifier for the R-4C, based upon the LM383T audio amplifier IC. Resistors are ¼ watt. As pointed out in the text, all ground returns must be made through the connection to pin three to eliminate hum and feedback problems. The 1-μF monolithic ceramic capacitor *must* have 19-mm (¾-inch) leads (see text).

the dual control.) Do not allow the board or heatsink to touch any other part of the receiver ground. Next, add a small wire between the audio gain-control braid terminal and a close chassis ground. Disconnect the existing wire from the gain pot center wiper terminal, and connect the new amplifier input to this lug. Connect V_{cc} to the original audio-strip printed circuit board terminal with the blue wire from the audio-output transformer primary.

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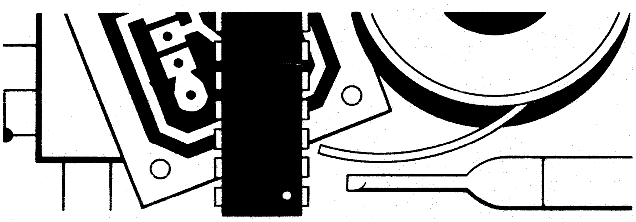
1. R. Sherwood, WBQJGP, G. Heidelman, K8RRH, "Present-Day Receivers – Some Problems and Cures," ham radio, December, 1977, page 10.
2. R. Sherwood, WBQJGP. G. Heidelman, K8RRH, "New Product Detector for the R-4C," ham radio, (ham notebook), October, 1978, page 94.
3. R. Sherwood, WBQJGP, G. Heidelman, K8RRH, "New Product Detector for the R-4C," ham radio, (short circuits), February, 1979, page 94.

ham radio

the weekender



the verti-loop a folded whip antenna for vhf mobile operation

Do you want a simple, inexpensive way to improve your 2-meter mobile performance in the fringe areas? How about a 314-wavelength vertical on your car roof? The concept sounds mind boggling — but what if it were only 1 meter (3 feet) long? Sound interesting? Try building one of these verti-loops, a name coined for a 314-wavelength vertical ground plane compressed to less than 112-wavelength long by folding the bottom section into a horizontal loop. The advantage is 2-3 dB gain over the 114-wavelength whip, and it still lets you keep the car in the garage. It's an easy weekend project and should cost less than $3.00 if you already have a roof-mounted antenna or an old discarded mag-mount CB whip. The results will be well worth the effort.

theory

Most vhf mobile operators know that a 1/4-wavelength whip is ideally located in the center of an unobstructed car roof. A compromise, because of physical size, is to use a 518-wavelength gain antenna on the trunk lid, where, because of obstructions on most cars, the 2-3 dB of potential gain is usually lost; results may even be inferior to those of the smaller roof-mounted whip.

A 314-wavelength ground-plane antenna is resonant, therefore nonreactive, and has a high-current feedpoint. The length of such an antenna may be calculated as half the length of a 312-wavelength dipole[1]:

$$L = \frac{149.95\,(N - 0.05)}{f} \qquad (1)$$

where L is the 312-wavelength dipole length in meters, N is the number of half wavelengths on the antenna (3), and f is the frequency in MHz (146).

$$L = \frac{149.95\,(2.95)}{146}$$

$$= 3.03 \text{ meters } (119.3 \text{ inches})$$

The vertical length would be half of this, or 1.52 meters (59.6 inches). The antenna impedance would be half of the 312-wavelength antenna impedance (105 ohms), or 52.5 ohms.

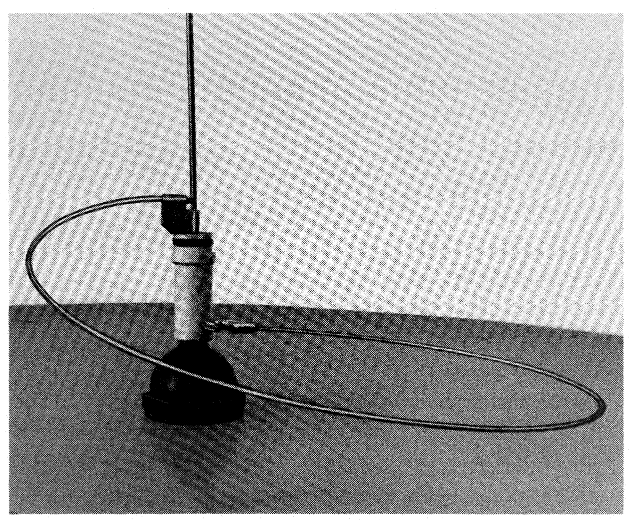
design

To shorten the 314-wavelength vertical to a reasonable size, it was divided into 114- and 112-wavelength sections, and the 114-wavelength section was folded into a loop. The result was the upside-down "halo" antenna depicted in **fig. 1.** The top section was shortened slightly to 91 cm (36 inches), and the bottom section was lengthened slightly to 61 cm (24 inches) for convenient fabrication sizes. When folded into a loop, the bottom section results in a 19-cm (7.5-inch) diameter.

construction

The antenna was constructed very simply using a 2.4 mm × 91 cm (3132 × 36 inch) stainless-steel welding rod (available at any welding supply shop) as the 112-wavelength section, and an ordinary tv uhf loop (available from Sears) as the folded 114-wavelength section.

To one end of a 5-cm (2-inch) section of plastic tubing (PVC works fine), a threaded stainless steel nut, which matched the threaded end of my rooftop 114-wavelength antenna, was epoxied. A piece of wood dowel (or plastic filler) was cemented into the other end of the tube and was drilled for a snug fit for



Close-up of the ¾-wavelength mobile vertical antenna. It resembles an "upside-down" halo.

By Herb Bresnick, WB2IFV, 16 Creekside Drive, Honeoye Falls, New York 14472

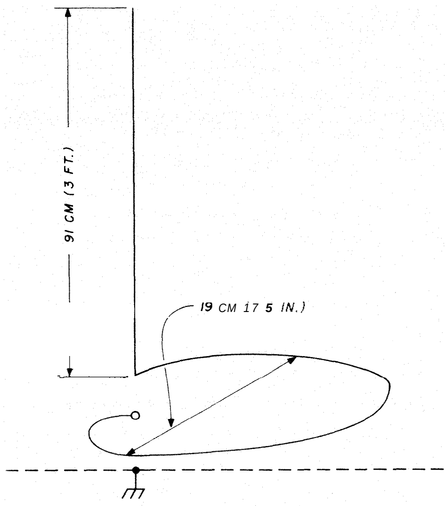


fig. 1. Design of the verti-loop antenna. Antenna is a %-wavelength vertical divided into ¼- and %-wavelength sections; the %-wavelength section is folded into a loop. Total extended length is ¾ wavelength on 2 meters. Antenna mounts on your car roof with a plastic base made of PVC tubing.

the stainless rod. A little epoxy will keep it from loosening **(fig. 2)**.

The Sears loop comes with hinged terminals, which were left on to permit fine tuning the antenna. One wire terminal was simply pulled out of its hinge, leaving a spring opening through which the stainless rod was forced. It makes for a snug fit, but permits sliding the loop down to where the rod joins the plastic tube.

The other wire terminal on the loop was cut off (leaving the hinge) and the wire remaining was passed through a small hole drilled in the side of the PVC tube just above the nut, where it was compressed by the threaded roof stud when the antenna was screwed on.

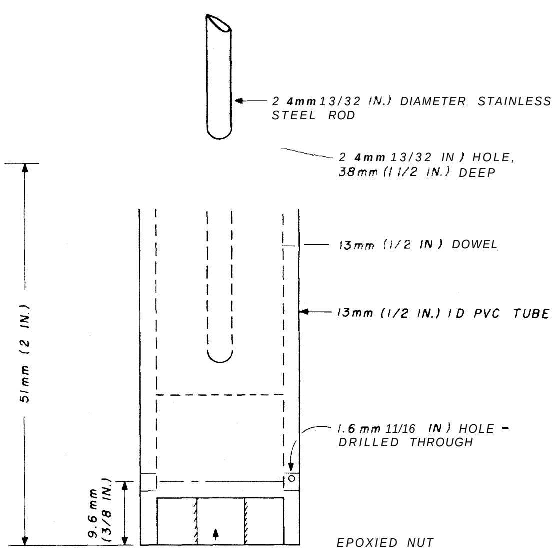
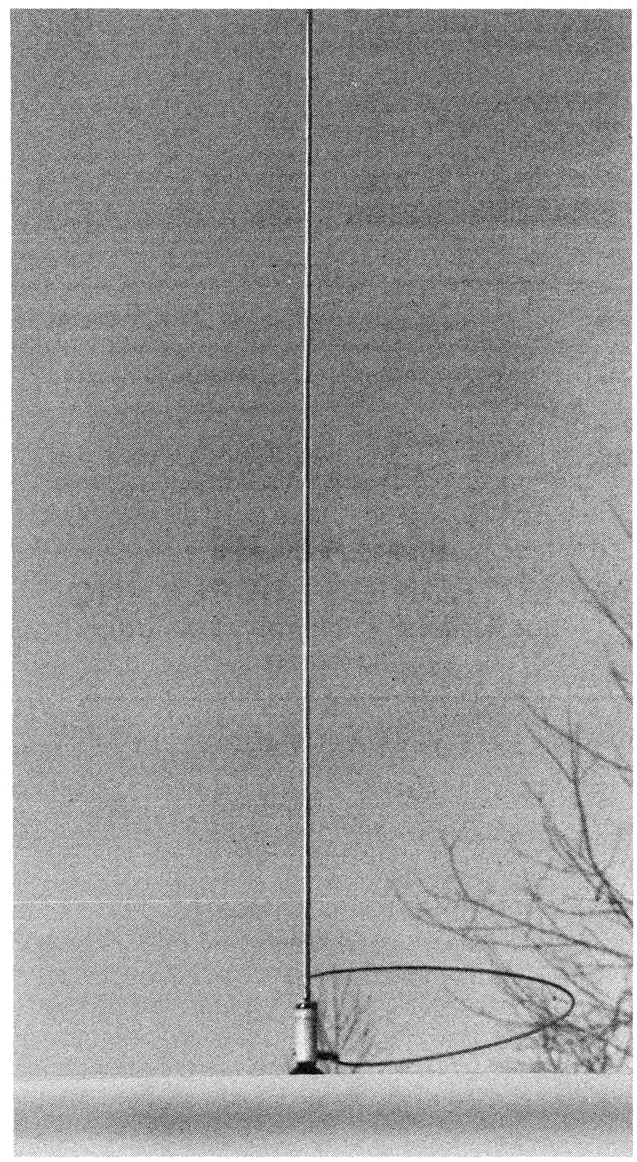


fig. 2. Details of the mounting base. An ordinary tv uhf loop was used for the %-wavelength section, or you can use stainless-steel wire.

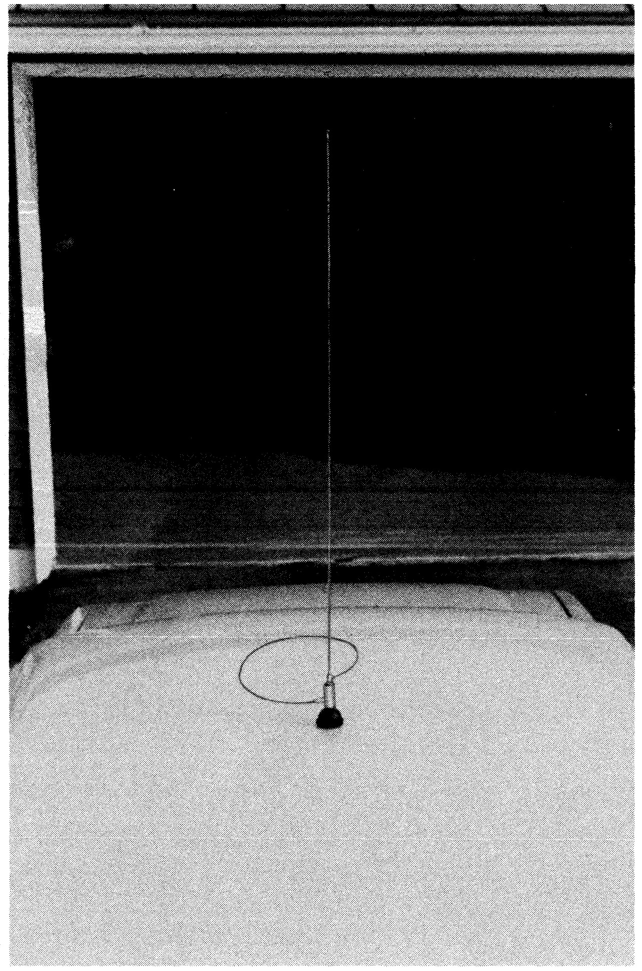
The antenna rod should be insulated from the stud except through the loop connection. The design can be modified easily to fit other base mounts or for new installations using a PL259 coax fitting, or even to a discarded CB magnetic mount if you don't want to drill a hole into your car roof.



View of the antenna at car-top level.

results

With the transmitter on an unused frequency and a SWR meter attached, the loop position was moved on its hinges for minimum SWR. It's not a critical adjustment if the dimensions are followed. I obtained 1.1 to 1.0 SWR with the loop about 2.5 cm (1 inch) from the car roof with the first try. The hinges are stiff enough to maintain the loop position even when



Author's antenna is a full-fledgzd %-wavelength vertical but fits under garage opening without dismantling. The stainless-steel radiator bends easily to 30-40 degrees from vertical.

driving over bumpy roads. I've built three of these antennas at a cost of $2.00-$3.00 each. All have resonated with practically no adjustment.

Actual dB measurements on the antenna have not been made; however, tests indicate a marked improvement over the 1/4-wavelength 'whip. One repeater that was barely readable is now almost full quieting into my receiver. Reports from friends are that noise on my signal (from a portable HT in the car) was all but gone and no evidence of mobile flutter occurred.

Best of all the car still fits into my garage, and, although the top of the antenna does hit the entrance, the stainless-steel rod is flexible enough to recover from a 30- or 40-degree bend. The antenna has a somewhat futuristic look to it — at least it looks different from your good buddy's 11-meter job!

reference

1. ***The ARRL Antenna** Book,* 12th edition, 1970, page 33.

ham radio

the jammer problem: some interesting solutions

The incidence of jammers is increasing — here's an interesting approach to the problem

In this paper we use the basic ideas of the scientific method in solving some interesting problems. It is important to have a clear understanding of, and agreement on, the nature of observation, cause and effect, hypotheses and testing of these hypotheses.

Science begins with observation. The scientific worker uses his mind to imagine how something might be constructed or how a problem may be solved. He knows that only by examining his observations can he determine if his imaginings correlate with the real world.

statement of the problem

The locale is a large city in the United States. In this city many Radio Amateurs abound. The problem we are about to address (and solve using the scientific method) concerns specific specimens. These specimens are Radio Amateurs (and others who are not legitimate Radio Amateurs) who persist in jamming the 2-meter repeater stations in this city. These specimens particularly delight in causing a disruption in radio communications over the repeater links by using various schemes.

The problem is to locate these specimens using three scientific methods: mathematical methods, methods from theoretical physics, and methods from experimental physics.

mathematical methods

Several mathematical methods are available to us to ferret out the offending 2-meter jammer. The mathematical methods we shall enumerate are easily seen to be applicable, with obvious formal modifications, to similar situations in other parts of the globe. As with other branches of knowledge to which mathematical techniques have been applied in recent years, the mathematical techniques of ferreting out 2-meter ham jammers has a singularly unifying effect on the most diverse branches of the exact sciences.

Hilbert, or axiomatic method

We place the culprit in an imaginary cage, which we have previously defined in time and space by using established techniques. We then introduce 'the following logic system:

Axiom 1. The class of jammers in the cage is non-void.

Axiom 2. If there is a jammer in the city, there is a jammer in the cage.

Rules of procedure. If p is a theorem, and p implies q is a theorem, then q is a theorem.

Theorem 1. There is a jammer in the cage.

Conclusion. Approach the cage with caution. Lift the jammer gently out of the cage, shave his beard, cut his mike cord, and put him in a playpen.

By Nuryev Sidelbandsk, UX3PU

method of inverse geometry

We place a spherical cage in the city, enter it, and lock the cage. We then perform a geometric inversion with respect to the cage. The jammer is then inside the cage and we are outside.

Conclusion. We proceed as in the method of Hilbert, above, except in this case we call in federal troops with Zen guns, who quickly dispose of the specimen. We then donate his equipment to a worthy organization.

"Call in the troops."

"Theorem 1 — there is a jammer in the cage."

method of projective geometry

Without loss of generality, we may regard the city as a plane. We project the plane into a line, then we project the line into an interior point of the cage that contains the jammer. We then perform a geometric involution of catharsis, which in turn projects the jammer into the city jail (without bail).

Bolanzo-Weierstrass method

In this method we bisect the plane by a line running north-south. The jammer is either in the east portion or in the west portion; let us assume he is in the west portion. We bisect this line with another line running east-west. The culprit is either in the north portion or the south portion. Let us suppose he is in the north portion.

We continue this process indefinitely, constructing a sufficiently strong force about the chosen portion at each step. As the volume of the chosen portion approaches zero, the jammer is ultimately surrounded by a fence of arbitrarily small perimeter. The specimen (jammer, if you will) is thus a victim of the confusion syndrome. He begins to believe he has, at last, become rational. It is now relatively easy to approach the jammer. We place a cowbell into the jammer's left hand and into his right hand we place a telegraph key, which is connected to a code-practice oscillator. It is now a very critical moment for the jammer. He begins to send messages to himself with the cowbell and answers himself with the telegraph key. We realize that the messages make no sense because the jammer is not comfortable with the Morse code.

Conclusion. To reduce the confusion factor in the jammer's mind, we gently tap him on the head with a small mallet. As we manipulate the mallet (which we do using the Morse code), the jammer begins to get the message, to wit: "Jamming is a no-no. Jamming is a no-no. Jamming is a no-no." I shall leave as an exercise for the student the decision as to how long we subject the jammer to this action.

Cauchy, or function-theoretical, method

In this mathematical method we consider a jammer-valued function, $f(z)$, where z is nondimensional but critical to resolving the problem.

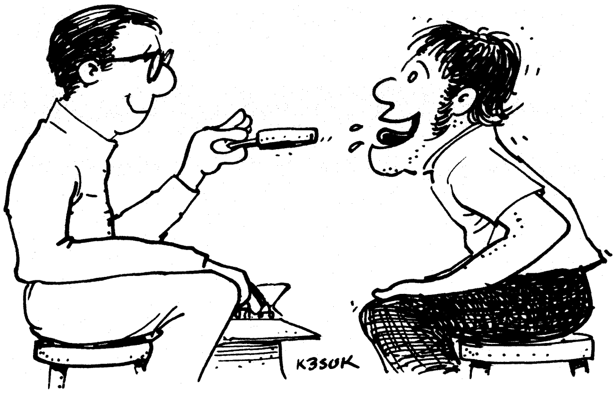
Let there be a function, $f(z)$, and let S be the cage

"The jammer is not comfortable with the Morse code."

enclosing the jammer. Consider the integral

$$\frac{1}{2\pi 1} \; {}_0\!\int^{C} \frac{f(z)}{z-s}\,dz \qquad (1)$$

where C is the boundary of the plane of encounter. Its value, after appropriate manipulation of eq. **1,** is



"Ply him with popsicles."

that of $f(S)$. Therefore, we have a culprit in the cage. We then dispose of him appropriately. In this case we ignore his stupid mouthings on the repeater but send a male nurse armed with ten-dozen Popsicles. The nurse plies the jammer with Popsicles until the jammer decides to throw up (a very effective problem-solving method).

methods from theoretical physics

We now examine some theoretical methods for dealing with the 2-meter jammer. These methods have not been proved in practice but offer some interesting food for thought for the theoretician.

Dirac method

We observe that jammers are known to use wild methods to attract attention. Therefore these jammers must be considered wild men. But, according to the scientific method, we must temper our observations with caution. Perhaps the jammer in question is not really wild but is somewhat tame with a rather limited mentality. Perhaps he needs help. Why punish him if, indeed, he's not really responsible for his actions: The author has a proven formula which, if applied to these situations, will produce good results in many cases. For those interested I shall be happy to furnish the method and appropriate action. Please send a self-addressed, stamped envelope for an immediate reply.

"Perhaps the jammer is not really wild."

methods from experimental physics

Using the methods from experimental physics, we are able to deal positively with the problem of 2-meter jammers. It should be noted that these methods are also valid for other inconsiderate jammers in the other Amateur bands.

thermodynamic method

We construct a semipermeable membrane, permeable to everything except jammers, and sweep it across the plane of interest (*i.e.*, the city in question). We also add a pot of hot water to the apparatus. As the semipermeable membrane is swept across the city, it picks up the jammers. The jammers are enmeshed in the membrane and, having nothing else to do, immediately start brewing coffee. We immediately detect the aroma of brewing coffee, descend on the culprits, and capture them in a loony net. We then enlist the aid of local authorities to dispose of the problem.

magnetomatic method

In implementing this technique we plant a large bed of popcorn arranged in an ellipse, whose major axis lies along the direction of the horizontal component of the earth's magnetic field. At one focus of this ellipse we place a cage (with strong iron bars). We then distribute over the plane in question large quantities of magnetized spinach (*spinacia oleracea*), which has high ferric content.

The magnetized spinach is eaten by various denizens of the locality, which are in turn eaten by the 2-meter jammers. The jammers, having partaken of commodious amounts of popcorn, are then oriented parallel to the earth's magnetic field. This causes a beam of jammers to be focused onto the cage. Since the jammers are full of our specialized popcorn and, by indirect ingestion, magnetized spinach, it is an easy matter to zap the jammers. The benefit of this method is that we can dispose of the culprits in large numbers.

ham radio

variable-frequency audio filter

Here is a tunable, inexpensive, two-IC audio filter that plugs into the receiver phone jack. It will drive headphones or a low-impedance speaker up to two watts if necessary. It operates from a single power supply that requires no regulation, thus it could be easily incorporated into some of the direct-conversion receivers used in QRP rigs.[1,2] It will also add to the selectivity of any receiver that lacks a sharp CW filter. Used with an ssb receiver, the tunability of the filter allows you to eliminate an unwanted signal by using the high or low skirt of the ssb filter and then peak the desired signal at whatever frequency results.

The circuit is that of the state-variable filter,[3] using three op amps (**fig. 1**). The LM324 IC used in this design has four op amps on the same chip, so the extra amplifiers don't add much to the cost or wiring complexity. Two identical integrator circuits control the frequency response, so only a two-section ganged pot is necessary for frequency selection. The Q, or sharpness, of the filter is controlled by a single resistor. The 3.3-megohm value shown in **fig. 1** was selected as a compromise between sharpness and ringing. Circuit Q comes out to 15. While no ringing is noticed, the noise definitely takes on the pitch of the frequency selected. The value of this resistor may be changed to suit your needs without altering the tuning range, The LM380 was added to the filter circuit to drive hi-fi headphones or a loudspeaker directly; the filter will drive almost anything that the receiver will drive.

With the values shown, the tuning range is from 800 Hz - 2 kHz. Voltage gain is close to unity overall, so the receiver gain control sets the signal level.

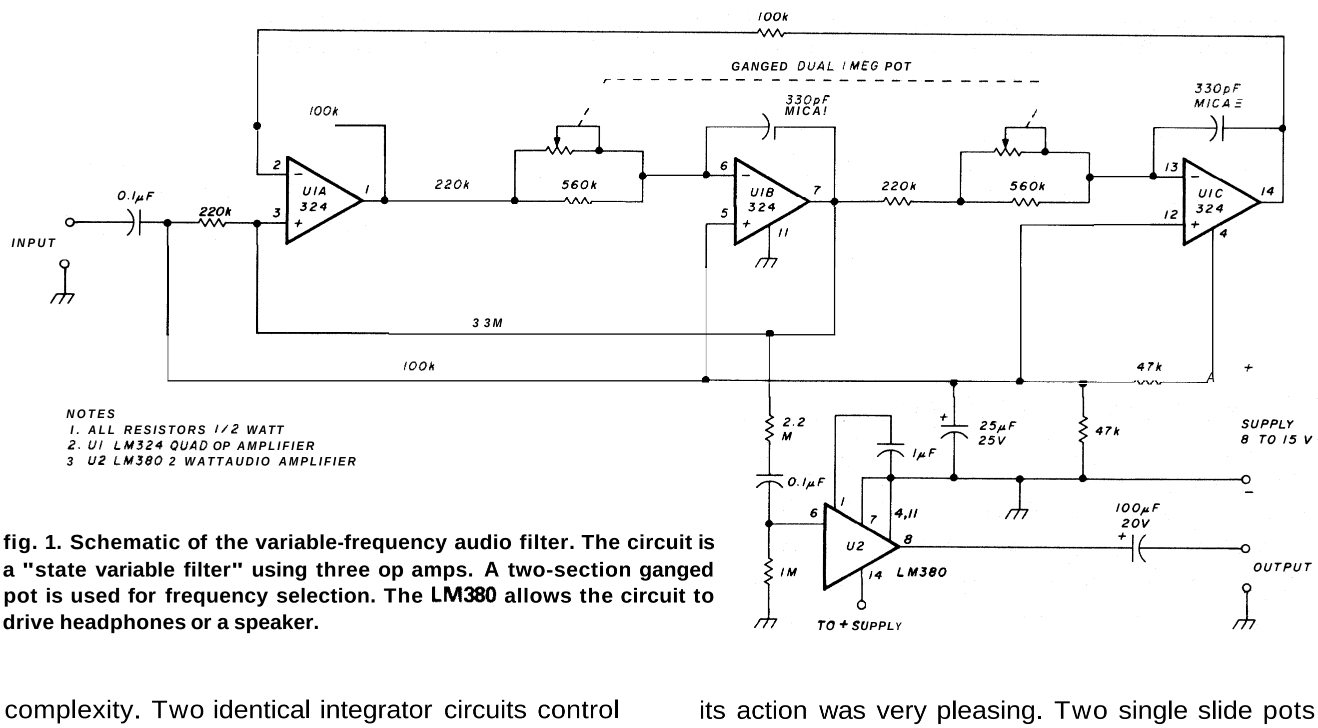


fig. 1. Schematic of the variable-frequency audio filter. The circuit is a "state variable filter" using three op amps. A two-section ganged pot is used for frequency selection. The LM380 allows the circuit to drive headphones or a speaker.

audio-frequency characteristics

The frequency-controlling network is designed to operate with a dual 1-megohm pot. In one model that was built, a dual linear-action slide pot was used and its action was very pleasing. Two single slide pots could probably be mechanically linked to give the same effect. The extra resistors connected in series and parallel with the ganged pots give audio bandspread by allowing the full resistance change of the ganged pots to alter the frequency over less than two octaves. They also make tracking of the ganged pots less critical. If a pot with a logarithmic taper is used,

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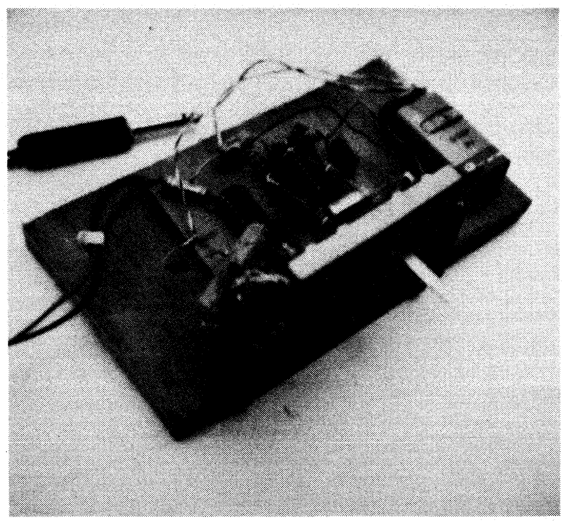

In this model of the variable-frequency audio filter, a dual, slide action linear potentiometer has been used. Because of the high current drain, the 9-volt battery will have to be replaced with an ac supply for permanent installation.

it should be connected for the most linear frequency variation.

If you want to design for a different frequency range, or use a different value of dual pot, the filter peak frequency is given by $f = 1/2\pi RC$. Plugging in the maximum and minimum resistance values will give the frequency extremes. For example, in **fig. 1** when the 1-meg pot is at a maximum, total resistance between pins 1 and 6 is 220k in series with 560k, paralleled by 1 meg. This is 578k ohms, and gives 812 Hz in the formula. When the 1-meg pot is at zero resistance, the 560k resistor is shorted and the resistance in the formula is 220k. This gives 2192 Hz. These values are all for capacitors of 330 pF.

The circuit draws about 11 mA at 12 volts with no signal. This current increases to about 35 mA for fairly loud signals. It would be possible to operate the filter from a 9-volt transistor battery, but not for long periods of time. The feedback loops in the circuit and the 47k divider resistors stabilize the operating point of the integrators at half the supply voltage, so regulation isn't necessary.

This filter is not the solution to all interference problems, but it will allow you to selectively boost one frequency 10 dB more than another only 200 Hz away. That can make the difference between a useful contact and a marginal one.

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1. Jay Rusgrove, W1VD, "A 20-Meter High-Performance Direct-Conversion Receiver." QST, April 1978.
2. "A Direct Conversion Kilogram," ***The*** ARRL ***Handbook,*** 1977.
3. Howard Berlin, W3HB, "The State-Variable Filter," QST, April 1978.

ham radio

split-frequency operation with th Drake R-4B receiver and the TR-4 transceiver

Modifications you can make to your Drake receiver and transceiver to enhance operating convenience and versatility

This article describes how I increased the versatility of my Drake TR-4 transceiver for split-frequency operation using a Drake R-4B receiver as an external VFO.

I purchased a used R-4B receiver, and the thought kept passing through my mind that it would sure help if I could use it to control the TR-4 transmit frequency (as in the R-4 TX-4 combination).

analysis

The Drake R-4 receiver and the TR-4 use different i-fs; therefore there's no direct interface as with the TX-4. After studying the circuit diagrams, it occurred to me that it should be very easy to duplicate the control circuit of the Drake RV-4 remote VFO and make use of only the VFO portion of the R-4B to control the TR-4.

The control circuit described was built and minor modifications were made to the R-4B to interface with the TR-4. The R-4B can be used as a separate receiver, then switched to remote VFO operation.

Circuit modifications don't interfere with R-4B normal operation. The added components can be removed in a few minutes at resale time to restore the receiver to its original condition.

The main problem in the project is the fact that there'sa 45-kHz difference in VFO operating frequency between the TR-4 and R-4B. This is apparently because of the 50-kHz second i-f of the R-4B. There's no reason why the VFO frequency can't be decreased 45 kHz to match the TR-4 VFO output; then the crystals in the premixer oscillator can be changed to oscillate 45 kHz higher to compensate for the VFO downward shift. I breadboarded the circuits to test the idea and everything worked as planned.

control circuit

I duplicated the control circuit of the RV-4 (fig. 1) and mounted it in a small minibox. This minibox sits on top or alongside the TR-4 to interface the R-4B. Only two critical items are in the control circuit. One is peaking coil L1 at the grid of V1 (12AU7); it must match the coax cable from the VFO to the 12AU7 grids.

The diagram in my manual didn't give a value for L1 so I experimented a bit. I reclaimed a vhf coil form from some surplus gear and wound 70 turns of no. 30 (0.25mm) enamel wire on the 6.5-mm (¼-inch) slug-tuned form. Too few or too many turns would not peak the 5-MHz signal, so the number of turns must be adjusted. A scope is a great aid in tuning L2.

The other critical item is the four-position special switch, A1. After tracing the functions of the Drake switch, I discovered that a standard four-position, five-pole switch would do the job. These parts could be ordered from Drake, but I used all junkbox parts.

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The 1-mH rf choke from TR1 collector is from an old TV set. Any coil will work as long as it has a good number of turns. Transistor TR1 turns the VFO B+ on and off through the rf choke and the coax line going to the R-4B receiver.

control switching

Position 1, separate receive. The 330-kilohm resistor in **fig. 1** is connected through S1D to B+, turning TR1 on, causing its collector to saturate. Thus TR1 collector goes low, pulling the 10-Vdc B+ line to the R-4B VFO toward zero, which turns the VFO off. The slide switch on the side of the R-4B must be forward to disconnect the control unit, thus defeating the above action. The VFO output is now sent to V8 (premixer) cathode. The R-4B operates normally.

Position 2, receive. The 330-kilohm resistor is connected through S1D and S1E to pin 5 of the Jones plug. This is receive cathode ground. During receive TR1 base goes low, cutting it off, and the B+ to the VFO goes high, which allows the VFO in the R-4B to operate during receive only. During transmit the receive cathode goes high, turning the R-4B VFO off. Note that the slide switch on the side of the R-4B must be toward the rear to switch the VFO output from V8 mixer to the control unit.

Position 3, transceive. The 330-kilohm resistor is connected through S1D and S1E to ground, turning TR-1 off, thus allowing the VFO B+ to rise to normal. This allows the VFO in the R-4B to operate on transceive.

Position 4, transmit. During transmit only, the 330-kilohm resistor is connected to ground through S1C and pin 3 of the Jones plug to the transmit cathode. TR1 is cut off allowing the VFO B+ to rise, enabling the R-4B VFO. During receive, the transmit cathode goes high, turning on TR1, which pulls the VFO B+ down so that the R-4B VFO can't operate.

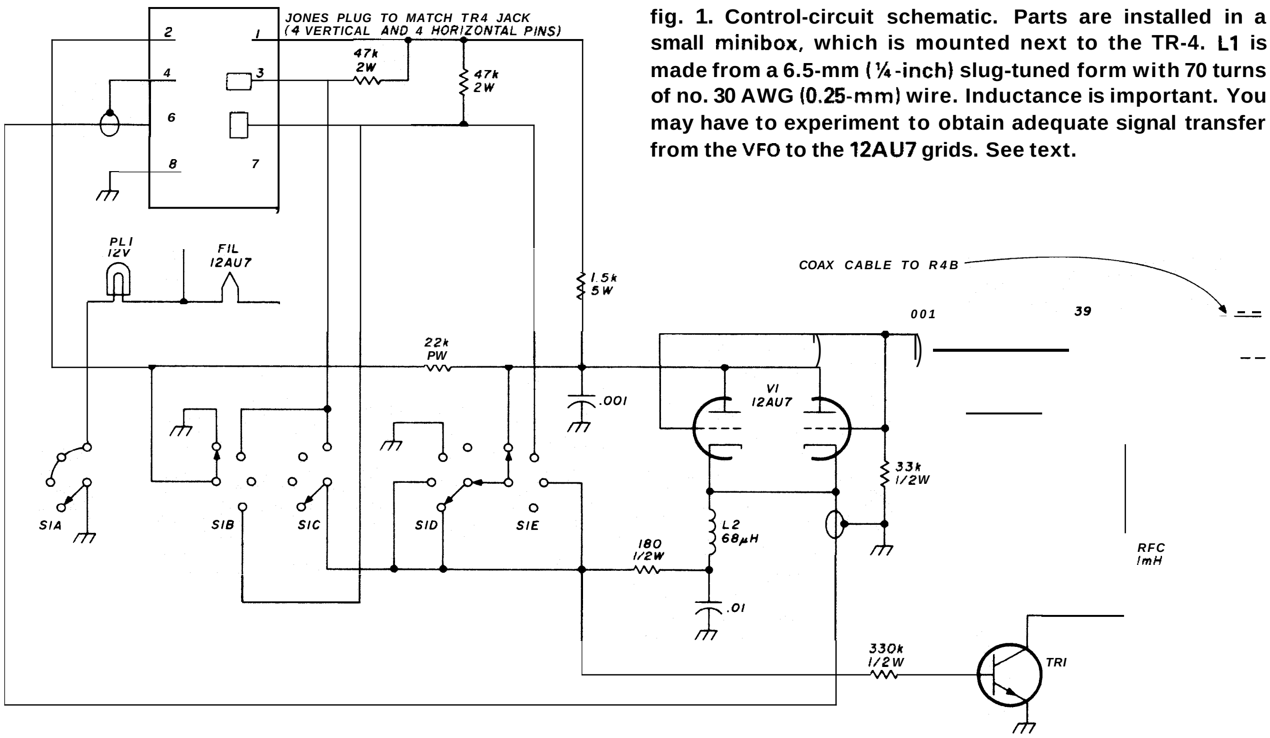


fig. 1. Control-circuit schematic. Parts are installed in a small minibox, which is mounted next to the TR-4. L1 is made from a 6.5-mm (¼-inch) slug-tuned form with 70 turns of no. 30 AWG (0.25-mm) wire. Inductance is important. You may have to experiment to obtain adequate signal transfer from the VFO to the 12AU7 grids. See text.

The 12AU7 is a cathode follower with both sections in parallel. It functions as an impedance transformer giving a low impedance output to feed the TR-4. At the same time, the 12AU7 acts as a switch to disconnect the signal from the remote VFO. The cathode is switched high or low at the appropriate time according to the switch position and the TR-4 transmit-receive relay.

control-circuit layout

Nothing is critical about the control-circuit layout, since it is primarily composed of dc switching lines. However, some attention should be paid to the rf components related to the 12AU7 grid. Leads should be short, and accepted rf-wiring practice should be used.

Transistor TR1 can be any silicon npn device with a voltage rating of about 40 volts. The coax cable from the R-4B to the control unit, and the cables

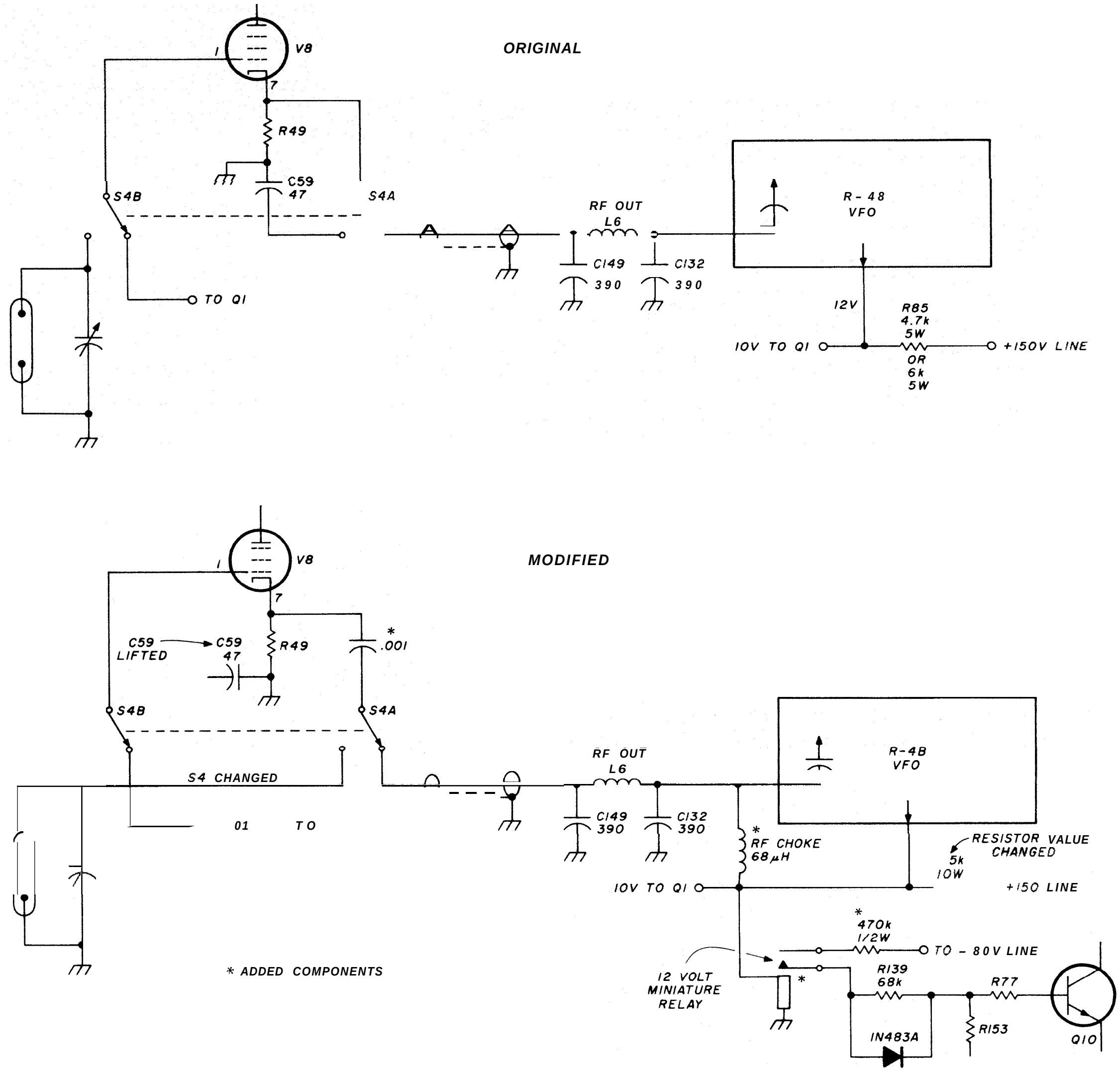


fig. 2. Before and after mods to the Drake R-4B receiver.

from the control unit to the Jones plug, should be as short as possible. *Caution:* Do not insert or remove the Jones plug in the TR-4 while the rig is on; otherwise the TR-4 transistors may be damaged.

R-4B modifications

Of the five modifications described, only the first two are needed to make the R-4B receiver function as a remote VFO. Modifications 3 through 5 are optional. The modifications to be described involve:

1. Changing the value of power resistor R85

2. Changing the position of three wires on S4

3. Adding four components to switch the neon VFO indicator lamp

4. Changing the VFO dial

5. Changing the crystals and adding padding capacitors if desired

Fig. 2 shows the R-4B circuit before and after modification.

Modification 1. Provide VFO output from the fixed-channel crystal jack on the side of the R-4B by following these steps:

1. Lift C59 from S4A. Bend C59 so that it doesn't touch anything. Leave it there for restoration at resale time.

2. Locate the jumper that comes from the fixed-channel crystal jack to S4B. Lift the connection from S4B and solder it to the lug on S4A where C59 was removed.

3. Locate the wire that comes from V8 pin 7. Lift if from S4A middle lug.

4. Lift VFO output coax center conductor from S4A and move it to S4A middle lug (vacated in **step** 3 above).

5. Place a small 0.001-μF ceramic capacitor in series with the wire removed in 3 above. Solder the other lead of the capacitor to S4A where the coax formerly was connected. This blocks the 10-Vdc control voltage, which we will place on the VFO output line, from reaching V8.

We have now modified S4 to provide the following switchings. Forward: reguiar R-4B receive mode. Back: VFO output to crystal jack, receiver disabled. The fixed-frequency crystal socket is now the VFO output jack. A length of RG-59 carries the R-4B VFO signal from this jack to the control box. I soldered the pins of a defunct crystal to the ends of the coax so I could plug the coax into the crystal socket.

Modification 2. Dc switching of the R-4B VFO is accomplished from the control unit by placing an rf choke from the 10-volt VFO B+ line to the rf output line from the VFO. A convenient spot is between the two solder tabs on the circuit board just behind the audio gain control. I used a 68-μH choke from my junk box. The value isn't critical as long as it presents a high reactance at 5 MHz. Check that resistor R85 can handle the additional dissipation when the 10-volt line is shorted to ground by the control box. R85 must drop the entire 150 volts from the +150-volt line. The control unit now enables the TR-4 to turn the R-4B VFO on and off in the same manner as in the remote RV-4 VFO.

Modification 3. Switching the neon VFO indicator lamp is optional but enhances operating convenience. I purchased a surplus 12-Vdc, 10-mA miniature relay and mounted it on a long bolt, which I installed where the audio output transformer is mounted. Your mounting will depend on what relay you have available. (See the comments on relay selection.) The following steps are necessary for this modification:

1. Connect the relay coil from the +10 volt line to ground.

2. Place a 470-kilohm resistor from the normally open contact (normally open with coil energized) to the – 80 volt line. A convenient spot is the solder tab that has a white wire with a green stripe (on my R-4B). This is the negative lead of C91, which is the 8-μF, 150-volt filter of the – 80 volt line. It's located on a small vertical board behind the notch-depth control, which is mounted on the right-side panel.

3. Locate the board containing 010. It's mounted below the VFO and just behind the front panel. Connect the other relay contact to the junction of R139 (68k) and D15 anode.

The relay functions as follows. When the TR-4 control box is set to inhibit a signal from R-4B VFO, the +10 Vdc line goes low (3 volts or lower). This action de-energizes the relay just installed, connecting negative cutoff bias to transistor Q10, which causes NE2 to be extinguished.

4. R85, 4.7-kilohm, 5 watts, should be changed to a 3.5-kilohm, 10-watt resistor if you use a 10-mA relay as I did. Extreme care should be used here. With the original 4.7-kilohm resistor, the additional current drain caused the normal 10-Vdc regulated voltage to drop below the zener regulation point. Reducing the value of R85 allows an additional 10 mA of current to be drawn, and the zener will still regulate at 10 Vdc. Do not operate the VFO with the lower resistance value without the relay coil connected, as the additional current will probably blow the 250-mW, 10-volt zener mounted inside the VFO enclosure. Easy does it!

If you use a relay with a coil other than 10 mA, 12 Vdc, adjust the value of R85 accordingly or perhaps provide a separate supply to drive the relay by a transistor. An alternative method would be to put a pre-regulator on the line (12 to 15 volts) with a zener of enough power-handling capability to do the job.

Modification 4. If you wish dial calibration on the R-4B identical to that on the TR-4, it will be necessary to order a TR-4 dial from Drake. (The price in June of 1978 was $2.00, plus $1.00 handling, plus postage.) It's installed in the following manner:

1. Remove all front-panel knobs (some slip on; others have a set screw).

2. Remove nut on the function selector switch.

3. Remove four screws at corners of the front panel. Be careful to catch the fiber spacers behind the panel (they're hard to find when they fall and roll across the floor).

4. Remove the two screws holding the metal shield over the neon bulb. Be extremely careful *not* to bend the neon-bulb leads, because then break very easily; many standard replacement neon bulbs will not fit. I had to cut a hole in the shield to allow the end of the bulb to stick out.

5. Very carefully remove the front panel without unduly bending the neon-bulb leads. Lay aside the clear Plexiglas sheet with the red line. Do not lose the pressure washer. Note how it came off so it can be replaced in the same way.

6. Remove two screws holding the pilot light shield mounted behind the panel.

7. Remove the two C-rings from the VFO shaft. You'll need a C-ring tool; it can be purchased in auto parts shops. Be careful! The C-rings are tight and are hard to remove. (I broke my tool and had to use a wheel puller to remove the first C-ring). The second C-ring wasn't so tight. *Important:* Note the position of these C-rings so they can be replaced in the exact spot.

8. Locate, identify, (make a drawing), and remove the three leads coming out of the VFO.

9. Remove the three nuts holding the VFO in place. Lift and remove the VFO assembly from the R-4B.

10. Turn the VFO shaft to find the final stop. Noting *exactly* where it was positioned (pencil mark),

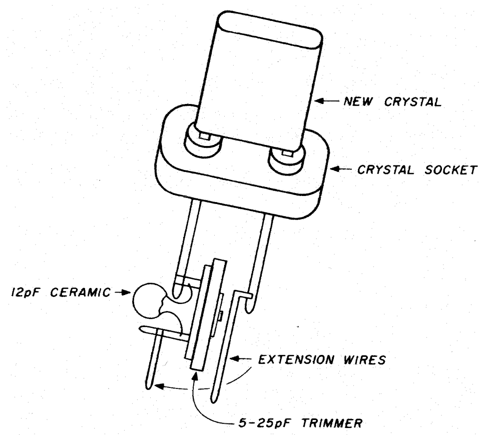


fig. 3. Construction of the crystal paddler, which plugs into the R-4B receiver crystal socket.

remove the dial by removing the large C-ring. Place the TR-4 dial in the same position.

11. Reassemble, following the steps above in reverse order.

Modification 5. To track the VFO for TR-4 use and separate receiver use, I ordered new crystals that are 45 kHz higher in frequency than the originals. For the 20-meter band to tune backward, as in the TR-4, I substituted a crystal frequency (14.645 MHz), which is added, rather than subtracted, to the VFO output to obtain the proper injection frequency.

This, by the way, is the same crystal needed for 80 meters. Therefore, by installing an appropriate jumper and one other change, the same crystal can be used for both 20- and 80-meter operation.

To keep expenses down I ordered only three crystals (my cost locally was $20.00 for the three crystals): one for 80 meters, which doubles for 20 meters; one for 15 meters, and one to receive WWV on 15 MHz. Thus I've covered my needs at the present. For the same crystal to be used for both 80 and 20 meters, the following steps are required:

1. Find C53 (68 pF) in parallel with R47 (1.5 kilohms). This network is the collector load for 01 at 25.1 MHz. Lift these two components from S5F and dress them out of the way so they don't touch anything but are handy for replacement at resale time.

2. Place a jumper from C55 (68 pF) to the switch tab vacated above. Solder the connections.

3. Place an insulated jumper wire from the 14.6-MHz crystal jack to the 25.1-MHz jack. (The 25.1-MHz crystal will no longer be used.) Put it away so it will be handy at restoration time." The jumper should be placed in the holes toward the front panel. The new crystal, 16.645 MHz, can now be plugged into the 14.1-MHz jack. Eighty-meter operation is as normal; 20-meter operation now tunes backward, as in the TR-4.

crystal padding

I found it necessary to put a trimmer in series with the new crystals to trim the R-4B receive frequency to exactly the same dial calibration as when using the R-4B as a remote VFO. I used a 12-pF ceramic cap in parallel with a small 5-25 pF variable, which I removed from some surplus equipment (see **fig.** 3). The cap is smaller than the standard-size trimmer and will fit between the solder tabs of a standard-size crystal socket. The capacitors are in series with one lead of a crystal socket adapter, which is plugged into the main crystal socket.

dial calibration

Dial calibration is as follows:

1. Calibrate the TR-4 dial as usual, using the TR-4 internal crystal calibrator.

2. Switch the control unit to receive from external VFO. Calibrate the external VFO dial (R-4B dial) in the same manner as the TR-4.

3. Switch the control unit to separate receive (don't forget to throw the slide switch for separate receive). The R-4B should now operate normally. Adjust the trimmer mounted below the new crystal to zero beat.

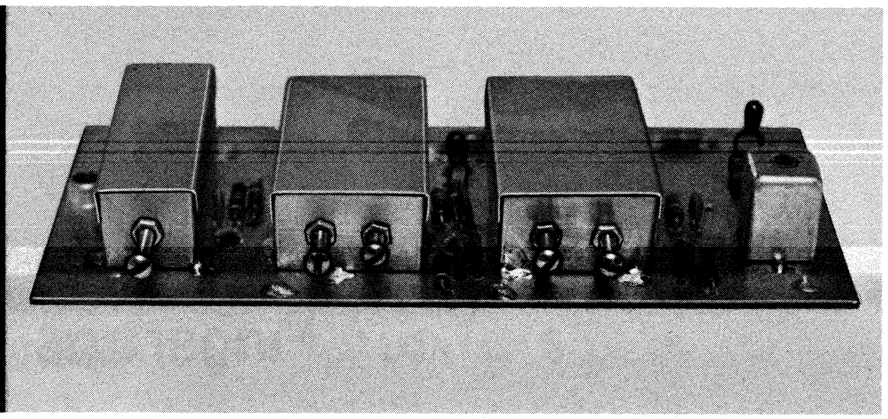
summing up

Now you can receive a station on the R-4B and then switch immediately to transceive on the same frequency by simply selecting transceive on the control box and moving the slide switch on the side of the R-4B. You're now transmitting with R-4B VFO control. It's a good idea to check the dial calibration from time to time to make sure that everything is still calibrated, otherwise you'll have a small frequency offset when switching to transmit from the R-4B.

You have greatly increased the versatility of the TR-4/R-4B combination, and, if your experience is like mine, you'll also have greatly increased your operating pleasure.

ham radio

"The steps in these modifications should be kept in a file with your equipment literature. Appropriate annotations to the steps will come in handy when you decide to restore your radio for resale. **Editor.**



high-performance 432-MHz converter

Complete construction details for a 432-435 MHz receiving converter that is easy to build and tune up

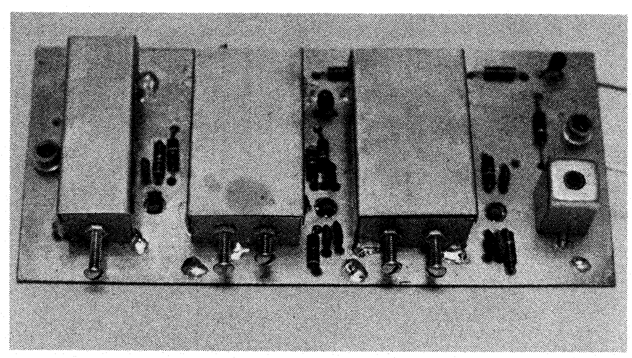
The 432-MHz receiving converter described here is the result of several month's work. As anyone who has attempted to build solid-state uhf gear knows, layout and construction technique is vital, not only to coming up with a good basic design, but also to building a converter that works properly.

Now that economically priced solid-state equipment and devices are available for use above 432 MHz, the cost of building uhf equipment has been reduced considerably. This converter uses three transistors (six if you build the oscillator chain for copying Oscar 8). Three Texas Instruments dual-gate 3CT225A mosfet transistors are used in the converter (90 cents each in small quantities). The 3CT225A transistors are rated at 900 MHz with a 4-dB noise factor; at 432 MHz with 12 Vdc drain voltage and 4 volts on gate 2, gain will be about 22 dB with a noise factor of 2 dB or less. These transistors, and the stripline tuned circuits, makes this a very good converter for weak-signal reception.

Two versions of the converter were built using coils in the tuned circuits, but the results were far below that achieved with striplines. A local-oscillator chain was not built on the converter board because the 404-MHz injection frequency was coupled from the oscillator chain in my 432-MHz transmitting converter. However, a diagram of an easy-to-build oscillator chain will also be described. I use it with the converter to copy Oscar 8 Mode J.

layout and construction

Component arrangement, component spacing, and wiring have been optimized for best results, based on several earlier versions. The spacing between the tuned circuits in the amplifier and mixer



Top view of the 432-435 MHz converter showing placement of the enclosures with the tuned lines. Input is to the left, output to the right. Transformer T1 is in the shield can near the output connector.

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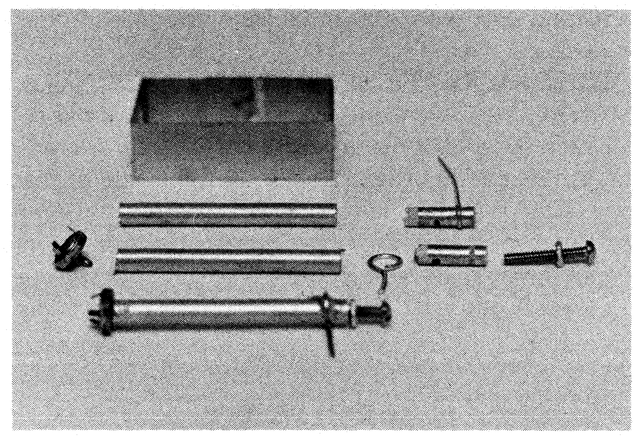
was laid out to provide satisfactory interstage coupling without the use of capacitors.

The circuit boards I used are glass epoxy. The oscillator board has foil on one side; the converter board has copper on both sides. Anyone can make the circuit boards, even if he has never etched boards before. Start by cutting the boards to size; clean the boards with steel wool and paint the copper. I used *Rustoleum* gray primer in a spray can — it protects the copper while in the etching solution, and it is easy to remove when you scribe the lines. After the painted boards have dried, using the pattern, trace out the lines; since there are no curved lines, scribing is very easy.

Use a small ieweler's screwdriver to scribe the lines. You can paint a spare piece of circuit board and practice getting the lines the right width; about 1/16 inch (1.5 mm) is best. Put the scribed boards in etching solution and agitate once in a while to speed up the etching process. After all the necessary copper has been etched off, thoroughly clean the boards with soap and water using steel wool. A final cleaning with abrasive household cleanser is recommended to remove any last traces of the etchant solution.

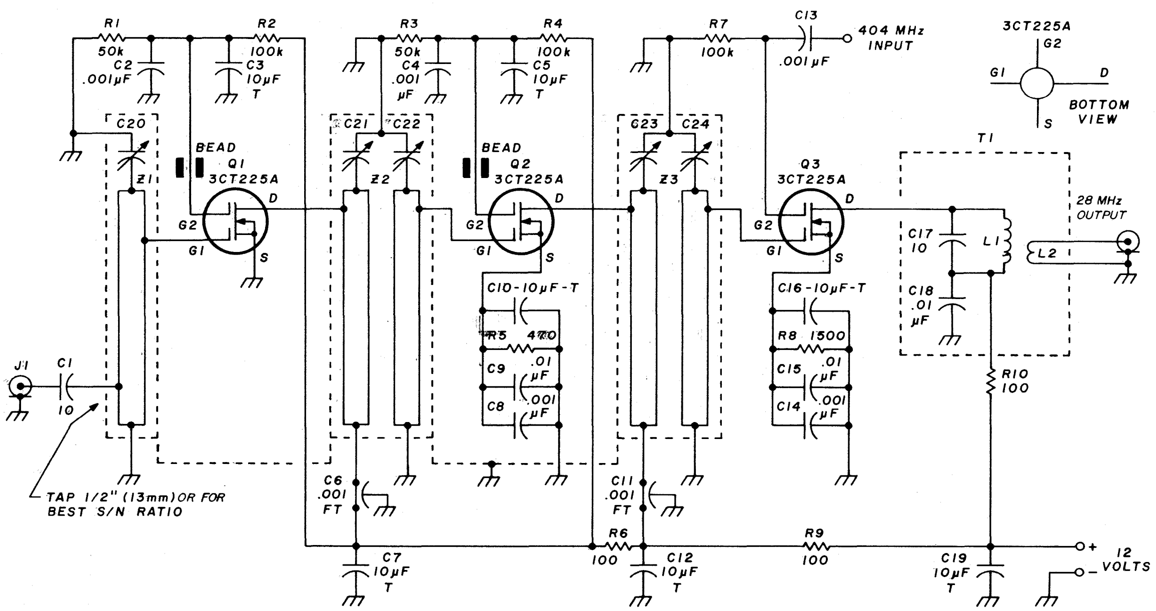
tuned-circuit construction

The enclosures for the striplines are made from brass strips 0.065 inch (1.5 mm) thick, 3/4 inch (19 mm) wide, and 12 inches (30 cm) long. These strips



Finished components for one of the tuned stripline circuits. The button mica capacitor is soldered to the bottom of the line; the piston capacitor is inserted in the top and soldered in place (see text).

and the 114-inch (6.5 mm) brass tubes can be purchased in many hobby shops. The dimensions for the second rf and the mixer stages are 1-1/4 x 2-1/4 x 3/4 inches (32 x 57 x 19 mm); the first rf stage is 2-1/4 x 314 x 314 inches (57 x 19 x 19 mm). These are inside dimensions, so, when bending, allow for the bend angle. One way is to mark off the first bend, bend in a vise, then mark the second bend, and so on. One side must be longer than the other so it laps and can be soldered. The spacing between the tuned lines is 1/4 inch (6.5

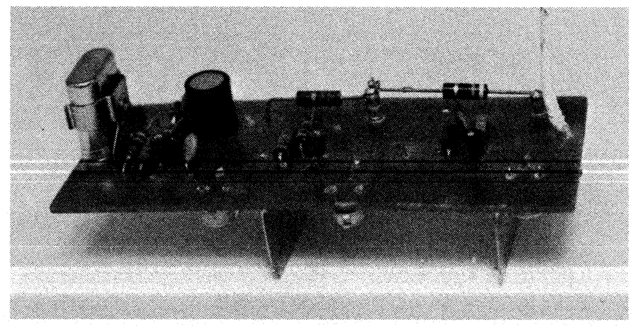


C20-C24 1 5-10 pF ceramic piston capacttors (Centralab 829-10)
T1 Hamtronics 7807 coil form and shield L1 is 22 turns no 26 (0 4 mm) close-wound, L2 is 4 turns no. 26 (0 4 mm) on bottom of L1

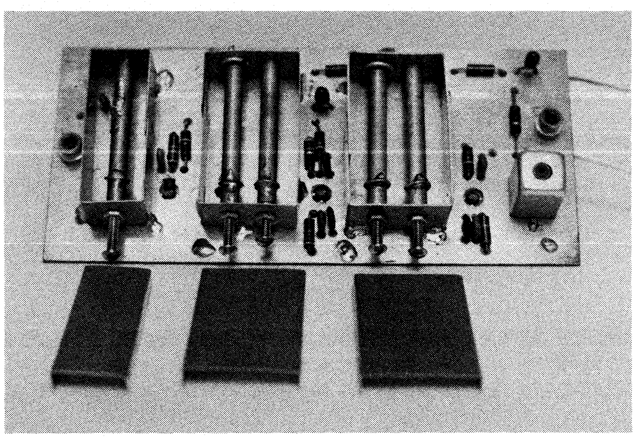
fig. 1. Schematic diagram of the high-performance converter for 432-435 MHz; companion local-oscillator chain is shown in fig. 2. Capacitors marked with FT are button mica feedthrough types; capacitors marked with Tare tantalum; all other capacitors are disc ceramic. Z1, Z2, and Z3 designate the stripline circuits.

mm) and 1/4 inch (6.5 mm) from the side of the enclosure.

The tubes for the tuned lines are 2-1/8 inches (54 mm) long; this allows 1/8 inch (3 mm) clearance between the end of the tube and the enclosure where the trimmer is mounted. Solder tin the inside of the tube about 1/2 inch (13 mm) down. Remove the wire pigtai! from the trimmer, and insert the trimmer in the tube; you may have to apply heat with the soldering iron to get it in. About 1/8 inch (3 mm) of the



Construction of the 404-MHz local-oscillator chain. Printed-circuit layout is shown in fig. 4.



Top view of the converter with the enclosure covers removed to show the placement of the tuned lines. Note the holes in the board for mounting the transistors.

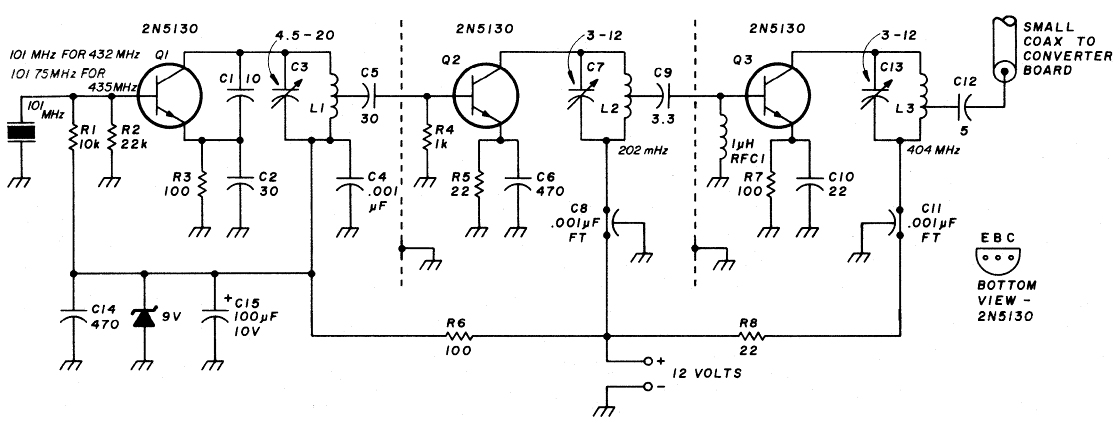
trimmer should protrude beyond the tube. Apply heat to the tube and sweat solder the trimmer in place.

Two of the tuned lines have dc blocking capacitors mounted at the ends; these are mounted like the trimmers, but the tubes will be shorter by the thickness of the capacitor. Form the piece of wire that connects the tuned lines to the PC board, but do not solder yet; this will be done later when the tuned lines are placed in the enclosures.

mounting and components

If you don't have a 404-MHz oscillator chain for injection, build the oscillator chain first. I built the oscillator board one circuit at a time, drilling the holes for the components as I went, testing each stage before starting the next. The schematics and photographs show all the details. After the oscillator board is complete and tested, lay it aside and start construction on the converter board.

Wiring the converter is straightforward; if you study the photographs and schematic you should have no problems. However, there's one precaution: Since the converter board has copper foil on both sides, the foil around the component holes on top of the board must be cleared to prevent shorts. To do this use a 118-inch (3-mm) drill and flatten the angle of the point so it will cut the copper from around the top hole but not touch the copper on the bottom of



L1 6 turns no. 16 (1.3-mm) wire, 114-inch (6.5-mm) diameter, 314 inch (19mm) long, tapped at one turn
L2 3 turns no. 16 (1.3-mm) wire, 114-inch (6.5-mm) diameter, 314 inch (19mm) long, tapped at 1-1/2 turns
L3 1 turn no. 16 (1.3-mm) wire, 114-inch (6.5-mm) diameter, with 112-inch (13-mm) leads

fig. 2. 404-MHz local-oscillator chain for the 432-435 MHz converter. Capacitors marked with FT are feedthrough types; except for the electrolytic (C15), all other capacitors are disc ceramic.

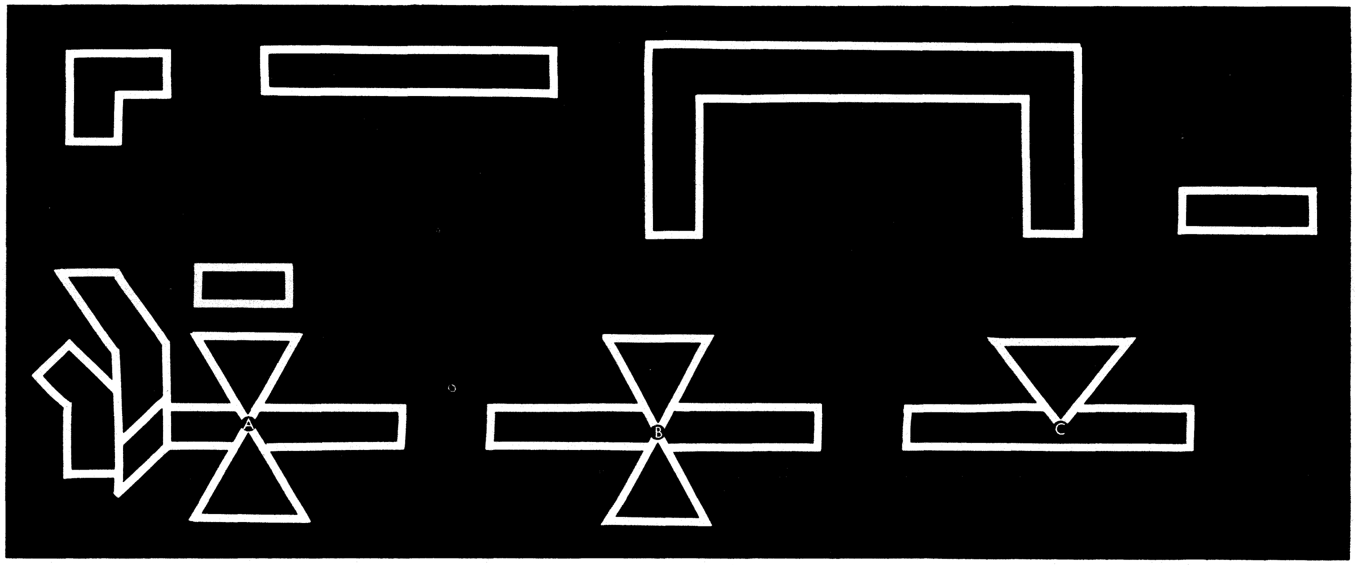


fig. 3. Full-size printed-circuit layout for the 432-435 MHz converter. Component placement for the converter is shown in the photograph. Transistor mounting holes marked A, B, and C are 5/16 inch (8 mm) diameter.

the board. Use an emery wheel to shape the drill bit for this task.

After the mixer wiring has been completed, connect the output of the oscillator to gate 2 of Q3 in the converter board with small coax cable; apply 12 Vdc and common to both boards. You should be able to receive the third harmonic of a two-meter transmitter that tuned to 144 MHz, or a weak-signal source if you have one. There are two excellent articles on weak-signal generators in past issues of *ham radio*.[1,2] When the mixer is working properly, proceed with the other two stages — testing as you complete each one.

You may be able to build the converter without using the 10-μF tantalum capacitors, but they do much to stabilize the circuits so they are well worth the extra cost. You may have noticed that on the schematic diagram ferrite beads are shown on gate 2 of each rf stage; the photograph shows only one on the first stage — my error, use the bead to improve stability.

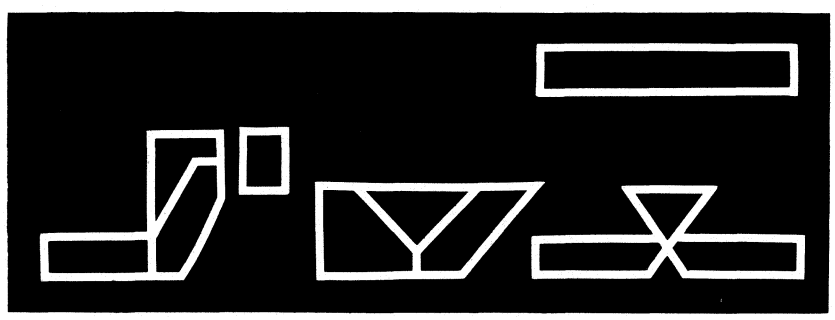
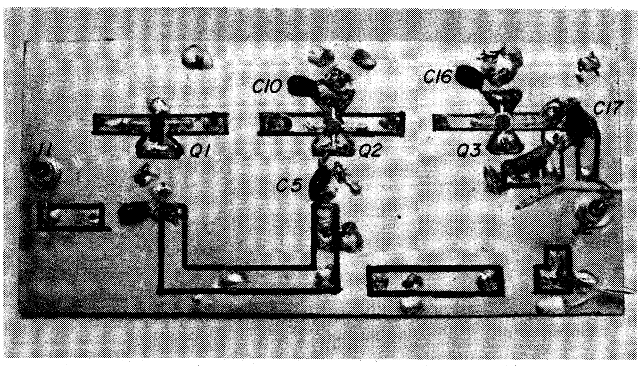


fig. 4. Printed-circuit layout for the 404-MHz local-oscillator chain. Component mounting and shield placement is shown in the photograph.

tune up

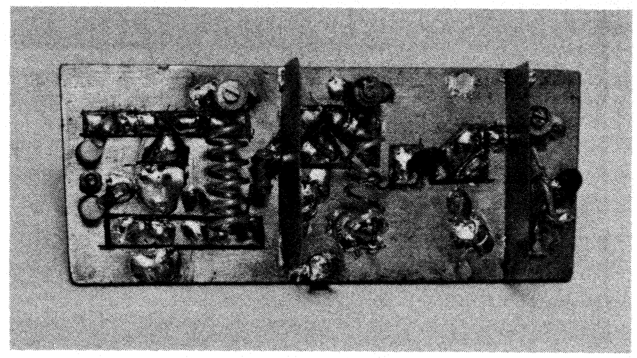
If a commercial signal generator is available, use it; if not, use a weak-signal source. You may find that the test signal will leak into the converter from places other than the antenna! Take the weak signal outside, as far away from the antenna as possible, or until you can't pick up the signal with the antenna disconnected. With the antenna connected you should receive the weak signal at S-5 or better. You can now peak everything up for optimum performance.

With my weak-signal source about 100 meters from the antenna I receive it at S-9 using a Kenwood TS-520. After the circuits have been tuned for maximum signal, tune away from the signal and check the noise level as indicated on the receiver's S-meter. The objective is to obtain the greatest signal-to-noise ratio. You can also adjust the converter for lowest noise figure if you have equipment for noise figure measurements.[3,4]



Component placement on the converter board (see fig. 3). In this view the input is to the left, output to the right. Designated capacitors are mounted on the bottom of the board. Small coax cable near C17 couples 404-MHz injection to the converter.

When using this converter with a TS-520 transceiver and no 28-MHz preamp, I have consistently maintained schedules on 432 MHz with stations over 100 miles (160 km) away, with reports ranging from S-4 to 20 dB over S-9 depending upon conditions. During band openings many stations have been heard 700-800 miles (1100-1300 km) out with S-meter readings over S-9. I have also used this converter to copy Oscar 8 and hear the spacecraft signals around S-6 without any special antennas. I



Component placement on the circuit board for the 404-MHz local-oscillator chain.

have built several 432-MHz converters from handbooks, magazine articles, and commercial kits, but this converter tops them all.

references

1. James Brannin. KGJC, "A Stable Small-Signal Source for 432 MHz," *ham radio,* March, 1970, page 58.
2. Bruce Clark, KGJYO, "A Stable Variable Output Weak-Signal Source," *ham radio,* September, 1971, page 36.
3. Louis Anciaux, WBGNMT, "Accurate Noise-Figure Measurements for VHF," *ham radio,* June, 1972, page 36.
4. Robert Stein, W6NBI, "Automatic Noise-Figure Measurements," *ham radio,* August, 1978, page 40.

ham radio

impedance measurements

using an swr meter

The simple SWR meter
and a handheld calculator
can be used
to measure
complex relationships
involving impedance
in tuned circuits

An SWR measurement can indicate when an rf source is properly matched to a load, but often it's desirable to know the actual impedance values for the load. Once these values are determined, it's then possible to design matching circuits or transformers to translate the load impedance to the desired impedance.

For example, as more solid-state broadband transmitters come into use, a matched load impedance becomes more desirable. While the high SWR of an antenna may not damage transistors rated to withstand that SWR, the output power may be reduced because of an improper load impedance. The lack of tuning and load adjustments prohibits rematching the load impedance to the final output stage for the rated transmitter performance.

Another example of the requirement for knowing the load impedance can be seen for a typical installation with an antenna cut for the 75-meter phone band, but with an occasional trip down in frequency to check into an 80-meter CW net. When an antenna tuner isn't available it's possible to construct a matching circuit so the transmitter still sees a load impedance within its rated SWR, even though the actual antenna SWR is somewhat higher. Knowing the actual impedance values of the 75-meter antenna while using it at 80 meters will yield a starting point for construction of such a circuit.

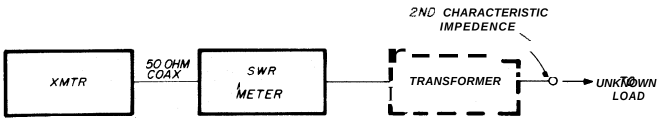


fig. 1. System for measuring characteristic impedances other than 50 ohms. The transformer can be a lumped-constant element, broadband transformer, or a transmission line such as a quarter-wavelength of 75-ohm coaxial cable at the frequency of interest.

measurement procedure

Many instruments are available to measure impedance values, but only a few fortunate Amateurs have them. An instrument that is available, or at least not too expensive or difficult to construct, is an SWR meter.

Once the SWR of an impedance is known, half the battle is over. If the reflection coefficient electrical angle, θ, is known, the impedance can be quickly determined by using a Smith chart. (This is discussed later.)

Information that's available is the SWR based on a characteristic impedance, Z_{01}, of 50 ohms. This means the source, usually a transmitter, looks like a 50-ohm source, and the SWR meter has a 50-ohm characteristic impedance. However, the use of a Smith chart isn't the only method of determining an impedance when the SWR and the reflection coefficient electrical angle are known. The impedance can be calculated outright by knowing the SWR, the reflection coefficient electrical angle, and the characteristic impedance of the measuring system.[1]

If another SWR measurement could be made with a different characteristic-impedance-measurement system, Z_{02}, the same impedance values would be found. Of course a different SWR reading would exist, as well as a different reflection coefficient electrical angle. But the impedance values of the load will not have changed.

Since the impedance is constant for the two measurements, the equations relating SWR, Z_{01}, and θ, are set equal to each other, yielding:

$$(\tan \theta_1)(\tan \theta_2) = \frac{\left[\left(\frac{Z_{02}}{Z_{01}}\right)\right](swr_1) - swr_2}{\left[\left(\frac{Z_{02}}{Z_{01}}\right)\right](swr_2) - swr_1} \qquad (1)$$

$$\frac{\tan \theta_1}{\tan \theta_2} = \frac{\left(\frac{Z_{02}}{Z_{01}} \cdot swr_1 \cdot swr_2\right) - 1}{(swr_1 \cdot swr_2) - \frac{Z_{02}}{Z_{01}}} \qquad (2)$$

The SWR is a measured value, and Z_{01} is known

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from the characteristic impedance of the test equipment.

This means that a second characteristic-impedance-SWR-measurement system must be devised. Such a system is easily obtainable by using the existing 50-ohm source and an SWR meter and transforming this impedance to some other impedance, as shown in **fig. 1.** The transformer can be a lumped-element, broadband transformer[2,3] or a transmission-line-type, such as a quarter-wavelength of 75-ohm coax at the frequency of interest. (This quarter-wavelength transformer results in a characteristic impedance of 112.5 ohms.)

The procedure is as follows. It's simpler to perform than to describe. Using the 50-ohm system, measure and record the SWR of an unknown impedance. This measurement yields swr_1 and $Z_{01} = 50$ ***ohms.*** Then connect the transformer at the output, or antenna side, of the SWR meter. Connect the unknown load

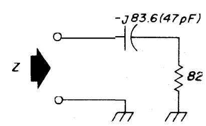


fig. 2. Impedance test circuit.

to the output port of the transformer. Again, measure the SWR. This measurement yields swr_2 and Z_{02}. These four values, swr_1, Z_{01}, swr_2 and Z_{02}, are substituted in **eqs. 1** and 2, resulting in two equations and two unknowns. The two unknowns are $\tan\theta_1$, and $\tan\theta_2$, where θ_1 and θ_2, are the reflection-coefficient electrical angles.

Unfortunately, more than one solution exists, giving more than one value for $\tan\theta_1$, and $\tan\theta_2$. But the resulting impedance calculations yield the same reactance magnitude using either values; whether the reactance is inductive or capacitive can't be determined directly. Another test must be conducted to make this determination; such tests are described later.

From the measurements, all the information necessary to calculate the actual impedance has been found. Substituting swr_1, $\tan\theta_1$, and Z_{01} into **eq. 3** gives the impedance. Likewise, substituting swr_2, $\tan\theta_2$, and Z_{01} into **eq.** *4* gives the same impedance:

$$Z = Z_{01}\left[\frac{1-j(swr_1)\tan\theta_1}{swr_1 - j\tan\theta_1}\right] \qquad \textbf{(3)}$$

$$Z = Z_{02}\left[\frac{1-j(swr_2)\tan\theta_2}{swr_2 - j\tan\theta_2}\right] \qquad \textbf{(4)}$$

test results and example

To test this procedure, I constructed a test load using an 82-ohm resistor in series with a 47-pF capacitor. A randomly selected piece of 75-ohm cable was found to look like a quarter wavelength at 40.5 MHz. At this frequency the resistor and capacitor look like 82-j83.6 ohms **(fig. 2)**.

The reactance of the 47-pF capacitor is found from:

$$X_C = \frac{1}{2\pi f_C} = \frac{1}{2\pi(40.5\times 10^6)(47\times 10^{-12})}$$
$$= -j83.6$$

The one-quarter wavelength of 75-ohm coax transforms the 50-ohm source impedance to 112.5 ohms:

$$\frac{Z_L^2}{Z_1} = Z2 = \frac{(75)^2}{50} = 112.5\ ohms \qquad \textbf{(5)}$$

Now the two SWR measurements are made. Using the 50-ohm system ($Z_{01} = 50$ ***ohms***), swr_1 of the capacitor and resistor in series is measured as 4:1. The quarter-wavelength, 75-ohm transformer is inserted between the SWR meter and the unknown impedance, and another SWR reading, swr_2, is measured as 2.5:1; Z_{02} is 112.5 ohms. These values are now substituted into **eqs. 1** and 2 to determine the values of $\tan\theta_1$ and $\tan\theta_2$. Note that calculating the values of θ_1 and θ_2 isn't necessary unless the reflection coefficient electrical angle is desired; only the $\tan\theta_1$ and θ_2 values are needed. These calculations yield:

$$(\tan\theta_1)(\tan\theta_2) = \frac{\left(\frac{112.5}{50}\right)(4) - 2.5}{\left(\frac{112.5}{50}\right)(2.5) - 4} = 4 \qquad \textbf{(6)}$$

$$\frac{\tan\theta_1}{\tan\theta_2} = \frac{\left(\frac{112.5}{50}\cdot 4\cdot 2.5\right) - 1}{(4\cdot 2.5) - \frac{112.5}{50}} = 2.77 \qquad \textbf{(7)}$$

Eq. 7 yields: $\tan\theta_1 = 2.77\tan\theta_2$. Substituting into **eq. 6** yields:

$$2.77\tan^2\theta_2 = 4$$
$$\tan\theta_2 = \pm 1.2$$

Solving for $\tan\theta_1$ by substituting the value of $\tan\theta_2$ into **eq. 7** yields:

$$\tan\theta_1 = 2.77(\pm 1.2) = \pm 3.32$$

The positive value for $\tan\theta_1$, Z_{01}, and swr_1 are now substituted into **eq.** 3 to give Z, the unknown impedance:

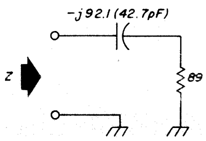
$$Z = 50\left[\frac{1-j(4)(3.32)}{4-j3.32}\right]$$
$$= 50\,[2.56\angle -46^\circ]$$

$$Z = 50\,[1.78 - j1.84] = 89 - j92.1$$

This gives the resistance as **89** ohms and the capacitive reactance as − j92.1 ohms, resulting in a capacitor value at 40.5 MHz of:

$$C = \frac{1}{2\pi f x_c} = \frac{1}{2\pi(40.5 \times 10^6)(92.1)} = 42.7\,pF$$

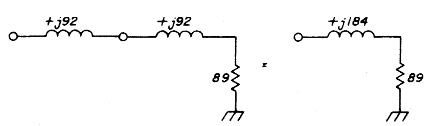
The measured circuit is then:



This compares favorably with the component values used in the circuit of **fig. 2.**

Two pertinent factors should be noted here. The values for Z_{02}, swr_2, and $tan\ \theta_2$ could just as easily have been substituted into eq. **4** to obtain the same impedance. The negative values for $tan\ \theta$ could have been used, giving a positive value for the impedance reactive component. Often, observation of the circuit can indicate the proper sign of the reactive component. If this isn't possible a second measurement can be made at another frequency, and the direction of change for the reactive component can be observed. For instance, a higher frequency should yield a lower capacitive reactance. If the reactance is inductive, the higher frequency should yield a higher value of reactance.

Another method of determining the sign of the reactance would be to insert a known reactance value in series with the measured impedance and make another measurement. For the example given above, an inductor of approximately +j92 ohms at **40.5** MHz (0.36 μH) would cancel the − j92 ohm reactance, leaving only a real component of 80 ohms. If the reactance happened to be positive in the first place, the second measurement would yield:



Actually, a second calculation isn't necessary, since it can quickly be seen that 89 + j0 would yield a much lower SWR than 89 + j184.

theory

The derivation of this concept follows directly by equating eqs. 3 and **4.** This is reproduced here for reference:

$$Z = Z_{01}\left[\frac{1 - j(swr_1)\tan\theta_1}{swr_1 - j\tan\theta_1}\right] = Z_{02}\left[\frac{1 - j(swr_2)\tan\theta_2}{swr_2 - j\tan\theta_2}\right]$$

$$(1 - j\,swr_1 \tan\theta_1)(swr_2 - j\tan\theta_2) = \frac{Z_{02}}{Z_{01}}(1 - j\,swr_2\tan\theta_2)(swr_1 - j\tan\theta_1)$$

Equating the real and imaginary terms yields:

$$(A)\ swrp - swr_1\tan\theta_1\tan\theta_2 = \frac{Z_{02}}{Z_{01}}(swr_1 - swr_2\tan\theta_1\tan\theta_2)$$

$$(B)\ swr_1\,swr_2\tan\theta_1 + \tan\theta_2 = \frac{Z_{02}}{Z_{01}}(\tan\theta_1 + swr_1\,swr_2\tan\theta_2)$$

From A,

$$\left[\left(\frac{Z_{02}}{Z_{01}} \bullet swr_2\right) - swr_1\right]\tan\theta_1\tan\theta_2 = \left(\frac{Z_{02}}{Z_{01}} \bullet swr_1\right) - swr_2$$

$$\tan\theta_1\tan\theta_2 = \frac{\left[\left(\frac{Z_{02}}{Z_{01}}\right)(swr_1)\right] - swr_2}{\left[\left(\frac{Z_{02}}{Z_{01}}\right)(swr_2)\right] - swr_1}$$

From B,

$$\tan\theta_1\left[(swr_1 \bullet swr_2) - \left(\frac{Z_{02}}{Z_{01}}\right)\right] = \tan\theta_2\left[\left(\frac{Z_{02}}{Z_{01}} \bullet swr_1 \bullet swr_2\right) - 1\right]$$

$$\frac{\tan\theta_1}{\tan\theta_2} = \frac{\left[\left(\frac{Z_{02}}{Z_{01}}\right)(swr_1)(swr_2)\right] - 1}{\left[(swr_1)(swr_2)\right] - Z_{02}}$$

Exactly what transpires can readily be seen by looking at the Smith chart shown in **fig.** 3. The first SWR measurement, $swr_1 = 4$, yields a reflection coefficient, ρ_1, of 0.6. The electrical angle of the reflection coefficient, θ_1, is found from $\tan\theta_1 = \pm 3.32$. Therefore, $\theta_1 = \pm 73.2°$. Using only the positive value and knowing that the total distance around the Smith chart is 180°, one-half wavelength, the ratio of 73.2' to 180° gives a reflection coefficient at 0.203λ. This is shown as Z_1 and corresponds to the normalized impedance of 1.78 − jl.84 for the 50-ohm measurement system. When the 112.5-ohm system is employed, swr_2 (2.5) results in a value of ρ_2 = 0.43. The reflection coefficient electrical angle, θ_2, is 50.2°, found from $\tan\theta_2 = 1.2$. For this system the reflection coefficient in wave-lengths is 0.14λ. These values are also plotted on the Smith chart, but it must be remembered that the center of the chart represents 112.5 ohms + j0. This normalized impe-

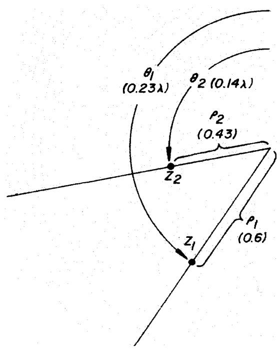


fig. 3. Test results shown on a Smith chart for the example described in the text.

dance is 0.79 – j0.82 and is shown as Z_2. Actually, both impedances, Z_1 and Z_2, are equal; only the characteristic impedance of the measurement system is different.

One other point should be mentioned. Suppose the load is actually 82 – j83.6, but a matching circuit is designed to match the source impedance to the measured value of 89 – j91.1. How suitable is this? A new reflection coefficient must be determined from:

$$\Gamma = \rho \angle \theta = \frac{Z - Z_0}{Z + Z_0} \tag{8}$$

where $Z_0 = 82 - j83.6, Z = 89 - j92.1$

Making these substitutions yields a reflection coefficient of ρ = ***0.045***, which results in an SWR of 1.09:1.

conclusion

While there are many methods of measuring impedance, the procedure described is straightforward and requires only one simple instrument, the SWR meter. Also, since so many handheld calculators are available, the calculations using the measured values can be accomplished very quickly. Other means of calculating can be employed, although more time is required. The results are acceptable for most practical applications.

references

1. Reference Data for Radio Engineers, International Telephone and Telegraph Co., 4th edition, page 587.
2. Ben Lowe, K4VOW, "A 15-Watt Output Solid-State Linear Amplifier for 3.5 to 30 MHz," QST. December 1971, page 11.
3. The Radio Amateur's Handbook, ARRL, Newington, Connecticut, April 1975, page 581.

ham radio

ITAL TECHNIQUES

flip-flop internal structure

Previous parts of this series have shown the three basic flip-flops, the latch, and clocked flip-flops JK and D. Only the RS latch has been examined for timing. Timing is crucial in proper operation of JK and D flip-flops, so it's worthwhile to examine some typical internal structures for timing relationships.

Fig. 1 shows a NAND-gate equivalent of a master-slave flip-flop with waveforms and "1 and 0" state notation. The master-slave term comes from using one latch (G3 and G4) to control the other latch (G7 and G8). State feedback makes each latch dependent on the other. The 1 and 0 notation is useful for scratchpad state analysis, and the state is that of each gate after the clock edge has passed.

JK flip-flops are commonly found in master-slave form, and **fig. 1** has both J and K control inputs held high for a divide-by-two function. A negative clock edge will change output state, so the symbol would use an inversion bubble at the clock pin.

Initial conditions assume all inputs high and Q low. Master latch G3, G4 could be in either state but is chosen with G3 initially high. All gate states are stable in the left-hand 1 and 0 notation.

The first negative clock forces G1 high. G5 then goes low since both inputs are high (the NAND *RULE*) and forces G7 (Q output) high which, in turn, makes G8 ($\overline{Q}$ output) low. Waveform arrows show the sequential state change. The other four gates remain in the same state as long as the clock is low.

Returning the clock high doesn't affect output but will set up conditions for the next negative clock edge. G2 goes low and forces intermediate latch G3 and G4 to change state. G5 returns to a high, and output latches G7 and G8 remain stable (both G5 and G6 are now high).

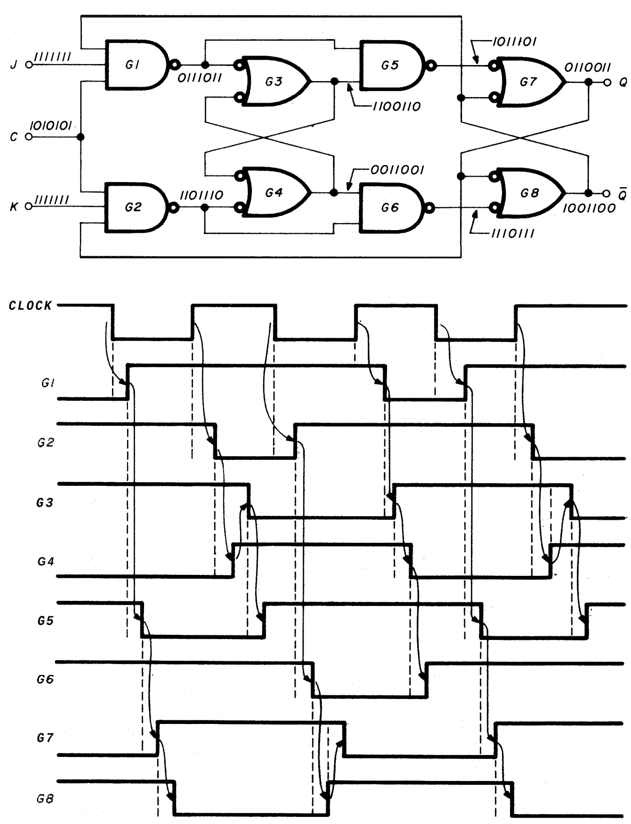


fig. 1. Master-slave flip-flop with J and K inputs held high.

By Leonard H. Anderson, 10048 Lanark Street, Sun Valley, California 91352

A second negative clock will make G2 high. G6 now goes low and flips the output latch. Q output changes state. Returning the clock high will flip the intermediate latch and set up conditions for another output state change.

An important thing to note is that both clock edges affect internal states and that each edge causes a sequence of four gate state changes with attendant delays. A high clock level must persist long enough to set up conditions for an output change. A low clock level must persist for a time sufficient to ripple-through state changes to toggle the output. Each time will limit maximum clock frequency.

changing control inputs

Fig. 2 has the same circuit but input K is held low and only J is changed. Initial conditions have Q low and J low. Since K is low, G2 will always remain high. Note that if both J and K were held low, there would be no output change at all since input gates G1 and G2 would be held high constantly. The intermediate latch has G3 low.

The first low clock does nothing. Returning it high has no effect either, since J is held low. When J goes high with the clock high, the intermediate latch flips from the low state of G1. G6 goes high and the flip-flop is set up for an output change.

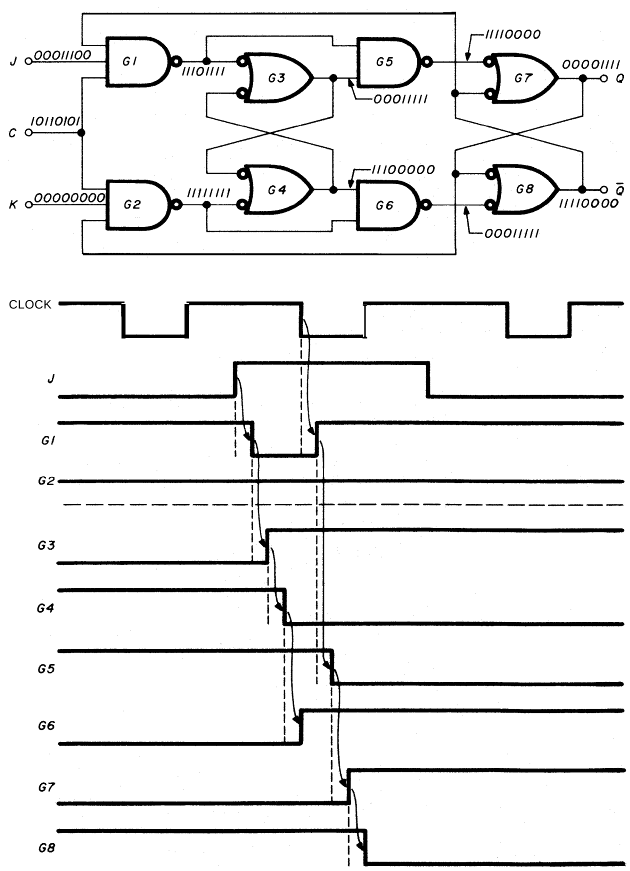


fig. 2. Master-slave flip-flop with K held low and J input switched.

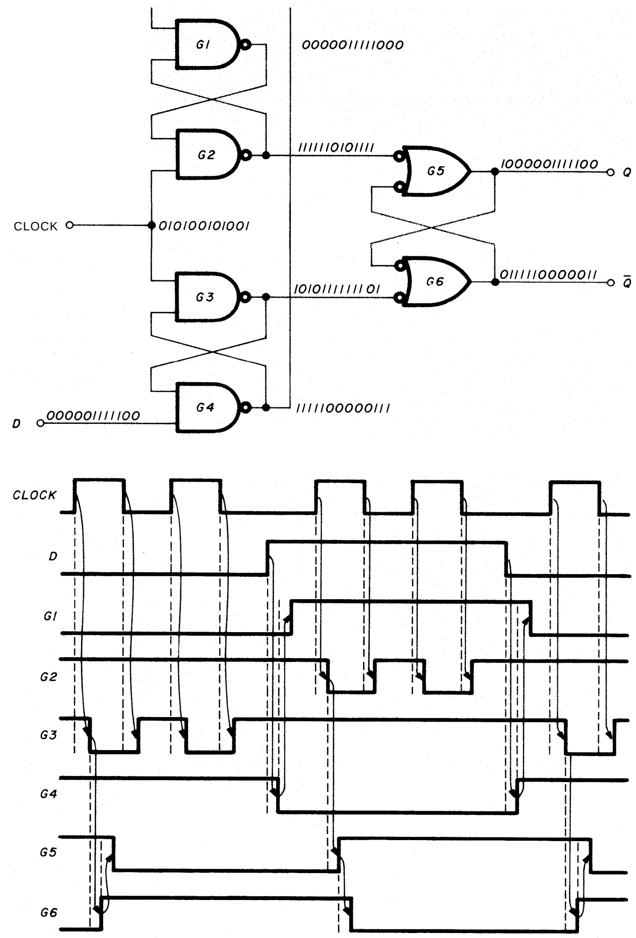


fig. 3. Type-D flip-flop.

The second low clock, while J is held high, will return G1 high and make G5 low. The output latch flips and Q goes high, matching the J input state. Again, four gate delays are required for both set-up and hold.

Once Q has toggled high, neither J nor the clock will have any effect. Why? The answer lies in state feedback from G8 ($\bar{Q}$) to G1 input. The low G8 state will inhibit any inputs from changing G1 (the *NAND* RULE again).

Holding J low and changing only K will result in similar action, except that Q will eventually go low and remain in that state. Try this out on some scratchpaper.

type D flip-flop

A NAND gate equivalent is given in **fig.** 3. This version changes output on a ***positive*** clock edge. Gates G1 through G4 are "almost" latches with an *AND* symbol shape; their action is best observed by following input changes.

Initial state has Q high and D low. With both D input and clock low, G2 through G4 are held high. G1 is held low by state feedback from G4. Output latch

G5, G6 remains stable since all inputs (from G2 and G3) are high.

The first high clock will force G3 low; its other input from G4 is also high. G3 then forces G6 (Q) high to flip the output latch. Clock return to low will change only G3, as does a second clock input. G3 stays high on a low clock.

Changing D input high while the clock is low will force G4 low, which, in turn, forces G1 high. Input gates are now set up to change the output. A third positive edge clock will make G2 low and flip the output latch to a Q high state. As long as D remains high, further clocks will not change Q.

Returning D low while the clock is low will force G4 high, then G1 low to finish a setup. The cross connections of G1, G2, and G3, G4 appear to make two latches. These, plus G4 to G1 connection, make one large latch, not two, with G1 and G4 acting as inhibits for G2 and G3.

This structure is faster than the master-slave JK. Setup requires only two gate delays, output change only three.

direct set and reset

A direct set or reset will override the clock and any control inputs. The master-slave JK may be modified by increasing the latch gate inputs from two to three. A SET (active low) is connected to both G3 and G7. A RESET (again, active low) is connected to both G4 and G8.

The D flip-flop is also modifiable for active-low set or reset but is more complex. A RESET is made to G2, G4, and G6, a SET made to only G1 and G5. The number of gate inputs must be increased for either.

Actual devices may have direct set and reset (sometimes called preset and clear, respectively) either active low or active high. Check for inversion bubbles. Some dual devices have either or both common. Check thespecsheet. In any situationan unused set or reset must be made inoperative. An unused active low should be tied high; unused active high tied low.

Removing a set or reset will restore a JK or D to normal clocked operation. Some time is required for this restoration, similar to setup and hold times. The spec sheet for a particular device should be checked for this and proper time alloted in circuit operation.

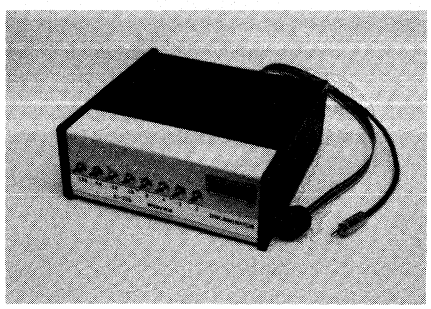
The next part of this series will go into interfacing analog signals and present one-shots.

ham radio

the ham notebook

external frequency programmer

The beauty of the IC-22S is that it offers precise frequency selection by programming with low-cost silicon diodes, rather than crystals with trimmers. This advantage though becomes overshadowed when channel selections ne[illegible] to be chang[illegible] out comes the soldering iron to redo the diode matrix. To solve this problem, I built an external programmer containing eight switches, eight diodes, and a zero-center microammeter for discriminator output readings.



External frequency programmer enclosed in a Ten-Tec style box. The connections to the transceiver are made via nine-conductor ribbon cable.

The internal speaker was disconnected from the external speaker jack and reconnected directly to the audio output. The discriminator output was then connected to the external speaker jack. The eight diode/cathode connections and the 9-volt common pad of channel 22 on the matrix board were connected to the nine pins of the accessory socket via a nine-conductor ribbon cable (see **fig. 1**).

The external programmer was built in a Ten-Tec cabinet. The discriminator meter was connected via a shielded cable and miniplug. Nine-conductor ribbon cable was used to connect the programmer to the accessory plug.

With the programmer, any frequency can be selected at the flip of a switch and at a considerable savings over the cost of a fully programmable 2-meter rig.

Hugh Pearl, WB9VWM

using the IC-22S below 146 MHz

As it turns out, the IC-22S is not restricted to 146.01 MHz and above. The IC-22S is, in fact, usable without modification from 145.350 MHz through 147.990 MHz. That's an additional 44 free channels! These additional channels do accommodate some of the new repeater frequencies as well as five Oscar frequencies. By keying the push-to-talk line you can also enjoy the excitement of Oscar 7 and Oscar **8.**

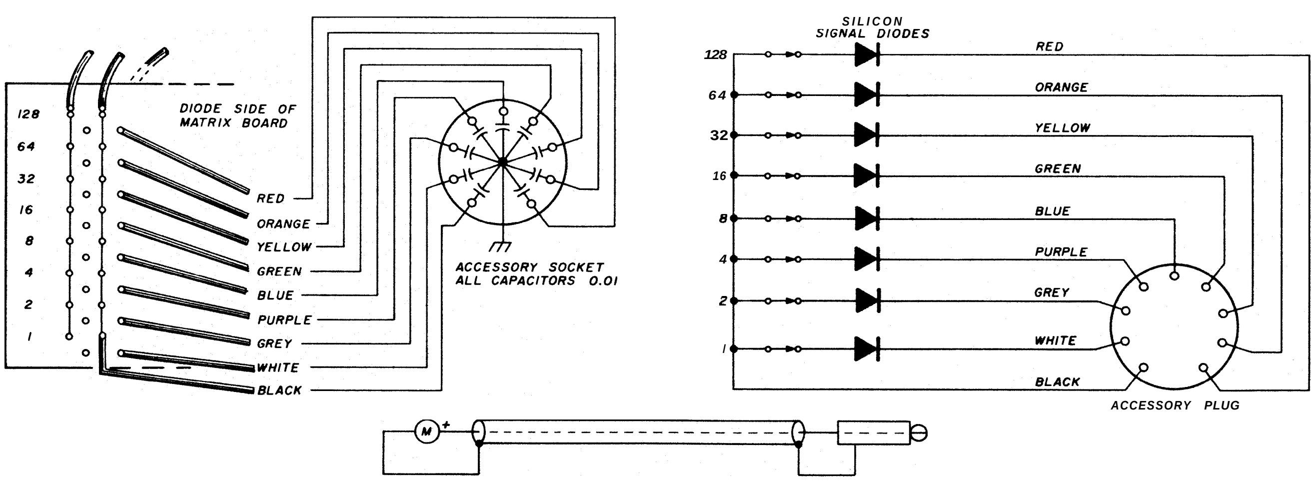


fig. 1. Diagram of the connections to the diode matrix board and switch connections for the external frequency programmer. The discriminator meter is connected to the rewired external speaker jack.

table 1. Diode programming matrix for frequencies below 146 MHz.

frequency MHz	total N	128 D7	64 D6	32 D5	16 D4	8 D3	4 D2	2 D1	1 D0
145.350	64		*						
145.365	65		*						*
145.380	66		*					*	
145.395	67		*					*	*
145.410	68		*				*		
145.425	69		*				*		*
145.440	70		*				*	*	
145.455	71		*				*	*	*
145.470	72		*			*			
145.485	73		*			*			*
145.500	74		*			*		*	
145.515	75		*			*		*	*
145.530	76		*			*	*		
145.545	77		*			*	*		*
145.560	·78		*			*	*	*	
145.575	79		*			*	*	*	*
145.590	80		*		*				
145.605	81		*		*				*
145.620	82		*		*			*	
145.635	83		*		*			*	*
145.650	84		*		*		*		
145.665	85		*		*		*		*

frequency MHz	total N	128 D7	64 D6	32 D5	16 D4	8 D3	4 D2	2 D1	1 D0
145.680	86		*		*		*	*	
145.695	87		*		*		*	*	*
145.710	88		*		*	*			
145.725	89		*		*	*			*
145.740	90		*		*	*		*	
145.755	91		*		*	*		*	*
145.770	92		*		*	*	*		
145.785	93		*		*	*	*		*
145.800	94		*		*	*	*	*	
145815	95		*		*	*	*	*	*
145.830	96		*	*					
145.845	97		*	*					*
145.860	98		*	*				*	
145.875	99		*	*				*	*
145.890	100		*	*			*		
145.905	101		*	*			*		*
145.920	102		*	*			*	*	
145.935	103		*	*			*	*	*
145.950	104		*	*					
145.965	105		*	*					*
145.980	106		*	*				*	
145.995	107		*	*				*	*

The programming of the IC-22S diode matrix is governed by the equation:

$$programmed\ number\ (N) = \frac{desired\ frequency - 144.39}{.015}$$

The resulting number must be an integer (no fractional part). The decimal number computed from the equation is then translated into a binary number and programmed accordingly into the diode matrix. For convenience, **table 1** presents the additional frequencies and the corresponding diode positions in a format similar to that in the instruction manual. The simplex-offset switch functions the same as for other frequencies below 147 MHz.

Steven Holzman, W1IBI

75S ( ) CW sidetone

The CW sidetone provided by my 32S-1 transmitter is of sufficient amplitude to give an adequate monitoring level while using headphones. Even when low-impedance headphones (stereo types with the sections paralleled) were used, no problems were encountered here.

In an attempt to increase the sidetone level while using a speaker, the value of R99 (2.2 meg ohms) was altered. This change proved insufficient, and, if carried too far, vox operation on ssb was hampered. The best balance between speaker, headphones, overall receiver volume, and CW sidetone level was obtained in the following manner.

Remove the two wires connected to the SIDETONE jack on the receiver rear lip, solder them together, and insulate. Using a piece of shielded wire (RG-174/U in my case), connect one end to the SIDETONE jack and the other end to the center (wiper) terminal of the AF GAIN control. The cable should be neatly routed around the chassis following the path of the existent harness. Refer to **fig. 2** for the complete wiring changes. The result is a sufficient sidetone level for both speaker and phones.

Paul Pagel, N1FB

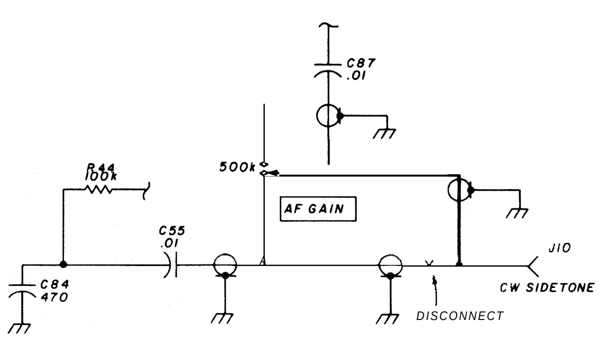


fig. 2. Sidetone modification for the 75S() series receivers. In this case, the level of the sidetone is controlled by the AF GAIN, providing a better balance between signal and sidetone levels. The heavy line indicates the added shielded cable.

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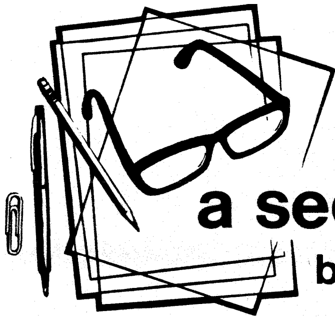


a second look

by Jim Fisk

Whether on or off the air, there's probably no Amateur Radio subject discussed more, scrutinized more carefully, or analyzed more qualitatively than antennas — the engineers analyze, the DX operators scrutinize, and the rest of us simply talk. And with the coming of spring and summer, interest in antennas peaks as Amateurs start working on their antenna systems for the coming DX season. If you have been thinking about updating your station antenna, you may find a useful design in this, our annual sntenna issue. If you don't plan to install a new antenna this year, this is a good time to take a close look at your existing system to make sure it's operating up to par. Now that the ravages of winter are over, check all the electrical connections for oxidation, inspect the feedline for damage, and if your system SWR is not as low as it used to be, now's a good time to find out why.

If you look at the record, there have been few important breakthroughs in Amateur antenna design since the delta-loop beam and the quagi, and many readers would argue that both of these antennas are actually extensions of existing antenna theory (quads and Yagis). Many specialized antennas have been developed for military and space communications in recent years, but most have little or no application in Amateur Radio. One example is the log periodic or LP, first designed as a microwave antenna, later as a high-frequency wire beam, and finally, for TV reception. The LP has never been especially popular with Amateurs, probably because we operate on segmented bands and the LP is a broadband device. However, there is renewed interest in the LP for DX work on 80 and 40 meters, and if Amateurs are allocated any new high-frequency bands at WARC 79, it's a sure bet that a good many three-band beams will be replaced by rotatable log periodics.

Unlike other areas of electronics and radio communications, where there have been occasional quantum leaps forward, improvements in basic antenna design have been slow and evolutionary, beginning in 1887 with Hertz' classical center-driven dipole. Marconi's first successful wireless experiments used essentially the same antenna: Two large copper plates excited by a spark gap. To increase the distance of his transmissions, Marconi put up larger and larger quantities of wire. Early Amateurs did the same — because of the low frequencies then in use, the DX performance of a station was proportional to the size and height of the station antenna. In fact, many of the radio engineering textbooks of the early 1900s included a mathematical equation attributed to Marconi which showed that communications distance was directly proportional to the square root of antenna height.

The first advances in gain antennas were the result of experiments by commercial radio companies which wanted to improve the reliability of their overseas service. First came the long wire, the vee, and the rhombic; then the lazy-H, Sterba curtain, and Bruce array, followed closely by the Yagi beam and the 8JK flat-top; W9LZX's cubical quad was introduced a few years later. Each new antenna design was eagerly tested by Amateurs who were trying to improve the performance of their stations.

The perpetual need to improve station performance has become something of a tradition in Amateur Radio, and whether it's the result of ever greater band crowding or the competitive Amateur spirit, I don't know, but it certainly is a fact. Single-sideband provided the needed improvement in the early 1960s, and compact kilowatt linears made their contribution in the late 1960s; that left the antenna system. The fact that there were no antenna breakthroughs did not dilute the compelling urge to enhance station performance — if you couldn't improve basic antenna layout and design, you could certainly increase antenna size, and hence, antenna gain.

As an example of this trend, consider the great interest in phased arrays for 40 and 80 meters, sparked primarily by W1CF's *QST* articles and his big signal on 75 meters. It wasn't too many years ago that most of us were content with a ground-plane antenna on 7 MHz and a dipole (usually not too high) on 3.5 MHz. In those days some of the more serious 40-meter buffs had two element-vertical arrays, and you might have read in one of the magazines about a few daring souls who had 40-meter Yagi beams, but you almost certainly had never actually seen one. In recent years, however, big low-frequency arrays and full-sized beams have become relatively commonplace, and the same rationale for greater antenna height and size has begun to permeate 10, 15, and 20 meters. As the late Sam Harris, W1FZJ, used to say when talking about vhf antennas, "If it lasted through the winter, it wasn't big enough." Based on the large number of antennas which came tumbling down this past winter, perhaps we've hit the practical limit!

Jim Fisk, W1HR
editor-in-chief

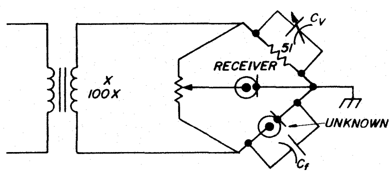
comments

RX noise bridge

Dear HR:

I found Ted Hart's letter in the August 1978, issue of *ham radio* both interesting and informative. I believe Pappot was the first to describe the R-X Noise Bridge *(ham radio,* January 1973), but he did not reference the source for the original, resistance-only noise bridge which he improved.

Hart's design, which is marketed as the Model TE 702, is technically interesting but would have an awkward capacity readout range if modified as indicated. Using the resistance values



he suggested, calling the fixed and variable capacitors C_f and C_v and letting X represent the resistance from the pot wiper to the upper corner of the bridge, the conditions for balance become:

$$R_u = \frac{51(100-X)}{X}$$

$$C_u = \frac{X \cdot C_v}{100-X} - C_f$$

where R_u and C_u are, respectively, the parallel resistance and capacitance of the unknown impedance. As Mr. Hart stated, the resistance range is zero to infinity. Furthermore, with a linear taper pot, the low end of the resistance scale would be spread out and the upper end compressed. For many applications, that would be more desirable than the linear and finite range of the Pappot design.

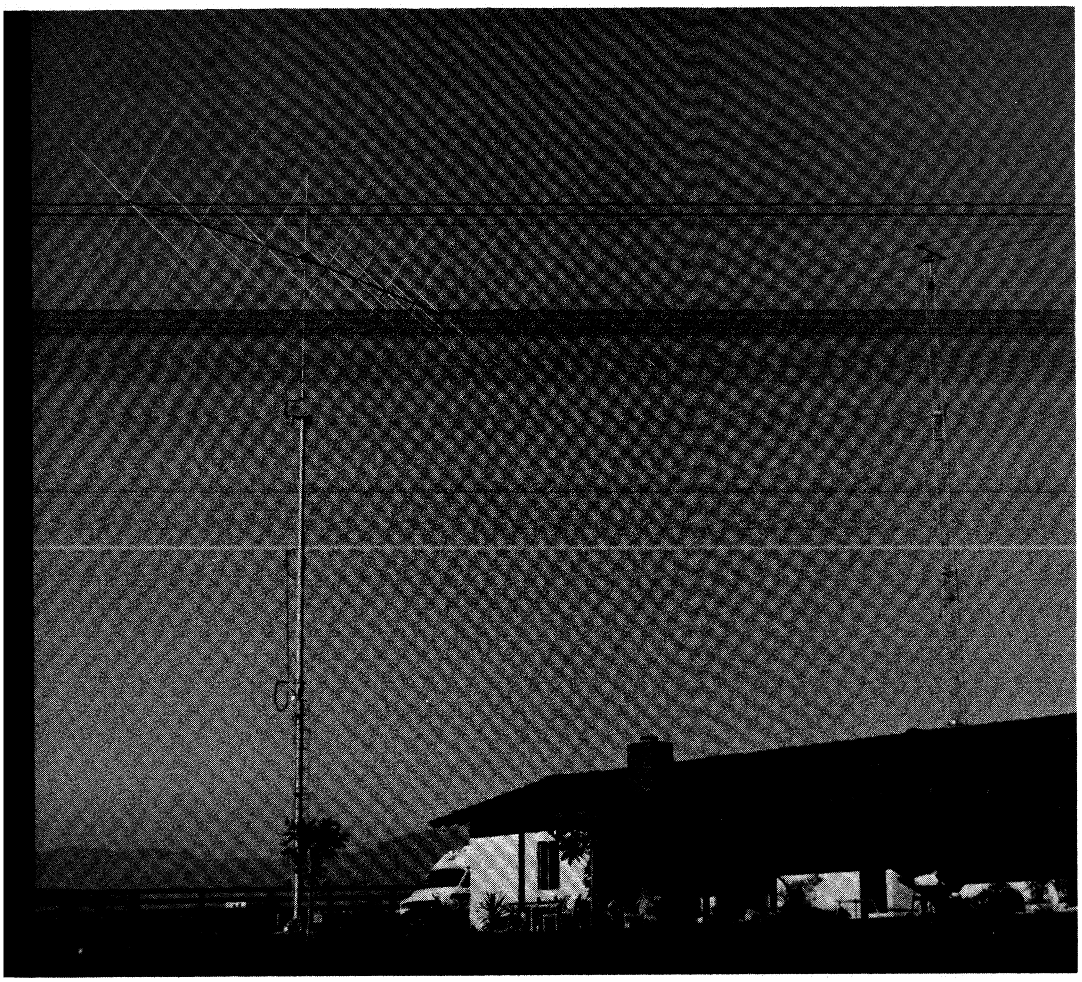
Conversely, the reactance or capacitance range of Hart's proposed reactance bridge would be awkward or unusable. Note that the estimated value of the unknown capacitance is affected both by the setting of the 100-ohm pot *and* the variable capacitor. The effective capacitance range is thus dependent upon the resistance measured. Suppose, for instance, that C_v is a standard broadcast variable with a range 20-365 pF. With C_f a 180 pF capacitor, the range for capacitance balance with $R_u = 51$ ohms would be $-160\ pF \leq C_u \leq 185\ pF$, which is similar to Pappot's R-X bridge as modified by Gehrke *(ham radio,* March 1975). But suppose R_u were 5 ohms. Because of the interaction between pot setting and reactance balance, the capacitance range would now be $20\ pF \leq C_u \leq 3470\ pF$. One could not even measure a pure resistance! Similarly, with R_u equal to 250 ohms, the capacitance range becomes $-176\ pF \leq C_u \leq -107\ pF$ and again does not include a purely resistive impedance. Of course, other values could be chosen for the capacitors, but the effect described above would still cause the reactance range to be severely distorted for R_u values away from 50 ohms.

Mr. Hart also makes the point that his model TE 702 works to over 250 MHz and that the reactance measuring version thereof would yield "a lot more accuracy — over a wider frequency range." It is well known that, with careful design, resistance-only noise bridges can be made to work well up into the vhf range. Adding the fixed and variable capacitors, however, severely reduces the upper frequency limit. The R-X noise bridge Doting and I described works accurately to 30 MHz, and data to substantiate this claim have been published (*ham radio,* February 1977). The two commercial R-X noise bridges claim upper frequency limits of 100 MHz. However the manual for one of these units suggests it may require recalibration for use at the higher frequencies; neither makes any claim for accuracy.

Quite independently of how the transformer may be wound, the upper frequency range of the R-X bridge is limited by residual errors in the bridge components themselves. The pot has stray capacitance to ground which changes with the pot setting, and the interconnection wires have reactances which cannot be ignored; most important of all, the variable capacitor has residual series inductance. 250 MHz is well above the series resonant frequency of any reasonable-size variable capacitor. Severe resistance and reactance measurement errors would start to appear well ***below*** the resonant frequency. These constraints are recognized by commercial equipment designers; where accuracy is required in vhf impedance-measuring equipment, bridge techniques using discrete components are typically abandoned.

I congratulate Mr. Hart for his invention, the Antenna Noise Bridge. The R-X Noise Bridge pioneered by Pappot was certainly inspired by this invention. Refinements to the Pappot unit which allow calibration and compensation without access to lab standards make the R-X bridge a very useful device for accurate measurements in the high-frequency range.

R. A. Hubbs, W6BXI
Orange, California



quads vs yagis revisited

The author attacks an old question in a new way by actually measuring the relative gain of high-frequency quads and Yagis

Which is the better antenna – a cubical quad or a Yagi? That question has aroused Radio Amateurs' emotions for more than thirty years now, and remains controversial in spite of the great volume of published research and analysis.

It has been a decade since Jim Lindsay published his classic work on quads and Yagis,[1,2] presenting both theoretical and empirical data to support his conclusion that the quad was in many ways the better antenna. He concluded that a quad of any length would outperform a similar size Yagi by 2 dB in forward gain. Basing his findings on model quads and Yagis operating at 440 MHz, Dr. Lindsay said that it was necessary for the Yagi to have 1.8 times the boom length of a quad to equal its forward gain.

At the time of its publication, Lindsay's study seemed conclusive, but in the years since the issue has refused to die. The question has now been subjected to computer modeling, which produced data

By Wayne Overbeck, N6NB, Pepperdine University, Malibu, California 90265

supporting Lindsay;[3] this was followed by a rejoinder from the owner of a highly successful Yagi-based contest station who pointed out sigificant errors in the modeling techniques.[4] That was followed immediately by a reply from the authors of the computer modeling study in which they conceded their errors, but stood by other parts of their work.

Meanwhile, Hardy Landskov, an antenna experimenter with a very pragmatic goal (to build a cubical quad that would be truly competitive in radio contests), spent many hours experimenting on antenna range owned by the United States Navy.[5] He came up with a design that the first user rated "good" on 20 meters and "excellent" on 15 and 10, in comparison to Yagis with which he'd "slugged it out" in pile-ups. During his antenna range work, Landskov built one 336-MHz Yagi and tested it against his quad design. He found the two equal in gain, although the Yagi was 1.65 times the quad's length.

popularity contests

The quad-versus-Yagi question has been argued with anecdotes, testimonials, and hunches, as well as with theory and quantification. Intuitively, some quad builders have concluded that their antennas were just not 2 dB better than similar-size Yagis. Others have reached the opposite conclusion by trusting the same intuitive sixth sense.

Moreover, one need only survey the antennas used by the consistent DX contest winners to see that Lindsay's conclusions remain controversial. Although there are notable exceptions, the great majority of big winners in DX contests today are using Yagis, not quads. If a quad is equal to a Yagi nearly twice its size, how could this be? Obviously, anecdotes, testimonials, and popularity contests, taken alone, prove nothing about the merits of the two antennas. But they do suggest that the issue is unresolved and deserves further study.

In an attempt to resolve this issue, I've spend hundreds of hours measuring the performance of quads and Yagis in both the high frequency and vhf spectrum and have encountered considerable evidence that the cubical quad is *not* inherently superior to the Yagi.

This seems to be particularly true below 100 MHz. However, there are potential variables that may sometimes bias the results of uhf modeling in favor of the cubical quad.

what happens at uhf?

My quads-and-Yagis project really began as an attempt to develop an efficient antenna for the 432-MHz Amateur band. In 1970, I bought a 432-MHz Yagi of the only brand then commercially available, and optimistically took it to an antenna gain measuring session at the West Coast VHF Conference. The gain was measured at 6.4 dBd (dB over a dipole) — a shocking figure compared with the manufacturer's claim of 13 + dBd. What was wrong?

The element lengths and spacing seemed acceptable, if not optimum, and the SWR was below 1.5:1. Nevertheless, I finally replaced the driven dipole with a quad-type full-wave loop. The gain improved to almost 10 dBd (as measured at the next West Coast VHF Conference).

That led to the next logical step. I set up a backyard antenna range and set out to design a good 432-MHz cubical quad. After many frustrating hours, I decided I couldn't come up with a combination of element lengths and spacing that would beat the hybrid quad-driven Yagi at 432 MHz. Up to four elements, the quads were competitive, but beyond that the quad-driven Yagi was better.

Note that I said I couldn't design a quad that would beat the *hybrid.* A mere 4-element quad was superior to the original 11-element Yagi with its "stock" driven element on 423 MHz, even though the quad was less than half the Yagi's length!

Finally, of course, I gave up on the quad and set out to optimize the quad-driven Yagi. Many hours and many designs later, the design that has come to be known as the *Quagi* emerged. Since the original publication of that design,[6,7] the Quagi has been duplicated all over the world, with the design published in Amateur journals in such diverse countries as the Soviet Union, South Africa, and India!

At about the same time, a Danish scientist working in an anechoic chamber also observed the phenomenon that led to the Quagi antenna. He was working with antennas that combined loop driven elements and reflectors with three types of parasitic directors — loops, Yagi-type rods, and flat plates.[8] He reported that for long-boom arrays (over two wavelengths), the loops were the *least effective* directors, although the difference was never greater than 1 dB.

free dB?

When the Quagi design was published, many Amateurs assumed its main advantage was a "free" 2 dB of extra "loop gain," and that the gain of any Yagi could be improved 2 dB by replacing the driven element with a full-wave loop. This is not the case, unless the original dipole is less than adequate.

For instance, while the gain of the 432-MHz Yagi previously mentioned increases dramatically when the dipole is replaced by a loop, the same is *not* true of the 144-MHz version of that antenna. Replacing the two-meter model's dipole with a loop results in almost no change in gain. At that frequency, it would appear the gamma-matched dipole is still a reason-

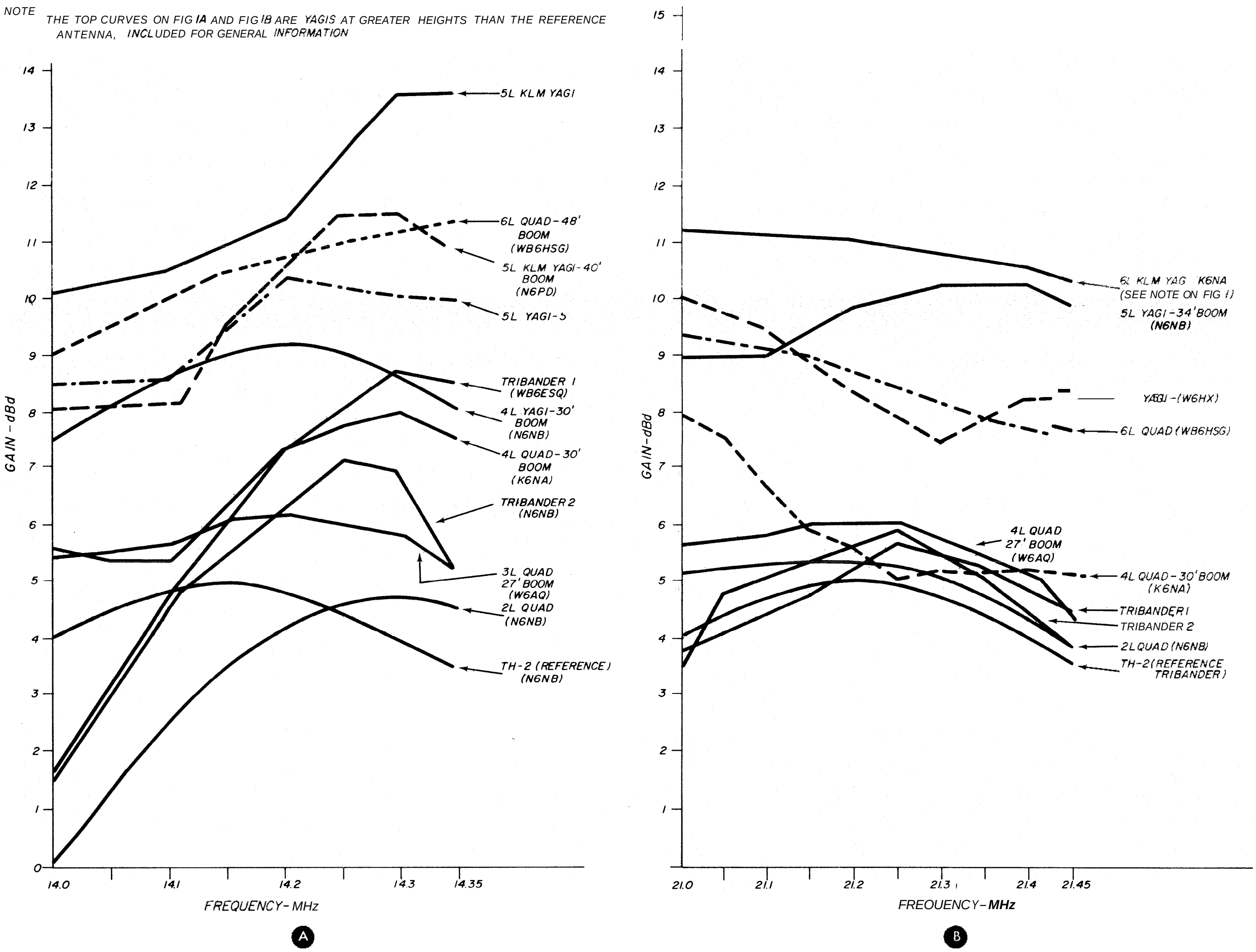


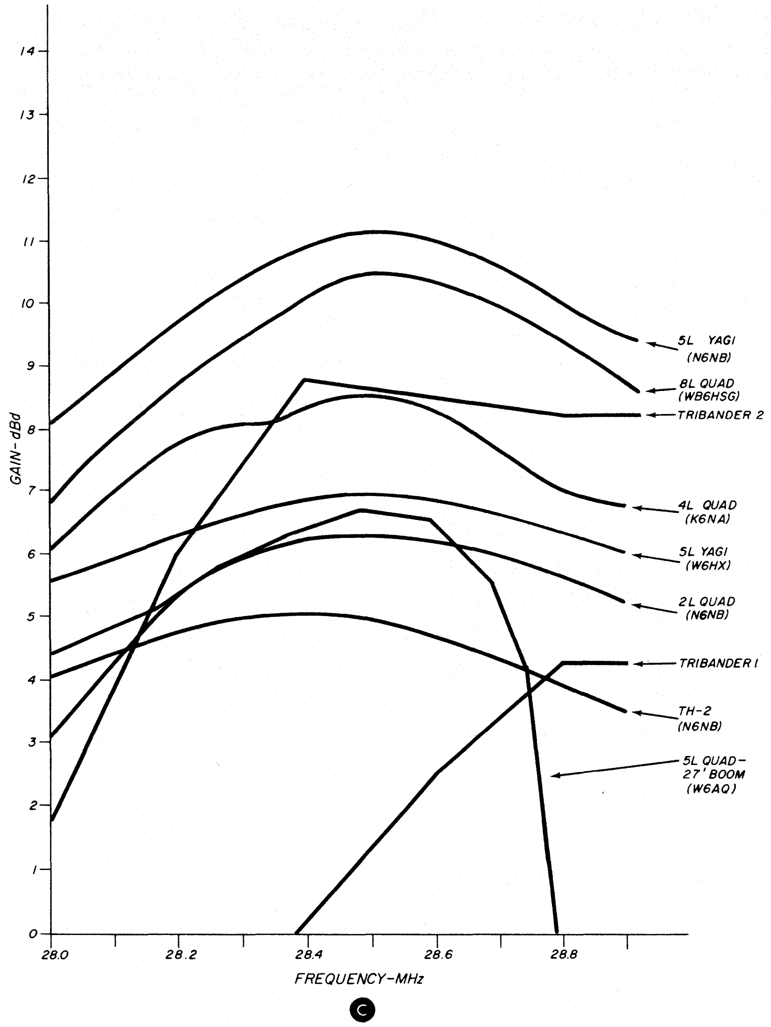
fig. 1. Relative gain measurements for 20, 15, and 10 meters. (A) shows 20 meters, (B) 15 meters, and (C) 10 meters. The top curves on 20 and 15 meters are for a Yagi which is at a greater height than the reference antenna. They were included only for general information, not for direct comparison.

ably efficient driven element. Nor does the Quagi's driven loop automatically ensure it 2 dB more gain than similar-size Yagis that have solved the feed problem in other ways. At 432 MHz, two of the 15-element Quagi's toughest competitors in antenna gain competitions are the KLM 16-element *log-periodic-driven Yagi,* and the F9FT 21-element array, which has a large oval-shaped folded dipole sometimes called a "flattened quad." What the vhf Quagi offers is the gain of a long-boom Yagi with an *efficient driven element* that can be easily duplicated using tools and materials readily available to Amateurs. It *does not* offer 2 "free" dB.

When most Radio Amateurs, and some commercial antenna builders, set out to build a conventional Yagi for uhf, the result may well be disappointing. Anyone who has attended a few antenna gain measuring sessions at vhf conferences has seen seemingly well-built 432-MHz Yagis measure very low gains, despite all indications that they should work well (*i.e.*, good SWR, good E- and H-plane patterns, and so forth).

At 1296 MHz, building an efficient Yagi is even more difficult, and no commercial manufacturer currently attempts to do so for the Amateur market. But there is a commercial "loop-Yagi" available, an effective antenna of quad-type full-wave loops. For whatever reasons, quads seem to retain near-high-frequency efficiencies at higher frequencies than do conventional Yagis.

Nevertheless, in previous comparisons of quads and Yagis, much has been made of the results of uhf modeling, with authors sometimes citing the sophistication of their antenna ranges as evidence of the validity of their results. Now no one would question the principle of uhf modeling for designing antennas.



All of my antennas were experimentally optimized that way. One can derive good element spacings and lengths, and scale those values to lower frequencies, as long as frequency-related variables such as changing length-to-diameter ratios are not overlooked.

However, quad-versus-Yagi studies are a little different. Here someone is generalizing about the *comparative* performance of *two different antennas* at high frequencies on the basis of the comparative performance of uhf models of those antennas.

You can not consider it good engineering practice to compare two power transistors, for instance, at one frequency and then generalize about their relative performance at another frequency an order of magnitude removed. But our assumptions about quads and Yagis have been based largely on just such extrapolations! Those assumptions would be valid if quads and Yagis were equally efficient at high frequency and uhf (or declined in efficiency at the same rates). However, with antennas that exist in the real world, as with other rf-handling components, there may well be frequency-related variations in efficiency.

All one can really say, after a series of uhf experiments comparing quads and Yagis, is something like this: "These particular quads outperform these particular Yagis (or vice versa) in this frequency range." The fact that uhf modeling (including mine) tends to produce pro-quad results does not necessarily settle the matter on the 10- 15-, and 20-meter bands.

high frequency experiments

With these questions about comparative antenna data from 400 to 14 MHz in mind, I felt it was time to attack the quad-versus-Yagi question in a new way, by actually measuring the gain of quads and Yagis in operational high frequency installations.

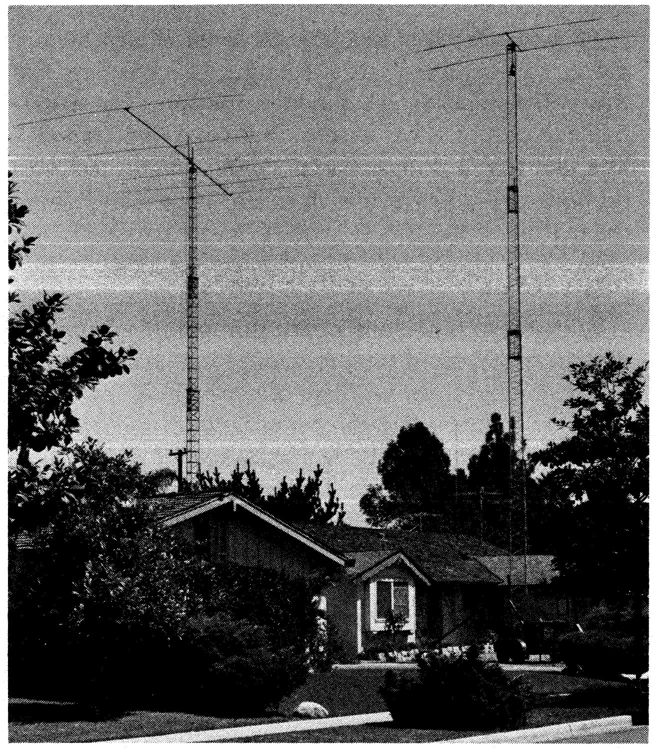
Using the measuring techniques described in a previous article,[9] I have rated the gain of several large 20-, 15-, and 10-meter antenna systems in **fig. 1.** Briefly, the technique involves placing a directional reference antenna alongside each antenna to be tested (at the same height) and reading the strength (in dB) of a nearby signal as received by each antenna. The signal source is a directional antenna of the same polarization as the test and reference antennas, located at least 40 wavelengths, but not more than about 5 km distant. The receiver agc is disabled and its audio feeds a Hewlett-Packard 400L audio voltmeter.

How do you place a reference antenna beside various antennas that may be 70 or 80 feet in the air? The answer, of course, is a self-supporting crankup tower on a trailer. I just happen to own not one but two such monsters; they were built for portable contest operating.

As a reference antenna, I selected a 2-element triband Yagi (Hy-Gain TH-2) because reference dipoles sometimes "see" reflections that can produce misleading results when a more directional antenna will not. All results are relative to the 2-element tribander. Its gain was assumed to be 5.0 dBd at its center frequency on each band. If that gain rating is too high (or low), all the other numbers will be incorrect on an absolute scale, but *correct in relation to each other,* which is what really matters for our purposes.

20-meter results

If the cubical quad antenna was to look good anywhere, 14-MHz turned out to be the place. Among the antennas I measured was the 6-element quad (8 elements on 28 MHz) at WB6HSG. It has a 14.6-meter (48-foot) boom. Patterned after Landskov's design but with a second reflector (added because quad pioneer C.C. Moore advised WB6HSG that a second reflector would surely provide more coupling, and hence more gain and front-to-back



A 5-element, log-periodic-style, KLM 20-meter monobander. The author's reference tribander is positioned beside the antenna, which is owned by NGPD. This KLM monobander had the highest measured gain of any 20-meter antenna (tested at the same height as the reference) over much of the phone band, but trailed the larger WB6HSG quad on CW.

ratio, than another director), the antenna is impressive in both appearance and performance.

As fig. 1 shows, this quad outperformed any Yagi tested (except a much higher one) over much of the 20-meter band. (The higher antenna is K6NA's 5-element log-periodic Yagi at 32.3 meters (106 feet). It was included for general information, but not considered a part of the study. Note how it compares with an identical Yagi owned by N6PD which was at the same height as the reference antenna.)

It should be pointed out that only one Yagi, an array surrounded by wires and other towers at W6HX, was as big as the quad at WB6HSG. If Lindsay's thesis were correct, WB6HSG's quad would be expected to beat all of these Yagis by at least 2 dB. In fact, the 12.2-meter (40-foot) log-periodic Yagi that was 21.3 meters (70 feet) high topped the bigger quad in much of the phone band.

Aside from the performance of WB6HSG's quad on 20 meters, these tests consistently produced data favoring the Yagi design. In no other case did the long-boom quads even equal the gain of comparable size Yagis, let along exceed their gain by anything like 2 dB.

An interesting example is the 4-element quad at K6NA, which has a 9.1-meter (30-foot) boom and uses Lindsay's spacing. K6NA (ex-W6MAR) has long been concerned about the quad's seemingly poor performance. In fact, its apparent mediocrity provided inspiration for Landskov's attempt to optimize the quad design. This is the quad Landskov specifically cited as an example of a poor performer in his article.

After Landskov's research was completed, K6NA took the quad down and moved to a new home. When put back up, the quad retained the original spacing but used Landskov's dimensions for the elements (which differ from Lindsay's mainly in that the percentage differences between the driven and parasitic elements are greater). For whatever reason, the data in fig. 1 show that it still performs poorly. Both Landskov's larger quad design (as adapted by WB6HSG) and a number of Yagis proved to be substantially better than this quad.

On 20, even my 2-element quad was disappointing, comparing unfavorably with the 2-element trap tribander (on a 2-meter [6-foot] boom) used as a reference!

15-meter results

On 15 meters (fig. 1B), the results were much the same, except that my 2-element quad squeaked past the reference tribander by 0.25 dB across most of the band. Neither big quad matched my 5-element Yagi on a 10.4-meter (34-foot) boom. At one embarrassing point (21.250 MHz), K6NA's 4-element quad fell to within 0.1 dB of the gain of the 2-element trap tribander.

Both WB6HSG's and K6NA's quads appeared to be resonant below the bottom of the band. Since both were cut to Landskov's dimensions and also provided their best SWR at 21.0 MHz, it appears his dimensions are too long on this band. WB6HSG has since shortened all his 15-meter elements and reports a much better SWR and front-to-back ratio in the phone band.

As on 20 meters, K6NA's higher Yagi was tested and plotted on 15, but the results are not considered comparable. Still, an antenna like that helps to explain K6NA's success in contests, doesn't it?

10-meter results

On 10 meters (fig. 1C), my 2-element quad finally behaved like an antenna without traps, topping the little tribander by up to 1.5 dB. Nevertheless, fig. 1C could not be used to support Lindsay's thesis. The top curve is a commercial 5-element Yagi on a 7.9-meter (26-foot) boom, little more than half the boom length of WB6HSG's big quad. Incidentally, this Yagi was measured directly against the big quad. After I compared the quad against the reference tribander, I lowered the portable tower and put up the 10-meter Yagi. The Yagi had the edge at every point in the band where we took a reading (each 100 kHz).

K6NA's quad was consistently 2 to 2.5 dB down from the smaller Yagi. It did, however, outperform a 5-element Yagi at W6HX, but that antenna was known to be defective before my measurements, perhaps due to driven element problems. For contest work W6HX uses another 10-meter Yagi considered to be "1 to 2 S-units better," but that antenna could not be measured due to logistics problems.

front-to-back ratios

Another claim often made for the cubical quad is a superior front-to-back ratio. The emphasis in the previous tests was on forward gain. We did measure the front-to-back ratio of WB6HSG's long quad on 10 meters, since he felt it was particularly good on that band. It varied from 18.2 dB at 28.4 MHz to an excellent 28.1 dB at 28.8 MHz. For comparison, the source antenna for those tests, a Hy-Gain TH-6 tribander at N6MB, was rotated. Its front-to-back ratio varied from 19 to 25 dB across 10 meters.

While this test did not show the long quad to be dramatically superior to a tribander in front-to-back ratio, comparing my 2-element quad against the reference 2-element tribander did reveal a substantial edge for the quad. On all three bands, the quad exhibited in excess of 20 dB while the tribander never exceeded 14 dB.

high and low quads

When this data was first presented at a meeting of the Southern California DX Club, two questions were raised that led to additional research. First, someone pointed out that virtually everything I had done was at a height of about 21.3 meters (70 feet). "A quad really comes into its own is at low heights," someone suggested.

Almost everyone who has written about quads has denied that a quad provides a lower angle of radiation than a Yagi.[10] Nevertheless, I set up two 21.3-meter (70-foot) towers side by side near a body of salt water (at the Ventura marina) to test the relative performance of a quad and a Yagi at high and low heights. The over-salt-water site was selected to eliminate the possibility of actual ground being so far below the earth's surface as to invalidate the experiment.

The result was conclusive; the relative performance of the quad and Yagi was exactly the same at both 7.6 meters (25 feet) and at 21.3 meters (70 feet)! This was true on zero-angle line-of-sight signals, high-angle "short skip," and on long-haul DX signals. There was no height at which the quad's relationship to the Yagi changed. The curves rating the two antennas could have been derived at any height between 7.6 and 21.7 meters (25 and 70 feet).

table 1. Test results for different 50-MHz monoband quads vs a 5-element Yagi.

antenna	gain over reference
4-element long-boom quad, 5.3-meter (17.5-foot) boom	– 0.65 dB
4-element Machin "optimum" quad, 4.5-meter (14.75-foot) boom	– 0.85 dB
5-element quad, Landskov design, 2.7-meter (9-foot) boom	– 0.90 dB
3-element quad, 0.21λ spacing, 2.5-meter (8.33-foot) boom	– 1.05 dB
4-element quad, Lindsay design, 2.6-meter 18.5-foot) boom	– 1.55 dB

antenna height — 1.5 wavelengths

source — 3-element Yagi at various points on a nearby hill, with vertical angles ranging from 0 to 9 degrees up from the site

reference — 5-element Yagi on a 4.3-meter (14-foot) boom, rated at 0 dB reference

relative strength variation from 0 to 9 degrees vertical angle — ±0.3 dB

relative strength variation when reference and test antennas are placed on opposite towers — <0. 15 dB

main-lobe centered tests

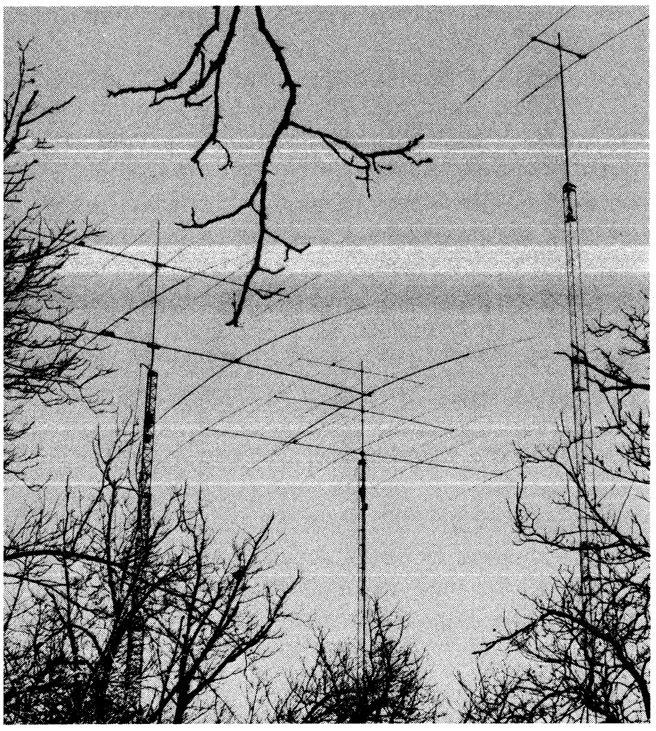
The other major question raised by the SCDXC members was really twofold: 1) my results might be invalid because I had been testing triband quads against monoband Yagis; and 2) my antenna ranges were, in effect, "ground-reflection" ranges, but I was measuring signals at zero degree angles rather than up in the middle of the main lobe.

To determine if these factors were significant in my research, I set up another experiment. I selected a cleared construction site where the soil had just been leveled, a site on flat land about 500 meters from the foot of a hill. The vertical angle to the hilltop was about nine degrees, according to AA6DD's transit. Roads traversed the hill at several elevations, permitting tests at other radiation angles between zero and nine degrees.

I chose 50 MHz for these tests in a tradeoff between the desire for results applicable to the high frequency spectrum and the desire to work with antennas having manageable dimensions. Since the construction methods, element and boom diameters, and wire sizes were essentially the same as those used at 14 MHz (as a percentage of the respective wavelengths), the results here would be expected to be similar to those in the upper high frequency bands.

I prevailed upon AA6DD and K6LPF to spend an afternoon and evening on the hilltop. Then AA6DD spent another day moving up and down the hillside, always with a 3-element, 50-MHz Yagi mounted on his car as a signal source.

These 50-MHz tests were interesting in several



Leafless trees give this photo of the antenna farm at W6HX an eerie feeling. At the upper right is the reference tribander positioned to match the height of one of the W6HX antennas for the gain comparisons.

respects. First, there was virtually no difference (±0.3 dB) in the comparative performance of the quads and Yagis tested at the various vertical angles from zero to nine degrees. In view of what has been previously written (and what I found) about the similarity of the vertical angle produced by quads and Yagis at any given height above ground, this result would be expected. It indicates that measurements can indeed be made with a local signal source (arriving at or near zero degrees) to produce results valid at higher angles, as long as the test and reference antennas are at the same height.

Perhaps the most interesting thing about the 50-MHz tests was the opportunity to compare a variety of well-known quad designs in a monoband configuration with a good 5-element Yagi as an on-site reference.

The results are summarized in **table 1.** The Lindsay and Landskov designs are those described in the articles already cited. The Machin "optimum" quad is a design recently published with the claim that it represents the true optimum spacing for quad arrays.[11] The Machin spacing is 0.15 λ reflector to driven element, 0.30λ driven to first detector, and 0.30λ driven to second director. The longest 4-element quad is an outgrowth of my 432-MHz experiments. Its spacing is conventional (about 0.21λ) except between the first and second directors, where it is 0.49λ! I used Landskov's element dimensions for this antenna. The 3-element quad is spaced 0.21λ and uses Landskov's dimensions.

In all cases, the element lengths were scaled from the original designs by determining the percentage differences from the driven elements and then applying those percentages to an experimentally derived correct length for a 50.1-MHz driven loop.

50-MHz summary

In these tests, monoband quads with short booms (up to about 0.5 wavelength) did indeed approach (but not equal) the gain of a longer Yagi. However, increasing the boom length still did not produce any quad that equaled the reference Yagi.

Neither Machin's long-boom quad nor mine delivered enough gain to justify its size. For all practical purposes, Machin's quad is identical in gain to the far smaller Landskov design, casting doubts on his claim that he has discovered the unique optimum design for a multi-element quads.

In fact, for an Amateur who could not accommodate a long-boom Yagi on his real estate, the Landskov quad design would seem to offer an excellent compromise. In this monoband version, it fell only 0.9 dB short of the 5-element Yagi. For comparison, a 20-meter Landskov quad is 9.75 meters (32 feet) long, while the Yagi would be 15.85 meters (52 feet) long.

Another good design appears to be the relatively wide-spaced 3-element quad shown in **table 1.** On 14 MHz it would fit on a 9.1-meter (30-foot) boom; it delivers excellent gain for its size and boom length.

It should be emphasized again, however, that these are monoband quads. The high frequency study indicated that the same results are difficult to attain in multiband configurations. And given the mechanical problems inherent in stacking monoband quads for several bands (with an acceptable stacking distance between the top of one quad and the bottom of the next higher one), stacked monoband Yagis still look pretty good for high performance on a variety of high-frequency bands.

about trap tribanders

If this field research suggests that long Yagis are the most consistent high-performance antennas for serious DXers and contest operators, where does that leave the thousands of Amateurs who use trap tribanders? How good are these multiband Yagis?

To find out, I set up two manufacturers' top-of-the-line tribanders side by side. Using the same test procedures as in the other high-frequency experiments, I ran gain curves on the two big tribanders. Then I replaced one tribander with the original reference TH-2 and repeated the experiment so the gain of the two big tribanders could be plotted on the same chart as the other antennas tested.

Perhaps the most notable conclusion, and the least controversial, was that these big tribanders sacrifice bandwidth for multiband coverage. Both were adjusted for phone band operation, using the factory-specified dimensions, and both exhibited drastic gain fall-off in the CW bands.

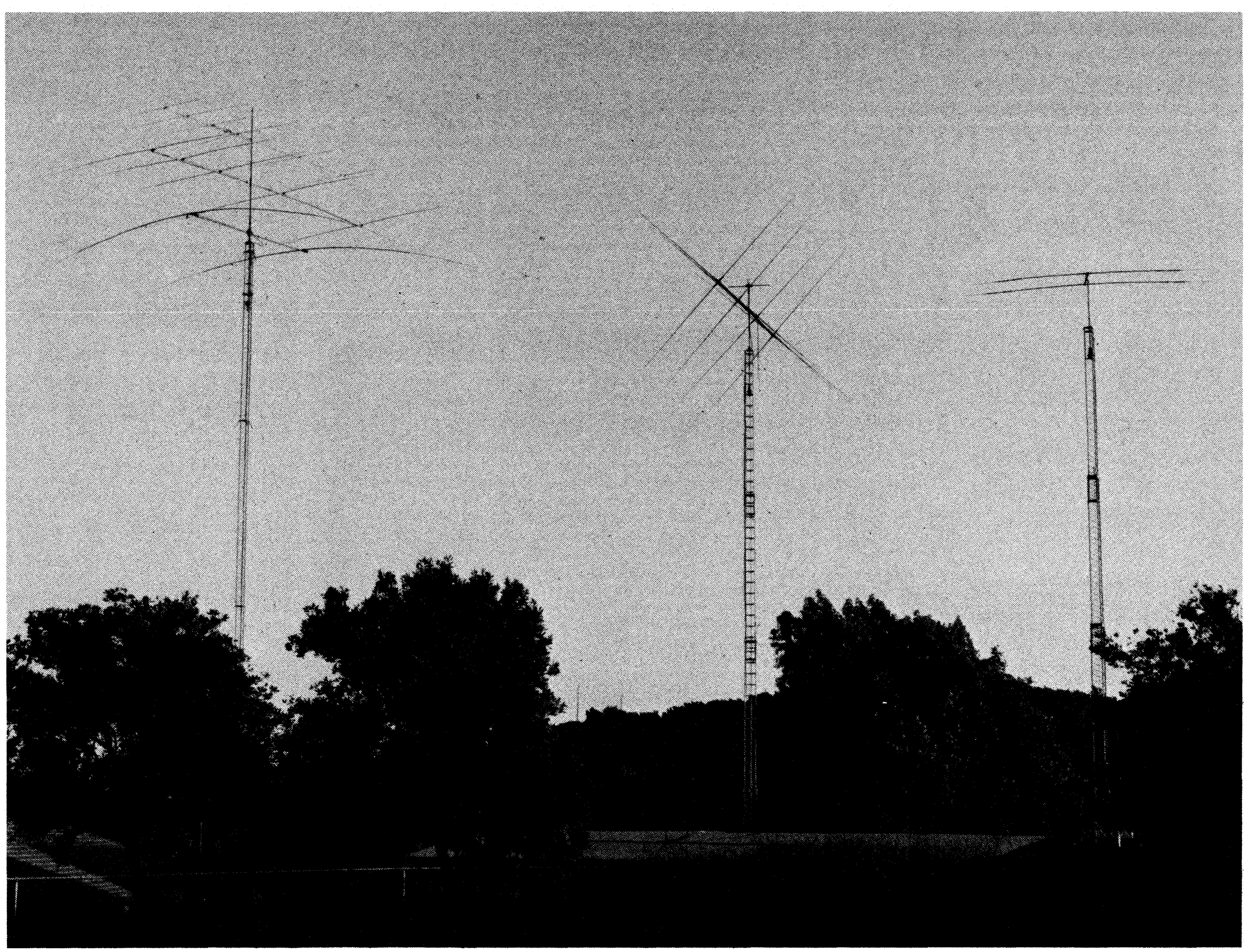
The results of the comparison produced no clear-cut answer for the Amateur who asks, "Which should I buy?" Tribander number 1 was much better on 20 meters, but slightly inferior on 15 and very inferior on 10. The poor showing on 10 can be explained in part by the owner's choice of the "high phone" length settings, but even at 29.4 MHz it was only 1.4 dB better than the reference TH-2, which was tuned much lower in the band! In fact, except for a couple of instances, the big tribanders exhibited very little gain above the two-element reference tribander! And both tribanders were markedly inferior to the monoband antennas, even in that very narrow range of frequencies where the tribanders hit their peak performance on each band.

None of this is new or surprising information, of course. But a look at the antenna gain charts could provide a dramatic illustration of the compromises involved in trap tribanders! If your goal is to keep in touch with friends and do some casual DXing, a tribander is fine. But if you want a first-rate signal, a trap tribander isn't the answer!

conclusions

These field experiments are by no means the last

Another source of big signals, N6NA (ex-W6MAR). At the left is the stack of Yagis which look as good on the gain plots as they do in the air. The 20- and 15-meter Yagisare considerably higher than the reference tribander, rendering the gain figures on these big beams invalid for comparison purposes. In the center is the 4-element quad, a source of much head scratching because of its questionable performance.



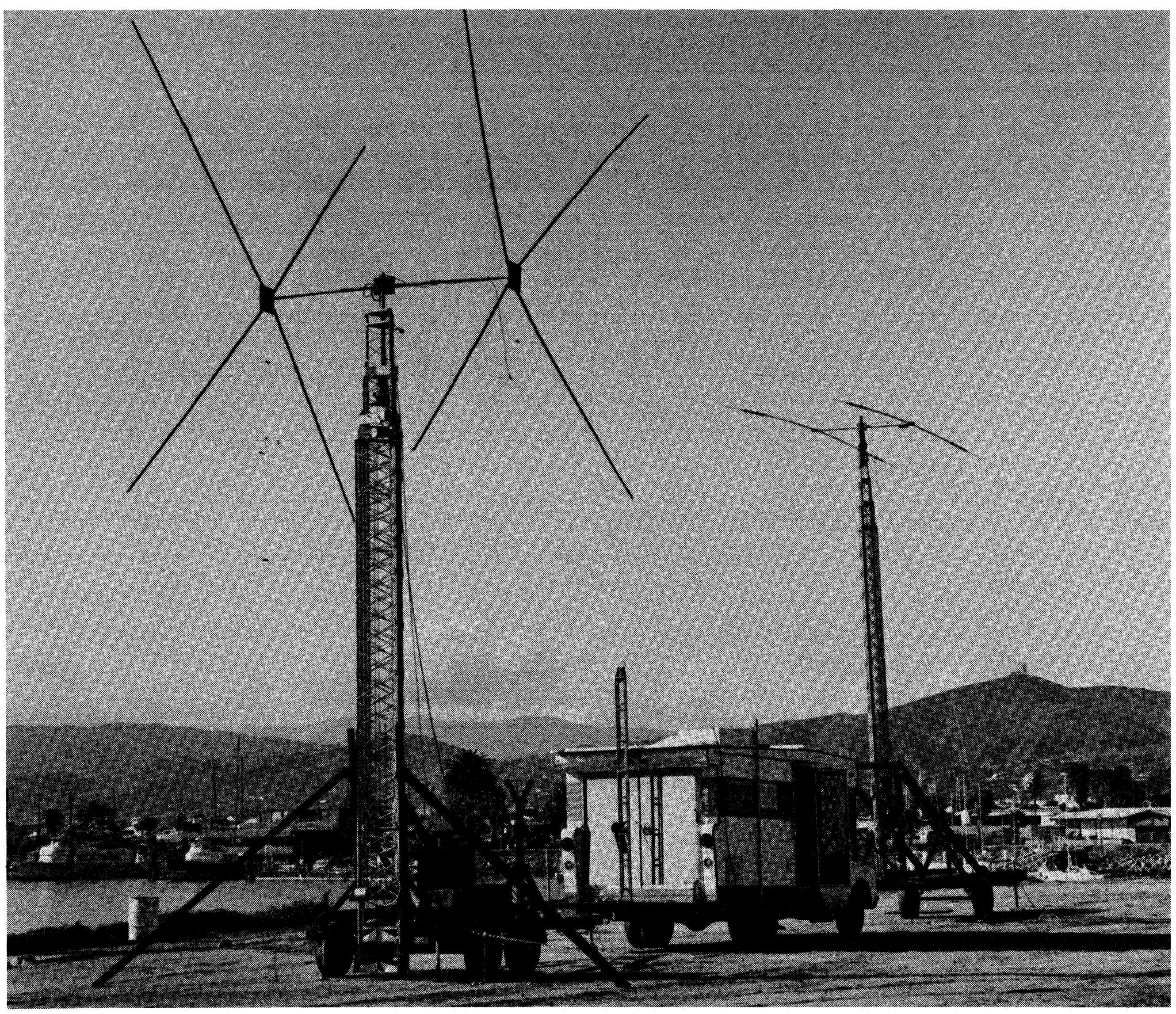
ones I'll undertake to compare working quads and Yagis on the high-frequency bands. Any judgement at this point is necessarily tentative, subject to revision as more data is produced, but the following conclusions are justified:

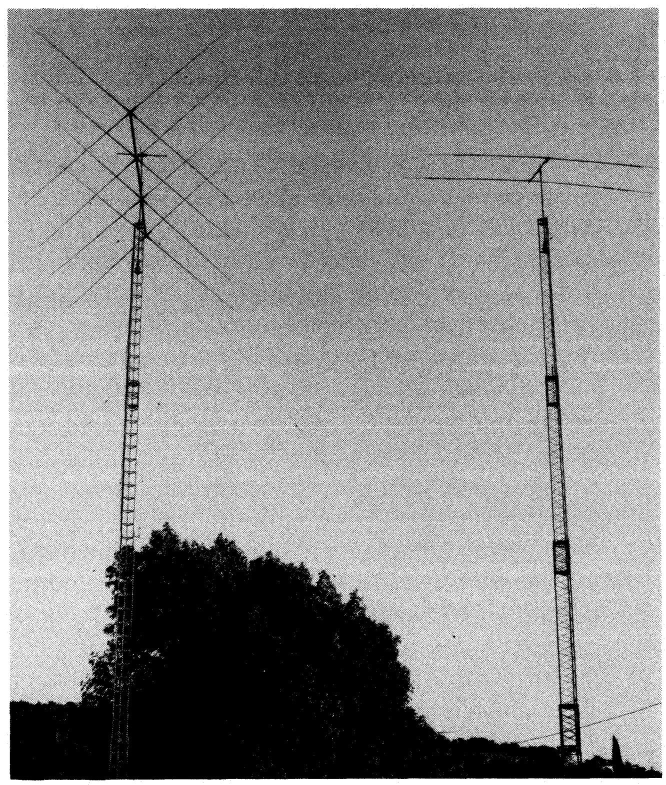
1. Cubical quads do *not* "come into their own" at low heights. At any given height, the vertical angle of radiation of quads and Yagis is virtually identical. The old idea that quads are better iow-height performers than Yagis should be recognized as the myth that it is.

2. As the frequency is increased into the uhf region, the performance of cubical quads and Yagis may not deteriorate at exactly the same rate, given the mechanical differences between the two designs, particularly their driven-element configuration. This creates difficulties that must be accounted for if you wish to generalize about the relative performance of the two antennas at the high frequencies on the basis of uhf modeling.

3. While it may be possible to design a high-frequency cubical quad with a long boom that will outperform a similar size Yagi by 2 dB as Lindsay suggested, no quad I tested approached that level of performance. In few cases did a quad of more than two elements even equal a comparable-size Yagi. The 50-MHz tests illustrate the difficulties of achieving gain increases, when a quad's boom is lengthened, that equal the gain increases normally found when a Yagi's boom is increased. In fact, the quad

High and low antennas by the sea. Each antenna is on a trailer-mounted crankup tower beside a salt-water marina, ensuring an electrical ground near the surface. Regardless of the height, the relative performance was identical.





The enigmatic quad. This 4-element quad at KGNA, because of its mediocre performance, inspired Landskov's efforts to develop a competitive long quad. Now modified to Landskov's dimensions, it remains so marginal that the reference 2-element tribander nearly topped its gain midway across 15 meters. Were the reference beam a 3-element trap tribander, it would have probably matched or beaten this quad across all three bands.

seems to be at its best in two- and three-element designs.

The data presented here would support a conclusion not far afield from the assessment of quads versus Yagis in Bill Orr's original work on the subject some 15 years ago.[12] In that first edition, Orr said that a 2-element quad was superior to a 2-element Yagi, but that larger quads were inferior to comparable-size Yagis.

In his 1970 edition, Orr qualified that conclusion considerably, and his latest antenna book, published in 1978,[13] says flatly that quads are 2 dB better than Yagis. The new work has a "truth table" that simply rates the 4-element quad at 12 dBd and the 3-element quad at 10, with the corresponding Yagis 2 dB lower!

Neither our tests at 50 MHz, nor the 10-, 15-, and 20-meter tests at KGNA (both tests involving quads almost identical to the 4-element design in Orr's new book), would support that conclusion. I have yet to see a 4-element quad deliver 12 dBd gain or anything close to it on any frequency, or to outperform a well-designed 3-element quad by 2 dB, for that matter.

Perhaps there really is a high-frequency cubical quad out there somewhere that delivers 2 dB more forward gain than a well-designed Yagi of the same boom length, a quad that does what Orr, Lindsay, and others say a quad will do. Perhaps, but I tested quads built to the most popular published dimensions by respected Radio Amateurs and none of their antennas came close to beating a comparable-size Yagi by 2 dB. In many instances, even trap tribanders were comparable to quads of similar boom lengths, with full-size Yagis far better than either!

To those who still believe in the superiority of the quad antenna, I offer a challenge — bring me a high-frequency quad of four or more elements that you believe outperforms a comparable-length Yagi. I'll provide two towers in an open field for the side-by-side tests. If your quad really delivers 2 dB more than my Yagi, I'll publicly recant the conclusions presented in this article.

acknowledgments

Many Amateurs cooperated in the time-consuming process of gathering data for this report. In particular, I would like-to thank W6AQ, K6AV, WAGBTX, W6CG, AA6DD, WB6ESQ, WAGFVR, WB6HSG, WGHX, WAGIJZ, KGLPF, NGMA, NGMB, KGNA, NGNT, WGOXY, and NGRS for their help. Without their field work, the charts and tables would be blank.

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ham radio

rf impedance bridge measurement errors and corrections

How to determine the effects of stray reactance in rf impedance measurements and how to compensate for them

The current interest in various commercial and homemade impedance-measuring devices which measure both the real and imaginary components of impedance certainly represents progress toward higher-performance antenna systems. To attain the level of accuracy that these devices can deliver, however, it is necessary to appreciate the sources of error than can creep into a test setup and then correct for these errors. Unfortunately, few Amateurs seem aware of these problems, and consequently they accept their bridge readings as gospel.

Basically, there are two sources of error: those external to the measuring device (usually stray capacitance, which will cause any impedance-measuring device to read low), and those within the measuring device itself. Even the prestigious General Radio 91611606 family of rf bridges has systematic errors which must be corrected for if accurate results are to be obtained. I was unpleasantly surprised when I finally got around to working out the predictable errors in my measurements.

benefits of bridge corrections

A user of an impedance-measuring device may ask, "Why go to all the trouble of calculating correction factors?" First of all, it does not take that much additional effort. If you want accurate measurements, you are going to spend a fair amount of time and effort putting together a good test setup; the additional time and effort to punch numbers into a calculator is minor.

In addition, you can get a great deal of personal satisfaction when your measured data falls right where the textbooks say it should. When this happens to me, I feel as though I really understand whatever I am working on and am really the master of it. This feeling of accomplishment is probably why I experiment in the first place — that and curiosity. Finally, when you are sure of your measurements and then get unexpected results, you can explore the device with a lot more confidence.

In the material that follows, I will discuss in detail the error caused by stray capacitance and give a calculator program to correct this error. I will briefly discuss instrumentation errors and give a procedure for calibrating a measuring Instrument; discussions of calibration will necessarily be in outline form since the many types of instruments in general use will each require slightly different calibration procedures.

stray capacitance errors

The effect of any stray reactance, either capacitive or inductive, shunted across an impedance is to lower the apparent value of that impedance. Consider the impedance $Z_x = R_x + jX_x$ in **fig.** 1 with a reactance jX_a shunted across it. In the discussion that follows, I will assume that all reactances are inductive (positive), as this simplifies the problem of algebraic signs. When a conclusion is reached, I will explain the changes, actually very minor, which are necessary for negative reactances.

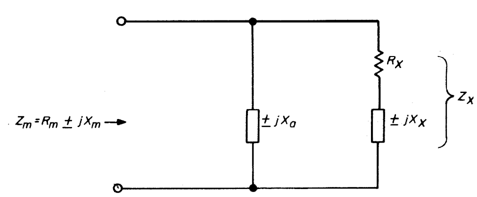

fig. 1. Diagram showing the stray reactance $\pm X_a$ shunted across an unknown impedance $Z_x = R_x \pm jX_x$. The text presents equations for compensating for the effects of $\pm X_a$.

In **fig.** 1, R_x and X_x are the real and imaginary parts of the unknown impedance. The shunt reactance is represented by jX_a and is presumably known, or at least estimated. The resulting impedance formed by $Z_x || jX_a$ (the parallel bars | | should

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be read "in parallel with") is given mathematically by the well-known "the product divided by the sum" rule:

$$Z_m = \frac{jX_a(R_x + jX_x)}{R_x + j(X_a + X_x)}$$

After rationalizing the denominator by multiplying both numerator and denominator by the conjugate of the denominator, the separation of the results into real and imaginary parts followed by equating the real and imaginary parts gives

$$R_m = \frac{R_x X_a^2}{R_x^2 + (X_a + X_x)^2} \tag{1}$$

$$X_m = \frac{X_a[R_x^2 - X_x(X_a + X_x)]}{R_x^2 + (X_a + X_x)^2} \tag{2}$$

In actual practice, R_m and X_m will be read from the bridge, so will be known. The stray reactance will also be known, or at least estimated. R_x and X_x are the unknown quantities. It will therefore be desirable to solve **eqs. 1** and **2** for R_x and X_x. The algebra is rather involved, so I will just give the results:

$$R_x = \frac{R_m}{\left(1 - \frac{X_m}{X_a}\right)^2 + \left(\frac{R_m}{X_a}\right)^2} \tag{3}$$

$$X_x = \frac{X_m - \frac{R_m^2}{X_a} - \frac{X_m^2}{X_a}}{\left(1 - \frac{X_m}{X_a}\right)^2 + \left(\frac{R_m}{X_a}\right)^2} \tag{4}$$

If the stray reactance is capacitive, X_a will be negative, so that a negative sign should be used in the above equations with X_a; the same holds true for X_m. If there were no stray capacitance, X_a would be infinite, so that $R_x = R_m$ and $X_x = X_m$.

Eqs. 3 and **4** do not require great mathematical ability to solve, but the algebra is tedious when done by hand, even with a pocket calculator. A program can be written for a programmable calculator, however, that eliminates much of the burden (and many chances for mistakes). An HP-25 program to solve these equations is shown in **fig. 2.**

It will be seen that while the external corrections may be minor when working with low-impedance networks at low frequencies, the errors increase at higher impedances and higher frequencies. The illustrative example given in the HP-25 program shows that a correction of about 5 per cent in resistance and the correction of the reactance term completely changes sign from negative to positive. This example is based on an actual measurement I made on an experimental balun. If I had tried to correct a negative reactance in the balun when the reactance was actually positive, it is obvious I would have been wasting my time! A situation like this can be very misleading.

Two final comments are in order: First, while shunt reactance has been portrayed as a source of error, the same shunting effect can also be used to advantage to measure an equivalent series impedance that is above the range of the bridge. The unknown impedance is shunted down by placing a reactance, usually capacitive, in parallel with it. The unknown impedance is then calculated by **eqs.** 3 and **4,** preferably using the calculator program. Capacitors between 35 and 200 pF are usually satisfactory; the value should be no larger than necessary to bring the unknown impedance within range of the bridge.

The second comment is that **eqs.** 3 and **4** should be used only with impedance-measuring devices which measure unknown impedance in terms of its series equivalent impedance; *i.e.*, $Z_x = R_x \pm jX_x$. If the measuring device gives results in terms of admittance, as does the GR-821, or as parallel resistance and reactance, as does the Hewlett-Packard RX meter, **eqs.** 3 and **4** are not applicable.

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM.

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	23 02	STO 2					
02	21	$x \leftrightarrow y$					
03	24 00	RCL Ø					
04	61	X					
05	15 22	1/X					
06	23 03	STO 3					
07	71	÷					
08	24 01	RCL 1					
09	21	$x \leftrightarrow y$					
10	41	−					
11	23 04	STO 4					
12	15 02	X^2					
13	21	$x \leftrightarrow y$					
14	24 03	RCL 3					
15	71	÷					
16	23 05	STO 5					
17	15 02	X^2					
18	51	+					
19	23 06	STO 6					
20	71	÷					
21	74	R/S					Display R_x
22	34	CLR X					
23	24 05	RCL 5					
24	61	X					
25	24 02	RCL 2					
26	24 04	RCL 4					
27	61	X					
28	41	−					
29	32	CHS					
30	24 06	RCL 6					
31	71	÷					
32	74	R/S					Display X_x
33	13 01	GTO 01					
34							
35							
36							
37			SAMPLE CALCULATION:				
38				C_a =	6 pF		
39				R_m =	106 ohms		
40				f =	30 MHz		
41				X_m =	−1.7 ohms		
42							
43			ANSWER:	R_x =	104.895	ohms	
44				X_x =	10.896	ohms	
45							
46							
47							
48							
49							

MEMORY REGISTERS	
R0	$-2\pi C_a \cdot 10^{-6}$
R1	1
R2	$\pm X_m$
R3	X_a
R4	$1 - X_m/X_a$
R5	R_m/X_a
R6	denominator
R7	

fig. 2. H-P 25 program for solving the stray reactance effects given by *eqs. 3* and 4. The program also calculates the shunt reactance for a fixed value of stray capacitance and the variable parameter, frequency (see STO 0 and steps 3, 4, 5, and 6). Note that the sign STO 0 must be negative because the stray reactance is capacitive. Before running the program store $-2\pi C_a \cdot 10^{-6}$ in STO 0 and in 1 in STO 1. To run program, key in R_m (ohms), press ENTER, key in frequency (MHz), press ENTER, key in X_m (ohms with proper sign), and press R/S. Calculator displays R_x at step *21*, and X_x at step 31.

instrumentation errors

The shunt reactance error discussed above is external to and independent of the impedance-measuring device. This error will occur whether you use a laboratory-type instrument costing several thousand dollars or a homemade device, built with parts from the station junkbox.

I will now discuss in a very general way errors that occur within the measuring instrument itself. The GR-91611606 instruments will be used as a vehicle, but I will try to present the material in a manner that makes it applicable to any similar instrument, commercial or homemade.

All impedance-measuring devices (that I am aware of) have internal errors, especially at the extreme ends of their frequency or impedance ranges. The sources of these errors can be very subtle. For example, in the GR-91611606 family, the principal error at high frequencies is caused by the changing effective series inductance of the RESISTANCE capacitor; *ie.,* the variable capacitor connected to the resistance dial. This inductance causes the effective capacitance to increase as the frequency increases, thereby causing the resistance dial reading for a given resistance value to read low. For the GR-916, this error alone can be as large as 30 per cent at 60 MHz and 100 ohms, for a typical instrument.

At the low-frequency end, the dielectric loss in the REACTANCE capacitor causes an effective series resistance that increases with higher REACTANCE dial settings and low frequencies, again causing the resistance dial to read low under some conditions.

Both these errors are predictable, systematic errors which exist in addition to any random, unknown errors caused by manufacturing variations or operator error. Graphs for correcting both types of errors in the GR-91611606 are given in the instruction manuals and should be used if you want good accuracy. In addition, an equation for correcting the high-frequency error in a typical instrument is provided along with a procedure for obtaining the fudge factor for any particular instrument. I have programmed this equation on an HP-25 pocket calculator, but am not including the program here because of its limited interest. However, I will be glad to provide a copy to interested readers on receipt of a large, self-addressed, stamped envelope.

I feel certain that all impedance-measuring instruments have some systematic error. Just because the instruction manual for an instrument does not mention internal errors does not mean that there are none; it just means either that the manufacturer did not know how to determine it or he could not afford to work it out for the price at which he is selling the instrument. Actually, working out the instrument correction factors is not too difficult — and it is absolutely essential if you want to obtain accurate measurements.

In calibrating an impedance-measuring instrument, the most important item is a known value of impedance. For this, I use a Hewlett-Packard 906-A 50-ohm termination. This termination is an accurate 50-ohm resistor with negligible reactance well into the GHz region This model uses a type-N connector; the cost is in the $25-30 range. Though this may seem expensive for a 1-watt resistor, the cost is small compared with what you already have invested in a commercial bridge. If your bridge is homemade, you may need a precision 50-ohm resistor more than you realize!

A second reason for using this type of calibrated load is that since the load has negligible reactance itself, you can accurately determine the stray reactance which must be known before accurate test data can be obtained.

If you want to avoid the expense of a laboratory-type termination, I suggest the use of the RN-55 family of Mil-spec resistors. These seem to have the lowest reactance of any family I have tried. They also have the advantage of being available in values other than 50 ohms.

The calibration procedure for a typical instrument is very simple; although it does not apply to the GR-916, it can be used with a GR-1606. Measure the impedance of the termination at various frequencies through your range of interest. Since it is necessary to know precisely the resistance and reactance presented to the bridge terminals, it is necessary to make corrections using **eqs. 1** and **2.** To do this, you must estimate the shunt capacitance, and this is where a reactance-free load comes in handy. Use **eq.** 2 first; assume values of X_a corresponding to shunt capacitances of, say, 1, 2, 3 . . . pF, *etc.* at your highest test frequency. Set $R_x = 50$ *ohms* or whatever resistance value you are using, $X_x = 0$, and calculate the values of X_m for each capacitance. Compare these values of X_m with the X_m you measure, and when your calculated values of X_m have bracketed the measured values of X_m you have also bracketed the actual value of C_a. Now back up, say, 0.5 pF, and determine C_a as accurately as you wish. Remember that since X_a is capacitive, a negative sign should be used in all of the correction equations.

When you accurately know X_a, solve **eq.** 1 to calculate the resistive component of impedance appearing at your bridge terminals at each frequency; compare these against the measured resistance of the load and determine the correction factor for your instrument at each frequency. These corrections can then be plotted.

One difficulty in using a chart is that a chart cannot be programmed into a computer. It would be convenient to determine the equation of the correction curve so that it could be programmed into a calcula-

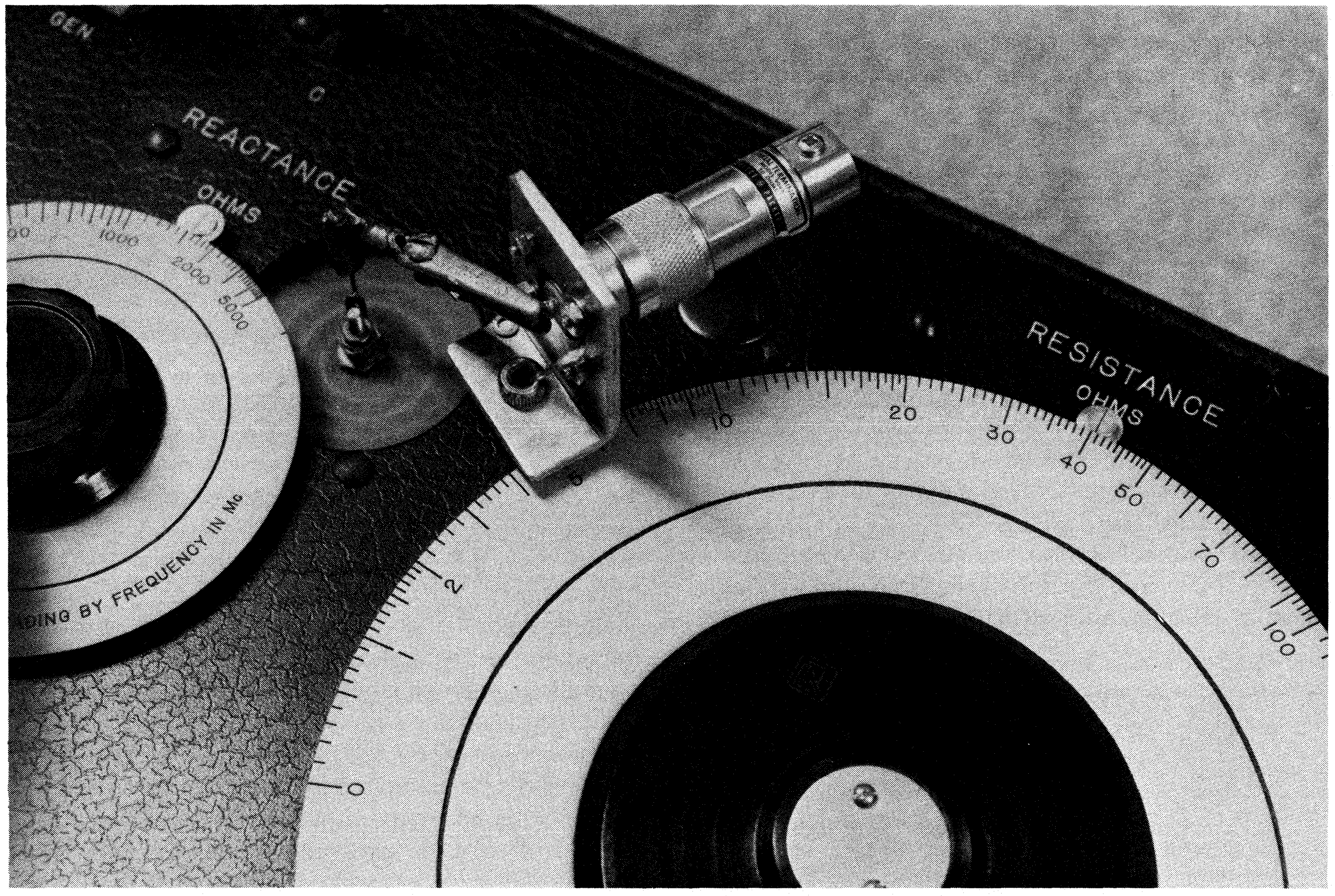


fig. 3. Photograph of K4KJ's test setup for calibrating a GR-916 rf impedance bridge using a precision 50-ohm termination.

tor. It should be possible to do this and those interested may refer to a textbook on the subject.' Thus both the bridge correction and the shunt reactance correction can be worked out in one calculation.

The advantage of having a reactance-free resistance can be seen. Your impedance-measuring device can only measure the reactance across its terminals; it cannot distinguish between stray reactance or a reactive component in the unknown impedance.

A note of caution concerning the order in which the corrections are made: when measuring an unknown impedance, the usual application, the bridge correction *must* be calculated *first.* **Eqs.** 3 and **4,** which are used to calculate the stray capacitance correction, assume that the measured values are accurately known. On the other hand, when calibrating the bridge using a known value of resistance, the stray capacitance correction *must* usually be made first because you are trying to determine the actual impedance across the bridge terminals. The GR-916 appears to be an exception; I do not know why.

examples

At this point some examples should help clear up the details. The first example will demonstrate the use of a reactance-free termination to determine the shunt capacitance and show how this capacitance can differ from the measured value.

I was in the process of calibrating my GR-916 using an H-P 50-ohm termination. A photograph of my setup is shown in **fig.** 3. The stray capacitance to ground of the test lead and type-N coax connector measured 2.95 pF using a GR-821 admittance bridge.

At 54 MHz the impedance of the 50-ohm termination measured 40.3-j3.333 ohms on the GR-916 bridge. For a high quality termination, this appears way off, but let's correct it. Start with **eq.** 2. Set $R_x = 50$ ohms; assuming the stray capacitance to be 2.95 pF as measured, then at 54 MHz, X_s will equal -990.894 ohms. Solving **eq.** 2 gives $X_m = -2.49603$ as compared with -3.3333 ohms measured. Hence the stray capacitance must be more than 2.95 pF.

Next try, say, $C_a = 4.0$ *pF,* and calculate $X_m = -3.3774$ ohms, again compared with $X_m = -3.3333$ ohms measured. Back up a little and let $C_a = 3.95$ *pF* and calculate $X_m = -3.3355$, which is very close. Since the test fixture measured 2.95 pF, the stray capacitance in the bridge must be 1 pF.

Assuming the stray capacitance is 3.95 pF, it is interesting to calculate the equivalent impedance looking into **fig. 1.** Use **eqs. 1** and 2 and obtain $R_m = 49.776\ ohms$. The difference between this value and the 40.3 ohms measured by the bridge is the bridge error. The reason why the shunting effect is so small is that the shunt capacitance and resistance level are both relatively low.

I am not going to pursue this example further because the remainder is unique to the **GR-916** and would have relatively limited interest, but I did want to demonstrate the value of having a high quality termination and how to use it in estimating stray capacitance.

My second example should be of special interest to users of baluns, particularly W2AU balun users. In this case I was measuring the input impedance of a W2AU 1:1 balun at the unbalanced end with various values of resistance connected across the balanced terminals. **Table 1** gives the uncorrected bridge readings, the impedance with bridge corrections only, and finally bridge and stray capacitance corrections. A photograph of this test setup is shown in **fig. 4.**

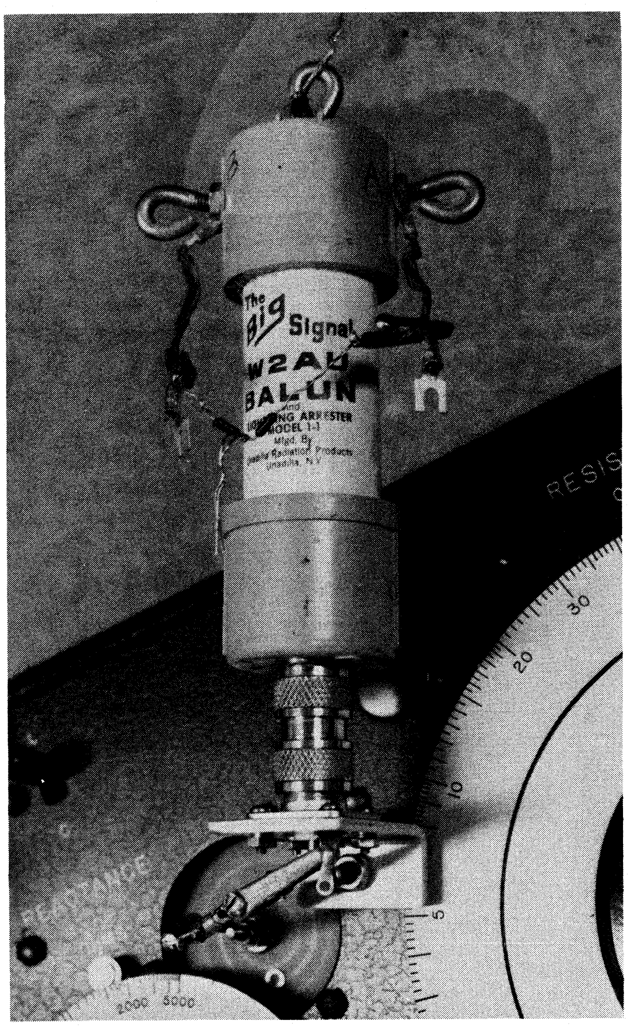


fig. 4. Photograph of K4KJ's test setup for measuring the input impedance of a W2AU 1:1 balun with various resistance values across the balanced terminals.

table 1. Measured performance of a W2AU 1:1 balun with various values of load resistance across the balanced terminals.

load resistance ohms	original bridge reading	with bridge corrections only	with bridge and capacitance corrections
54.0	70 - j14.00	74.64 - j14.00	76.56 - j7.34
68.0	*56* - j22.07	59.61 - j22.07	62.53 - j18.12
102.9	47.1 - j43.33	50.09 - j43.33	55.45 - j42.20
201.0	25 - j54.17	26.52- j54 17	30.26 - j56.88

The stray capacitance of the test fixture with a type-UHF connector and a male-to-male adapter measured at 5.3 pF; adding 1 pF for the bridge capacitance gives 6.3 pF. I am giving only the 30-MHz measurements.

Comparing the original bridge reading with the fully corrected data shows as much as 10 per cent correction, approximately equally divided between instrumentation and stray capacitance effects.

Comparison of the load resistance and final data columns will be of interest to those using 1:1 baluns. It shows the necessity of matching the balun impedance to that of the load and transmission line. This is a need that is just beginning to be recognized. I believe this data is typical of ferrite-rod baluns, but that's another story.

conclusion

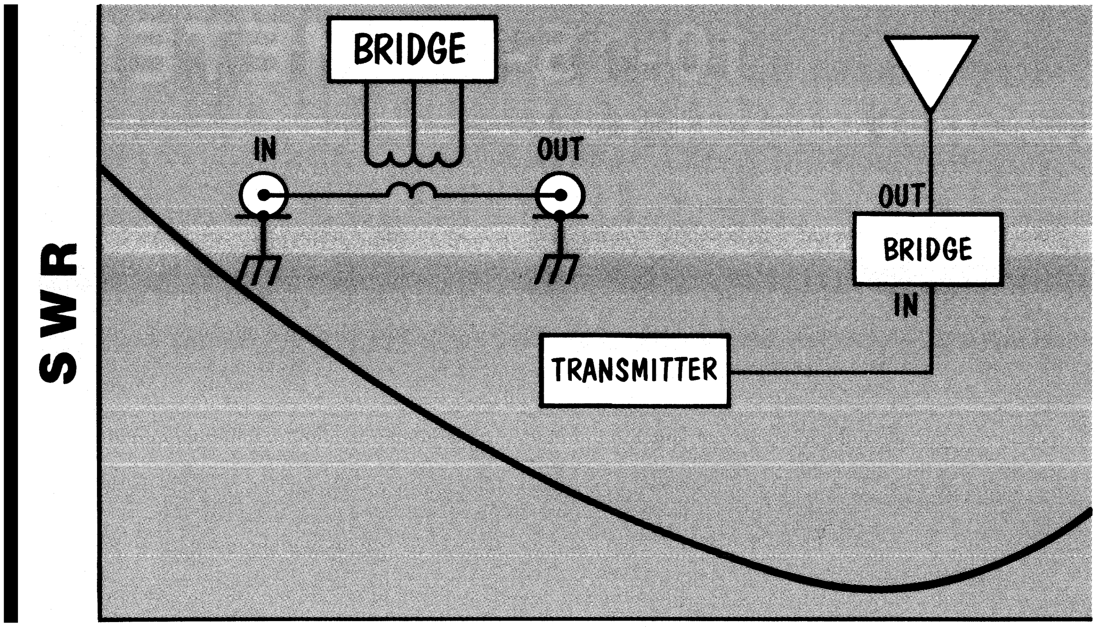
Impedance-measuring devices can be very useful for matching antennas as well as in many other applications around the ham shack. Their results, however, must be treated with caution because even the best instruments and test setups are subject to errors. The most predictable source of error caused by the test setup is stray capacitance; this effect can only be minimized, not eliminated. Equations have been presented for calculating the true impedance in the presence of stray capacitance. Possible instrument errors have been discussed in a general way, and a possible method described for correcting these errors as well.

The program for computing the true impedance in the presence of stray capacitance on an H-P 25 programmable calculator is probably usable on other H-P calculators, although the keystroke numbers may be different. I do not have programs for calculators of other manufacturers.

reference

1. C. R. Wylie, Jr., *Advanced Engineering Mathematics*, second edition, McGraw-Hill, New York, 1960, Chapter 5, "Finite Differences," pages 130-193.

ham radio



broadband reflectometer and power meter

Construction details for a combination peak-reading power meter and broadband SWR bridge

The reflectometer design described in this article uses a coupler technique that has not received attention in the Amateur literature worthy of its versatility, simplicity, and useful characteristics. Apart from being easy to construct, this design covers a three-decade frequency range, from 100 kHz to 100 MHz, and can be constructed with a power sensitivity as low as 500 mW or as high as 500 watts.

The problem with most high-frequency reflectometer designs, in the experience of the authors, is that they generally cover only a frequency range of a decade or less – usually 3 to 30 MHz. Many are quoted as covering well into the vhf range (usually to 150 MHz), but, in practice, they suffer from rather extreme errors in accuracy due to construction discontinuities as well as from large sensitivity excursions across the range.

The predominant technique employed in most published Amateur designs, and in many commercial designs, is to have one or more secondary "coupling" lines inserted into a short length of coaxial transmission line. The well-known *Monimatch* uses this principle. Two insulated wires are simply slipped under the braid of a short length of coax to form

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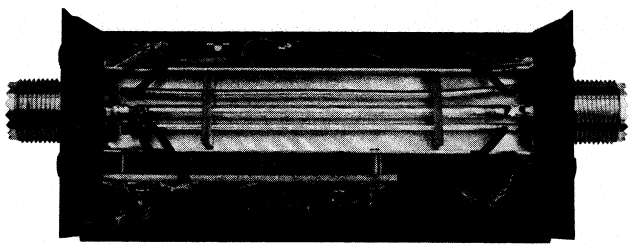
coupling lines which sense the forward and reflected components of standing waves on the transmission line.

A similar method, popularly used in commercial designs, is to construct a section of coaxial transmission line from a sheet metal trough with two short coupling lines supported parallel to the center conductor.



Apart from construction discontinuities, the technique suffers from two serious drawbacks. The coupling coefficient of the secondary lines varies with frequency, being least at the low frequencies and rapidly increasing as the length of the coupling lines becomes a significant fraction of a wavelength. Secondly, the technique suffers from ever-decreasing accuracy at the higher frequencies for similar reasons, and poor sensitivity at the lower frequencies reduces the accuracy at low standing-wave ratios.

Properly engineered and constructed, reflectometers using this technique can have excellent characteristics and accuracy across a bandwidth of as much as several octaves. Getting them to perform consistently across a decade or more is tantamount to magic — you rapidly run into the law of diminishing returns.

design points

Although the technique employed in the coupler of this reflectometer has been around for a number of years, it has received inadequate attention in the literature. The basic requirement for a reflectometer is that it generates two voltages which are proportional to the forward and reflected voltages or currents existing in the transmission line under measurement. Techniques employed to fulfill this requirement use either voltage or current deflectors coupled to the transmission line to produce two out-of-phase signals, since the forward and reflected components on the transmission line are 180 degrees out of phase.

This reflectometer employs a current transformer constructed as follows. A short length of coaxial cable is passed through a ferrite toroid forming the primary. The braid, or outer conductor of the coax, is connected to form an electrostatic shield. The secondary consists of a winding around the circumference of the toroid which couples to the magnetic component of the leakage field of the short length of coax cable.

The load on the current transformer secondary is center-tapped so that out-of-phase signals appear at each end of the secondary. The load center-tap is returned to a voltage divider which samples the rf signal on the transmission line to perform the addition and subtraction across the load, yielding the forward and reflected components.

From this, the SWR may be computed from the following equation:

$$SWR = \frac{E_f + E_r}{E_f - E_r}$$

where

E_f = forward voltage components
E_r = reflected voltage components

Although a current transformer is employed, for

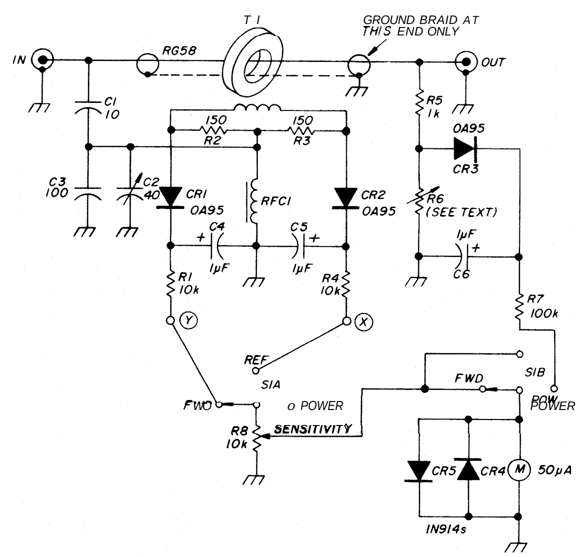


fig. 1. Schematic diagram of the broadband reflectometer and power meter. A specially configured switch was used for S1 (C&K type 7211); otherwise a double-pole, triple-throw switch will be necessary. The meter has a basic scale of 50 μA, with an internal resistance of 2000 ohms. The sensitivity pot (R8) should have a logarithmic taper. Transformer T1 is wound with 40 turns of number 35 AWG (0.14 mm) enameled wire over a Neosid type 28-511-31, F14 material toroidal core (initial permeability of 220). The measurements (outside diameter, inside diameter, thickness) of this toroid are 12.7 x 6.35 x 3.18 mm (0.5 x 0.25 x 0.125 inch). It is available from Neosid Limited, 10 Vansco Road, Toronto, Ontario, M8Z 5J4, Canada. C2 is a small mica compression trimmer.

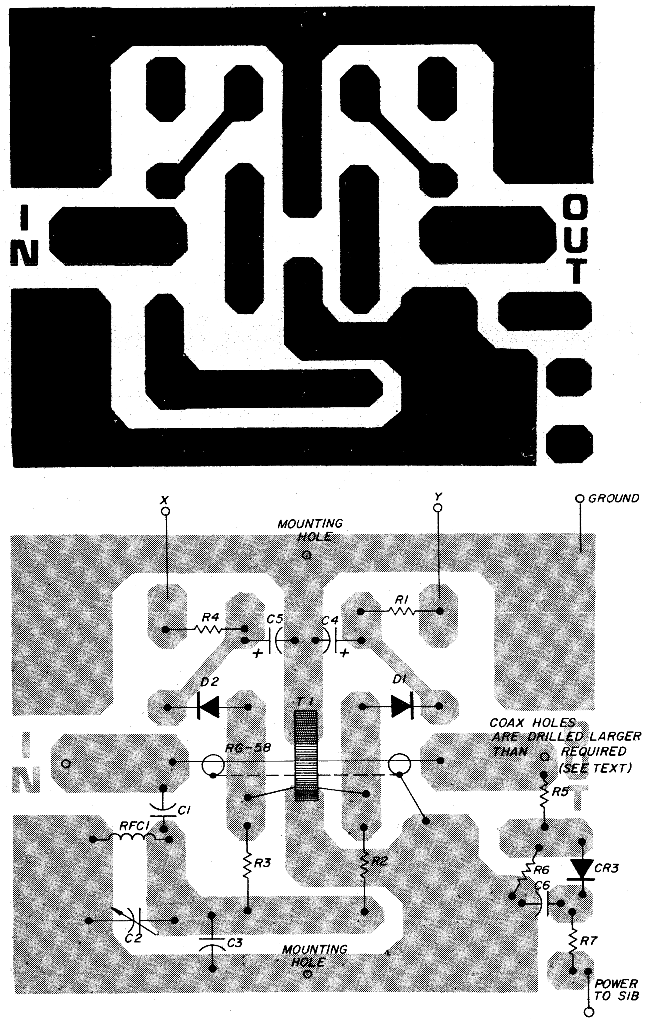


fig. 2. Printed circuit board layout (above) and parts placement diagram (below) for the reflectometer and power meter. Note: In this project the components are mounted on the foil side of the circuit board.

convenience the detected components are presented as voltages.

circuit description

The secondary of the current transformer drives a center-tapped resistive load (R2, R3) connected to a voltage-sampling network (C1-C2/C3) across the rf input such that sum and difference voltages will appear across the ends of the transformer (see **fig. 1**). The two diodes, CR1 and CR2, rectify the sum and difference voltages, with rf and audio bypassing being provided by C4 and C5. A dc return for the diodes is provided by RFC1.

Power measurement is made by rectifying a portion of the rf voltage tapped off the line by the resistive divider, R5, R6. CR3 and C6 form a peak detector, as the load, R7 and the meter, is very light. CR4 and CR5 provide protection for the meter.

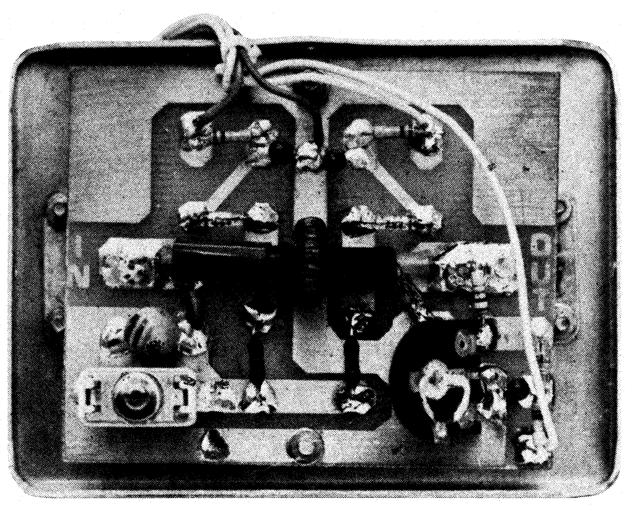
Some published designs use a resistive tap across the line to sample the rf for the current transformer load. At low powers, however, the low-resistance dc return of RFC1 provides better sensitivity.

The power rating is limited principally by the voltage breakdown characteristics of the coax line through T1. Standing-wave ratios greater than 3:1 generate substantial voltages across the transmission line, and any high-power operation with high SWR values should take this into account.

The sensitivity bandwidth is limited by the permeability of the toroid and the number of turns on the secondary winding. If this reflectometer is constructed for use at higher power levels, the sensitivity bandwidth is considerably improved.

Construction is very straightforward. The printed circuit board design shown in **fig. 2A** is recommended; otherwise, variations in construction may affect performance, particularly at the higher frequencies. As can be seen from **fig. 2B,** the components are mounted on the *copper side* of the board. Once all the components are soldered into position, the board is attached to the coax sockets and mounting bolts.

The toroid current transformer's secondary should be wound first. Following that, the coax primary should be assembled. Cut a 45-mm (1-3/4 inch) length of RG-58/U (single shielded), stripping back the braid and insulation as illustrated in **fig.** 3. Refer also to the component overlay and photographs. This operation is not all that critical, but it is advisable to follow the diagrams. Slip the toroid over the short length of coax and solder the coax and T1 leads to the printed circuit board as illustrated in **fig.** 2. Position the toroid centrally and attach it to the coax and board with a small amount of pliable rubber cement. Follow this by mounting all the other components.



View of the printed circuit board used in the reflectometer showing all the components and the general construction technique.

The printed circuit board is mounted in a metal box on which two suitable coax connectors have been mounted. The printed circuit board is soldered to the center pins prior to securing the assembly with the two mounting bolts. To avoid undue stress, which can cause problems following assembly, coaxial plugs (assembled with cables) should be inserted into the two coax sockets when soldering the board to the center pins. Make sure that a good fillet of solder secures the pins to the printed circuit board IN and OUT pads.

The board is mechanically secured by two mounting bolts (shown in **fig. 2** and the photos). A nut is placed under the board on each mounting bolt and a second on top of the board. This also serves to ground the board groundplane. Solder the top nuts.

The drawings show mountings for two SO239 sockets, although any of the other popular series of connectors may be used. However, center-to-center spacing of the sockets should be maintained at 60 mm (2-3/8 inches). The reflectometer was mounted in a 100 *x* 75 *x* 50 mm (4 *x* 3 *x* 2 inch) aluminum box. The meter used has a 50 x 50 mm (2 *x* 2 inch) face, but a larger unit, offering better accuracy, may be used, although this would necessitate a larger box.

calibration

A suitable rf source, a dummy load, and an rf voltmeter or an accurate rf power meter will be required for calibration.

SWR Scale. The instrument is connected between the rf source and the dummy load, the rf to the IN socket and the dummy load to the OUT socket. The sensitivity control should be set fully counterclockwise. The switch should be set to read forward power.

With the rf source on, rotate the sensitivity control clockwise until the meter reads full scale. Now, switch to read reverse power; adjust the trimmer, C2, to obtain minimum meter reading, increasing the sensitivity as the meter reading decreases. You should be able to reduce the meter reading to zero or very near.

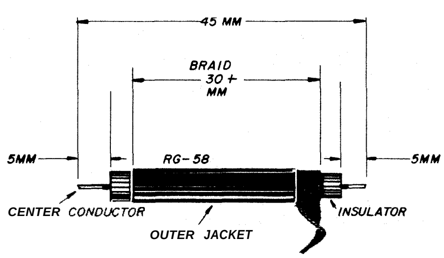
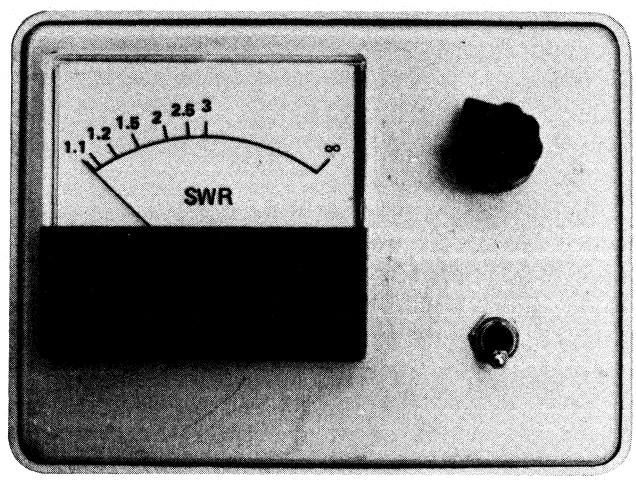


fig. 3. Dimensions for the section RG-58 of coaxial cable used as the primary of the transformer.

Once this operation is completed, the reflectometer section is calibrated. The photo shows the calibrated meter scale we attached to the meter face.



Any meter scale can be calibrated by using **table 1** or the SWR equation.

table 1. Meter readings to calibrate the new SWR scale.

SWR	scale reading
3.0:1	0.5 full scale
2.5:1	0.42 full scale
2.0:1	0.34 full scale
.5:1	0.2 full scale
1.2:1	0.1 fullscale
1.1:1	0.005 full scale

Power. Fig. 1 shows a divider network which Samples the rf voltage on the transmission line. The lower divider resistor is shown as a trimpot. A deposited-carbon type was used, but a fixed resistor may be substituted. The trimpot was set so that the full-scale meter reading corresponds to a particular peak power measured by another method (a borrowed power meter or rf voltmeter across the dummy load).

R6 values for particular full-scale power readings are given in **table 2.** The power scale should be calibrated to suit the individual instrument, as it will be nonlinear, the nonlinearity depending on the particular diode used for CR3.

table 2. Resistance value for R6 to change full-scale power reading.

peak power watts	R6 value
500	6.8 ohms
200	two 33 ohms in parallel
100	33 ohms
50	68 ohms
20	two 330 ohms in parallel
10	330 ohms
5	680 ohms
3	1k and 100 ohms in series (linearity suffers)

performance

The current transformer response is essentially aperiodic, with the -6 dB points (compared with midband) from 200 kHz to 40 MHz (see **fig. 4).** This unit has a midband full-scale sensitivity of 500 mW (sensitivity control at maximum). At 50 MHz less than 5 watts is required to carry out measurements, even with low SWR values,

The inherent impedance of the unit was measured using a 5-watt TEK dummy load and a Hewlett-Packard vector impedance voltmeter. The results are illustrated in **fig. *5.*** The impedance discontinuities displayed are well within the accuracy limitations of the meter movement. The real (or resistive) component of the reflectometer's impedance is within 5 per cent of the nominal 50 ohms — most of this variance probably is due to connector and construction discontinuities.

The variation in the real part of the impedance is within ± 1 ohm across the frequency range and can be essentially ignored. The reactive component is negligible up to 30 MHz, at which point it begins to become slightly capacitive. This is largely immaterial. The overall impedance decreases rapidly above 100 MHz.

The short length of coax through T1 is "short" compared with the wavelength at 100 MHz, and it is physical discontinuities that contribute to the inaccuracies measured, becoming significant around 100 MHz. The SO239s used aren't constant-impedance connectors, and they probably contribute as much as construction to the upper-frequency limitation.

possible modifications

For higher powers, the sensitivity of the reflectometer may be varied by one or several of the following

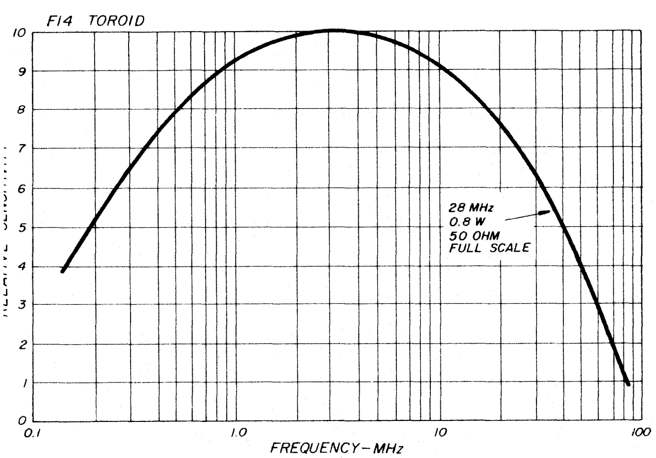


fig. 4. Relative sensitivity of the reflectometer portion of the meter. It takes 0.8 watt across 50 ohms to produce a full-scale reading at 28 MHz.

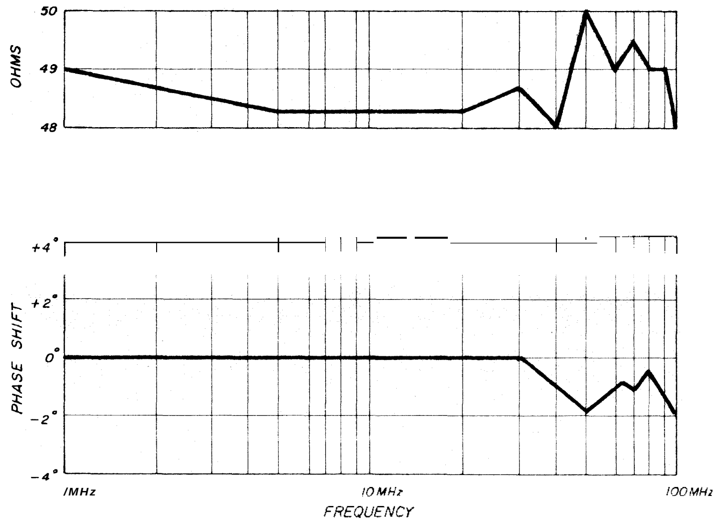


fig. 5. Impedance characteristics of the broadband reflectometer, as measured with a Hewlett-Packard vector voltmeter while terminated with a 50-ohm load.

methods. For power around 20 to 50 watts, R2 and **R3** should be reduced to 47 ohms. For powers above this, the number of turns on the toroid should be reduced. As a guide, twenty turns on T1 with R2 and R3 down to 47 ohms should prove adequate for powers in excess of 150 watts.

The basic reflectometer construction is so simple and inexpensive that several could be built to provide remote monitoring of individual antenna installations. Protection circuitry for transceivers and power amplifiers may be simply implemented using the basic reflectometer circuit driving protection circuitry from the outputs of CR1 and CR2. It is possible to use the reflectometer for swept VSWR measurements by using the differential output of CR1 and CR2.

Accurate measurement of VSWR values below 2:1 can be made by driving an expanded-scale differential voltmeter circuit as described in reference 2. This type of measurement is useful when measuring the VSWR performance of an antenna over a narrow bandwidth, particularly narrow-band loaded whips used for mobile applications.

The basic sensitivity bandwidth may be shifted up in frequency by a decade or more such that it rolls off around 1 MHz at the low end and above 50 MHz at the high end by using a toroid having a permeability of 50 rather than 220.

references

1. **P.G.** Martin, "Frequency Independent Directional Wattmeters," ***Radio Communications,*** (England), July, 1972.

2. H.C. Gibson, G8GCA, "Test Equipment for the Radio Amateur," Radio Society of Great Britain, London, 1974.

ham radio

new approach to measuring SWR at high frequencies

A new type of sampling bridge for measuring power and SWR provides linear response and high directivity

In the past, VSWR measurements were made mainly to check antenna matching characteristics. Therefore, these measurements were performed at fairly high levels, with a directivity of 26 dB being sufficient. The new circuit design in this article provides low-level VSWR measurements between 30 watts and 30 mW incident or reflected power with a directivity of 40 dB.

applications

After the introduction of highly linearized transistors, wideband transformer techniques to cover 1.5 to 30 MHz became increasingly important. Wideband power stages, with exceptionally low intermodulation distortions, are required in test instruments, antenna distribution amplifiers, and ssb transmitters. Because there is higher feedback in transistor circuits than in vacuum tube designs, the return loss of the matching transformer plays an important role. Slight matching changes will degrade the intermodulation distortion performance. Therefore, it has become vital to measure the properties of wideband transformers over a large frequency, power, and directivity range. So far, no instrument has become available that will easily measure these parameters.

circuit description

The initial bridge arrangement for measuring the forward and reflected voltages (power) on a transmission line was invented in Germany by Dr. Buschbeck more than 30 years ago (see **fig. 1**). The obvious disadvantage of this circuit is that the two output ports have a fairly high impedance and are very sensitive to loads. Resistive loads below 10 kilohms will significantly disturb the directivity as well as the frequency response. This requires special diodes for high resolution. Also, this type of power meter suffers from the disadvantage that the power reading is highly nonlinear and depends very much on the mechanical configuration. The other disadvantage caused by the high impedance is that there is always fairly high cross-talk between both output ports, which effectively limits the directivity to about 26-30 dB.

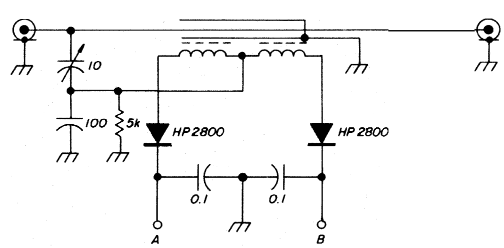


fig. 1. Schematic diagram of a conventional bridge arrangement for measuring SWR. The load across points A and B, if sufficiently low, will significantly affect the directivity as well as frequency response.

In an effort to overcome the rectification nonlinearity, a special version of the tunnel diode, called a

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"backward diode," was used. Backward diodes* have virtually no threshold voltage and exhibit an extremely good approximation to a square law characteristic between a few μV and several mV. However, back diodes used as rectifiers have an extremely low impedance and therefore will load the circuit.

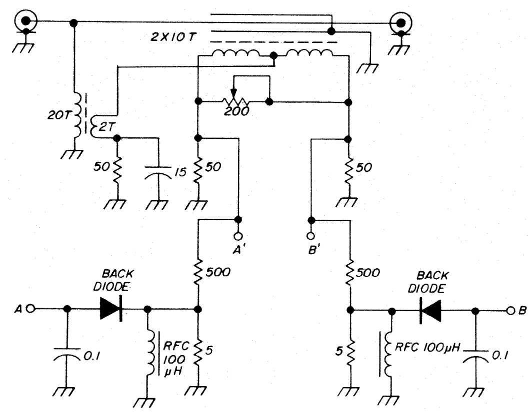

fig. 2. Diagram of a new bridge arrangement that employs back diodes and a transformer voltage divider. This configuration provides wide frequency response and high directivity. The transformers are wound on TC9-type ferrite cores. This is the circuit used in the Rohde & Schwarz NAUS 80 rf power meter.

Because of this loading effect, an attempt was made to use a compensated RC voltage divider, as used in oscilloscopes. This was not successful, however, since the required flatness of output voltage tracking and frequency could not be achieved.

Analyzing Dr. Buschbeck's circuit, it is obvious that the primary limitation is a result of the capacitive voltage divider, which, as a side effect, limits the cutoff frequency at the lower end of the frequency range. To overcome this limitation, a transformer was employed as a voltage divider, which led to the circuit shown in **fig.** 2.

The advantage of **fig. 2** is that both outputs are now terminated in 50 ohms, providing the necessary low impedance for the back diode rectifier. The inductive voltage divider provides a significantly flatter frequency response. In addition, since all impedances are now 50 ohms, the cross-talk is also reduced, allowing directivity of 40 dB.

This new technique can be used effectively to measure the characteristics of wideband transformers and power amplifiers under actual operating conditions. Network analyzers presently on the market are not capable of measuring at this low frequency range and high power level.

"Another description of the back diode can be found in the Hints and Kinks section of *QST* for April, 1978.

ham radio

folded umbrella antenna

An effective, easy-to-build, all-band amateur antenna based on the principles of the folded unipole used in the broadcast service

A survey of amateur antennas would probably show that 95 per cent of them fall into one of four general categories:

1. Horizontal dipole (including inverted vee)
2. Vertical (ground mounted and ground plane)
3. Yagi (multiband and monoband)
4. Quad (multiband and monoband)

The folded umbrella falls into none of these groups — yet, when you see what it can do you'll wonder, "Why not?" This article describes a versatile antenna that is:

1. Broadband on all frequencies, 1.8 through 30 MHz
2. Easily tuned
3. Fully effective without ground radials
4. Without critical dimensions
5. Simple, inexpensive, and easily erected by one man
6. Space saving
7. DC grounded
8. Adaptable to your tower

It seems that most homes lack the space required for a 3.5-4 MHz dipole, or for the ground radials needed to operate a conventional series-fed vertical antenna efficiently. The folded umbrella has evolved from the effort to overcome these space problems.

The first step was to consider an antenna used in the commercial-broadcast field and known as a folded unipole. Shown in **fig. 1**, it is a *grounded* broadcast antenna tower, with steel arms across the top which are connected electrically to the tower. Wires are connected to the ends of these arms and dropped to the bottom, forming a cage around the tower that is insulated from the tower at all points except the top. The cage wires are tied together at the bottom and fed directly at that point with 50-ohm coaxial cable. The advantages claimed for this antenna are as follows:

1. Broadband performance
2. Low radiation-angle
3. Elimination of expensive base insulator
4. Elimination of approximately 6100 meters (20,000 feet) of copper wire
5. Elimination of expensive matching network and weatherproof housing

These features are important to the operator who's trying to put a broadcast station on the air with limited funds.

the result

The folded umbrella is simply the result of several approaches to the development of an antenna which is, roughly, the *electrical equivalent* of the folded unipole. The outcome of the evolution process is shown in **fig. 2.**

design considerations

In the interest of simplicity and economy, this antenna was built around a 12-meter (40-foot), four

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section, telescopic or pushup TV receiving-antenna mast. In the development process, it soon became evident that the cage wires could best be supported by the nylon guys. Thus, the wires were pulled much farther away from the center mast – a fortunate accident as will be seen later.

Also, a 4.6-meter (15-foot) aluminum top-loading whip was added at the 12-meter (40-foot) level, making the antenna 16.8 meters (55 feet) high. The antenna will work well enough without the top loading, but the whip lowers the radiation angle and improves efficiency, especially at frequencies below 7 MHz.

Note that this antenna is *not* to be confused with the conical monopole or the discone. It is not within the scope of this article to discuss the differences, but they are well covered in reference 1.

tuning and matching

The commercial broadcast version of the folded unipole requires no variable tuning or matching arrangement, since it's designed for a single frequency. However, to match the folded umbrella on any frequency across all six high frequency amateur bands, a tuned open-wire line and transmatch unit are used. **Fig.** 3 shows four different transmatch and feedline combinations. **Fig. 3A** is the basic combination, using an unbalanced transmatch. In this configuration, be sure that the side of the feedline grounded at the antenna is the *same side* that's connected to the ground terminal on the transmatch.

Fig. 3B illustrates the use of a balanced transmatch. In this case, a 4:1 balun is used at the antenna to maintain a balanced condition on the feedline.

Fig. 3C shows an arrangement which, theoretically, should unbalance things and bring unwanted rf fields into the shack. However, it has been tried and found successful in some cases. As a matter of fact the experimental antenna, which has drawn so much favorable mail, is operated in this manner. Note the *balanced* transmatch, and *no balun* at the antenna.

Fig. 3D shows the use of a short length of coax when it's inconvenient to bring open-wire line into the station. Keep this coax *as short as possible.* The open-wire part of the feedline should be, ideally, about 20 meters (65 feet) long (or a multiple thereof). When using other lengths it may be difficult to obtain a 1:1 swr on some frequencies, especially the higher frequencies. If this happens, experiment with slightly different line lengths of plus or minus 0.9-3 meters (3-10 feet) until it becomes easy to obtain 1:1 swr on virtually any frequency.

performance

Performance of the folded umbrella is quite gratifying. When operated properly, the swr should be 1:1, and your rig should see a 50-ohm resistive load on all hf amateur frequencies. On 1.8 MHz, the folded umbrella substantially outperforms a half-wave inverted vee, whose apex is 15 meters (50 feet) above ground. For all other bands, the comparison antenna is a multiband inverted vee, 40 meters (130 feet) long, 15 meters (50 feet) above the ground, with tuned feeders and transmatch.

From 3.5-4 MHz, the inverted vee is generally better up to 805 km (500 miles) because of the high radiation angle of the inverted vee or the low radiation angle of the folded umbrella. Between 805-1610 km (500-1000 miles) the superiority of the two antennas alternates depending on propagation conditions. Beyond 1610 km (1000 miles) the folded umbrella takes over, its superiority improving with increasing distance.

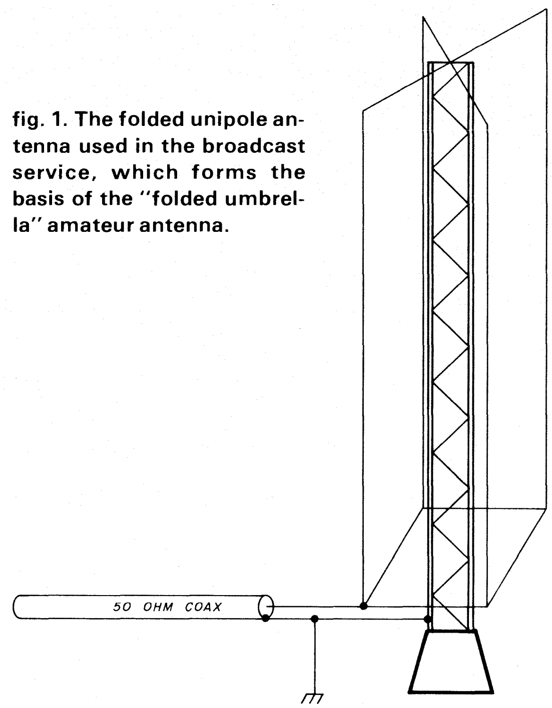


fig. 1. The folded unipole antenna used in the broadcast service, which forms the basis of the "folded umbrella" amateur antenna.

From 7-7.3 MHz, the folded umbrella definitely outperforms the inverted vee at any distance. It appears that the diamond shape of the wire cage begins to provide a measure of cross-polarization from 7 MHz up resulting in the following:

1. A diversity effect, which minimizes fading on the transmitted signal
2. A much better snr on receive because of the closed-loop design and because the antenna is dc grounded
3. Broadband performance. For example, if the folded umbrella is tuned for a 1:1 swr at 7.150 MHz and the transmatch is left untouched thereafter, the maximum swr observed at either 7.000 or 7.300 MHz is 1.2:1

The experimental model was not used much between 14 and **30** MHz since a quad is generally used on these bands, but performance is comparable to that of a dipole on these frequencies. Other builders have reported excellent DX results on the upper bands.

operation without transmatch

The foided umbrella is basically a 3.5-4 MHz antenna. If it is fed directly with coax and no transmatch, the swr is less than 2:1 at both **3.5** and **4** MHz. From **1.8** to **2** MHz the coax-fed antenna shows an swr of approximately **2:1**. On the 7-MHz band, swr is in the 2.5:1 range. At **14.2** MHz, swr is 5:1, at **21.3** MHz, swr drops back to 1.5:1, and on **30** MHz, swr is about 3:1.

Since none of these swr numbers is really high, a transmatch and tuned feedline can easily subdue them, resulting in *1:1* swr across all six bands.

using your tower

By applying unipole principles, you can use your tower. Just start by connecting one wire at some arbitrary point on the tower, say **12** meters (40 feet)

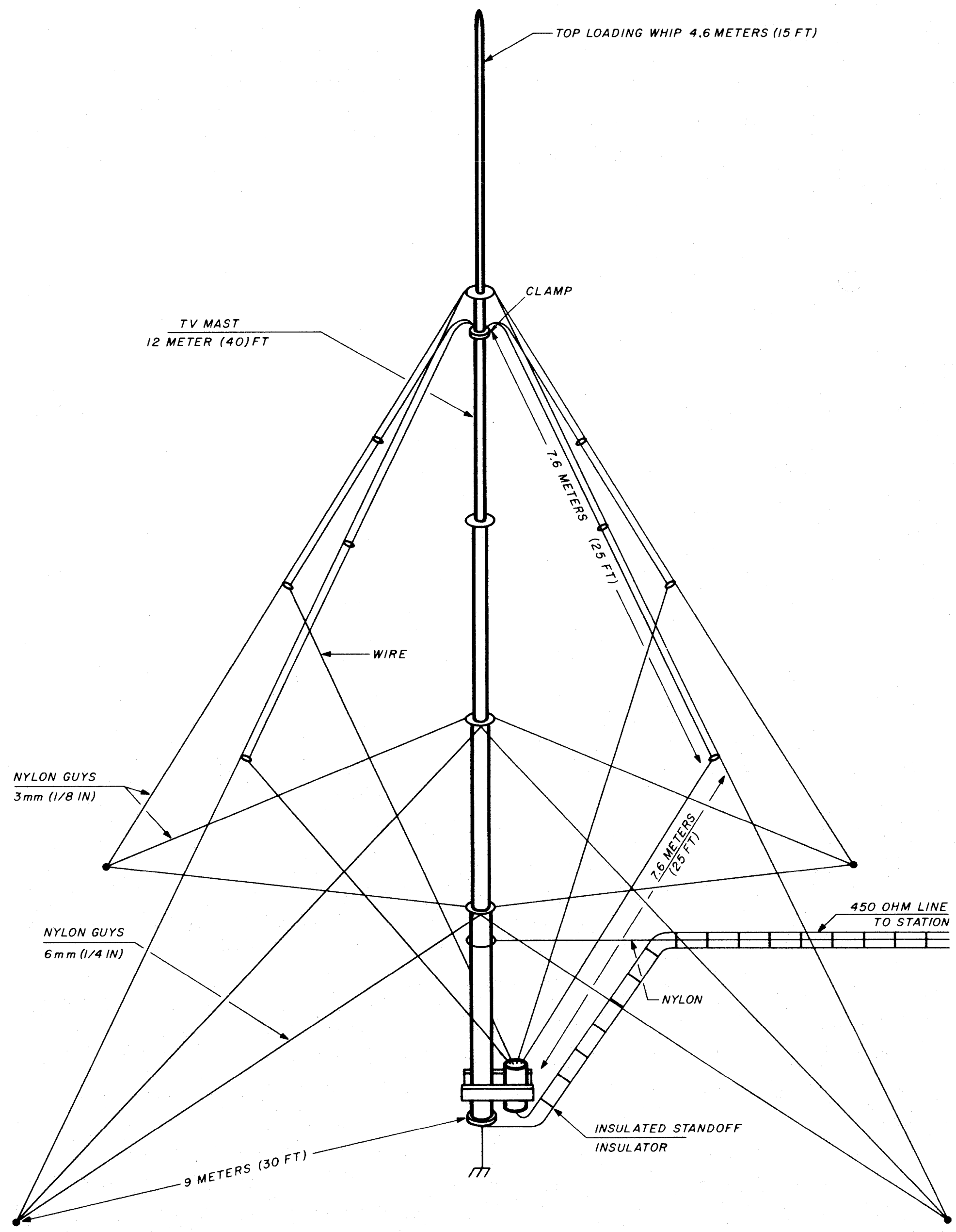


fig. 2. Construction details of the folded umbrella antenna, which is, roughly, the electrical equivalent of the folded unipole. Dimensions are approximate and are not critical.

above the ground. Use nylon line to pull the wire away from the tower. Then, bring the wire back to the bottom of the tower, as shown in fig. **4.**

Use an exciter/vswr meter and a bridge or a grid dipper to check the resonant frequency. If it falls within the 3.5-4 MHz range, add the other three wires and proceed as with the folded umbrella. Your beam should provide adequate top loading. Be sure to ground the tower well.

construction

1. Set up the collapsed pushup mast, using 6-mm (1/4-inch) nylon guys on all four sides. Use the closely woven nylon, not the loosely braided type. The former is much stronger and will not unravel. Do not

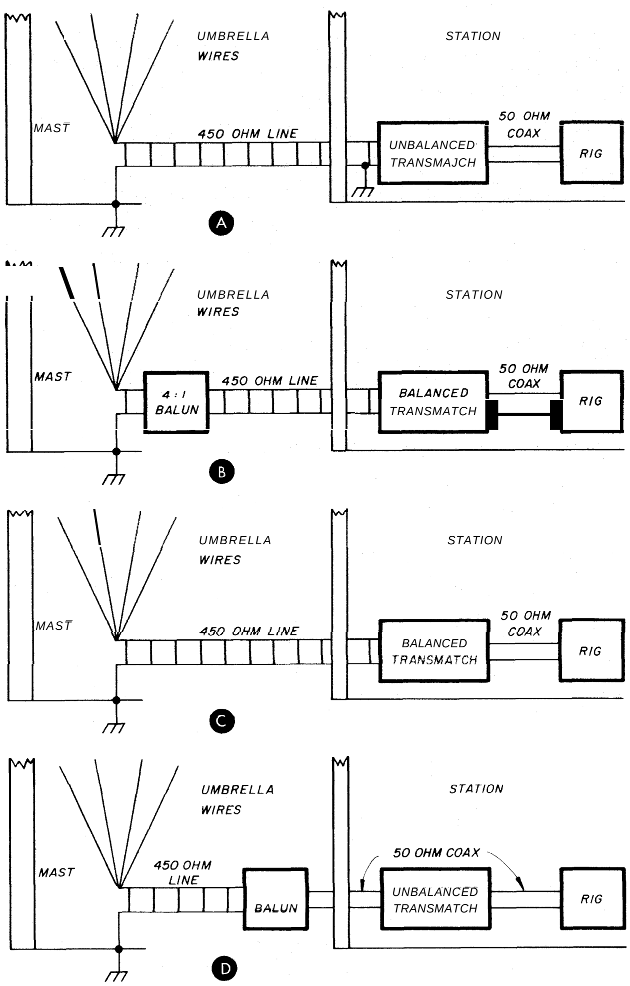
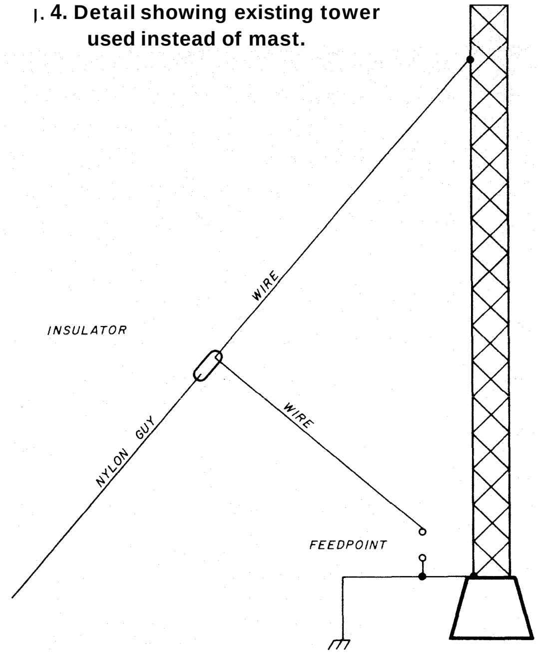


fig. 3. Transmatch and feedline configurations for the antenna. Sketch (A) shows the basic arrangement using an unbalanced transmatch. Note grounding. In (B) a balanced transmatch is used. A 4:1 balun is required. Sketch (C) shows an arrangement as in (B), but without the 4:1 balun. Transrnatch is balanced (no ground). Sketch (D) shows how to use a short length of coax cable between the open-wire line and the station. The coax cable length should be as short as possible.

|. 4. Detail showing existing tower used instead of mast.



use other synthetic line. Some of the other kinds of line become brittle with exposure to sunlight and will eventually break, while nylon retains its strength indefinitely. Nylon will stretch, but the close-woven type requires tightening only a few times initially.

2. Place the guy anchors at least 9 meters (30 feet) from the base of the mast (or use the house or trees where possible). Tighten the bottom guys firmly and place your ladder against the mast for further work.

3. Metal rings are supplied on the TV mast for attachment of guys. Be sure these rings are now in place before proceeding.

4. Insert 4.6 meters (15 feet) of aluminum tubing into the top of the pushup mast. This tubing can be made up of two or three telescoping pieces, if desired, just so long as the bottom (largest) piece fits snugly inside the top section of the pushup mast. The tubing can be secured by drilling through and bolting or by slotting the pushup mast and using a clamp (see fig. 5).

5. Attach four nylon (close-woven) guys to the top guy connection ring (fig. **5**).

6. Pull the top section up out of the collapsed mast assembly about 0.6 meter (2 feet) to facilitate connection of the umbrella wires.

7. Tie two small loops in each of the four top nylon guys as shown in fig. **2.** These loops should be about 4 and 8 meters (12.5 and 25 feet) from their attachment points on the mast.

8. Install umbrella wires.* Use approximately 16.8 meters (55 feet) of wire for each of the four elements. Clean and tin about 5 cm (2 inches) at one end of each wire. Clean and sand the mast just under the guy attachment ring. Using a stainless-steel hose clamp, attach the umbrella wires, spacing them equally around all four sides of the mast (fig. **5**). Smear silicone rubber cement (GE or Dow-Corning) liberally over the hose clamp and bare wire ends to prevent corrosion of the electrical connection.

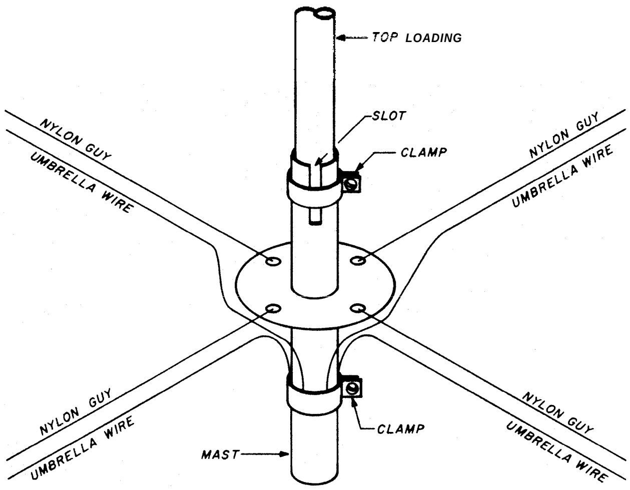


fig. 5. Detail showing attachment of top guys, "umbrella" wires and top loading whip.

9. Thread one of the antenna wires through the two loops on each of the guys. You are now ready to begin raising the mast.

10. Use heavy leather gloves and be very careful when pushing the mast up. Whenever you stop the mast part way up, be sure to tighten the locking bolt very firmly, using pliers or a sufficiently large crescent wrench. If the mast should slip unexpectedly, it can pinch and cut your hand most painfully. This is the reason for the heavy leather gloves.

11. As you raise the mast, temporarily secure the bottom ends of the umbrella wires to their respective guys with tape. Then, keep these four wire/guy assemblies tied away from the mast as you raise it. This prevents the wires and guys from becoming tangled during the erection of the mast.

12. Now, push the top section up to its full length. There will be a hole at the top of the next section below. When the bottom of the top section is pushed high enough to clear this hole, push a large cotter pin (provided in the mast hardware) through this hole. Spread the ends of the cotter pin only enough to secure it, which will make it easy to remove at some later time. Now, tighten the locking bolt firmly. Repeat this process on the other sections.

13. Push the next section up and secure as above.

14. Attach four more 3-mm (1/8-inch) nyion guys to the ring at the top of the section just above the bottom section. This set of guys will end up, on full erection, 6 meters (20 feet) above the ground (fig. **2).**

15. Push this last section up and secure, as above, and temporarily secure all guys.

16. Now, provide a bracket at the base of the mast for the purpose of connecting the four umbrella wires together while insulating them from the mast. This is the feedpoint for the antenna. One suggested method of anchoring and insulating the feedpoint is shown in figs. **2** and **6.** However, as with other parts of this project, there are many different mechanical arrangements which will produce the same electrical results. Use your ingenuity. Remember, none of the dimensions are really critical, since the tuned feeders compensate for physical variations. If you're following fig. **6,** however, pull the umbrella wires through the PVC tubing, tighten the clamp below the tubing, and trim off excess wire, leaving several cm of wire below the clamp. Now, bare the wires below the clamp and solder them together.

17. Drive three or four 2.4-meter (8-foot) ground rods around the base of the mast and connect them

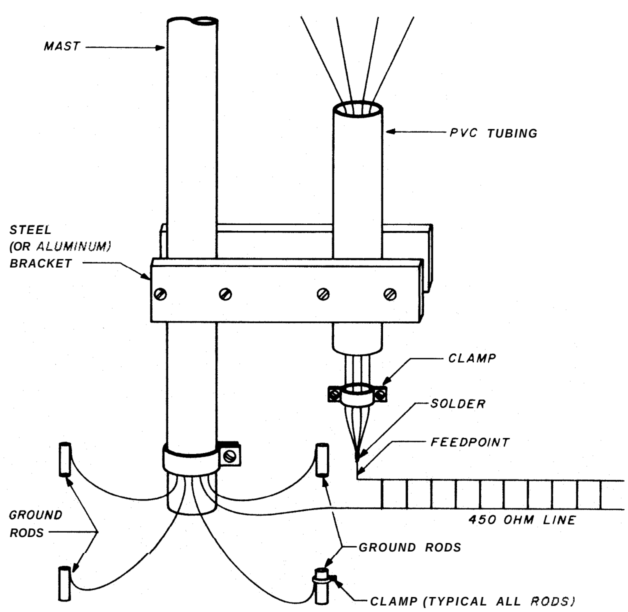


fig. 6. Mechanical details at the bottom end of the TV mast. The ground rods are essential for good performance.

*Any available type of copper wire up to **2** mm (no. 12 AWG), solid or stranded. The experimental model uses surplus 1.3 mm (no. 16 AWG) stranded, insulated wire.

securely to the mast with heavy copper wire, tubing or strap **(fig. *6*).**

18. Attach feedline, soldering one side to the feedpoint and connecting the other side to ground. The most practical line to use is the 450-ohm plastic ladder line made by Saxton. It is strong, flexible, and much easier to use than bare open wire while just as effective. Furthermore, it's affected very little by ice,

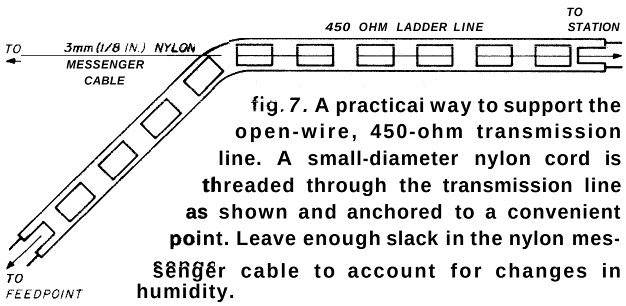


fig. 7. A practical way to support the open-wire, 450-ohm transmission line. A small-diameter nylon cord is threaded through the transmission line as shown and anchored to a convenient point. Leave enough slack in the nylon messenger cable to account for changes in humidity.

snow, or rain, as far as tuning is concerned. **Figs. 2** and ***7*** show a simple, practical way of supporting the line. For the experimental model, it was necessary to run the line 18 meters (60 feet) to the house, keeping it at least 3 meters (10 feet) above ground. To relieve tension on the line, 3-mm (1/8-inch) nylon was threaded through the holes in the line and stretched from the mast to the house, serving as a "messenger" cable. Here again, use your own ingenuity to fit your situation.

19. After tuning and testing, tighten and secure the guys, using a plumb level to be sure the mast is vertical.

closing remarks

Here is an antenna that can do many things for you. Don't expect it to perform like a beam – but it will more than hold its own against conventional dipoles and trapped verticals at all frequencies and on all high frequency amateur bands.

I have received literally hundreds of inquiries from people who were impressed with what they heard when I was using the umbrella antenna. All inquiries have been answered, and those who have built the antenna report equally gratifying results.

The writing of this article was deliberately delayed pending receipt of data from others to provide the reader with authentic, reliable information.

reference

1. Paul H. Lee, K6TS, "The Amateur Radio Vertical Antenna Handbook," Cowan Publishing Company, 1974, pages 49-52 (discones and monopoles); pages 75-80 (folded unipoles).

ham radio

80-meter broadband antennas

A discussion of different approaches to broadbanding 80-meter antennas, including theoretical calculations and actual model measurements

Judging by the horrendous congestion that can be found on the 80-meter band almost any evening, it is certainly one of the most popular bands. But all of its users share the same problem with antennas. The ratio of the upper limit of the band to the lower limit is 1.143:1, larger than that of any other high-frequency Amateur band. Building an antenna which has adequate bandwidth for the entire 500 kHz may seem like an impossibility.

Years ago, transmitters had a very large range of matching capability. This luxury disappeared when the compact transceiver appeared on the scene over ten years ago. The problem has been compounded by the introduction of transmitters with solid-state power amplifiers requiring a load very close to 50 ohms. Protection circuits are normally employed to reduce the driving power to avoid damage to the output transistors. The use of a transmatch will eliminate the mismatch problems, but takes away the convenience of frequency changes without retuning.

The antennas shown in this article exhibit acceptable impedance matches over much more than the usual bandwidth. Although these designs may not be the perfect solution in every installation, they should provide starting points for further experimentation.

current designs in use

One of the classic ways to increase the bandwidth of a simple dipole is to enlarge the effective diameter of the conductors in the radiating elements. The cage antenna is one example of this, in which each half of the dipole is made of several conductors. The multiple conductors are suspended on spreaders at two or more places. This simulates a conductor of roughly the diameter of the spreader. The fan dipole, or bow-tie, antenna is a simplification of the cage arrangement. In this design, each half is composed of two wires joined at the feedpoint. The wires fan out from this point and are held apart by a single spreader at the far end.[1] Antennas of this type provide a noticeable increase in the bandwidth over which a transmitter may be adjusted for a conjugate match, but they have apparently not enjoyed much popularity. Perhaps it is because these antennas are considered somewhat unsightly. The antennas I'll discuss have wider bandwidths for the same level of complexity (and lack of beauty).

antenna models

To provide a more convenient means of attacking the problems associated with broadband antennas, two types of antenna models were used to test the designs. Simple mathematical models were used to

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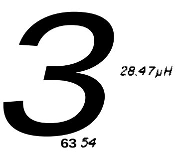
allow prediction of the approximate antenna impedance match vs frequency, before they were constructed, whenever possible. An equivalent circuit for a simple wire dipole appears in **fig. 1.** This model accounts for the change in reactance at the feedpoint as the frequency is changed, but it does not allow for the normal tendency of the resistive component of the feed impedance to rise with increasing frequency. This tends to be a second-order effect and doesn't seriously affect the results. It does, however, greatly simplify the calculations.

If the resonant frequency of this mathematical model is checked, it wili be found to be 3742 kHz. It is not 3750 kHz, which is the arithmetic center of the band. Like all resonant circuits, the model exhibits *geometric symmetry.* The resulting effective center of the band is $\sqrt{3500\ kHz \times 4000\ kHz} = 3742\ kHz.$ This frequency is used as the band center throughout the calculations to maintain band-edge symmetry when plotting the SWR. In real antennas, the many other variables involved make the 8-kHz distinction unimportant.

The plot of the calculated impedance for this mathematical model is shown on the Smith chart in **fig. 2.** The accompanying SWR plot is given in **fig.** 3.

fig. 1. Approximate equivalent circuit of a standard wire dipole antenna.



This is typical of several antennas I have used in the past. These antennas were mostly inverted vees with center heights of about 12 meters (40 feet) and the ends about 3 meters (10 feet) high. A 5:1 SWR is considered intolerable by many Amateurs and much worse than a typical antenna would measure. This is probably because many SWR meters read closer to 3.5:1 for such an antenna. Close inspection with a noise bridge or other impedance-measuring equipment will show a typical SWR meter to be an incurable optimist. This defect is especially obvious when attempting to read fairly high mismatches at lower power levels.

Calculations of predicted antenna impedances and matches were facilitated by the use of programmable scientific calculators (HP-55 and TI-59) and an IBM 370 running the SPICE circuit analysis program written at the University of California, Berkeley. The antennas were also physically modeled by scaling them up in frequency and down in size by a factor of almost nineteen. This moves the resonant frequency to 70 MHz, which is the center of the i-f passband used in most microwave equipment designed for

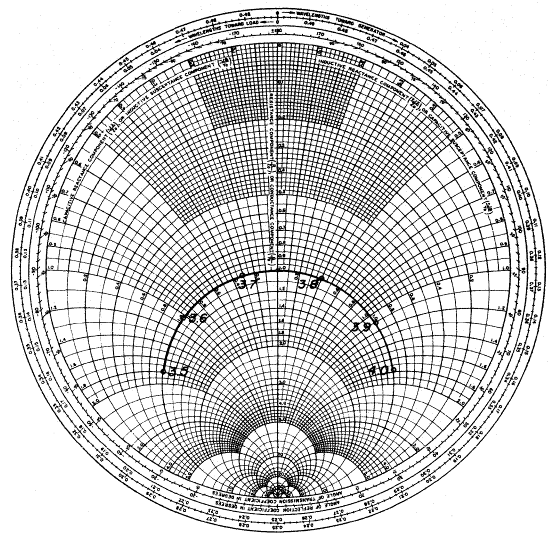


fig. 2. Calculated input impedance of the dipole equivalent circuit, normalized to 50 ohms.

telecommunications service. Microwave link analyzers cover the range from 50 to 90 MHz and include the capability of accurately measuring return loss. Return loss is just 20 times the logarithm of the reflection coefficient, expressed in dB. Although the Hewlett-Packard 3702B/3710A and its companion return loss hybrid are designed for 75-ohm measurements, the use of a minimum-loss resistive matching pad permits its use with nominal 50-ohm loads. The loss of this pad was accounted for in the measurements made.

The scaled antennas were suspended from the ceiling inside a laboratory for the measurements. There were many reflections from the metal framework of the building, but they were probably on a par with those one could expect from power lines and the exteriors of buildings in the vicinity of an 80-meter antenna. All of the plots in this article show the equivalent scaled-down frequencies corresponding to the measurements made on the scale models. All of the necessary corrections have been made.

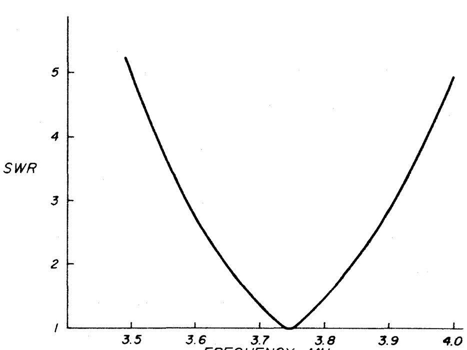


fig. 3. Calculated SWR of the dipole equivalent circuit.

parallel antennas

This antenna configuration was suggested by the common use of paralleled dipoles for several bands. Others have apparently tried paralleling two antennas cut for different parts of the same band to widen the overall bandwidth with little success. The secret for

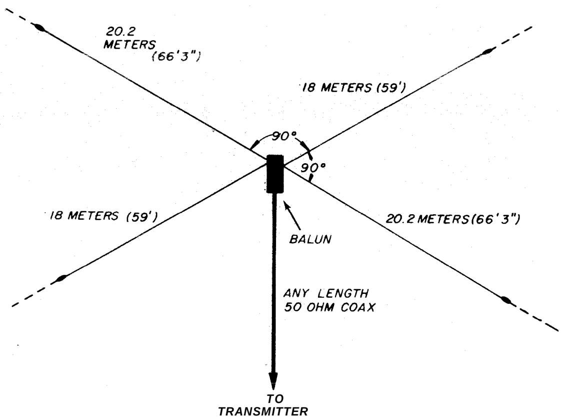


fig. 4. The parallel antenna is formed by two stagger-tuned dipoles, mounted at right angles to each other and fed at a common feedpoint.

proper operation is to mount the two antennas in such a way that they do not couple to each other. This can be done by mounting them at right angles as shown in fig. **4.**

One of the two dipoles is cut for 3530 kHz and the other to 3966 kHz. The higher frequency is 1.06 times the geometric center of the band (3742 kHz). The

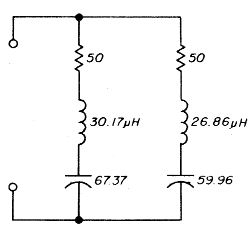


fig. 5. Approximate equivalent circuit of the parallel antenna. Individual resonances are at 3530 and 3966 kHz.

lower frequency is the center frequency divided by 1.06. The equivalent circuit of this arrangement is given in fig. **5.** It is just the parallel combination of two dipole equivalent circuits which have been adjusted up and down in frequency by multiplying and dividing the reactance values by the 1.06 factor.

It is recommended that this antenna, and the others shown as well, be fed through a balun. Currents flowing on the outer conductor (shield) of the coaxial feedline can cause undesired coupling between the antennas, which can restrict the bandwidth.

The calculated impedance of the antenna is shown in fig. ***6.*** The loop in the curve is interesting. The

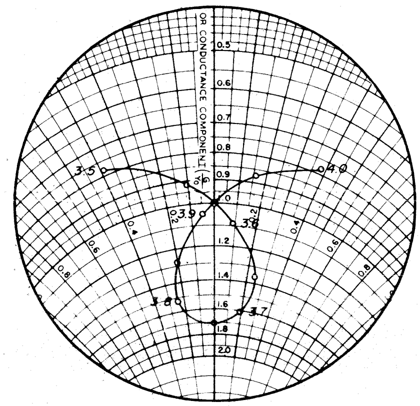


fig. 6. Impedance calculated for the parallel antenna equivalent circuit, normalized to 50 ohms.

antenna impedance is capacitive at low frequencies just like the single dipole, crosses over through a perfect match, then becomes inductive. The impedance then becomes capacitive again as the frequency increases, goes through perfect match once again, and finally becomes inductive once more at the top of the band. The calculated SWR is shown by the solid line in fig. ***7.*** Measurements on the scale model of the antenna are also shown on the same plot by the dashed line. The measurements show good agreement with the calculations (most of the discrepancy between the two curves results from the individual dipoles never having a perfect match to 50 ohms). Despite this, the SWR is better than 2:1 across the entire band. If you can make dipoles that have a wider bandwidth than the ones assumed here, your antenna could be even better than this.

One of the interesting features of the parallel antenna is its tendency to be omnidirectional. Like the turnstile antenna, the currents in the two dipoles are out of phase. Over the middle of the band, at least, the currents in the two dipoles are roughly equal. Fig. **8** shows the calculated phase difference between the two dipoles as a function of frequency. The ratio of the two currents, given in dB, is plotted in fig. **9.** Even though the currents can differ by more than 10 dB at the band edges, the normal nulls in the

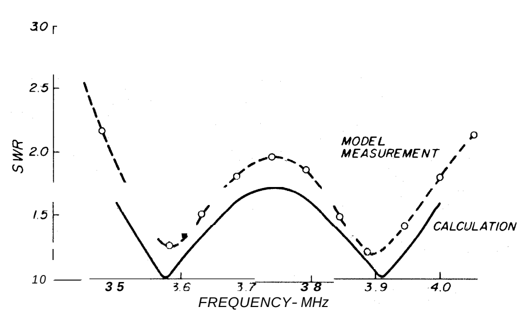


fig. 7. Calculated and measured SWR of the parallel antenna.

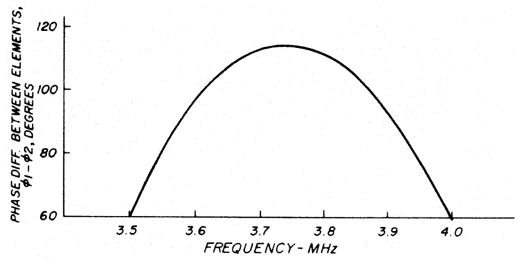


fig. 8. Calculated phase angle between the currents in the two stagger-tuned dipoles of the parallel antenna.

radiation pattern of a dipole will be filled to a large extent. Signals radiated at a high angle will be elliptically polarized. This is of some benefit when receiving. The antenna will respond fairly well to linearly polarized signals which arrive parallel to either antenna from high elevation angles. Since many of the signals received on the 80-meter band arrive at high angles, the elliptical polarization is probably of much more value than the omnidirectional characteristics observed toward the horizon.

It is worth noting what happens when the resonant frequencies of the two dipoles are separated by a different ratio. Calculations were made at two other pairs of resonant frequencies. When the ratio is 1.05 instead of 1.06, the SWR at the band edges is about 2.5 to 1 and the midband SWR is about 1.4 to 1. If the ratio is reduced to 1.04, the SWR at the middle of the band barely rises to 1.05 to 1, yet the SWR remains about 3.1 to 1 at the band edges. The loop in the impedance plot on the Smith chart shrinks, but it does not disappear until the two dipoles are tuned to the same frequency.

Most Amateurs probably don't have a house lot of the right shape to permit placing two such antennas at right angles to each other. It may be desirable to erect the antennas as inverted vees instead of as conventional dipoles, Increasing the coupling between the antennas will reduce the bandwidth of the system, but the use of inverted vees has a minimal effect

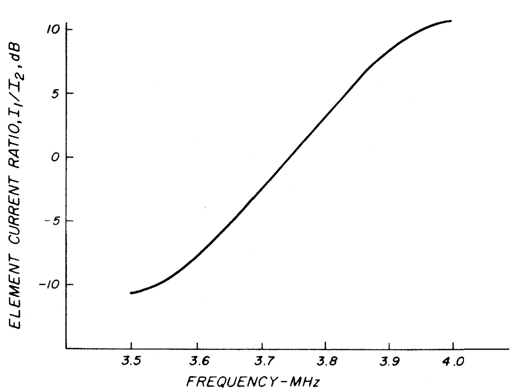


fig. 9. Calculated current ratio between the two dipoles of the parallel antenna.

on the coupling. The angle between the dipoles may deviate from 90 degrees by about 15 degrees without significant degradation. If the angle decreases below 60 degrees, the match at the high end of the band will deteriorate. As the angle is decreased even more, the lower frequency antenna seems to dominate, and the tuning of the higher frequency dipole has less

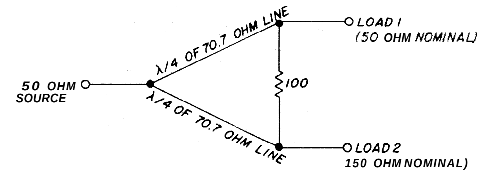


fig. 10. Schematic representation of the two-way Wilkinson hybrid with the ideal transmission line impedance.

and less effect on system operation. When the antennas are nearly parallel, the SWR curve appears to be that of the lower frequency antenna alone.

Wilkinson hybrids

Before looking at the next antenna configuration, a discussion of a very interesting circuit is in order. The Wilkinson hybrid is very commonly used as a

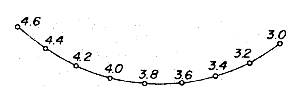


fig. 11. Calculated input impedance of the Wilkinson hybrid terminated by two 50-ohm loads. The quarter-wavelength lines are assumed to be 75 ohms. Data are normalized to 50 ohms.

power divider or combiner at microwave frequencies.[2] The arrangement of the circuit is shown in **fig. 10.** This is the simplest form of the circuit. Other variations allow driving more than two loads. Modifications providing very wide bandwidths have also been developed.[3]

Fig. 11 shows the calculated impedance of a Wilkinson hybrid terminated with 50-ohm loads. The

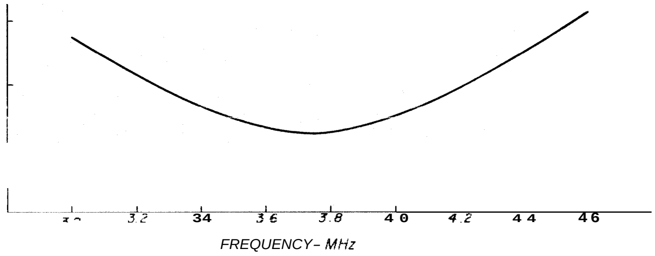


fig. 12. Calculated SWR at the input of the Wilkinson hybrid using 75-ohm cables. Each leg was terminated in a 50-ohm resistive load.

quarter-wavelength lines are assumed to be 75 ohms instead of 70.7 ohms to indicate the performance to be expected with available cables. For this reason, the input impedance is never exactly 50 ohms. The corresponding SWR is shown iii **fig. 12.**

Like other hybrid circuits, the Wilkinson provides isolation between the two loads, This occurs because a signal entering the *Load 1* port can travel to the *Load2* port by two paths. One path is directly through the 100-ohm resistor; the other path is through one of the quarter-wavelength lines to the source and then back up the other quarter-wavelength line. Because the second path totals one half-wavelength, the signal traveling this route is 180 degrees out of phase with the signal through the resistor. As it happens, the amplitudes of the two signals are equal, and complete cancellation occurs.

No circuit is perfect, however, and several imperfections can cause a reduction in the isolation obtained between the two load ports. It should be obvious

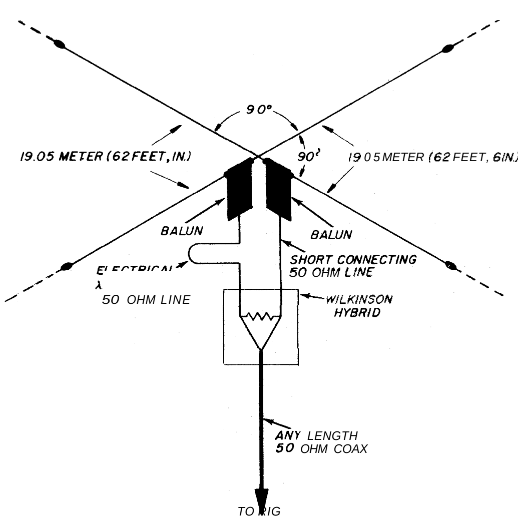


fig. 13. Diagram of the turnstile antenna driven with a Wilkinson hybrid. Both dipoles are tuned to the band center and are mounted at right angles to each other.

that the rejection is perfect only at the design frequency, since the transmission lines give the right phase shift only at that frequency. Even so, the isolation will be better than 20 dB over the whole 80-meter band. The other major source of less-than-ideai performance is the possibility that the driving source is other than 50 ohms. Most Amateurs fail to realize that the output impedance of a power amplifier designed to drive 50-ohm loads is seldom 50 ohms. This effect could easily cause a 10-dB degradation of the isolation between the output ports.

One clever use of the Wilkinson hybrid with Amateur antennas has been in driving phased vertical arrays.4 The isolation of the hybrid is useful in preventing interaction of the antennas via the phasing

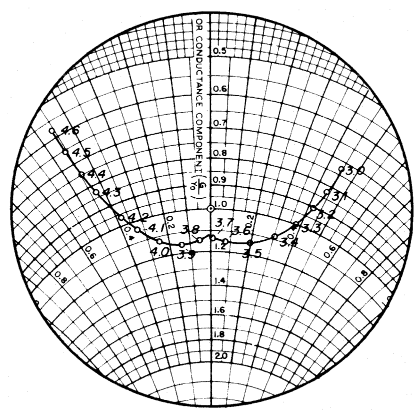


fig. 14. Input impedance, calculated using SPICE program, of the turnstile antenna driven by a Wilkinson hybrid with 75-ohm lines. Data are normalized to 50 ohms.

harness, which makes designing and adjusting such an array much easier.

Since the power from the input of the hybrid is equally split, the loss is 3 dB between the input and each output. Power coming in either of the two load ports is attenuated by 3 dB because half of the power goes back into the source port and the other half is dissipated in the 100-ohm balancing resistor. This has an interesting effect on the reflected power — it is cut in half! Unfortunately, when the reflected power is burned up in the load, it cannot be re-reflected and travel again out to the antenna or other load where the power is wanted. This causes a drop in system efficiency.

A special case occurs when two identical loads are driven by the hybrid: one load driven directly from the hybrid output and the other load driven through a quarter-wavelength of 50-ohm cable. If the loads are not 50 ohms, there will be a reflection at the load back toward the source. Since one of the loads is 90 degrees farther away, the reflection coming back will be a total of 180 degrees out of phase with the reflection from the directly connected load. The reflected waves are thus equal and out of phase. The power summing action of the Wilkinson hybrid will cause them to cancel out. Where did the power go? It is all dissipated in the 100-ohm balancing resistor. None of the reflected power will arrive back at the source, and the apparent match is perfect. This leads directly to the next antenna system.

turnstile plus Wilkinson

A turnstile omnidirectional antenna is shown in **fig. 13.** The Wilkinson hybrid is used to drive the two antennas with equal power. An extra electrical quarter-wavelength of 50 ohm coax is connected

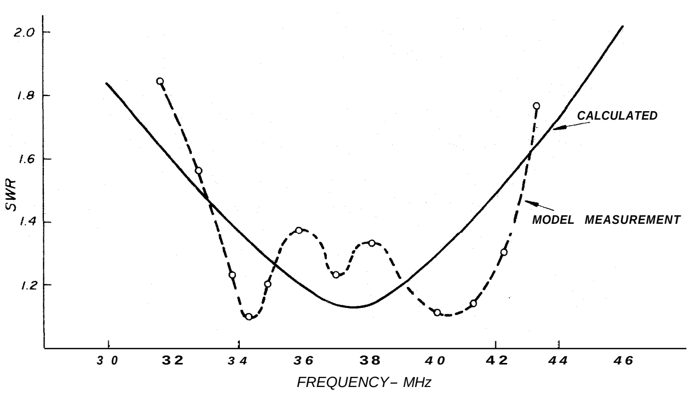


fig. 15. Calculated and measured SWR of the turnstile antenna driven with a Wilkinson hybrid.

between the hybrid and one of the antennas. The hybrid is made of two electrical quarter-wavelengths of 75-ohm coax and a high power noninductive 100-ohm resistor. The bandwidth of this antenna system is unbelievable! See **fig. 14** for the calculated impedance of the system. The calculated and measured values of SWR appear in **fig. 15.** Many dummy loads don't match this well. It is perhaps an apt comparison, unfortunately; the bad news is in **fig. 16.** This plot shows the equivalent loss as a function of frequency. At the edges of the 80-meter band, about 2.5 dB of transmitter output is turned to heat in the 100-ohm resistor. If your kilowatt power amplifier has 65 per cent efficiency, almost 285 watts must be dissipated at 3.5 or 4 MHz, leaving only 365 watts to be radiated.

These calculations are still based on the original dipole equivalent circuit with the 5:1 SWR at each band edge. The amount of loss in the resistor is reduced if the antennas fed by the hybrid are wide in bandwidth.

Baluns are almost mandatory for this antenna configuration. The cables in the phasing line and the

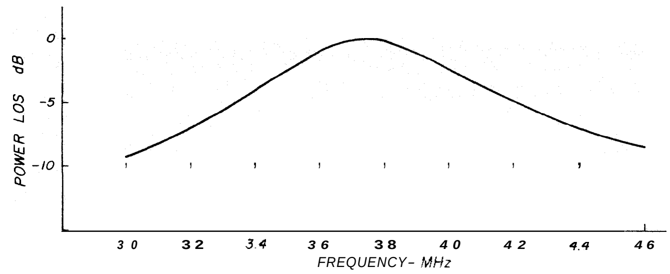


fig. 16. Calculated equivalent loss in the balancing resistor for the turnstile antenna driven by a Wilkinson hybrid.

hybrid will cause undesirable resonances to affect the impedances of the two dipoles if currents are allowed to flow on the cable sheaths. Whether a balun is used or not, the connections at the feedpoints *must* be insulated from each other. If no balun is used, this means the two shields of the coaxial cables at the feedpoints cannot be allowed to make contact, since these points are *not* at zero potential with respect to each other.

If a solid-state transmitter is in use and a transmatch is considered undesirable, an antenna of this type might be a good compromise. The power lost in the balancing resistor in the hybrid might be less than you would lose when the automatic drive reduction circuit is activated by another antenna with a poorer match.

conventional turnstile

An obvious question occurs immediately. Why not remove the 100-ohm resistor? This would transform the antenna into a conventional turnstile antenna. The Smith chart in **fig. 17** shows what happens

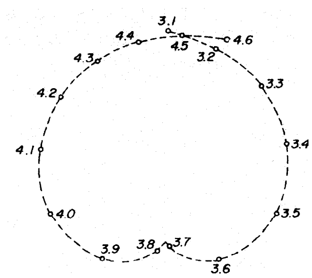


fig. 17. Impedance calculated for a conventional turnstile antenna (no 100-ohm balancing resistor), normalized to 50 ohms.

when this is done. The corresponding SWR curves are shown in **fig. 18.** The results are not that impressive, but are included here to satisfy possible curiosities. If the necessary space for all of the wire is available, a better choice would be the parallel antenna. This would eliminate ail of the coaxial transformers and provide an increase in bandwidth.

coaxial trap antenna

When the wondrous ciaims about the "double bazooka" antenna came forth, it was hard to resist trying one. Even if only 10 per cent of the claims were true, it still had to be amazing. After the disillusion-

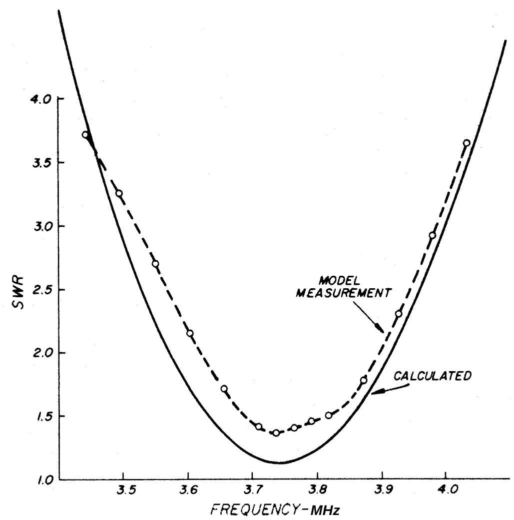


fig. 18. Calculated and measured SWR **of the conventional turnstile antenna.**

ment,[5] those strange lengths of RG-58/U cable cried for use in an antenna. The resulting design is sketched in **fig. 19.**

The shorted quarter-wavelength sections of coax look like parallel resonant traps that decouple the short end-sections at the high end of the band and make the antenna think it is shorter. At the lower end of the band, the shorted pieces of coax look inductive, slightly loading the antenna and allowing the antenna to be somewhat shorter physically than would otherwise be required. The larger diameter of the coaxial traps also contributes to the increased bandwidth.

The synthesis of a mathematical model for this antenna was not attempted. However, both full-size and scale models of the design were built. The full-size antenna used RG-58/U and the scale model used RG-174/U, a miniature 50-ohm coax. The scaling of the conductor diameters is not in proportion to the frequency ratio, which certainly results in some errors.

A Boonton model 250-A RX meter was used

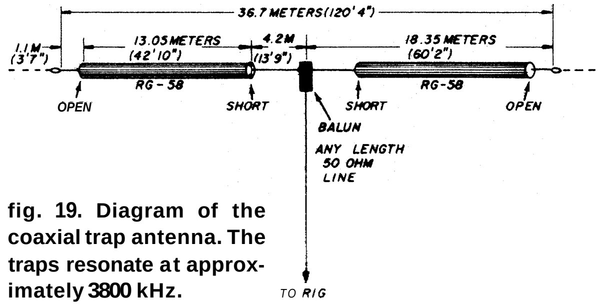


fig. 19. Diagram of the coaxial trap antenna. The traps resonate at approximately 3800 kHz.

through a known length of 50-ohm cable to measure the impedance of the scale model. The scaled down frequencies and impedances are plotted in **fig. 20**. The equivalent SWR curve is shown in **fig. 21.** Notice that the shape of both plots is similar to those of the previously mentioned parallel antenna, although the overall match is not as good. This antenna is worth consideration when the necessary acreage is not available for two full-size dipoles.

Tuning the coaxial trap antenna can be difficult. The first thing to do is to grid-dip the traps as shorted half-wavelength sections at about 7.6 MHz. A noise bridge could be used to look for a zero impedance at this frequency instead. When this is done, the traps will be quarter-wavelength at about 3.8 MHz. If a noise bridge or grid-dipper is not available, cutting to the specified length shown may be close enough.

If the coaxial cable used for the traps has foam-polyethylene insulation, the traps must be lengthened to approximately 15.85 meters (52 feet) and the inner wires shortened to about 1.4 meters (4 feet, 7 inches). For Teflon-insulated cables, the appropriate dimensions are 13.75 meters (45 feet, 1 inch) and 3.5 meters (11 feet, 6 inches).

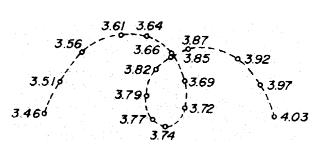


fig. 20. Measured impedance of the scale model of the coaxial trap antenna normalized to 50 ohms. Frequencies are scaled down to 80-meter equivalents.

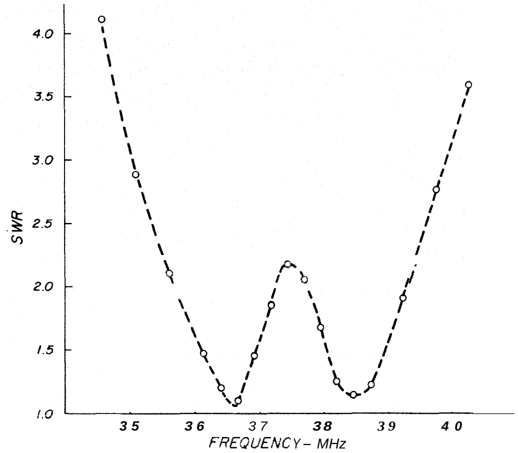


fig. 21. Measured SWR of the scale model coaxial trap antenna.

Once the coaxial traps are cut, connect the wires to them that go from the shorted end of the traps to the feedpoint. Leave these pieces of wire about 30 cm (1 foot) longer than indicated to allow for trimming. Hoist the antenna into position without the end wire stubs connected. Adjust the length of the inner wire sections to obtain resonance around 3800 kHz. Then connect the end stubs to the center conductor at the open-circuited end of the coaxial traps. Trim their length to optimize the match at the low end of the band.

It may be necessary to go back and make minor adjustments to the lengths of the wires at both ends of the coaxial traps to center the SWR curve as desired. Remember to keep the lengths of all sections the same on both sides of the antenna. It is quite possible that your dimensions may come out quite different from those shown, due to effects from the surroundings.

comments please

I hope that some of the ideas presented here have been thought provoking. I would be interested in hearing of any productive modifications to the designs presented in this article. Comments on both successes and failures in working with these antennas are also welcome.

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ham radio

matching complex antenna loads

to coaxial transmission lines

A simple method for determining the correct length and stub position when stub-matching antenna systems

Matching problems in vhf coax systems can be a stumbling block for the Amateur interested in advanced or large-scale antenna arrays. Commercial antennas match closely to a %-ohm system and provide little challenge. When several antennas are stacked, however, the matching or feed system becomes more complex. If the antenna is homebrew, even one antenna can cause matching problems by providing other than a 50-ohm nonreactive load.

There are two simple methods of matching a 50-ohm system to a load of another impedance: the quarter-wavelength transformer section and the matching stub. Each method has its good and bad points; each is perfect for some applications and much less than perfect for others.

The quarter-wavelength transformer section is well described; it is simply a quarter-wavelength section of coaxial cable used between the load and the rest of the feedline. The Z_o, or characteristic impedance, of this section is an interim value between the load and the feedline impedances. It is this impedance value and the quarter wavelength that are critical to proper transformer function. Since the formula for transformer section Z_o requires the desired impedance at both ends of the section, these values must be known.

The stubbing method involves a physically short, open-circuited section of coaxial cable connected into the feed line with a T connector (see **fig. 1**). The

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positioning piece places the stub at a point on the line where the resistive component of the impedance is equal to Z_o and the reactive component is inductive. The stub itself is simply a capacitor which is adjusted to cancel the inductance. This makes the stub attachment point appear to be resistive and equal to Z_o; in effect, it becomes a perfect load. To use a stub assembly you must know the following:

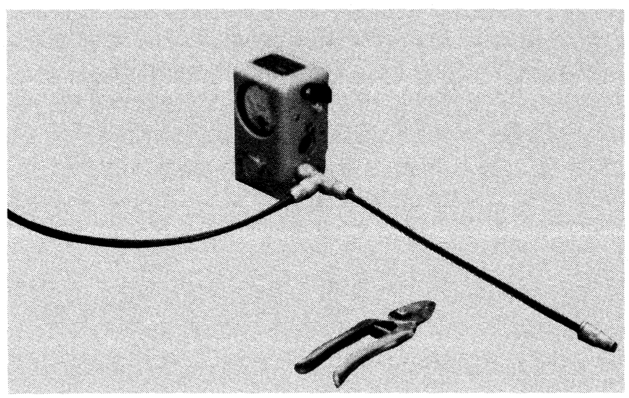
1. VSWR of the unstubbed system (measured at the antenna)

2. Wavelength in the line

3. Actual position of the standing wave voltage minimum point nearest the load

which system?

The strong points of the systems dictate their best uses. When the load impedance is unknown (and probably complex), the stub is the easiest choice. When paralleling antennas of known impedance, the quarter-wavelength section is perfect for the impedance conversion, since the impedance values for both the antenna and transmission line are known. If multiple homebrew antennas are used, of course, each antenna is stubbed to 50-ohms and paralleled with quarter-wavelength sections.



Basic stubbing assembly. The load attaches to the connector at the right. The shears are used to trim the stub for the lowest reflected power.

In general terms, however, unknown complex impedances can easily be reduced to a purely resistive load which is equal to the feedline Z_o with a stub, while transformations from one known impedance to another are more easily done with transformer sections because the construction is easier.

stub system

When using a stub matching system, I use a capacitive stub because the end of the stub is left electri-

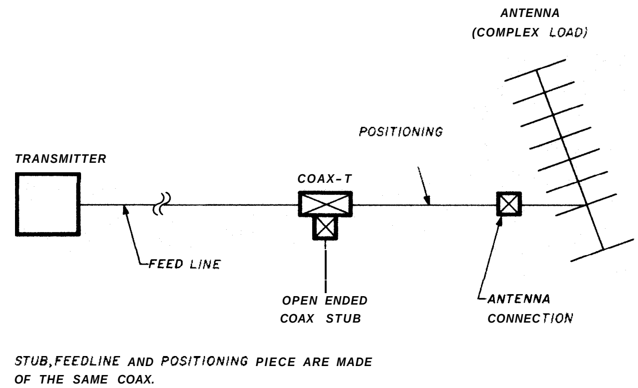


fig. 1. Diagram of a stub-matching system to match the complex load present by the antenna and the fixed output impedance of the transmitter. In this example, the feedline, positioning piece, and stub are made from the same coax.

cally open. This allows easy adjustment of length and permits the whole feed system to be checked occasionally with a leakage meter to check dielectric condition.

The stub is installed in the inductive region of the feedline at a point within 90 degrees toward the generator from the voltage minimum nearest the load (see **fig. 2**). The distance from the load to the first voltage minimum point is part A of the positioning piece length and should be found by direct measurement. A slotted line may be used if its velocity constant is identical to that of the cable in use. Details of its use may be found in **reference 1.**

If an appropriate slotted line can't be found, a simple fluted line may be constructed from a 60-cm (2-foot) length of coaxial cable. The cable is marked every 2.5 cm (1 inch). A small (1-cm, or 318-inch)

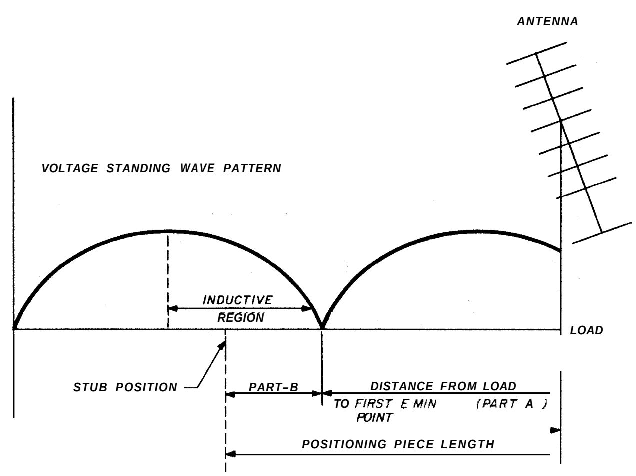


fig. 2. The stub is located 90 degrees toward the generator from the first voltage minimum. This places the stub within the inductive region, allowing an open stub to be used.

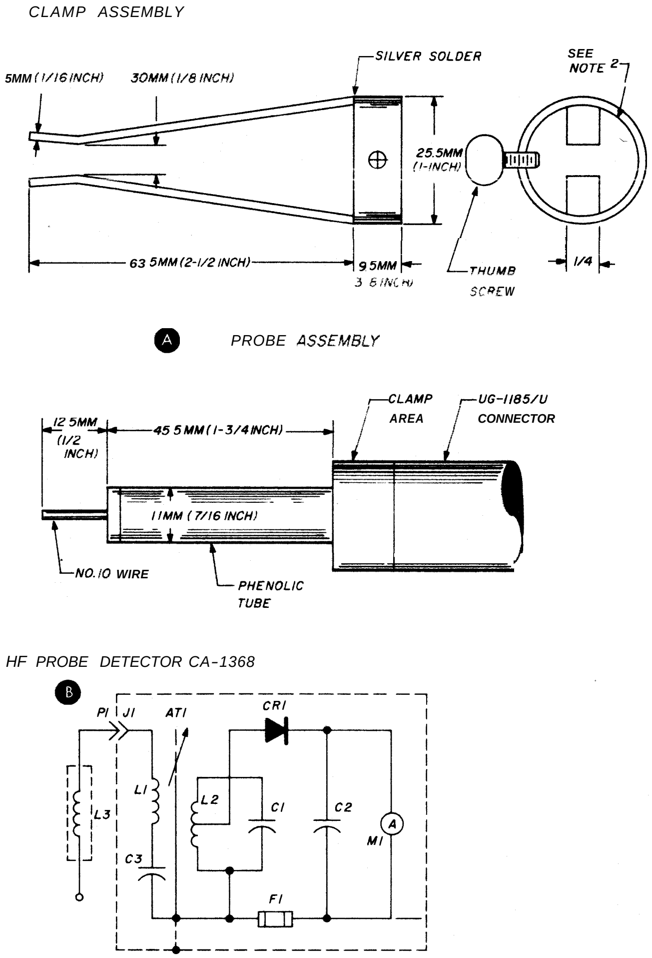


fig. 3. Detailed diagram of a detector assembly which can be used with a homemade fluted line (A). The clamp assembly slides over the probe assembly and is secured with a thumb screw in the clamp area. The schematic diagram shown in (B) is suitable for use with the probe. This detector has the capability, by use of the attenuator, of determining the VSWR on the line. If other means of measuring VSWR are available, the attenuator can be deleted. The tuned-circuit can be changed to measure the desired frequency, acting as a simple rf voltmeter.

area around each mark is stripped of outer insulation and the braid is quickly tinned with a large soldering gun. Using a drill press and V block, drill a small hole through the tinned shield and inner insulation to expose the inner conductor as shown in the photograph. A suitable detector probe is shown in **fig.** 3.

This arrangement allows rf voltage measurements along the line every 2.5 cm (1 inch). By using interpolation the voltage minimum can be located quite accurately between holes.

When part **A** of the positioning piece length has been determined, part B is calculated with

$$\theta(electrical\ degrees) = Tan^{-1} \frac{1}{\sqrt{VSWR}} \qquad \textbf{(1)}$$

To convert θ in electrical degrees into part B in cm, first measure the distance between the voltage minimum points on your fluted line. Multiply this figure by two to find the line wavelength. Then

$$part\ B\ (cm) = \frac{\theta \times line\ wavelength\ (in\ cm)}{360} \qquad \textbf{(2)}$$

If you're working with inch dimensions, simply substitute those dimensions at the appropriate places in the above formula.

Line wavelength can also be calculated from

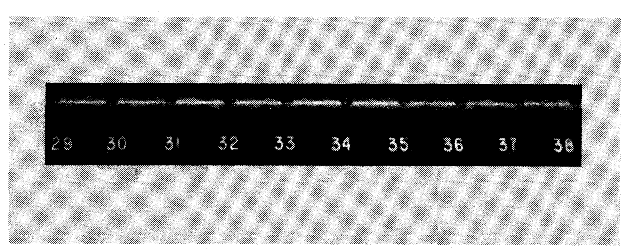
$$Line\ wavelength = \frac{29980\, v_p}{f_{MHz}}\ (cm) = \frac{11803\, v_p}{f_{MHz}}\ (inches)$$

where

v_p = velocity factor of the line (0.686 is typical for coax with polyethylene dielectric)
f_{MHz} = operating frequency
part B = the distance from the voltage minimum to the stub location
VSWR = the standing wave ratio of the unstubbed system measured at the antenna
θ = part B expressed in electrical degrees
Line wavelength = wavelength measured on the fluted line in cm (or inches) or computed.

The total length of the positioning piece is part **A** (the measured part) plus part B (the computed part). A section of feedline cut to that length (including connectors) is connected as shown in **fig. 4** with a variable capacitor substituted for the stub.

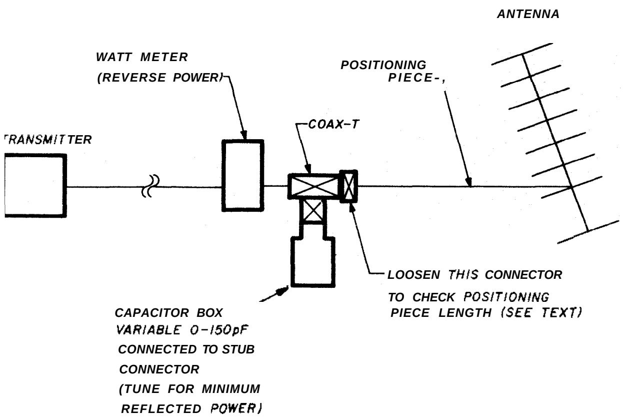
As the capacitor is adjusted, the reverse power indicated on a VSWR meter will dip to a minimum. If the positioning piece is the correct length, the minimum will be zero. If you don't obtain a zero reading, loosen the connector on the stub end of the positioning piece as shown in **fig. 4** to slightly lengthen the piece (don't take it off). If the reflected power dimin-



A commercial fluted line section.

ishes as the connector is loosened, the positioning piece is too short and should be lengthened in 1-cm (318-inch) increments until a minimum is obtained (representing a VSWR of 1.1 or less). If loosening the connector increases the reflected power, the piece is too long and should be shortened in 1-cm (318-inch) increments.

When the positioning piece is the correct length, the minimum reflected power should be noted and an open-circuited coaxial stub with a connector on only one end should be substituted for the capacitor



ig. 4. After the exact length of the positioning piece has een determined, a capacitor box can be substituted for the oaxial stub. Varying the capacitor should reduce the reverse power to a very small value. If not, the length of the ositioning piece will have to be adjusted until the reverse ower is negligible. Once the value of the capacitor has een determined, the equations in the text can be used to ind the physical length of the coax.

ox. The stub length in electrical degrees is deternined with the following formula

$$\phi = tan^{-1}\frac{VSWR - 1}{\sqrt{VSWR}} \qquad (3)$$

vhere

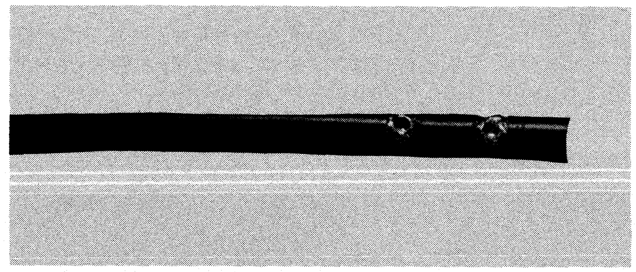
ϕ = electrical length of the stub in degrees

his is converted to stub length in cm by this ormula:

$$stub\ length = \frac{\phi(line\ wavelength,\ cm)}{360}$$

The stub should be cut at least 2.5 cm (1 inch) too ong. After it is in place use garden pruning shears to ut off first 1 cm sections, then smaller ones as the eflected power approaches the previously noted ninimum; try not to cut off one section too many!

When the stub is properly installed and pruned, our antenna will show a virtually perfect 50-ohm oad to the feed system and the VSWR on the feedne will be nearly 1:1.



Sample section of a homemade fluted line.

quarter-wavelength matching system

When using a quarter-wavelength transformer section there are only two computations to make. First the quarter-wavelength length is found either by direct measurement with a fluted or slotted line section or by the following formula

$$transformer\ section\ length =$$

$$\frac{29980\ v_p}{4f_{MHz}}\ (cm) = \frac{11803\ v_p}{4f_{MHz}}\ (inches)$$

Next, the characteristic impedance of the transformer section is determined from

$$Zo = \sqrt{Zs \bullet Zr}$$

where

Zo = impedance of the matching section
Zs = impedance at one end of section
Zr = impedance at the other end of the section

A section of the appropriate cable is then chosen, cut to the computed length, equipped with connectors, and installed in the system (see **fig. 5**).

The real problem with this method is finding the appropriate impedance cable, of course. Referring to the example, how far would you have to look to find commercial 61-ohm cable? The problem can be alleviated if the system is designed so the required characteristic impedance is near commercially available

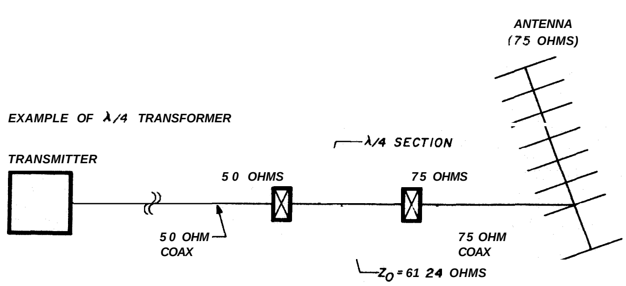
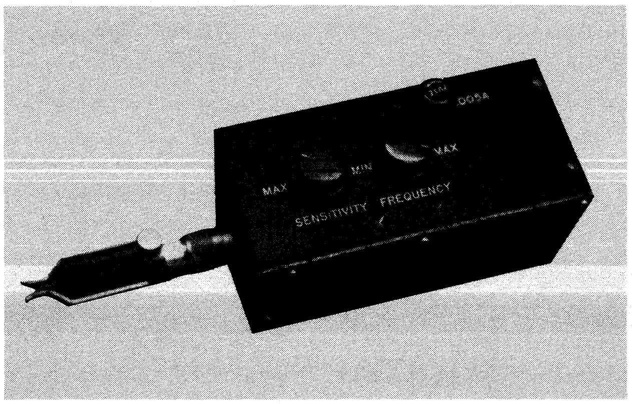


fig. 5. Example of a quarter-wavelength matching transformer. To match between 50 and 75 ohms requires a coaxial section with a characteristic impedance of 61.24 ohms.



Complete fluted line detector showing the probe and detector.

values, but this isn't always possible. You can also construct the cable yourself in rigid form, or forget it.

quarter-wavelength sections as power dividers

If careful attention is paid to the values of impedance available for conversion at different points in a proposed antenna system, some natural combinations can be found. One example is the parallel combination of 50-ohm antennas with 75-ohm matching sections as shown in **fig. 6.**

In this example 50-ohm loads are connected to 71-ohm matching sections which transform the impedance to 100 ohms. The feedline then sees two 100-ohm loads in parallel, or a total load impedance of 50 ohms. In addition, since the input points to the 100-ohm section are equal, the current (and power) will split evenly between the antennas.

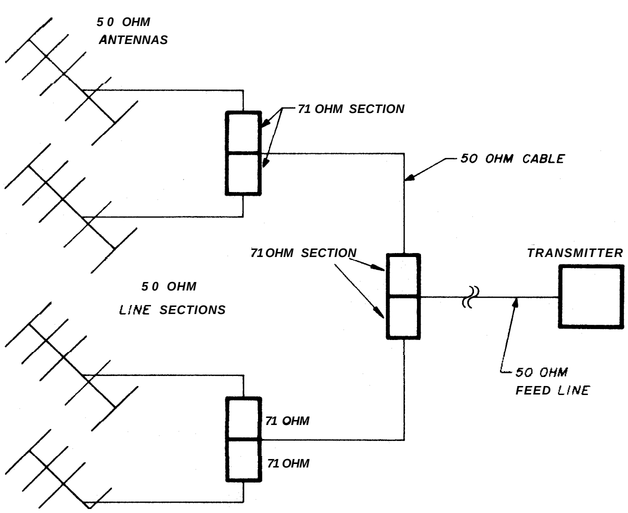


fig. 6. As seen in this diagram, four antennas can be matched to the transmitter by using quarter-wavelength sections of 70-ohm coax. Each antenna's impedance is transformed to 100 ohms and combined with another antenna to get back to 50 ohms.

Further, loads with the same impedance can be fed with different power levels by using different cable values in the sections (although this brings back the probiem of noncommercial impedance values). In this case the desired power levels are converted to currents. The input impedances to the transformer sections are then selected to obtain the desired currents as shown in **fig. 7.**

phasing

Another factor to remember when operating multiple antennas is phasing, or timing the arrival of rf energy at each antenna. Generally, if the antennas

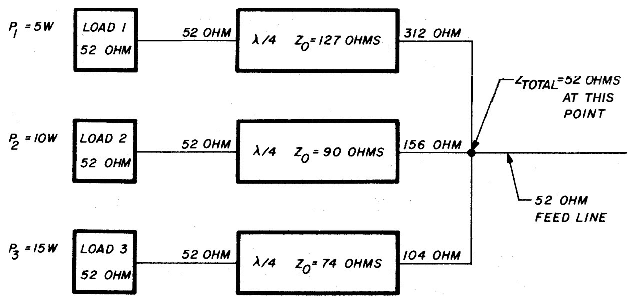


fig. 7. By selecting the impedance of the matching sections, an unequal power division can be obtained. This procedure is useful when trying to create a particular radiation pattern from a group of antennas.

are simply stacked together, you want them to operate in phase; all feedlines from the antennas to the matching transformers must be equal length.

If the antennas are not to be in phase, the appropriate delays can be easily obtained by adding additional cable in the indicated antenna feedlines. Added cable length increases the phase delay; the amount of delay can be computed from

$$\text{degrees of delay per cm of cable} = \frac{360}{\text{line wavelength, cm}}$$

Stubbing and matching sections are not really the bugaboo that they may seem. A good stubbing job can be done easily in an hour; a matching transformer takes less time. The frequency changes over the two-meter band, for example, have negligible effect. These matching tricks can be a big help in making efficient, custom-tailored Amateur arrays.

reference

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ham radio

multiband antenna system

A different approach to triband-beam design results in performance comparable to that of large, single-band, multielement Yagis

Until the early 1950s, a person interested in serious DX had to resort to stacked Yagis, fittingly called a Christmas tree, when faced with the limitation of a single rotor and mast. For most of us, however, it is difficult enough to build and tune a single Yagi beam, much less three stacked beams. During the record sunspot peak of 1956-58, several Amateurs tried to solve the multiband antenna problem, resulting in a number of new beams, especially mini types.

Most people are familiar with the W3DZZ trap antenna either as a dipole, ground plane, or Yagi tribander. The disadvantages and difficulties of this antenna, compared with a single-band, full-size Yagi, are that at 14 and 21 MHz the element is less than full size, causing reduced gain and bandwidth (both SWR and F/B ratio bandwidth). It is a major problem to seal the traps (tuned circuits) so that moisture and the polluted atmosphere do not cause corrosion at element, coil, and capacitor contacts, especially if dissimilar metals are used. In addition, a compromise between trap Q and bandwidth had to be chosen. For a triband element, four traps are required per element, with the contact resistance in the traps causing losses. This form of the triband Yagi is now the most widely used Amateur DX antenna, and it is manufactured in several countries.

G4ZU tribanding

Substantial initial interest, except in the U.S., earned G4ZU British patent number 790,576 (12-2-1958) for his multibanding method. Here are some claims and rebuttals for his version:

1. A beam element that was self-resonated near 21 MHz was made to resonate near 14 MHz by inserting a loading coil or twin boom hairpin loop in the middle of the element.

Fact. The coil actually used had only about half the inductance a coil would need to act as claimed.

2. An "automatic-switching stub" in the form of a piece of open-ended twin lead or coaxial cable was connected in parallel with the loading coil. The stub was to act as an electrical short when the antenna was used near 21 MHz. This was because the stub alone resonated at this frequency, electrically eliminating the tuning effect of the loading coil at 21 MHz.

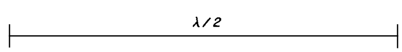
Fact. An open quarter-wavelength stub acts, under matched conditions, like a near short, but G4ZU had a different case and insisted that the stub cable had to have a very special "velocity factor" (*i.e.*, capacitance per unit of length). It appeared to me that the cable capacitance, in conjunction with the parallel inductor, performed the two-band tuning — and not the stub, as claimed.

3. The 28-MHz tuning was not explained by G4ZU. In private correspondence, the inventor stated that the coil-to-mounting-channel capacitance did the trick together with a part of the stub.

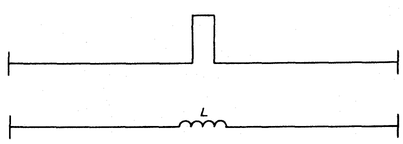
Fact. By placing the stub cable inside the element or double-boom tubing, the resultant coupling of distributed L and C caused the 28-MHz resonance to occur, along with others.

The experiments which demonstrate these facts can easily be repeated. Start with a dipole resonating at

$$f_{MHz} = \frac{300}{\lambda(meters)} \quad or \quad \frac{492}{\lambda(feet)}$$



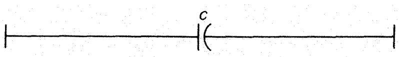
A dipole can be tuned to a lower frequency by inserting a loading inductance, L.



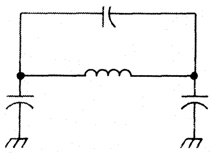
This article first appeared in Amateur Radio, the journal of The Wireless Institute of Australia, in April, 1978.

By Hans F. Ruckert, VK2AOU, 25 Berrille Road, Beverly Hills, 2209 Australia

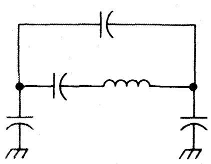
Or, to a higher frequency with a capacitance, C.



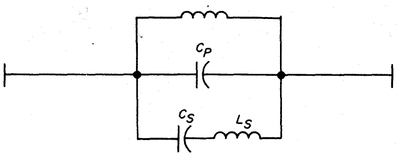
By looking at the distributed L and C components, the antenna can be reduced to a parallel-tuned circuit,



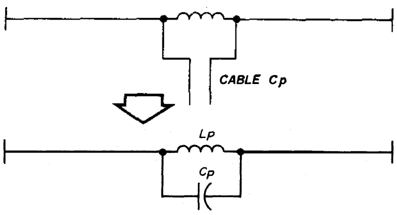
with the dipole itself acting as a series-tuned circuit,



resulting in the combination looking like two resonant circuits.



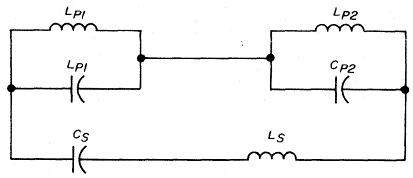
The paralleled parallel-($L_p + C_p$) and series-tuned circuits ($L_s + L_s$) form the "multiband tank" used in transmitters to cover 3.5 to 30 MHz without coil switching. There are always two resonances occurring at the same time, 3.5 to 8 MHz (L_p and C_p) and 7 to 30 MHz (L_s and C_s), depending on the values of C_p and C_s



By replacing the series-tuned circuit (L_s and C_s) with the dipole half-elements, and the cable capacitance by a lumped capacitor of the same value, you obtain (in either case) a two-band element (for example 14 and 21 MHz, 21 and 28 MHz, or even 70 and 180 MHz). The cable (stub) resonance and the velocity factor of the cable used are of no consequence — only the cable capacitance matters. L_p may be a coil, a hairpin loop, or a double boom with shortening bar. Bringing the cable (C_p) near the element creates more resonances, due to the distributed L.

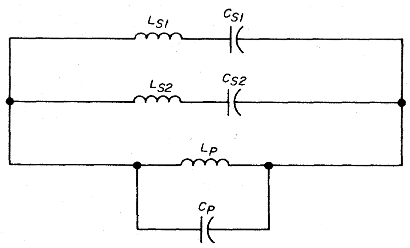
This form was too difficult to tune, and unwanted resonances occurred as well.

After understanding what made the G4ZU beam work on three bands, I looked for a three-frequency circuit that was tunable (and controllable) and could be converted into a triband antenna element. To obtain resonance on three different frequencies (14.15, 21.25, and 28.6 MHz) at the same time without switching inductors or changing capacitors, you need three inductors and three capacitors arranged as follows:



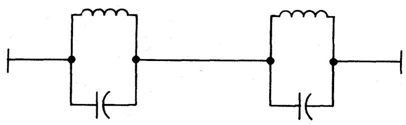
A

OR

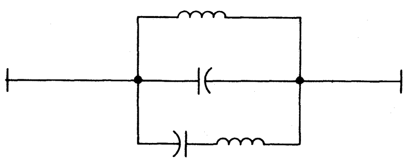


B

These two circuit versions fulfill this requirement. By adjusting the three L and C values, the three simultaneous resonances can be moved over a wide range.

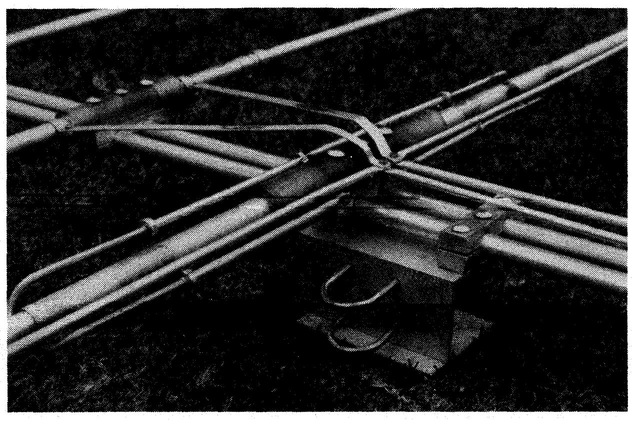


A



B

As previously described, one series-tuned circuit can again be replaced by a dipole to obtain a triband element. In **A** above, you then have two differently tuned, parallel-tuned circuits in series; in B, a series and a differently tuned parallel-circuit in parallel in the middle of the element. This "triband element" may be any Yagi-type radiator, director, or reflector, the groundplane radiator, or a cubical quad element.



Close-up view of the center element showing the hairpin loops and interconnecting straps. Note that a double boom is used in this antenna.

The dipole may have any length, from one quarter wavelength to a full wavelength. The resonant circuits are not tuned to the antenna operating frequencies, and should not be confused with dipole traps (W3DZZ type).

Since 1960, the **A** version has been built in Yagi, groundplane, and quad form by Amateurs in several countries. I've described these antennas in VK, ZL, DL and W-land Amateur literature. Other Amateurs (JA, ZS, DM, OK) have also described their experiences with this system. Unfortunately, antenna manufacturers showed no interest. This is a true triband-antenna element, where the full element is used on all three frequency bands, unlike the unused dipole ends of the W3DZZ system.

DJ2UT version

DJ2UT was particularly successful in using this tri-band-tuning system, and he asked the writer for permission to produce this antenna and to call it the VK2AOU beam (see **fig. 1**). He had continued the antenna's development, later extending the 14-MHz elements to full size to be competitive with other full-size Yagis. On 21 MHz, the element is 0.75 wavelength and on 28 MHz a full wavelength long (collinear), resulting in excellent gain and bandwidth on the 15- and 10-meter bands. The front-to-back ratio, and so the reflector gain and bandwidth, were improved by feeding the reflector via a crossed-phasing line. This resulted in more concentrated radiation in the vertical plane. Mechanically, the reflector to driven element and driven eiement to director spacing is only 2 meters (6 feet), producing a short beam. For strength, a twin boom (25 x 3 mm [1 x 1/8 inch] Al-Mg-Si corrosion-resisting tubing) is used. All clamps are cast from an aluminum alloy. Only stainless-steel screws, bolts, and nuts are used, to avoid electrolysis and corrosion at contacts between dissimilar metals.

feed system

Feeding with a single coaxial cable presented a number of problems because of the large impedance and phase changes, especially at 21 MHz, A T-match connected to both the driven element and reflector finally gave the desired, and easy to control, results. The 28-MHz match is improved by selecting a suitable L/C ratio for the tuned circuits. By placing proximity, or matching elements, for 21 MHz and 28 MHz in front of and near the driven element, the impedance at the T-match is also suitable for 21- and 28-MHz operation. At 21 MHz, the resonant frequencies of the 21-MHz match element and the radiator are above and below 21.25 MHz respectively, much like a band filter. At 28 MHz, the match element also acts like an additional director.

construction

The centers of the long elements and the 21-MHz match element have a polycarbonate center that seals and holds the coaxial cable capacitors (about 75

director	**8.6 meters**
28-MHz match element	**4.9 meters**
21-MHz match element	**6.7 meters**
radiator	**10 meters**
reflector	**10.6 meters**
radiator T-match	**each side 1 meter long**
reflector T-match	**each side 1.4 meters long**
radiator to 21-MHz match element spacing	**0.4 meters**
28-MHz match element to 21-MHz match element spacing	**0.4 meters**
director to radiator to reflector spacing	**2 meters each**

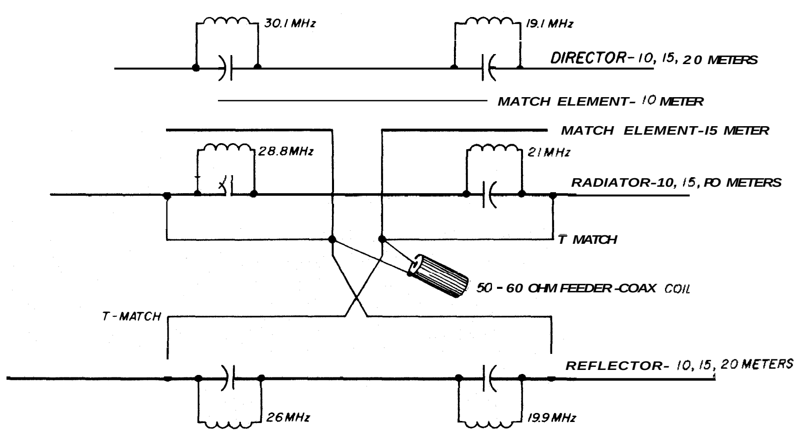
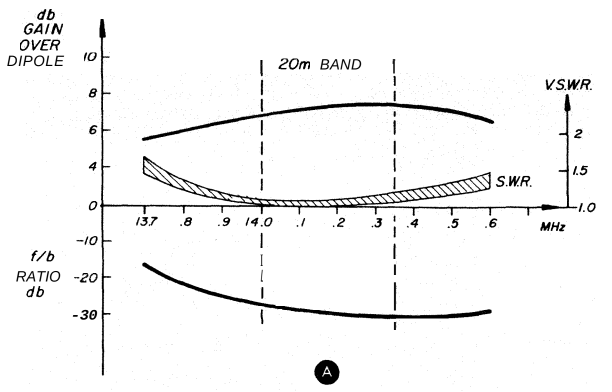


fig. 1. Physical layout of the VK2AOU multiband beam as manufactured by DJ2UT. The capacitors are formed by short lengths of coax inserted in the center of the element, while the inductances are small hairpin loops.



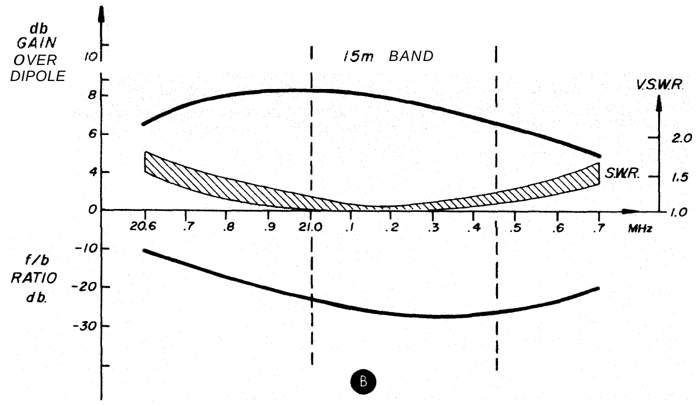
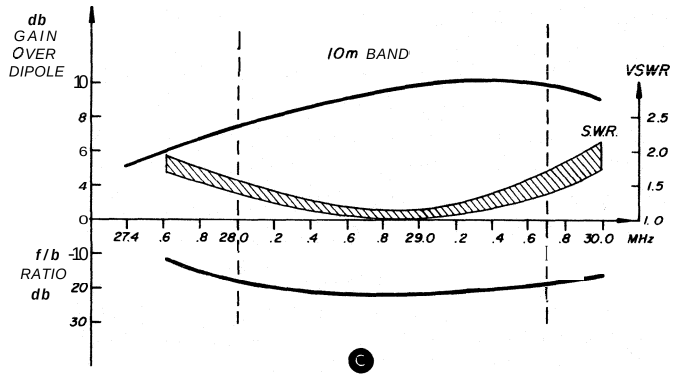
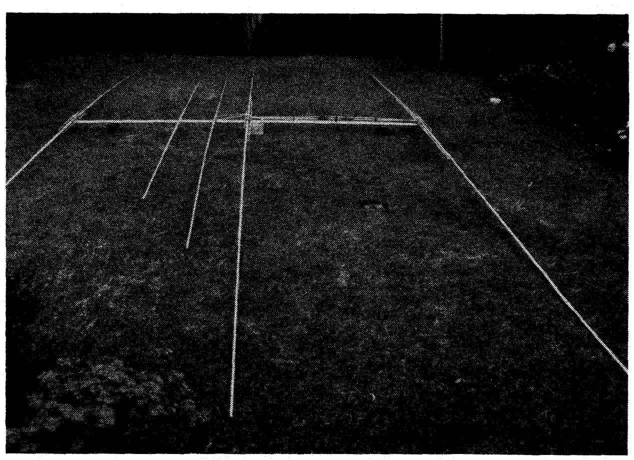


fig. 2. Performance characteristics of the VK2AOU beam. The test dipole was located at the same height, though ten wavelengths distant.



and 100 pF), the stubs for the hairpin loops, and the pieces of 30 x 2 mm element tubing. Two-part clamps and three bolts hold each of the five elements to the boom, and also the boom to the mast-mounting bracket. An insulated wire and clamp for the mast extension are used to support the boom and avoid sagging. The weight with the original tubing amounts to 23 kg (50 pounds). The turning radius is 5.8 meters (19 feet) and the antenna area is 0.65 square meters (7 square feet).

The antenna can handle a continuous rf power of 2.5 kW. The tuning elements are not at high rf voltage points as in trap beams. Galvanized copper solder lugs are used to attach the RG8U coax. The use of the popular 1:1 ferrite balun is not recommended for obtaining symmetrical feeding of the beam halves because it was discovered that the same degree of coupling was achieved without this core. DJ2UT recommends using 3.5 meters (11.5 feet) of coaxial cable in the form of a closely wound, six-turn cylindrical coil near the beam feed point to achieve a balanced feed.



The complete triband beam prior to installation.

performance

When compared with a W3DZZ-type triband beam on a 7-8 meter boom, this antenna exhibits a superior forward gain and reflector bandwidth (see **fig. 2**). On 20 meters, the performance is better than that of a two-element quad or three-element, full-size Yagi. On 15 meters, due to the extended elements, performance is similar to that of a four-element, full-size Yagi. On 10 meters the performance is due to the collinear (double length) elements and the 10-meter match element, and is comparable to the performance of a five- or six-element Yagi.

The VSWR curves are shown as a band because nearby objects cause a change depending upon proximity.

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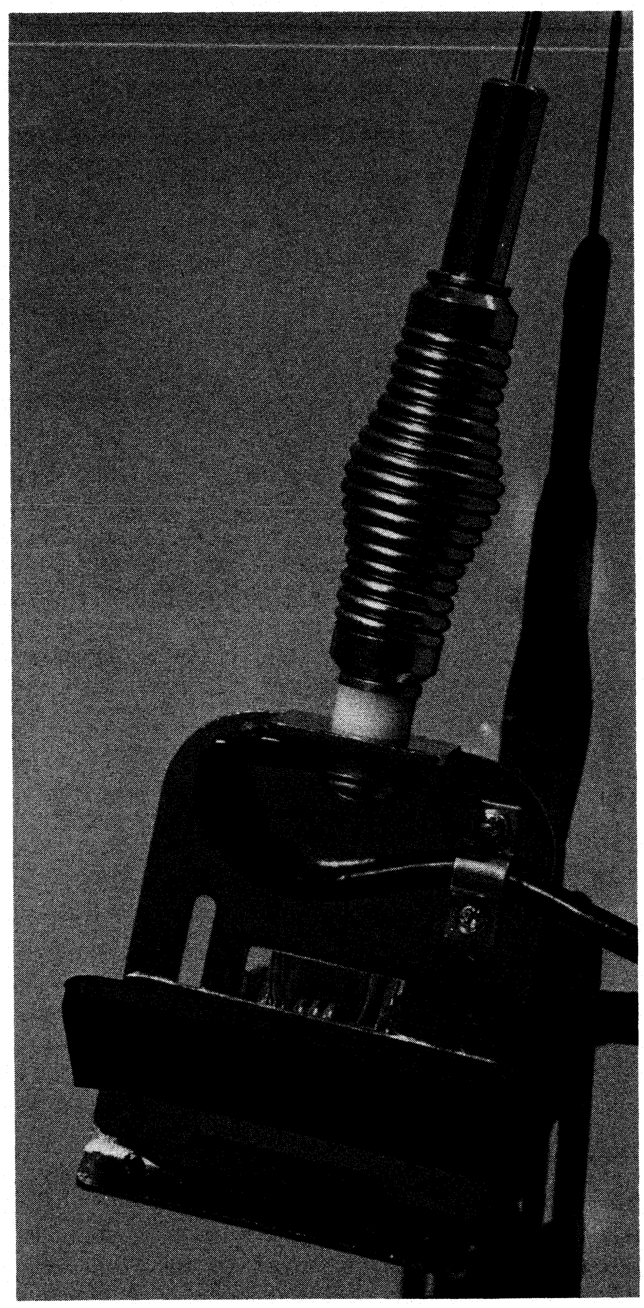
ham radio

homebrew 2-meter mobile antenna



There's no **doubt that** the Citizen-Band boom is here. There are millions of CBers and millions more CB radios and accessories. Hams, always ready to make something out of nothing, are discovering many ways to use CB components for Amateur-Radio use. This article is about another use: a 2-meter, %-wavelength, mobile gutter-clip antenna made from CB antenna parts. Although I designed it for 2 meters, the same techniques could be used for any vhf or uhf band by making appropriate changes in antenna length for the desired frequency.

Starting at the base and working up, the parts used are these: a gutter-clip antenna mount (originally intended for use with a center-loaded CB antenna); a spring with an M6 (¼-20) threaded female base and a top that takes a whip (this is a part from a base-loaded CB antenna); and, finally, the whip itself, which is stainless steel. The whip can be a CB or Amateur Radio part.

construction

Here's how it all goes together. The gutter-clip mount that I found had RG-58/U coax and a PL-259 connector attached. If the clamp you find doesn't include the cable you'll have to make one. Obtain a 3-meter (10-foot) length of RG-58/U coax and solder a PL-259 connector (or the connector to fit your radio) to one end.

The other end of the coax goes to the antenna. Solder the coax center conductor to a solder lug large enough for a 6.4-mm (0.25-inch) bolt. Even if your clamp has coax with it, you may have to replace the solder lug with one that will fit. If you solder on a new lug, use heat-shrink tubing over the connection, part of the lug, and about 25 mm (1 inch) of the coax. This will help protect the connection and act as a cable strain relief.

Connect the coax outer conductor to the metal clamp. This is a little tricky. Make the ground connection at the U-shaped holddown clamp. To do this and still have shielding from there to the antenna base, carefully cut away a 3-mm (0.125-inch) piece of the outer insulation all the way around at the point where the coax will be under the holddown clamp. Be careful not to cut the braid. Then wrap a 25-mm (1-inch) length of bare wire around the coax braid. This wrap will protrude past the outer insulation to provide good contact with the clamp and bracket when tightened down.

You're now ready for the spring. The spring may have a 6.4-mm (0.25-inch) stud screwed into the bot-

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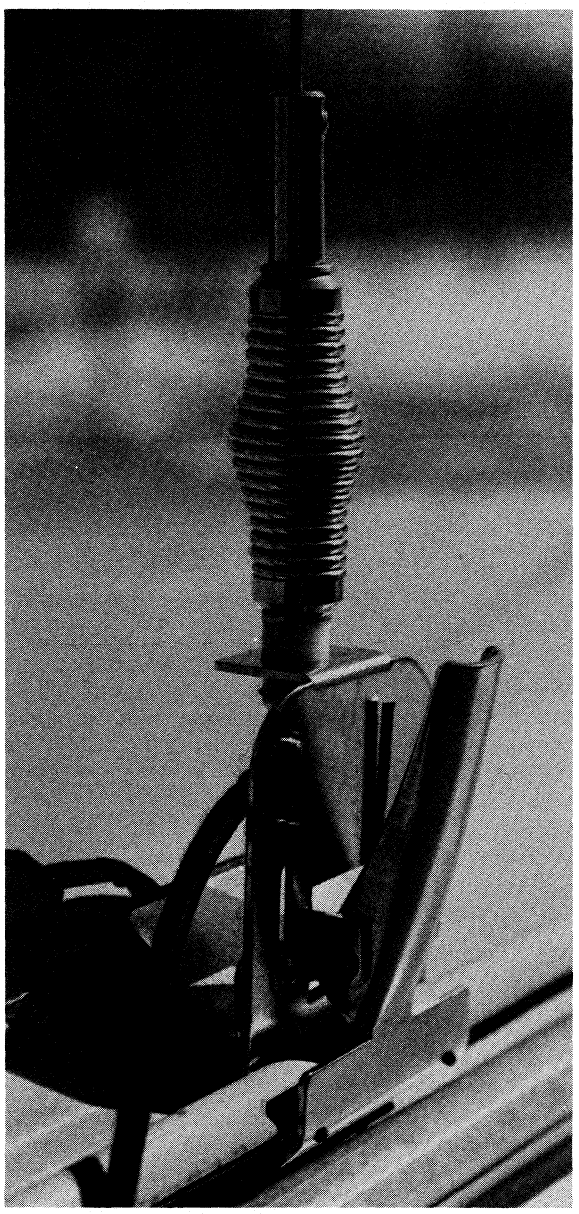
tom; if so, remove the stud. To attach the spring to the clamp, use an M6 (5/4-20) by 25-mm (1-inch) round head bolt through the solder lug, an insulating washer, the clamp, an insulating washer, a lock washer, and the bottom of the spring in that order, bottom to top. Note that some of the gutter clamps have insulating washers smaller than 6.4 mm (0.25 inch); if so, drill them out before assembling. Tighten the bolt.

radiating element

You're now ready for the antenna whip. It can be almost any stiff wire, but stainless steel designed for



Construction of the 2-meter, quarter-wavelength, mobile whip antenna. Parts were scrounged from CB supply stores and the author's junk box. The antenna is easy to unclip and remove when you leave your car unattended.

this purpose looks good and won't rust. You may be able to find a broken piece large enough to work, or it can be purchased at any electronics store that sells CB or ham equipment. A CB whip for a bottom-loaded antenna can be used and should be long enough to make two 2-meter antenna whips.

The whip goes into the top of the spring and is held in place with a setscrew (which may take a hex key wrench for adjustment). Position the whip in the opening as far as it will go, then back it out about 12.7 mm (0.5 inch). In this way you can loosen the setscrew and make minor changes in whip length without having to cut the whip. Antenna length is measured from the top of the whip to the bottom of the lockwasher under the spring. How do you find the correct length? Use the following formula:

$$\textit{length in mm} = \frac{74970}{\textit{frequency in MHz}}$$

$$\textit{length in inches} = \frac{2952}{\textit{frequency in MHz}}$$

For example, the middle of the 2-meter repeater band is 147 MHz:

$$\textit{length in mm} = \frac{74970}{147} = 510$$

$$\textit{length in inches} = \frac{2952}{147} = 20$$

operation

After the antenna is cut to the correct length you're ready to try it out. Clamp it on to the gutter mount and feed the coax through a window. A vhf swr bridge can be used to adjust the whip length for the lowest reflected power. This shouldn't be necessary, however, if you calculated and measured accurately.

In using the antenna, I found it to be very effective and convenient. Its performance is more than adequate for repeater use, and it's easy to unclip and slip through the car window when the car is parked, which is an excellent security feature. The antenna is small and weighs much less than the CB antenna for which the gutter clamp was intended, so you should have no problems with the clamp coming loose.

My total cost came to around $4 for parts, which were purchased from a "junk box" at a local Olson electronics store. Antenna parts from damaged CB antennas or surplus parts should be available free or inexpensively from a CBer friend or a CB store. Even if the parts are purchased brand new, the cost is still considerably less than for most commercially made units.

ham radio

sloping antenna array
for 80-meter DX

An array using wire elements that has given a good account of itself during the pileups — it's the W2LU sloper

Anyone who has put in an evening trying to work DX on 80 meters will understand the frustration that results in a desire to develop a better antenna system for this band. With 80-meter DX you soon learn that the criteria by which a "better" antenna system is judged would weigh directivity nearly equal to gain: *i.e.,* "You can't work 'em if you can't hear 'em." An increasingly popular way of obtaining a reasonable amount of both has been with the sloping dipole system. The sloper, as must be expected, has both positive and negative aspects.

the sloper antenna

Probably the sloper's biggest drawback is that, to get good azimuth coverage, it's necessary to have at least two radiators directed 180 degrees apart, with three radiators at 120-degree intervals being preferable. This arrangement requires a good-sized piece of real estate, typically a 61 by 53 meter (200 by 175 foot) lot for a full three-radiator system with reflectors.

The second drawback is that a big tower is highly desirable. Although functional slopers have been erected with tower heights of only 15 meters (50 feet), heights of 24 meters (80 feet) or more are preferable, with 31-37 meters (100-120 feet) being ideal.

On the positive side we have a number of points. Perhaps first is the sloper's excellent operating characteristics. The complete system as outlined in this article demonstrates a front-to-back ratio of 20-25 dB on signal paths on the order of 3200 km (2000 miles) or more. The gain of the system is estimated to be between 2 and 5 dB over a dipole, depending on the direction relative to the radiator.

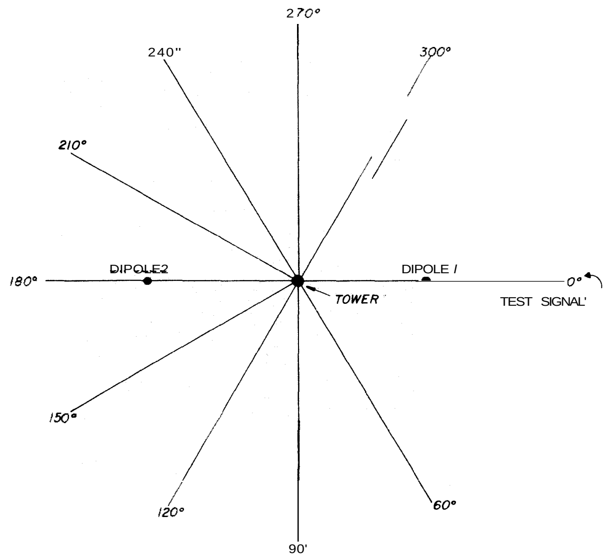


fig. 1. Plan view of the experimental sloper antenna. Two sloping dipoles were used. One was pointed directly at a vertically polarized test signal; the second was pointed 180 degrees away.

The other major advantage of the system, once the cost of the real estate and tower have been discounted, is its basic simplicity and its cost and effort effectiveness. The radiators are ordinary dipoles, and the reflectors are either plain wire or inductively loaded wires. Furthermore, the radiators, being dipoles, present approximately balanced feedpoints and don't require an extensive ground system for efficient operation. A system of radials tied in with a grounded tower would probably reduce ground reflection losses.

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experimental system

While developing this system, I performed experiments to determine the ideal angular spacing between the radiators and radial reflectors. The configuration used for these experiments consisted of two slopers; one was pointed directly at a vertically polarized test signal source approximately 0.8 km (0.5 mile away, and a second was pointed 180 degrees away. Two reflector wires were then placed radially, one to each side, so that the included angles between the radiators and reflectors could be varied in a symmetrical fashion as shown in **fig. 1.**

The tests were repeated several times and the results appeared to be consistent. **Table 1** indicates representative values.

Several conclusions were drawn from these tests. First, by comparing signals from the two dipoles, I found that an included angle of 60 degrees between the reflectors and the further dipole represented the best front-to-back ratio – 20 dB. An angle much smaller than this resulted in degredation. An angle of 60 degrees between the reflectors and the near dipole resulted in a forward gain of about 5 dB over a dipole. Therefore a system of three dipoles and three reflectors optimized both front-to-back ratio and forward gain. A four-dipole, four-reflector system would probably perform satisfactorily; however it's doubtful whether the added complexity would produce meaningful improvement in performance. On the other hand, dropping back to a two-dipole, two-reflector system would definitely result in degraded performance.

table 1. Test results from experiments to determine ideal angular spacing between radiators and reflectors in the experimental sloper antenna system.

reflector position (degrees)	dipole 1 (dB)	dipole2 (dB)	f/b (dB)	gain (dB)
no reflectors	38	28	10	0
60-300	43	32	11	5
90-270	41	28	13	3
120-240	41	21	20	3
150-210	40	27	13	2

A system was then installed using a three-dipole, three-reflector configuration with 60-degree spacing as shown in **fig.** 2. All reflectors were cut to 43 meters (140 feet) and all dipoles were made resonant at 3.8 MHz. Although direct 180-degree, front-to-back checks were no longer possible, a large number of on-the-air checks confirmed the expected results.

As a further refinement to the system, I decided to try adding additional parallel reflectors under each dipole as shown in **fig.** 3. In this installation the dipole and radial reflectors are supported by a 30.5-meter (100-foot) tower. The added reflectors were hung from the 21-meter (70-foot) point on the tower, resulting in approximately 0.1-wavelength spacing from the dipole. Each reflector was 26 meters (85 feet) long and resonated at 3.5 MHz, with about 30 turns of B&W 3905-1 inductance. Obviously, higher towers would allow one to take greater advantage of this concept by making possible greater spacing and less inductive loading. Although I made no direct comparative measurements with and without these reflectors, it seemed that front-to-back ratio improved by about 5 dB, and the theory seemed valid.

Tuning of the parallel reflectors was accomplished by grid dipping and noting the amount of rf energy in the element. With power fed to the radiator, I could

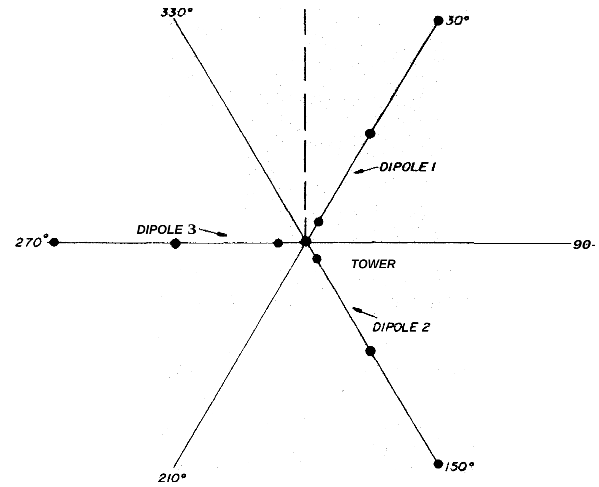

fig. 2. Sloper system using three dipoles and three reflectors with 60-degree azimuth spacing.

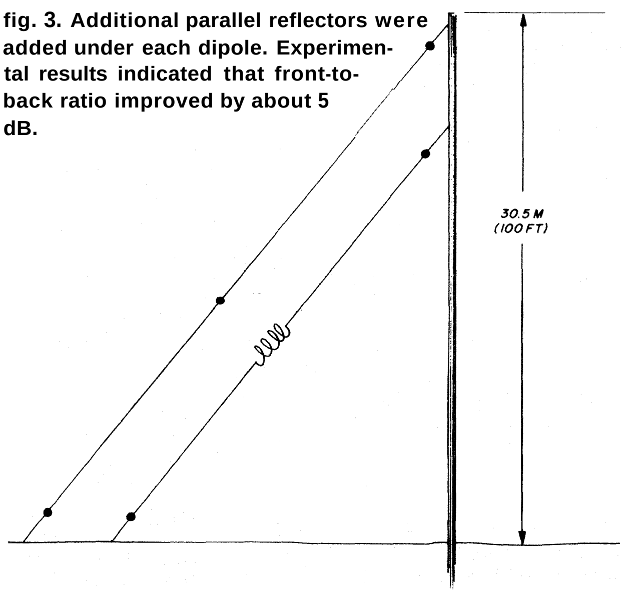

fig. 3. Additional parallel reflectors were added under each dipole. Experimental results indicated that front-to-back ratio improved by about 5 dB.

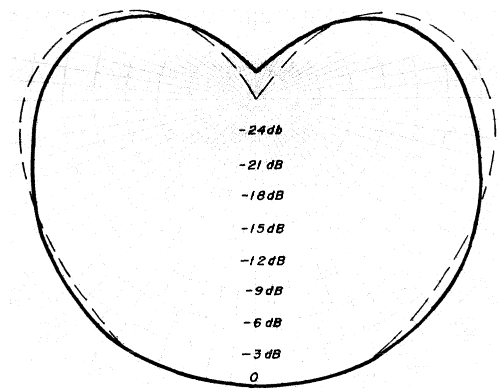

fig. 4. Beam pattern of the sloper using three radiators, three radially oriented reflectors, and three parallel reflectors. The data indicates response in the horizontal plane; that is, directivity, not gain.

draw an arc with a pencil from the low end of the reflector." I assumed that this arcing was greatest when the element was self-resonant. The element was then made to resonate about 10 per cent lower in frequency, or at 3.5 MHz.

Thus the system was complete — using three radiators, three radial reflectors, and three parallel reflectors. With the aid of a large number of comparative signal-strength readings — taken in the receive mode (using a Drake R-4A receiver) to eliminate the S-meter variable — a beam pattern was developed, as shown in **fig. 4.** Also shown in **fig. 4** is the ideal pattern for a two-element vertical array using quarter wavelength spacing. It can be seen that the sloper pattern compares quite favorably in horizontal directivity. Remember that this data indicates a comparison only in the horizontal plane, not vertical, and that it indicates directivity, not gain — gain being a function of directivity and efficiency.

The system has been used for several years and has provided excellent, consistent, and reliable performance. It's been a great asset in digging DX out of the QRM and has always given an excellent account of itself in the pileups.

ham radio

"Not a very good test from an engineering standpoint, but one used by hams for years. Use caution and make certain you are insulated from ground. **Editor**

antenna-performance measurements using celestial sources

A simple procedure for measuring the receiving figure of merit of large vhf and uhf arrays using celestial bodies as noise sources

High-performance vhf and uhf arrays for moonbounce, meteor, and tropo scatter are proliferating in the Amateur Radio ranks. This article describes a simple procedure for measuring the receiving figure of merit of large vhf and uhf arrays using radio stars, the moon, or the sun as a source. I developed the technique from the references and recently used it to measure the performance of the 25.6-meter (84-foot) moonbounce dish at K3NSS.

The method derives the ratio of the antenna gain, G, to the receiving system noise temperature, T, or G/T. G/T is a measure of system receiving sensitivity and applies directly in link calculations and serves as a standard for performance comparison. Knowing G/T and either the antenna gain (G) or the system noise temperature (T), the unknown parameter can be derived.

measurement procedure

G/T is measured from celestial sources by taking the ratio of the received signal* noise power to the noise power received from the cold sky. This ratio, called the Y factor, is compared with the known radiation flux density, S, of the source **(table 1)** to determine the receiving system figure of merit, G/T **(eq. 1)**. Units of G/T are decibels per degree Kelvin, dB/°K.

The advantages of this technique are simplicity

*The signal of these celestial sources has noise characteristics

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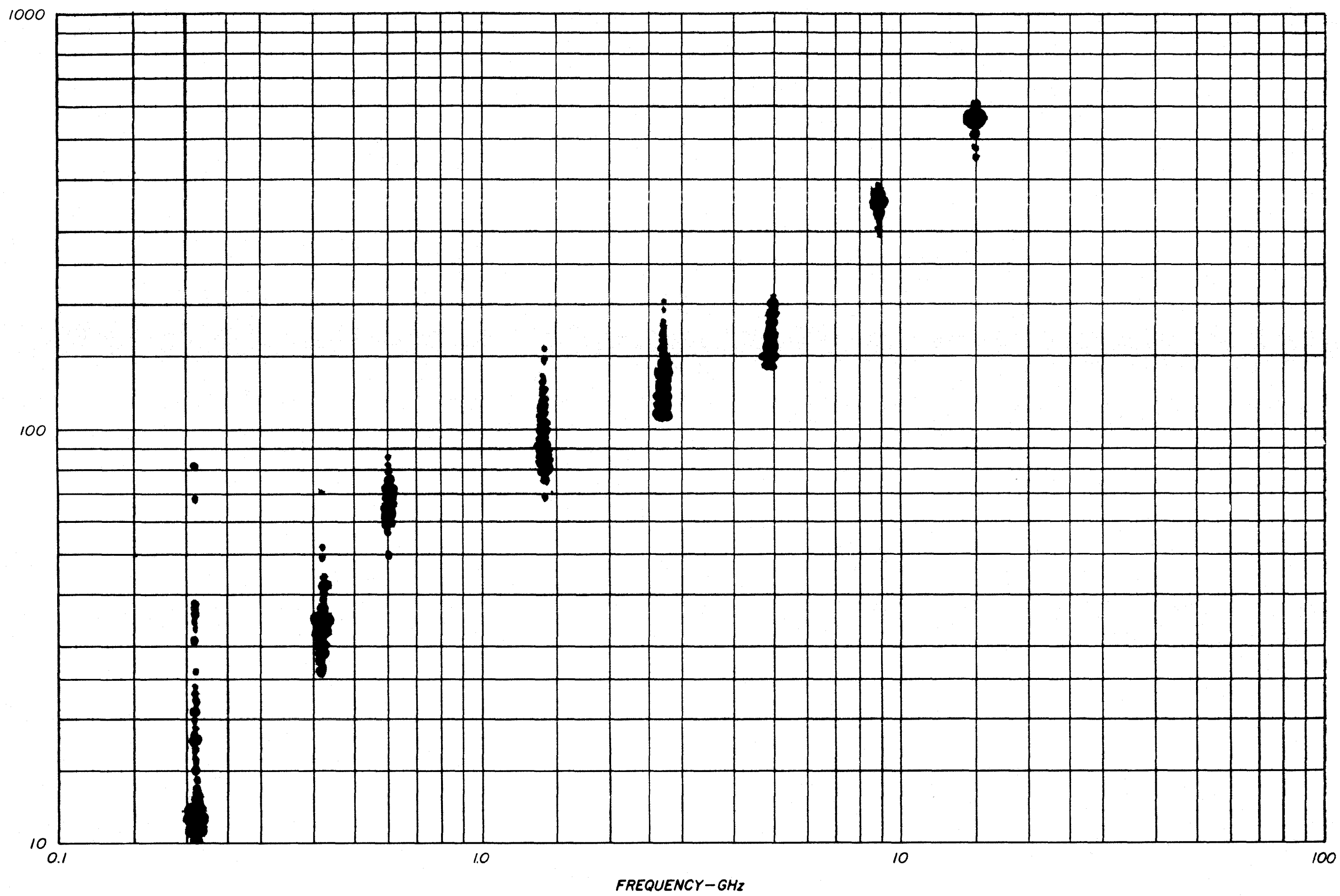


fig. 1. Solar flux density vs. frequency for July through December, 1978.

and accuracy. The technique is elegant because it requires no knowledge of either antenna gain, G, or system noise temperature, T. The critical measurement, Y, is calculated as a ratio so that measurement units and biases cancel. Finally, no special equipment beyond that for normal station operation is necessary. Only a suitable measure of receiver output power is required. This could be from the receiver S-meter, although better accuracy will result if a simple diode noise detector is used, as described in reference 1.

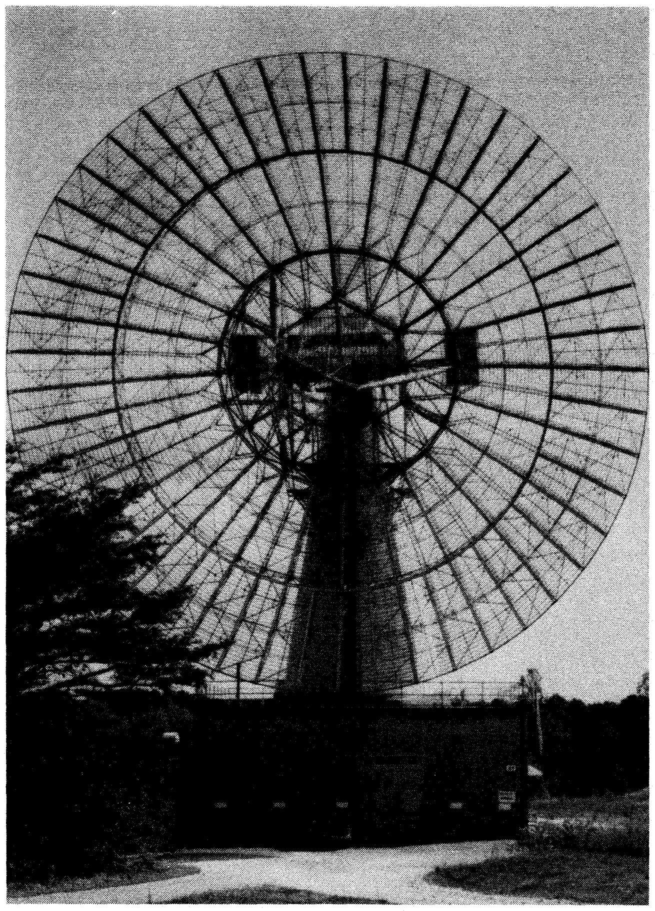
The qualifying minimum performance necessary to use this procedure is that the system must be able to detect sun noise, which is the strongest source. Measurement accuracy depends on the following factors:

1. Source used. The flux density, S, of the sun varies randomly by a factor of two or more. Therefore measurements from the sun may not be accurate within 3 dB unless corrected for actual solar flux at the time of observation (see **table 1** and **fig. 1**). If G/T is large enough to permit radio star measurements, accuracies to within 0.5 dB are attainable under favorable reception conditions.

2. Atmospheric/ionospheric attenuation. Source attenuation varies inversely with frequency and elevation angle. Observations at 144 MHz near the horizon may not be accurate within 3 dB. Observations of Cassiopeia A or Cygnus A at 1296 MHz, at elevation angles greater than 45 degrees, can be accurate to ±0.5 dB. The references contain additional information for those interested in corrections to obtain the greatest possible accuracy.

3. Number of measurements. Averaging a large number of observations, or applying linear regression analysis, will result in improved accuracy.

4. System noise floor. The more sensitive the system, the bigger will be the difference between signal noise power and cold sky noise power; ***i.e.,*** a larger Y factor. Reference 2 has an excellent chart of sky noise. It illustrates the relatively high noise characteristic of the galactic equator and the relatively low sky noise at the galactic poles. The more sensitive the system, the more attention must be paid to cold sky



The **K3NSS** 25.6-meter (84-foot) parabolic reflector in the stow position. A small group from the Southern Maryland Amateur Radio Club, including the author, restored the antenna to 432-MHz moonbounce. All rf systems, including the **feedlines** and feed antenna, were missing. The 432-MHz system was built and installed from individual and club assets. The restoration began in the fall of 1976, and the first contact was completed with LX1DB in April, **1977.**

temperature. For example, to take readings from the moon, which has a very weak S, care would have to be taken to ensure that the cold sky reading was from a low noise area. This isn't as important for sun measurements. The objective is to get large Y factor measurements.

measurements

The algebraic relationship of G/T to the Y factor and source intensity, $I(s)$, is given by:

$$G/T = \frac{Y-1}{I(s)} \qquad \textbf{(1)}$$

where:

$$Y = \frac{\text{signal noise power (source)}}{\text{cold sky noise power}}$$

$$I(s) = \frac{\lambda^2 S}{8\pi K} \qquad \textbf{(2)}$$

λ = wavelength (meters)

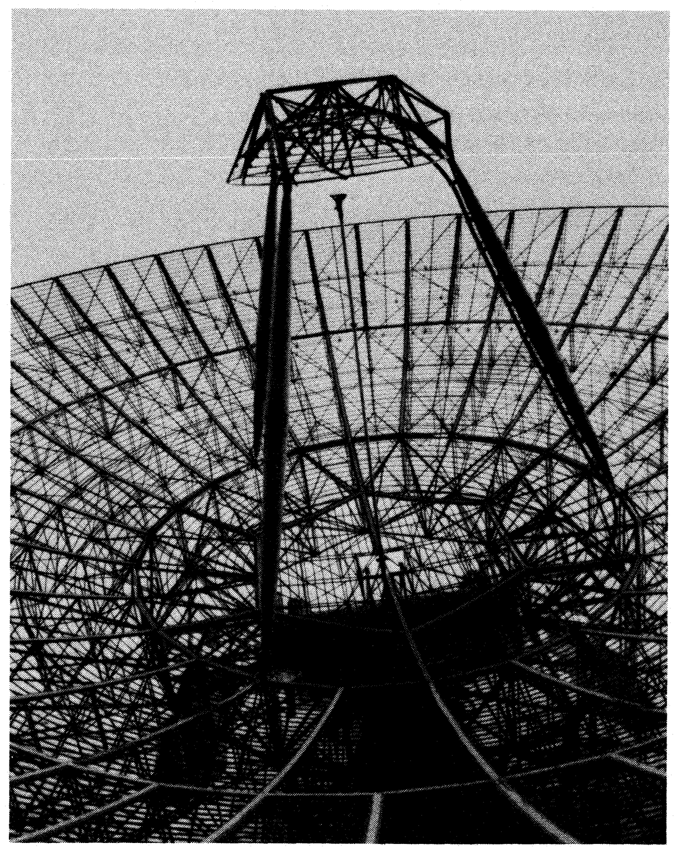
S = source flux density from **table 1** (watts per square meter per Hertz)

K = Boltzmann's constant
= 1.38×10^{-23} joules per degree Kelvin

Eq. 2 translates the source flux density, S, to units of equivalent temperature, T. It includes a factor (1/2) to account for the unavoidable loss resulting from the interception of a randomly polarized signal by an antenna sensitive to a single polarization. The factor $\lambda^2/4\pi$ accounts for the antenna gain relationship to wavelength (area factor).

The Y factor used here differs from that described by K2LMG and W2YBP in reference 1, but the principle is the same. WAØIQN, reference 3, describes a sun-noise measurement procedure using an estimate of antenna gain by use of horizontal and vertical

A view showing the antenna feed supporting the radiating element (dipole) covered by a polyethylene bag at the focus. The Gregorian reflector, supported by the 5.5-meter (**18**-foot) fiberglass tripod, was not usable in the restoration. Instead, the dipole feed was mounted on a feed tower, clamped at the base to a rotator for polarization control. The coaxial feed tower was made from 7.6-cm (3-inch) diameter aluminum irrigation pipe with a 3.5-cm (**1-3/8-inch**) diameter copper water pipe as a center conductor. The focal length of the dish is 7.6 meters (25 feet).



beamwidths, θh and θv. The procedure described here will yield better accuracy.

Eq. 1 is an algebraic ratio. The G/T value is more useful in logarithmic (decibel) units. Therefore the following conversion applies to the G/T ratio from **eq. 1**:

$$G/T\,(dB/^{\circ}K) = 10\,\log_{10}[G/T \text{ ratio from } \textbf{eq. 1}] \quad \textbf{(3)}$$

Remember that algebraic ratios usually result from, or are used in, equations with multiplication and/or division terms. Equations employing or yielding logarithmic values usually have addition and/or subtraction terms. **Eq. 1** written for logarithms is:

$$G/T\,(dB/^{\circ}K) = 10\,\log_{10}(Y-1) - 10\,\log_{10}[I(s)] \quad \textbf{(4)}$$

Typical G/T values for Amateur arrays range between -10 and $+15$ dB/°K.

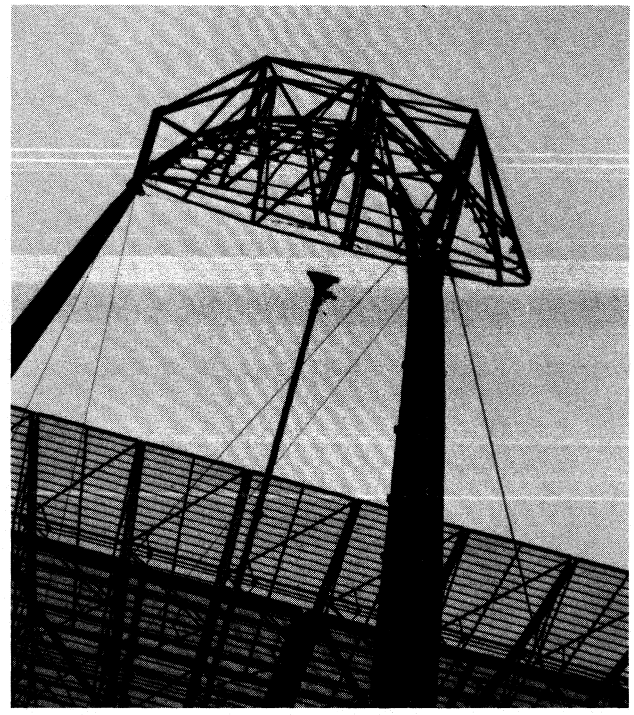
Those interested in the derivation and further explanation concerning radio astronomical measurements of antenna performance should consult references 4 through 8. Wait (reference 4) presents a clear derivation of **eq. 1,** and the remaining references contain extensive information dealing with every aspect of radio astronomy measurements.

measurement techniques

The ease and accuracy of measurements depend greatly on your ingenuity. The following observations are submitted as food for thought:

1. Receiver linearity. The receiver chain must be operated *linearly* over the dynamic range of the measurements. As a starter, the receiver agc must be defeated.

2. Receiver bandwidth. The i-f should be operated at the widest possible bandwidth, since noise-power measurements are involved.



Close-up of the feed dipole. The dipole is covered with a polyethylene bag to keep rain water out of the assembly. The feed tower is guyed by use of a slip-ring collar, visible as the bulge half-way down the feed-tower. Since this photo was taken, a reflecting element has been added to the feed dipole, giving a cardioid feed pattern.

3. Measuring the Y factor. An accurate measurement technique is to note the cold sky power, then acquire the source; then, by use of a precision variable attenuator, reduce the source signal to equal the previous cold-sky reading. The required attenuation is the Y factor. **Caution:** remember to convert decibel readings to algebraic ratios and *vice versa* as applicable.

table 1. Source flux density, S, as a function of frequency for selected celestial bodies. The data were derived from references 5 through 7. Values are in watts per square meter per $Hz \times 10^{23}$.

frequency (MHz)	sun (X 10^{-23}) note 1	moon (X 10^{-23}) note 2	Cassiopeia A [redacted] 10^{-23}) note 3	Cygnus *A* (X 10^{-23})	Taurus A (X 10^{-23})
144 note 4	52	0.01	15.0	10.0	1.80
432	280	0.08	5.55	4.00	1.25
1296	525	0.80	2.05	1.40	0.92

Notes, 1. The value for sun S varies randomly by a factor of two or more. The value listed is typical for a quiet sun.

2. Calculated by the author from the relationship: $S = (2KTm/\lambda^2)\,\Omega m$ where Ωm = solid angle subtended by the moon viewed from earth. See reference 7.

3. S decreases approximately 1 per cent per year. Value corrected by the author to 1978. See references 5 and 6.

4. 144 MHz values extrapolated by the author. Expect large measurement variations because of atmospheric and ionospheric effects

5. Daily solar radiation measurements are available from the Space Environmental Laboratory, 325 Broadway, Boulder, Colorado 80302. Data is available to remote computer terminals by contacting J. D. Schroede: at (303) 499-1000, extension 3780.

4. Power vs voltage measurements. Be sure to measure the Y factor as a power ratio. Most meters and chart recorders respond to voltage or current, not to power. Squaring a voltage or current ratio yields the corresponding power ratio.

5. Acquiring the source. Locating a source in the sky is a matter of determining its Greenwich hour angle, GHA, and declination for the time of observation. **Table** 2 lists celestial coordinates (right ascension and declination) for stellar sources. Reference (9) gives GHA and declination directly by hour for the sun and moon, as well as procedures for finding stellar GHA from right ascension.

table **2.** Location, advantages, and disadvantages of celestial sources for antenna-performance measurements.

celestial source	position (see note 1) right ascension (hrs.-mins.)	declination (degrees)	advantages	disadvantages
Cassiopeia A	23-22	N 59	Strong stellar source. Observable 24 hrs/day north of 32° n latitude	S decreasing about 1 per cent per year / Poor visibility in Southern Hemisphere
Cygnus A	19-59	N 41	Observable over much of the earth for substantial period each day	Some polarization evident in very precise measurements
Taurus A	5-33	N *22*	Visible for good portion of day over most of the earth	Third strongest stellar source
Sun	see reference 10		Strongest source; Easy to locate	Large, random variations in intensity
Moon	see reference 10		Easy to locate	Very weak source

Note 1. Right ascension and declination are coordinates on the celestial sphere. See references 9 and 10 to convert to antenna pointing coordinates for local time.

Polar antenna mounts are usually calibrated directly in GHA (or the equivalent measure, LHA) and declination, since these are used for the mounting axes. To use azimuth/elevation antenna mounts, conversion formulas, calculator, and computer programs are published in reference (10) and other sources.

6. More information. The Eimac compendium, reference (10), contains a wealth of information for anyone with a serious interest in large vhf/uhf arrays. It's free from Eimac.

measurement refinements

There's much more to be said regarding measurements of high-performance antenna systems using celestial sources. For example, the determination of antenna gain, G, by measuring the equivalent system noise temperature, T, is not yet adequately addressed in Amateur Radio literature. Several rather esoteric correction factors not covered in this article are available. These include corrections for propagation attenuation and source polarization assymetry (stellar sources). Those who strive for absolute precision may be interested in the more detailed discussions on factors affecting measurement accuracy contained in references **4** through 6.

acknowledgments

I'd like to thank Ruth Phillips, K3AGR; Howard Eich, W3HE; Jim Erickson, K3LFO; Dave Phillips, W3JPM; and Willie Mank, W1ZX; all made major contributions to the restoration of the K3NSS 25.6-meter (84-foot) parabolic reflector at Cheltenham, Maryland, to moonbounce operation, including the measure of performance from Cassiopeia A. That effort was the inspiration for this article.

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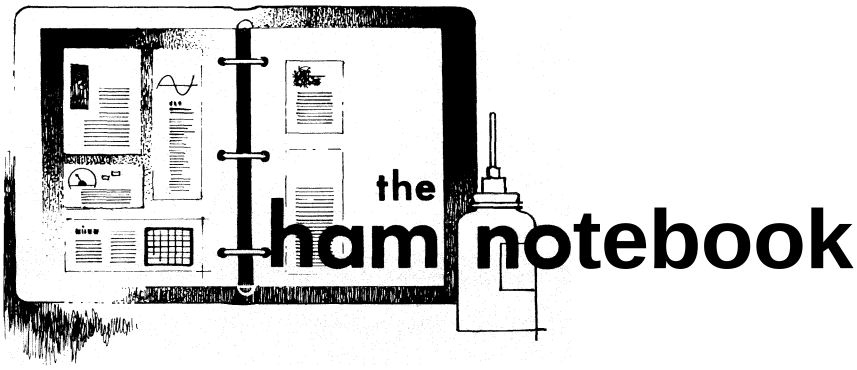
ham radio



self-gating the 82S90/74S196 decade counter

In many typical counter circuits, the clock input to the first decade counter is logically combined with the count gate prior to being applied to the counter. In some cases, the limiting frequency factor is then the external gate and not the first decade counter. Fig. 1 shows a method of using the internal circuitry of the 82S90/74S196 to perform the gating, thus allowing the IC to be used to its full 100-MHz capability without any degradation caused by external gating.

When the Count/Load line and the Rest line, pins 1 and 13 respectively, are high, the counter will operate in a normal clocked fashion. Since the count gate is connected to the Count/Load line, the counter will continue to function for the length of the gate. However, at the end of the counting time, the count gate goes low, shifting the IC into a mode where the counters actually act as latches fed from the Data Input lines. In addition, the counters also ignore any inputs on the clock lines. By tying the load lines to the output lines, the information is held, while it is processed through to the readouts. A negative reset pulse will reset all flip-flops to zero, ready for the next count period.

R. S. Naslund, **W9LL**

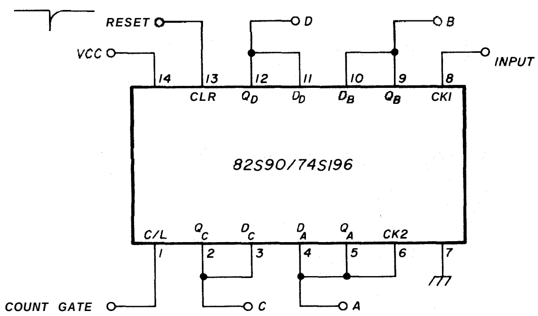


fig. 1. Schematic diagram of a method for using the internal gating in the 82S90/74S196 counter to eliminate any external clock gates. By using the count gate to control the Count/Load line, the IC will ignore any pulses on the clock line when the counters are used as latches.

metalized capacitors

Where the need exists for a very small size, low cost, stable, close tolerance capacitor in the range of 0.001 to 2.2 μF with good high-frequency characteristics, the Siemens MKM and MKH types are a good choice. They can be obtained in either metalized polycarbonate or metalized polyester, with 5 per cent tolerance, rated at 100, 250, or 400 volts. As an example of their compactness, the 0.47 μF, 100 volt capacitor measures a 1 x 1 x 0.5 cm. The feature that attracted me most was their very low parasitic inductance of 20 nH; this is due to the stacked foil construction and the package design which allows very short lead lengths. Several values were checked for series resonance with 3 mm leads; values for two popular ceramic values are included in the list below for comparison:

Ceramic	0.02 μF	12.0 MHz
Siemens	0.047 μF	10.0 MHz
Siemens	0.068 μF	8.0 MHz
Siemens	0.1 μF	6.5 MHz
Ceramic	0.1 μF	6.3 MHz
Siemens	0.47 μF	2.8 MHz
Siemens	0.68 μF	2.6 MHz

Stability of the Siemens capacitors, of course, is not as good as polystyrene or silver mica, but I find them very useful for timing circuits, active filters, and similar applications.

Bill Wildenhein, **W8YFB**

cure for the R4C backlash

Shortly after I received my new R4C, the dial started to develop a backlash problem. At first glance, it appeared that the clearance between the nylon gears was not correct. However, upon closer examination I determined that the backlash was caused by a binding of the outer edge of the plastic concentric dial closest to the front panel. In order to get a better look at the problem, I removed the white metal bracket that has the pilot lamp socket attached to it. This bracket is attached to the front panel by a machine screw on each end.

Each machine screw has a metal spacer between the bracket and panel. While removing the screw, hold the spacer with a long-nose plier or it will fall down into the bottom of the chassis necessitating the removal of the bottom cover in order to retrieve it.

After removal of the bracket, you will notice two felt pads on each end of the plastic index plate attached to the front panel. Remove these two pads completely. These pads were added by Drake to prevent the concentric dial from rubbing on the plastic index plate, which could warp with age.

Drake has since corrected the problem on their later production runs by relocating the index plate from the inside of the panel to the front of the panel. The felt pads have also been removed.

Bernard White, W3CVS

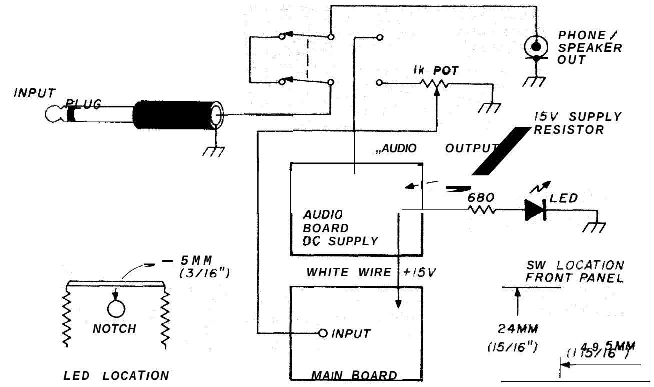


fig. 2. Diagram of the modifications to the QF1 audio filter. The power-indicating LED is ed in a 5-mm hole just above the NOTCH. To allow the operator to switch the QF1 in and out of the circuit, a miniature double-pole, double-throw switch was installed in the indicated position.

autek filter

The Autek OF1 ssb/CW active filter has proven itself to be a very useful device. Like any other "black box" in the typical ham shack, on-the-air use indicates certain modifications would make it far more useful.

Since the OF1 operates from the ac line, a low-drain indicator light would show when the power is on. This is easily accomplished with an LED mounted just above the NOTCH control on the front panel. A 15-Vdc supply is available, with the appropriate series resistor, you have a light, color of your choice.

The second modification allows ahe operator to switch the filter in or out of the circuit. Thus, the headphone/speaker can monitor either the filter or the direct output from the receiver. This is very easily accomplished with a miniature double-pole, double-throw switch installed just to the left of the existing ac switch. An equalizing control was included to prevent overloading the Autek Amplifier. Thus, the phones will hear basically the same level in or out of the filter. Since the pot requires very little adjustment, it was mounted on the rear, just below the three-lug strip. This three-lug strip is the tie point for the filter input, making modification very simple.

Bill Long, K6EVQ
Bob Landgrave. WA6WZQ

improved stability for the 32S transmitter

While operating CW "split" with my Collins 32S-1/75S-3 combination, I noticed a definite frequency shift on the transmitter signal. This occurred only on the first dash or couple of dots when using VOX CW keying, being most noticeable on 10 meters. When PTT was used, the drift was completely unnoticeable.

My transmitter is a modified 32S-1 which has been up-dated to an S-3.[1] A zener diode provides regulation for the B-plus supply to the HFO and PTO. The stock 32S-3 also regulates the voltage to these oscillators, but it is supplied by an OA2. Examination of the circuit revealed that the voltage for the regulators is derived from the +275 volt input via a set of contacts on the VOX relay, K1 (see **fig. 3**). This means that during receive mode, there is no voltage applied to the HFO or PTO, with the transition from zero to full operating voltage (regulated or not) causing the oscillators to shift frequency slightly.

By changing the circuit to maintain voltage on the regulators, and hence the oscillators during receive, this anomaly is avoided. Simply move the lead connecting R17 to the VOX and connect it to C137 on the power amplifier cage wall. This capacitor has +275 volts applied continuously.

After this change is made, VOX keying will be just as good as with PTT, with no shift discernible. However, the effectiveness of this modification to an unmodified 32S-1 has not been determined.

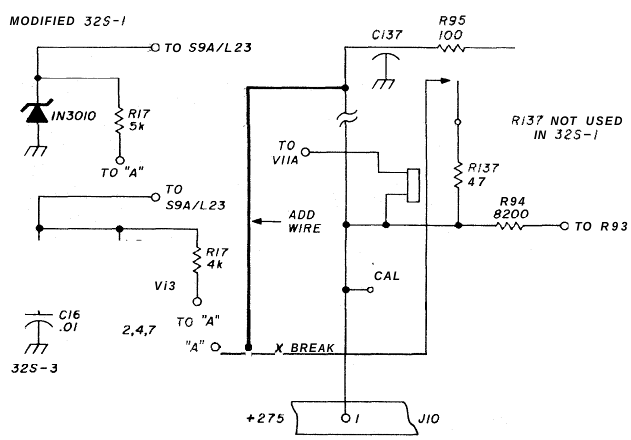


fig. 3. By changing the connection to R17, the oscillators are energized at all times. This eliminates the frequency shift caused by switching their plate voltage.

reference

1 Paul K Pagel, N1FB, "Updating the Collins 325 1," ham *radio,* December, 1978, page 50

Paul Pagel, N1FB

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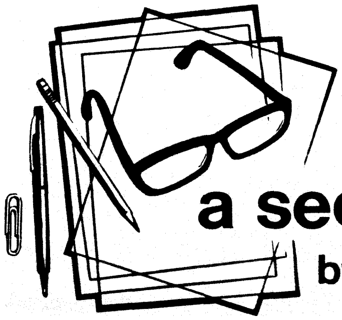


a second look

by Jim Fisk

Beginning with a small handful of amateur wireless operators in the early 1900s, the Amateur Radio population in the United States has grown to the point where it's now approaching 360,000. Last year the growth rate was about 8 per cent, down slightly from 1977, and this year it is expected to be about the same. And while *orderly* growth is healthy for the hobby, in many ways the Amateur Radio Service is like the proverbial "house that Jack built," with rooms added as they are required, with little thought to future construction – or indeed, to the esthetics of the architecture!

If you study the history of Amateur Radio, it's easy to understand why this happened: for years there were more licensed Amateur Radio stations in this country than all the other radio services combined, many top members of the FCC were hams, and the management ranks of most major radio-electronics firms were filled with licensed amateurs – many, in fact, began their careers as Radio Amateurs. With influential friends in high places who had a vested interest in Amateur Radio, most operators gave little thought to the future. The complexion of Amateur Radio has changed over the years, however, and it's obvious that we can no longer afford such a *laissez faire* attitude toward our future.

One matter that concerns many of the older hams is that in the past 25 years the character of Amateur Radio has evolved slowly away from being a technician's hobby, where much of the operating equipment was homebuilt, to an operator's hobby, where little or no technical expertise is required. This is not necessarily a problem because our activities are closely linked not only to a rapidly changing technology, but to a dynamic society that continually confronts Amateur Radio with new obstacles, challenges, and opportunities for providing useful public service. Nevertheless, more thought must be given to the impact of this trend on the long range future of Amateur Radio.

With a steadily increasing number of amateurs and greater government intervention in terms of changed licensing regulations, restrictive antenna covenants, and RFI requirements (not to mention WARC 79 and the proposed revision of the Communications Act), it's increasingly apparent that *all* of us must give some serious thought to where the Amateur Radio Service should be in the coming decade. While long-range planning is hardly an exact science, it is possible to anticipate some of the problems, to perceive certain distant opportunities, and to develop appropriate recommendations. If we put our collective heads together, we should be able to plot a positive future course for Amateur Radio – rather than drifting out of control as we have for the past few years, reacting to external events as they have occurred. Positive results, however, will require a substantial amount of effort on a continuing basis by a large number of concerned amateurs. Complaining about the current state of affairs or railing about the "system" in the press is neither positive nor constructive.

Those of you who have read my editorials for the past eleven years know that I have pointedly avoided the politics of Amateur Radio. Therefore, when I suggest that a possible focus for future planning activities is the ARRL's Long-Range Planning Committee, you know that suggestion is not politically motivated. For those of you who are not members of the ARRL, the Long-Range Planning Committee was established by the ARRL Directors in January for the purpose of "reviewing and making recommendations concerning programs which the ARRL is and should be providing to its members and to the Amateur Radio Service . . ."

At its initial meeting in February the members of the committee, according to one of those present, agreed upon several criteria which would govern the committee's activities:

1. The general welfare of the entire Amateur Radio Service was to be served, not just parts of it.
2. No fact of the ARRL's operation was exempt from scrutiny.
3. A subject as complex and far reaching as the future of Amateur Radio cannot be properly appraised without inputs from many different people — ARRL *members or not.*

If *you* have any comments or recommendations about the future of Amateur Radio, make it a point to let the Long Range Planning Committee (LRPC) have the benefit of your thoughts. A letter or card to Vic Clark, W4KFC (12927 Popes Head Road, Clifton, Virginia 22024), marked for the attention of the LRPC, will be acknowledged, and Vic will make sure that your comments are available to each of the members of the committee.

Jim Fisk, W1HR
editor-in-chief

comments

Longhand printed-circuit layout

Dear HR:

The article, "Printed-Circuit Layout Using the Longhand Method," which appeared in the November, 1978, issue of *ham radio* was especially interesting to me because I have been using the described technique for the past two years. However, the additional steps which I use will produce a better board and are desirable with high-density boards. First, lines on the paper pattern which represent the copper should be made red and the circuit components should be shown black. This makes reading and interpretation easier. Second, when the pattern is transferred to the copper foil surface by marking with a sharp point, the points should not be deep enough to produce burrs (burrs tear the pen point and eventually make it difficult to produce a clean fine line). Third, after the points are on the copper those which are connected should be joined by a pencil line. This is important on a high-density board because it permits inking the lines rapidly and the pen point does not have time to dry. Lines produced with a dry point must be retraced; a smeared line is nearly always the result.

Maximum ground-plane area is a requirement for rf circuits, especially for vhf circuits. Filling in all the blank area with a pen is tedious and it's difficult to make the area completely resistant to the etch solution. Tape can be used but covering small irregular areas is difficult. I find that an easy method is to outline the ground plane area first with the pen, then fill in all the enclosed area with the blue dye used by tool makers and machinists. I have tried many inks and paints for this step, and the blue machinists dye is superior to all others. It is easy to apply with a small brush and dries rapidly. The board may be etched minutes after application. A coat which appears too thin will resist the etchant even at elevated temperature. The ink wets the copper surface and flows easily but stops when it contacts the previously applied line.

The machinists dye is also useful for producing plug patterns on circuit boards. Coat a l-cm (½-inch) strip at the board edge and use a sharp point or jeweler's screwdriver to remove the ink from areas between the contacts. Use an old plug, placed against the board edge, for a pattern. It is easy to produce 22-contact plug patterns with this method.

I have used two types of the blue dye. One is called ***Dykem Steel Blue*** and is a product of Dyken Corporation of St. Louis, Missouri. Another is called ***Mike-O-Blue*** and is sold by Ashburn Industries of Houston, Texas. A four-ounce can is adequate for many boards.

This "longhand" method will produce high-quality circuit boards which have clean lines and ground-plane areas without pit marks commonly found on boards which have been prepared by other methods. Boards with 3-mil copper foil can be etched in less than ten minutes in a 50 per cent etch solution (ferric chloride) heated by placing it in a tray or plastic dish floated in hot water. Use only enough etch solution to cover the board about 1 cm (½ inch) and agitate during etching to provide a washing action. In addition, the copper surface can be seen during the process so the board may be removed when completed.

I have tried many types of ink pens and find that the ***Sharpie*** brand is best; their number 49 has the best point. Store the pen with the point down, this aids in keeping a generous supply of ink in the point and is always ready to use.

During the past years I have spent many hours trying to find an easy method for producing "longhand" etched circuit boards and have concluded that the technique described in *ham radio,* along with the additions indicated above, is the best.

Robert J. Grabowski, W5TKP
Houston, Texas 77005

Dear HR:

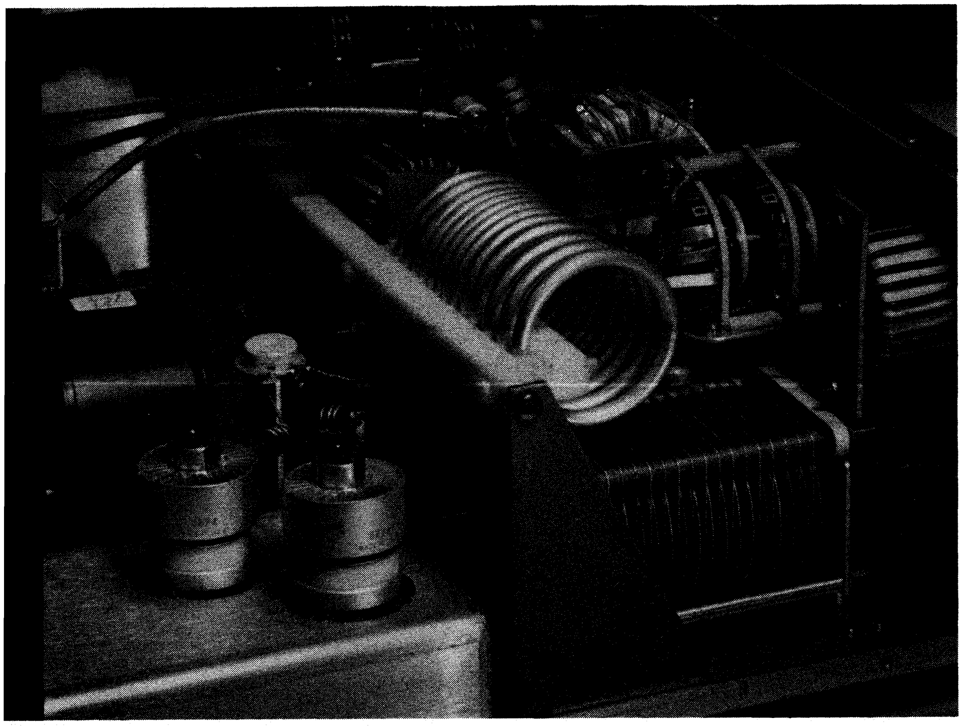
During a literature search for an electronics project I recently went through my file of *ham radio* — and was distracted for three evenings reading the fine articles in three years of issues! Yours is by far the highest quality journal of all those devoted to Amateur Radio; please don't compromise that quality.

Guy Rothwell, KH6JCD
Kailua, Hawaii

Dear HR:

I just breadboarded the CW processor described by Jones in the October, 1978, issue of *ham radio.* It's really sharp! However, the 555 oscillator interacts with the operation of the 576. To cure this problem, I've installed a 100-ohm resistor between the 5-volt line and pin 8 of the 555, and a 220-μF capacitor between pin 8 and ground. It makes a real improvement.

Jeff Davis, VE3CBJ
Grimsby, Ontario



design considerations for linear amplifiers

The first of several articles on practical construction techniques for hf power amplifiers

So you want to build a linear amplifier! So do many other Radio Amateurs. Without doubt, the most popular piece of home-built transmitting equipment (aside from small circuit board projects) is the high frequency linear amplifier. It can be built without having an advanced degree in solid-state technology and computer analysis.

Recent FCC decisions, moreover, have made a linear amplifier homebrew project more inviting to Amateurs, particularly those interested in 10-meter operation. For a period of time, in the early spring of 1978, it was nearly impossible to buy an off-the-shelf, commercial linear amplifier; most manufacturers had stopped production in view of the drastic redesign requirements imposed by the new FCC rules. Home-constructed amplifiers, happily, are exempt from the FCC straitjacket. And, more and more, Amateurs are discovering the fun of building their own amplifiers. It's not as hard as you might think! There's still fun in building and adjusting equipment, and the high-frequency linear amplifier

By William I. Orr, W6SAI, 48 Campbell Lane, Menlo Park, California 94025

described in this series of articles is a good project for the home builder, whether you're an old timer or newly licensed Amateur.

what to build

The builder of a linear amplifier undoubtedly has many questions that must be answerd before he can pick up a soldering iron or drill; what tubes to use, what plate voltage, drive level, harmonic suppression, TVI prevention, cooling, and packaging?

At this stage of the game, even the stoutest of hearts may falter. But cheer up, the overall problem is not complex if the project is approached on a workmanlike basis. The purpose of these articles is to provide a blueprint that will guide the builder through the design, construction, and checkout of a modern linear amplifier capable of operating on all Amateur frequencies between 3.5 MHz and 29.7 MHz. This first article covers design, selection of tubes and components, formulas that make the job easier, and practical construction considerations. A later article will cover the metal-work, assembly, and testing in detail.

Before you jump into the project, bending metal and soldering wires, you should know that there exists a vast amount of literature covering the design and construction of linear amplifiers. It would be foolish to ignore this storehouse of accumulated knowledge. At the end of this article is a list of suggested reading material, and you can learn a lot by observing what has happened in this interesting field of radio design. Since this series of articles cannot possibly cover every detail of building a linear amplifier, you can pick up a lot of very useful extra information if you scan some of the suggested reading material.

preliminary design

The first choice you will have to make concerns the tube (or tubes) to be used, the operating voltages, and the means of cooling the tubes so that their operating temperature will remain within the limits imposed by the manufacturer.

A word of warning is advisable on the subject of surplus or second-hand transmitting tubes. Large power tubes have a finite shelf life. The perfect vacuum has not yet been created, and old tubes (surplus World War II vintage in particular) are not to be trusted — they may have an imperfect vacuum. Surplus tubes marked JAN (which stand for Joint-Army-Navy procurement), such as JAN-813 or JAN-211, provide no warranty to the user, since the tubes are purchased by the military on a special contract with

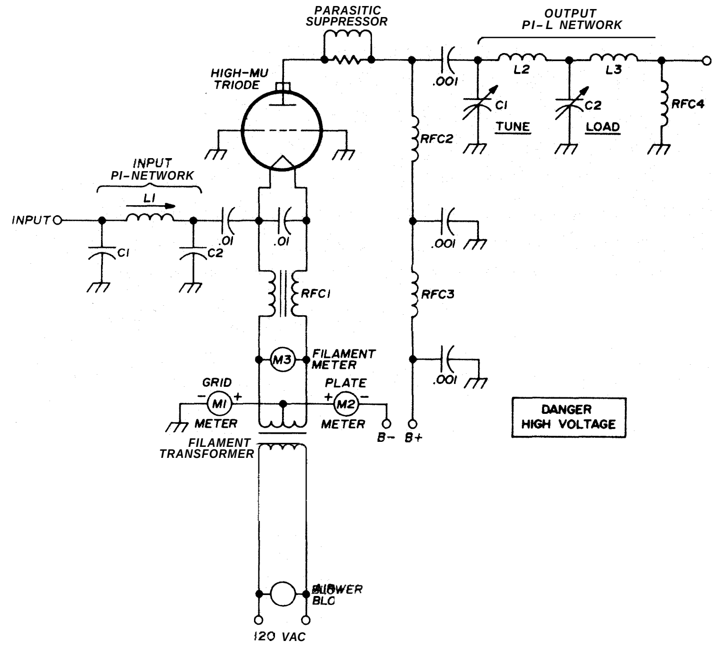


fig. 1. Schematic diagram of a basic grounded-grid amplifier circuit. A high-mu triode tube is used with the exciting signal applied to the filament circuit which has been isolated from the filament transformer and metering circuits by the rf choke, RFC1. A fixed-tuned pi-network circuit matches the output impedance of the exciter to the input impedance of the amplifier. A pi-L plate output circuit is used for maximum harmonic suppression, with a simple parasitic suppressor placed in the plate lead to dampen vhf oscillations. For safety, the metering is placed in the filament return circuit. The grid meter is inserted between the grid (ground) and the filament return, while the plate meter is in the B minus return lead to the power supply. The air blower is connected to the primary of the filament transformer. This circuit may be modified for two parallel-connected tubes by the addition of a second plate parasitic suppressor and increased air blower and filament transformer current capacity. Plate tank circuit components need not be modified if the two tubes run at the same voltage and current as one tube.

source inspection and no warranty return program. Thus, an Amateur who buys a JAN-labelled tube receives no warranty. New tubes purchased from a franchised dealer carry the manufacturer's full warranty. Dealers in surplus tubes, moreover, have no reliable facilities for testing transmitting tubes, which require a large, expensive, and exotic test console. Thus, the purchase of a surplus or second-hand tube may turn out to be penny wise and pound foolish.

To determine the tube type to be used, it is important to note that the most popular ham-type linear amplifiers seen in the various station descriptions and advertisements are capable of running 1-kW input on CW and 2-kW PEP input on ssb. The amplifiers use cathode-driven (grounded-grid) circuitry requiring a drive signal compatible with today's modern exciter (about 80 to 100 watts PEP output). A representative amplifier circuit is shown in fig. 1.

Further investigation shows that the modern amplifier concept employs low-profile styling and is designed for desk-top operation next to the exciter. In some cases, the power supply is an externai unit. In any case, the amplifier is capable of being controlled by the external VOX or push-to-talk circuit of the exciter. And it can be operated either from 120- or 240-volt primary service.

table 1. Typical rf linear amplifier service, cathode-driven (grounded-grid)

tube	plate **voltage**	zero signal plate current	maximum signal plate current	maximum signal grid current	maximum signal drive power	typical power output (PEP)
(2) 3-500Z	2500	260	800	240	80	1200
	3000	320	667	230	60	1250
3-1000Z	3000	240	670	220	65	1250
4-1000A	3000	90	670	270	125	1300
8877	3000	130	667	55	50	1150

Representative operating charactersitics of popular tubes suited for cathode-driven service. The 4-1000A is operated as a class-B triode with grid and screen tied together. It can be seen that in terms of efficiency there's not much difference between tube types. The 4-1000A requires the most drive power, the 8877 the least. The differences in power output are insignificant and are within error of measurement. Power output is a function of plate circuit loading and grid drive causing the values to be approximate.

power capability

Given these general specifications, the next step is to determine what goes into the black box that is to become your new linear amplifier. If the linear amplifier is to run at 1-kW in the CW mode and 2-kW PEP in ssb service, the choice of tubes to be used narrows. And since cathode-driven (grounded-grid) service is contemplated, selection is restricted to a few tubes which have the ability to sustain this power level with good linearity and low intermodulation distortion. Linear operation implies that the output signal is an exact replica of the input signal; low intermodulation distortion means that unwanted, spurious distortion signals are not generated in or near the signal frequency. When both of these criteria are met, the ssb signal is clean and no power is lost in *furry* sidebands or splatter.

Table 1 shows some practical tubes for linear service and their operating characteristics. From an engineering point of view, there's not much choice between using a single large tube or two smaller tubes in parallel in the high-frequency region. Multiple tubes are thought to be less efficient and generate more intermodulation distortion than one tube, but this is not the case. Power tubes designed for ssb service do not have to be matched pairs, as do the inexpensive TV sweep tubes used in some linear amplifiers.

Regardless of the tube or tubes chosen, grounded-grid triode operation implies class-B service (which you can find outlined in detail in the suggested literature). An important characteristic of this class of service is that tube efficiency is at maximum 66 per cent and usually runs close to 60 per cent. Inherent tank circuit losses in the amplifier reduce this a bit, so that the plate power output of a representative class-B amplifier may run about 55 per cent. However, in cathode-driven (grounded-grid) service, a portion of the driving power (feed-through power) appears in the plate output circuit and provides a measurable output efficiency of approximately 60 per cent for the stage.

Now, if your maximum input power level is known, as well as plate efficiency, it is easy to determine the power output of the amplifier as well as the power dissipation of the tube (generally known as plate *dissipation*).

If the 2-kW PEP power input condition is chosen and an overall amplifier efficiency of 60 per cent is assumed, the PEP output will be

$$2000 \times 0.60 = 1200 \text{ watts}$$

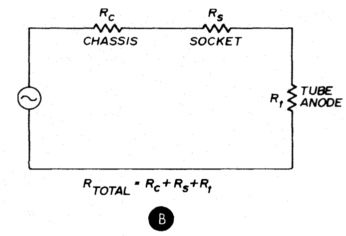
The remainder of the power ($2000 - 1200 = 800$ watts) is consumed in plate dissipation and circuit losses. Tube plate dissipation at 60-per cent efficiency runs close to 800 watts. Circuit losses run from 50 to 100 watts. These figures may add up to a little more than 800 watts of power loss, but a portion of this is accounted for by the plate dissipation attributable to the feedthrough power previously mentioned. So, without doing anything more complicated than a little grade-school math, the general operating parameters of the amplifier have been determined.

At the 1-kW CW condition, power input is 1000 watts and tube efficiency remains close to 60 per cent, provided certain circuit precautions are taken (these will be discussed later). You can estimate the power output and tube and circuit losses yourself for this power level.

Linear amplifiers and their power supplies have grown sleek and physically smaller in recent years. More efficient components are used and cooling techniques have improved, permitting the amplifier to be squeezed into a compact cabinet with high eye appeal. Some manufacturers and designers, however, have cheated, by skimping on the power transformer or by using an inadequate cooling system that allows the tubes to overheat during extended periods of operation. One old timer, when asked to judge the relative merits of two competitive, widely advertised linear amplifiers, replied, "That's easy! You don't need any test equipment at all. All you do is weigh the amplifiers. Unless one of them has lead fishing weights in it, the heavier amplifier is the toughest and best!"

There is more than a grain of truth in this remark. Attempting to cram a 2-kW PEP amplifier into a shoe box is a time consuming and complicated task, since the problem of getting rid of the heat caused by tube dissipation and circuit losses is a formidable one. Imagine a metal box the size of a 2-kW PEP linear amplifier with an 800-watt bulb burning inside of it! Or consider that a burner on an electric stove may be only 600 watts. This will give you a picture of the amount of heat that has to be removed from a 2-kW PEP linear amplifier during operation to prevent it from burning up.

The human voice, which is the usual modulating device in ham radio, luckily has a low average power level with quite high peak power. Thus, an amplifier designed for voice operation can have a power supply designed for low average power, yet be capable of sustaining full peak power for a short time interval. Many manufacturers count on this low average voice



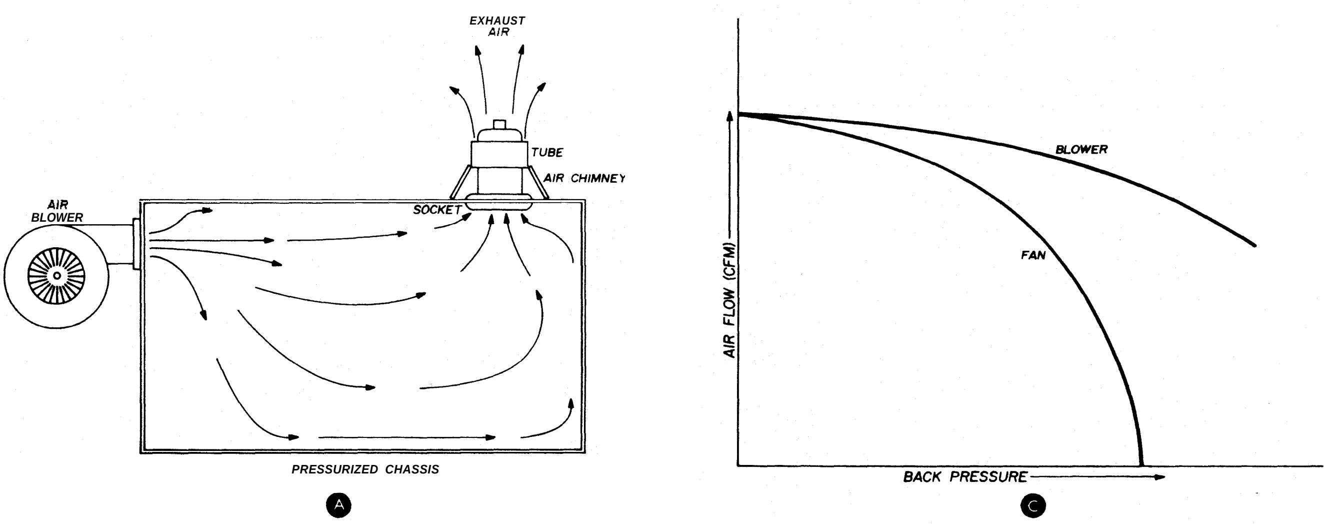


fig. 2. A representative air cooling system for a ceramic-metal power tube, such as the 8877. A forced-air cooling system is shown in (A); the blower is mounted on the chassis which acts as a plenum chamber. With the chassis airtight, the air is forced past the tube socket, tube base, and out the anode. The chimney is used to direct the air through the finned anode. An electrical analogy of the cooling is shown in (B). The blower is represented by a generator and the various back pressures by the voltage drops across the series-connected resistors. Total back pressure is the sum of the resistances. Representative fan and blower performance is illustrated in (C). The blower efficiency drops as the back pressure is increased, while the fan fails to deliver air at any appreciable back pressure. The ability to overcome back pressure is proportional to the speed of rotation of the blower or fan, plus the physical design of the blades. An inefficient fan allows air to slip around the ends of the blades. It is difficult to determine a good blower or fan, as opposed to a poor one, by intuition. Graphs of blower and fan performance may be obtained from the manufacturers.

power and skimp on the power transformer in an effort to squeeze their amplifier into a small cabinet. But what happens if speech processing is used to raise the average voice level and the amplifier is operated continuously during a DX contest? Or the amplifier is used for RTTY? The power transformer, amplifier, and tubes may not stand up under this added burden. Make sure your design is capable of tough, continuous operation. This is the least expensive approach for the long run.

tube cooling

Once the tube type has been chosen, the next item of business is to adequately cool the tube. The manufacturer's data sheet provides maximum tube temperatures and usually the amount of air required to do the job. What does this entail? An air cooling system is shown in **fig. 2.** It can be compared with a series electrical circuit, wherein the resistance to the flow of air created by the tube and accessories is equivalent to the opposition to current flow provided by resistors. The air resistance (back pressure) is equivalent to the voltage drop across the resistor, and the number of cubic feet of air per minute (cfm) required to overcome the back pressure can be compared with the voltage necessary to force current through the resistors. Back pressure is measured by a manometer and is expressed in terms of equivalent inches of water. Once back pressure and cfm are determined, the blower can be chosen that will force the required air through the system.

Air requirements for some popular transmitting tubes are listed in **table 2.** As an example, the 8877's maximum operating temperature is 250 degrees *C.* To hold this value, about 22 cfm are required to overcome a back pressure of 0.2 inch of water. This provides an anode dissipation of 1000 watts, more than sufficient for ssb operation at the 2-kW PEP level. The full anode dissipation rating of 1500 watts can be achieved with an air flow of 35 cfm, but at the price of a higher back pressure value of 0.41 inch of water. In passing, it should be noted that axial fans do not like working into high values of back pressure, as **fig. 2C** indicates.

A single 3-500Z tube requires 13 cfm air flow at a back pressure of 0.08 inch of water *per tube.* For two tubes, the *air flow requirement doubles to 26 cfm, but the back pressure remains the same.* Generally speaking, air flow is easy to obtain, but back pressure ability is hard to come by in simple, inexpensive, and relatively noiseless blowers. Amateurs like blowers that don't make noise. Unfortunately, movement of air creates noise, and the higher the back pressure requirement the more air noise that will be created. (This limits the size of a practical, air-cooled, transmitting tube to about 50 kW, above which it would probably require a Volkswagen engine to run the blower and would produce sufficient noise to drive the operator out of the station. Hence, the use of water or vapor cooling in the largest transmitting tubes.)

choice of blower

The most common air impellers are the centrifugal (squirrel cage) blower and the axial fan. The axial fan is the quieter of the two, but does not have the ability to work into a high level of back pressure. The ability of the squirrel cage blower to overcome back pressure is a function of the blower speed in rpm and the diameter of the wheel — the larger the diameter, the lower can be the rpm for a given amount of back pressure. Suggested blower specifications and fan information are included in the data for popular tube types.

It must be remembered, too, that the hypothetical Amateur living at an altitude of 1600 meters (5000 feet), in Denver, Colorado, for example, exists in a world of thinner air than that encountered at sea level and would have to increase the air requirements out-

table 2. Representative cooling requirements for various power tubes.

tube type	cfm	back pressure	blower diameter	rpm
3-5002	13	0.08	3	1600
(2) 3-500Z	26	0.083	3	3100
3-10002	25	0.43	3¾	3000
8874	8.6	0.37	2¾	3100
8875	[2]	0.16	[4]	2800
8877[1]	22.5	0.20	3	3100

Notes **1.** For 1000 watts anode dissipation
2. 1600 feet per minute from axial fan
3. In EIMAC SK-410 socket with EIMAC SK-406 chimney
4. Axial fan or blower

The listed values are given in cubic feet per minute and back pressure in inches of water. The impeller information is the centrifugal blower wheel diameter (inches) and motor speed in revolutions per minute. Low-speed blowers are attractive because they create less air noise, but they are unable to work into any appreciable amount of back pressure. As shown, these tubes, regardless of plate dissipation, require a blower speed of about 3000 rpm, except for the single 3-500Z. These data are for operation at sea level and the quantity of air should be increased about twenty per cent for operation at high altitude (Denver, Colorado, for example). The cooling requirements can be verified only by making temperature measurements on the tube seals and the anode. Glass tubes, such as the 3-5002, can be cooled from the side by an axial fan, but only after tests are made to ensure that the glass envelope temperature remains within specified limits.

lined in the illustration by about twenty per cent to achieve the same degree of cooling.

amplifier enclosure

Once the cooling requirements have been determined, all that remains is to get the cooling air into and out of the amplifier box. Why enclose the amplifier? Aside from cooling requirements, today's electrical specifications require rf harmonic suppression of a high order. This means that the amplifier must

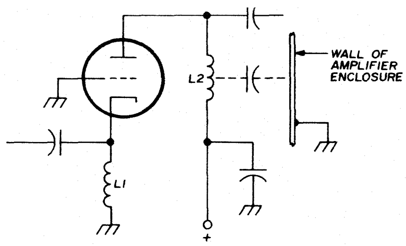


fig. 3. Both input and output chokes of a grounded-grid amplifier can form a parasitic oscillator circuit. Usually the cathode choke, RFC1, has more inductance than the plate choke. RFC2, but stray capacitance between the plate choke and the enclosure can lower the resonant frequency of the plate parasitic circuit until an uncontrolled oscillation can occur. This unwanted oscillation can be cured by removing turns from the plate choke, moving the choke farther away from the metal walls, or by placing a resistor either in series or in parallel with the choke. Means for detecting such parasitic are discussed in the text.

be placed in an rf-tight box and that connections to the amplifier be carefully filtered to prevent unwanted harmonic energy from escaping and blocking out Joe Sixpack's television receiver next door. All amplifiers generate and amplify harmonics of the driving signal; the task is to keep them from harm's way. Proper filtering will do the job.

Ventilation holes can be placed in an rf-tight box, provided they are properly screened. Wires can enter and leave the box provided they are properly filtered. A screened opening should be about twice the size of an unscreened opening to obtain the same air circulation, since the screening material represents nearly 50 per cent coverage of the area. A series of many small holes drilled in the top and bottom of an enclosure will provide ventilation without letting any great amount of rf energy escape, provided the holes are small compared with the harmonic frequency. For high frequency work quarter-inch holes are satisfactory. More smaller holes will work, too. Copper wire screening can be placed over the blower opening if the mating surfaces between screen, blower, and chassis are free of paint so that electrical continuity exists between the various metals.

indicating meters

Several meters are required to properly tune and operate a linear amplifier. At the very least, grid and plate currents should be monitored, and it is convenient to be able to read filament voltage. The grid and plate currents can be read on one meter switched between the appropriate circuits, but the use of separate meters is recommended for ease in tuning. All meters should be checked for accuracy before installation in the amplifier.

Placing the meters in the walls of the amplifier box is bad, since the rf energy can easily escape through the meter case and glass, invalidating the otherwise good shielding of the unit. It is wiser to place the meters on a separate front panel, with the amplifier box supported behind the meter panel. Meter leads are then brought out through appropriate filtering networks.

parasitic suppression

"You don't have to worry about shielding or neutralization. A grounded-grid amplifier just won't oscillate." Right? Wrong. A grounded grid amplifier makes a very good oscillator under certain conditions.

Low-frequency parasitic oscillations. Any amplifier can oscillate in the low-frequency region **(200-1500**kHz) by virtue of the interelectrode capacitances of the tube forming some resonant circuit with either the input or plate rf chokes **(fig.** 3). A sure cure for this problem is to change the type of choke, or else place a resistance in series or in parallel with the choke to inhibit oscillation. In the designs discussed here, the inductance of the input choke is very low compared with that of the plate choke, so that oscillation is improbable.

A low-frequency parasite can often be heard in a nearby broadcast receiver as an unsteady carrier or a rough buzz. Or, it can be found when the amplifier is operated with plate voltage (but no excitation) and the controls tuned at random. A small neon lamp is held near the plate lead. If a parasite is present the bulb will ignite with a bright ***yellow*** glow. The bulb should be held at the end of a dry wooden stick, as dangerously high voltage is present and exposed when the amplifier is operated with the cabinet shielding removed.

Vhf parasitic oscillations. Vhf parasites are treated by resonant circuits formed by connecting leads and interelectrode capacitances of the tube **(fig. 4).** They can be suppressed by loading the circuit until oscillation is impossible. A parasitic choke, composed of an inductor and resistor in parallel, will do the job. The suppressor is placed in the plate lead, but a second suppressor is sometimes required in the input circuit.

The suppressor represents a portion of the lead wound up into a coil and shorted by a resistor. At the parasitic frequency there is a large voltage drop across the coil. The resistor acts as an rf load for this

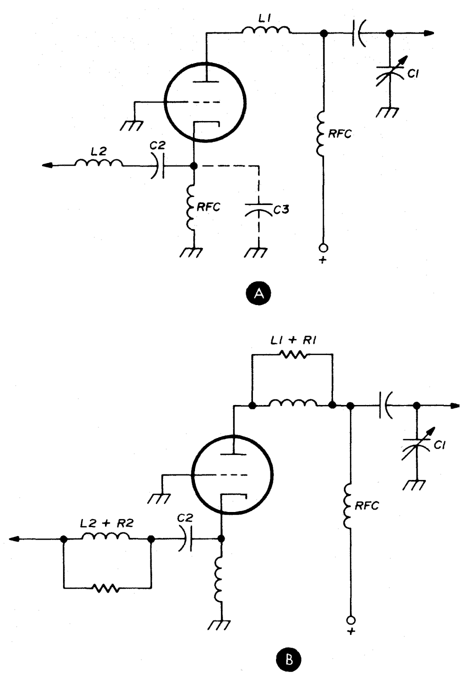


fig. 4. Vhf parasitic circuits in grounded-grid amplifiers are made up of stray input capacitance (C3) plus the inductance of the input and plate leads, L1 and L2, see (A). The parasitic oscillations may be suppressed, as shown in (B), by shunting a portion of the input and/or plate lead with a resistor to load the parasitic circuit. However, this choke must not be too tightly coupled to the plate circuit at the operating frequency or it will dissipate fundamental frequency power and overheat.

voltage drop. If the load is tightly coupled to the tube, oscillation will not take place, but if the load is tightly coupled at the operating frequency, the suppressor will probably overheat and burn up. The number of turns in the inductor must be determined by test so that sufficient inductance exists to do the job, but not enough inductance is used to couple too much fundamental energy into the resistor,

A vhf parasite can be determined by the neon bulb test. The bulb will glow with a bright ***purple*** color if oscillation is taking place.

High frequency parasitic oscillations. The grounded-grid amplifier can be turned into a splendid oscillator if the input circuit is detuned too far from resonance. The tuning range of the input circuit should therefore be quite restricted. It is a good idea to tune the input circuit to the middle of the Amateur band in use and then forget it. Actual adjustment of this circuit will be discussed in a subsequent article.

High frequency parasitic oscillation can also take place if output power from the amplifier finds its way back into the input circuits. Shielding and filtering (which also reduces vhf harmonic energy radiation) serves to reduce the possibility of high frequency oscillation. In some instances, the grounded-grid amplifier must be neutralized to achieve stability. Luckily, this is generally not required for amplifier operation below *30* MHz but the builder should be aware of the fact that high frequency oscillation at the operating frequency can take place in a grounded-grid amplifier if the proper precautions are not exercised.

summary

So far, we've slogged through a quicksand of physical design problems. Let's sum up what has to be done as far as this aspect of amplifier design is concerned:

1. The amplifier tube has to be chosen for the particular power level desired and must be properly ventilated and cooled.

2. The amplifier has to be placed in an rf-tight box for harmonic suppression and operational stability.

3. Circuit design must ensure that unwanted oscillations do not take place.

The next article in this series will discuss the electrical design of the linear amplifier.

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ham radio

AFC circuit for VFOs

Add a new measure of stability to an existing VFO by incorporating this AFC circuit

The search for frequency stability in Amateur equipment, both commercially manufactured and homebrew, has been one of the longest and most fervently pursued in the history of ham radio. In the earliest days, when King Spark reigned supreme, frequency stability wasn't much of a problem since the damped waves covered a considerable portion of the spectrum. However, as the state of the art advanced (and in those days it was an art), the vacuum-tube oscillator became *re rigeur,* receiver selectivity improved (necessary because of a more crowded spectrum), and the need for greater oscillator stability quickly became apparent. That need has been with us ever since!

In the early 1930s, crystal controlled transmitters became the way to go, with a powerful assist from the Federal Radio Commission (the ancestor of the FCC). You could actually find the same ham at the same spot on the dial (almost) night after night. After all, during the depression, who could afford more than one or two crystals? They cost about two days' pay (about $7.50) each! Crystal control was undeniably a great advance over the self-excited oscillators then in use for transmitter frequency control. By the standards of the day, transmitters became literally "rock stable." Unfortunately, receivers of that period didn't enjoy the same stability. The superheterodyne had become popular, and the instability of its high frequency oscillator (HFO) required a frequent touch of the tuning dial (in much homebrew gear) to keep the desired station audible. Manufactured gear was better, of course, but few Amateurs could afford it. By today's standards these oscillators were pretty crude, but we were dealing with a lot of CW and a little a-m telephony. With the broad bandpass windows in those old receivers, a-m was easily handled, and a changing CW beat note could be tolerated as long as the band wasn't crowded.

Use of the Amateur bands was rapidly increasing, however, and by 1939 or 1940 the trend was away from crystal control toward something known as the ECO, or electron-coupled oscillator. Great strides had been made in stabilizing the same old Hartley and Colpitts oscillators by using higher gain tubes such as the new pentodes, looser coupling between oscillator and load, and a myriad of other tricks that today we take for granted.

Rather than labor the points unduly, it is best to say that today, both VFO and crystal oscillators are in wide use in the ham bands, the more stable crystal oscillator being used for certain receiving functions, MARS frequency assignments, vhf repeater channels, and so forth. The VFO, of course, finds its major application in the HFO of the modern hf transceiver. A new state-of-the-art device, the frequency synthesizer, actually uses both techniques: a reference crystal oscillator of high stability and a particular type

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Front view of the VFO-stabilizing unit. The three LEDs are used to indicate the position of the control voltage within its total range.

of VFO known as the voltage-controlled oscillator (VCO) locked to the crystal oscillator by the phase-locked loop.

Still, the great majority of equipment in use today employs the VFO in one form or another, tube or transistor, Hartley, Colpitts, Vackar. They share one common failing; they *all* drift in frequency to some extent. Even the justly famed Collins PTO can be irritatingly unstable to a frequency-measuring nut.

Individual component variations in an otherwise sound VFO design can, in a production situation, become an annoying source of trouble. A more fundamental problem is that to thoroughly stabilize a VFO takes much time and use of temperature-compensating capacitors. The stabilizing procedure also requires a manufacturer to employ a trained technician for this task. The manufacturer attempts to arrive at a sort of stability "middle ground," or performance that will satisfy the majority of buyers.

what is a VFO

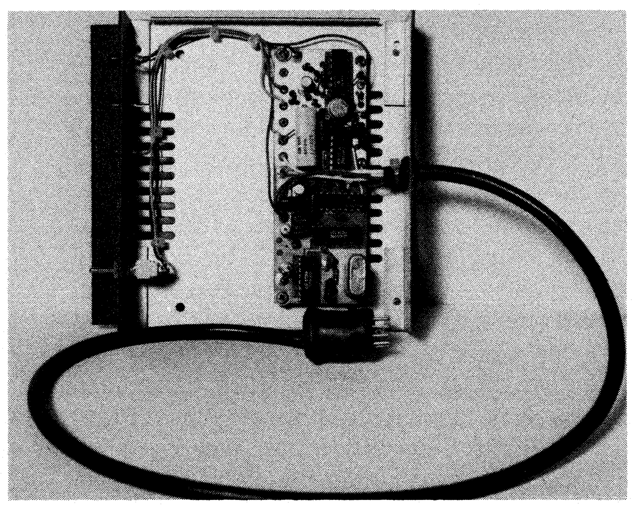
Let's briefly examine the device we're trying to improve so that a little more than the tip of this iceberg becomes visible. Contrary to opinions expressed by some sources, black magic does not play a major role in VFO design. The modern VFO is a marvel of construction. It is usually built in a very rigid box of steel or heavy aluminum, and almost invariably today is a solid-state device even though vacuum tubes may appear elsewhere in the radio. Use of solid-state devices removes from the VFO one of its worst enemies — heat. Most of the VFO circuits in production equipment today can trace their origin to the basic Colpitts circuit, probably because the coil tap in the Hartley introduces added switching complexity and cost.

Even more basic is the real heart of the VFO, the tank circuit or frequency-determining network. *Anything* that causes the oscillatory period of this network to vary (other than deliberate variation) will have an adverse effect on the stability of the VFO. The sole exception to this statement is the AFC circuit. However, this can be considered a form of operator control and, in fact, the operator can alter the VFO frequency through manipulation of the AFC circuitry, Undesirable factors include changes in temperature, change in the load on the VFO, changes in supply voltage, and vibration. An excellent measure of the designer's success is the stability of his VFO. From a manufacturing standpoint, add one more vital point. It must be reproducible! I once built a little VFO that was a marvel of stability. Three other people tried to reproduce it. Their versions oscillated and they were in the correct frequency range — but they drifted. The circuit was not easily reproducible. Therefore, it was useless except to me.

VFO improvement

Many papers have exhaustively covered the trials and tribulations of VFO design, both from the home-brew aspect and from the commercial viewpoint. It seems sufficient to note that most of the hints, techniques, design criteria, and other good and valid information presented in VFO design papers[1,2,3] are basically aimed at reducing frequency drift. We have relatively simple formulas for calculating the component values, but no one has yet come up with a magic formula to make the VFO stable.

The recent article by PA0KSB[4] suggests one solution to the stability problem. **Fig. 1** shows a stabilizing device in block diagram form — a single stage



Inside view of the logic portion of the control unit. The only other components are those installed inside the VFO enclosure. Additional information regarding the changes to the Atlas equipment can be obtained either from the author or by writing to Atlas Radio.

binary counter, a storage element (D-type flip-flop), a clock (low frequency crystal oscillator), and an integrator to drive the control element. **Fig. 2** shows the entire control element, a tuning diode, a bypass capacitor, an isolation resistor, and a small coupling capacitor. Room for these tiny components could easily be found in any VFO I've looked at. As you can see from the size of the coupling capacitor (1 pF), the effect of the VFO (dial) calibration won't be a major one. In all cases so far, minor readjustment of individual band trimmers has easily absorbed the small additional capacitance represented by the control element.

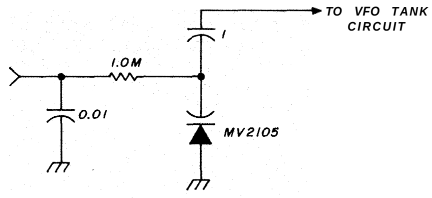


fig. 2. Schematic diagram of the control unit which is installed inside the VFO. **Since the added capacitance is very small, any change in frequency is easily compensated for by the VFO's trimmers.**

practical AFC package

The actual circuit of the AFC unit is shown in **fig.** 3. There are a few changes from the original. First, in this country at least, there are no 100-megohm %-watt resistors commercially available. To my knowledge, the highest value %-watt resistor available in quantity in this country is 22 megohms. A series pair of these resistors, along with a 2-μF capacitor (metallized Mylar or polycarbonate) will provide an 88-second integrator time constant, as compared with the 100-second value in the original circuit. The slightly shorter time constant has proven entirely adequate in practice. It must be noted that high-leakage capacitors, such as electrolytics, are unsatisfactory in this application. The added expense of the low-leakage capacitor has to be accepted to obtain a smoothly operating unit.

The circuit likes a low impedance input; this point is mentioned in the original article, but without particular emphasis. As a majority of the radios manufactured in this country have a VFO output impedance in excess of the 50 to 100 ohms which I (and apparently the AFC circuit) consider low impedance, a broadband autotransformer has been added at the input to help the interface problem. Use of this approach has allowed sampling of the VFO output voltage without detriment to either its output amplitude or intrinsic stability. This particular transformer has been examined with a Hewlett-Packard 250B Rx meter and found to be essentially flat from about 2 to over 23 MHz. This is more than adequate for any range of VFO frequencies yet encountered.

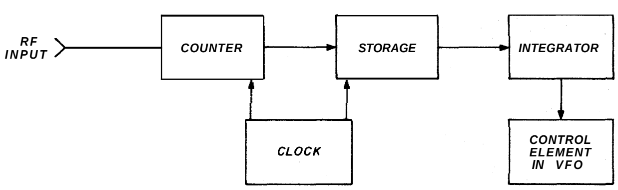


fig. 1. Block diagram of the AFC control unit. The clock controls both the counter and storage sections, determining when to count and when to transfer information to the next stage. The integrator serves to smooth out the information going to the control element, preventing the VFO **from jumping back and forth in frequency.**

Routine equipment turn-on generally gives a ramp voltage of about one quarter maximum. This, plus normal voltage change due to operation, necessitated some form of metering. An early approach with Atlas equipment involved switching the S-meter. However, this method has since been discarded in favor of three LEDs which, with the addition of a 7406, illuminate upper and lower limit red warning lights when the ramp voltage comes within one volt of either end of its range. In between, a green LED glows. The transition from red to green, at each end of the range, is reasonably abrupt. Only a few millivolts of overlap exist between the red and green LEDs. This metering system provides an adequate Go/No-Go indication which is particularly useful under mobile operating conditions. Trying to read a conventional analog meter under crowded freeway conditions is not exactly conducive to an extended life expectancy! Also, the components of the LED metering system are noticeably less expensive than a meter.

In addition to metering, this approach makes possible construction of the AFC circuitry in a completely separate box without a multitude of connecting wires. The length of the four-wire umbilical isn't critical, allowing the user to position the box for optimum convenience. This concept also reduces the amount of internal work necessary to incorporate the drift-cancelling circuitry. For example, in the case of the Atlas 1801210 series, the device plugs directly into the AUX VFO socket. A 100-ohm resistor at the radio end of the connecting cable prevents any interaction between the AFC system and the VFO in the radio. Internal work on the radio is limited to installation of the control circuitry within the VFO compartment and adding one wire from the control element to pin 1 of the AUX VFO socket (normally unused) to carry the ramp voltage. All other voltages are already present. The digital dial (if used) plugs into this same socket, but, because of the plug construction on

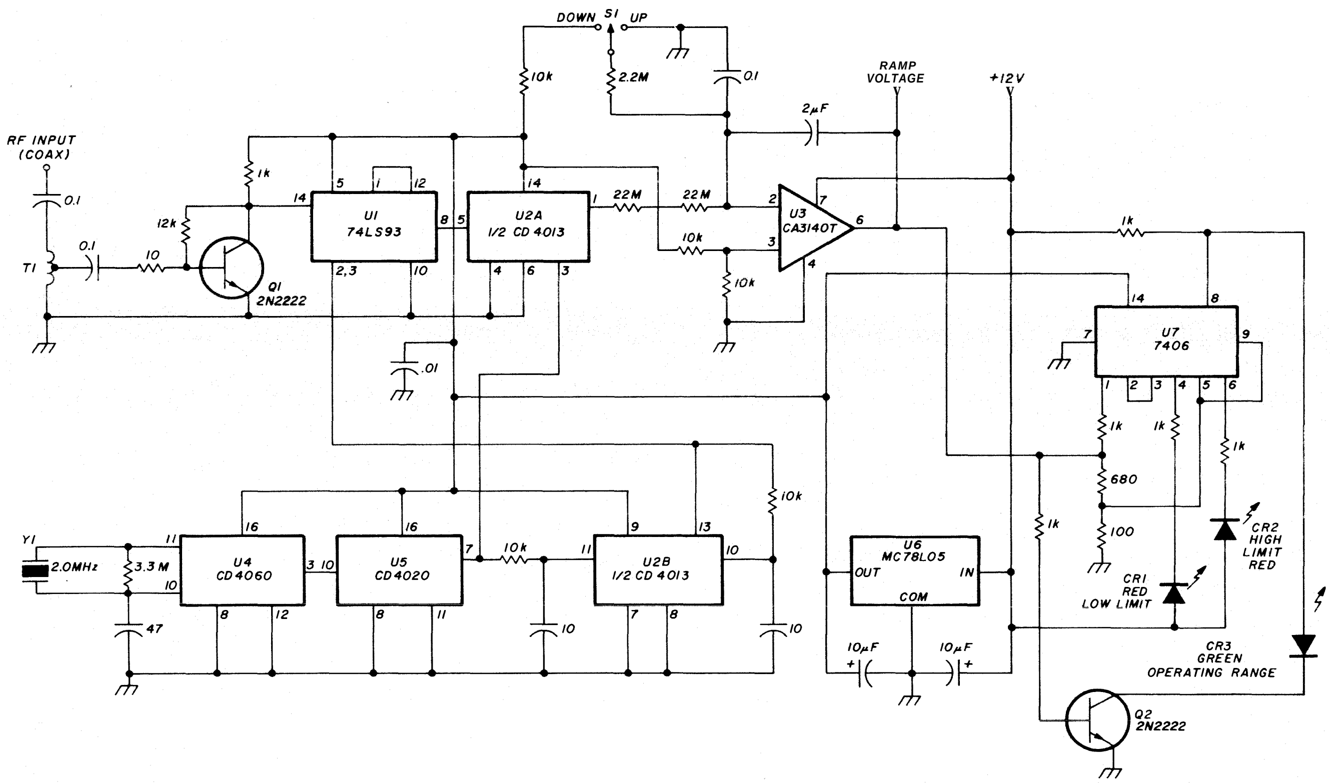


fig. 3. Diagram of the logic portion of the AFC control unit. The transformer is bifilar wound with nine turns of number 26 AWG (0.4-mm) enamel insulated wire over a Q2 ferrite toroid. Core should be 9.4 mm (0.37 inch) outside diameter, 5 mm (0.2 inch) inside diameter, and 4.9 mm (0.19 inch) thick.

both the digital dial and the AFC unit, they piggyback without trouble or interaction. In a transceiver as compact as the 210X, minimal work within the radio is a worthy consideration. Conversely, within the Atlas VFO compartment there is more than adequate room for the control-element components. After having modified several of these radios, total conversion time is less than an hour, including the minor recalibration required.

adaptability

The AFC unit will perform equally well on older tube-type VFOs. It has successfully been installed in Heathkit tube-type VFOs (SB400 series) with no problems. No adverse effects on dial linearity were noted. Since this VFO is not bandswitched, only one recalibration is needed. Even though the piston capacitor is of very low capacitance, more than enough range is available after adding the AFC control element.

Initial examination of an older Swan 350 proved enlightening. Although the schematic shows a solid-state VFO and emitter follower practically identical to the early Atlas units (as well as a tube-type VFO amplifier), maximum rf voltage was only 60 millivolts (at 7 MHz). Other bands weren't much different. This rf level doesn't allow resistive isolation in the pickoff line, because the input voltage to the AFC unit then becomes too low for reliable operation of the counter.

Even with a relatively poor VFO, AFC lockup will occur within a minute. It takes a little time for the integrator to do its thing. A good VFO will normally stabilize within about 15 seconds of turn-on. For example, the Atlas 350-XL with AFC settles down in ten to fifteen seconds, while a relatively poor home-brew unit, the first one on which the AFC approach was tried, took about a minute to stabilize. However, this VFO was so bad that without AFC it would drift noticeably during a two-minute transmission. I threw the whole thing together in a hurry several years ago and never got aroung to taming it. Consequently, it was a natural for AFC experiments. The first time the modified rig was used on the air, I was accused of having bought a new radio. "We don't have to chase you any more!" Constant attention had to be paid to the ramp voltage though, because that VFO never did settle down on its own. About once an hour the ramp voltage would have to be reset because it would be getting perilously close to its range limit. Conversely, the Atlas 350-XL can run all day and nothing has to be reset. The ramp voltage has more than enough range to keep this excellent VFO locked up indefinitely.

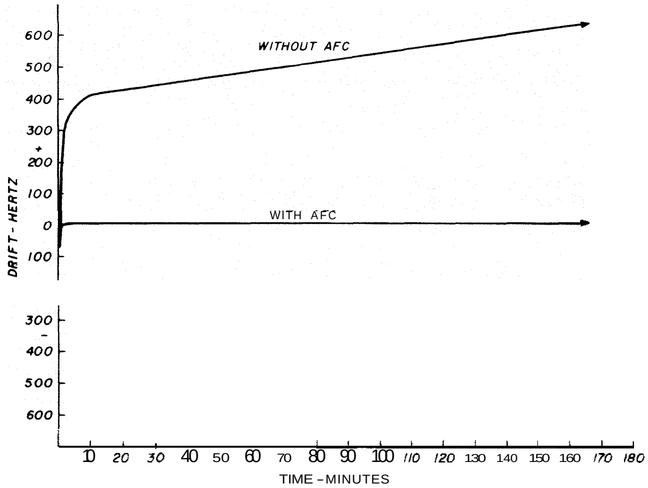


fig. 4. Examples of measured drift from VFOs with and without AFC.

The system of automatic frequency correction described in this article offers an order of magnitude stability improvement. It is not a universal cure for all the ills that may beset a VFO. For example, this circuitry cannot correct the frequency jump that is caused by sticking or misaligned mechanical tuning assemblies or from worn or corroded contacts on a bandswitch or variable capacitor shaft. Sudden frequency excursions of this type look to the circuit like "human intervention," a frequency shift too rapid to be corrected. One demonstration of AFC capability that invariably generates astonishment is connecting a counter capable of reading the VFO frequency to the nearest Hertz to a cold radio which is AFC equipped. When the radio is turned on, the counter will faithfully reveal the rapid initial warmup drift, the sudden stop, and then will remain for hours within about 5 Hz of the original lockup point (see **fig. 4**).

In the near future, buyers of Atlas equipment will be able to order this feature as an option; a retrofit program exists for older Atlas equipment. For those brave souls willing to invade the mysterious innards of their VFO, the system is available as a package; the modification really isn't all that hard to perform. If AFC of the VFO isn't the final answer to outstanding stability in older gear, it must be pretty close to it.

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ham radio

improving antenna accuracy in satellite tracking systems

A review of the problems encountered in Amateur tracking systems, with suggestions for improving accuracy and operating efficiency

Most users of the OSCAR communications satellites have some form of antenna tracking system for obtaining the greatest advantage from the available transmitter power and for increasing the strength of the received downlink signal. These systems range from simple, manually switched dipoles to elaborate arrays driven in azimuth and elevation under microprocessor control. But most setups fall somewhere in between – often with antennas driven in azimuth and elevation by rotators with separate control boxes manually operated by the same person carrying out the communications – similar to the arrangement shown in **fig. 1.**

This general arrangement is simple, inexpensive, and not particularly difficult to operate when the satellite is low in the sky. On these passes, neither azimuth nor elevation changes very rapidly, and it's a simple matter to update the antenna position from time to time. However, as the satellite gets closer and the elevation angle gets higher, things begin to change much more rapidly, which requires more frequent updating of antenna direction. On some passes activity gets downright frantic as the satellite passes close overhead and the operator finds that the elevation rotator is about to hit the stops and the azimuth angle has suddenly switched from south to north!

Even microprocessor-controlled systems are not entirely immune to these problems. What happens is that the azimuth angle begins to change so fast that the rotator simply can't keep up with it. Add to this the time lag in updating the antenna position in a manually controlled system, and you frequently find yourself in a situation where the elevation control has the antenna pointing straight up (or nearly so) as the satellite azimuth angle suddenly swings 150 degrees or more in a few seconds. So then you're stuck. The azimuth rotator slowly trundles around the 150 degrees (or 210 degrees in the *other* direction if the mechanical stop happens to be in the way), while the satellite merrily recedes into the distance.

Slew rate. The rate at which the azimuth angle changes is called the *slew rate.* What we'd like to do is find a way to anticipate excessive slew rates so we can reduce or eliminate the problem. The first order of business is to find the minimum elevation angle at which the slew rate can exceed the azimuth-rotator rotation speed. This, of course, depends on what kind of rotator you have, but most rotators turn a full circle (360 degrees) in one minute or 6 degrees/second, so we'll use this figure for the sake of argument.

If you ask a dozen different OSCAR users what that critical minimum elevation angle is you'll get as many different answers, but most will probably say it lies between 60 and 70 degrees. Experience in manually operated Az-El control systems does indeed seem to point to a number in this range; but surprisingly, the correct number for OSCAR 7 is an elevation angle of 87.3 degrees – a mere 2.7 degrees from the vertical! The mathematical procedure for arriving at this number is detailed in Apendix A*

*Appendix A is not included with this article but is available on request from the author. Please include a 229 x 305 mm (9 x 12 inch) self-addressed envelope with 28 cents postage. Overseas readers may send one IRC (5 for Air Mail) and omit the envelope.

By **Bill Johnston, N5KR,** 1808 Pomona Drive, Las Cruces, New Mexico 88001

along with the procedure for finding the maximum azimuth error (lag) between satellite and rotator as well as the lag duration. For most rotators the critical angle is extremely high. The faster the rotator, the higher the angle.

Obviously, then, the slew rate can exceed the capacity of the typical azimuth rotator for only a very short time on any given pass (in fact, for well under ten seconds at the *maximum).* So why does the slew rate cause so much trouble? We can best answer that by taking a look at just how high the slew rate can go, and in particular by looking at the special case where the satellite passes directly overhead.

Here we have the satellite approaching, let's say, from the southeast. Since it's going to pass overhead, it's coming straight toward us the whole time it's in view, and the azimuth angle doesn't change at all. (Actually, due to earth rotation, the satellite's ground track is slightly curved, and the azimuth angle *will* change a little.) At the instant the satellite passes through the zenith, the slew rate jumps to *infinity,* then becomes zero again as the satellite recedes to the northwest.

In other words, on an overhead pass, the slew rate is infinitely high if only for an instant. If the satellite passes very close by, but not directly overhead, the slew rate will become very high (but not infinite) and will remain high for a longer period of time.

Incidentally, the rate of change of the *elevation angle* is always quite low. For OSCAR 7, the fastest it can possibly change is a little more than a quarter of a degree per second, so this is never a factor we have to worry about.

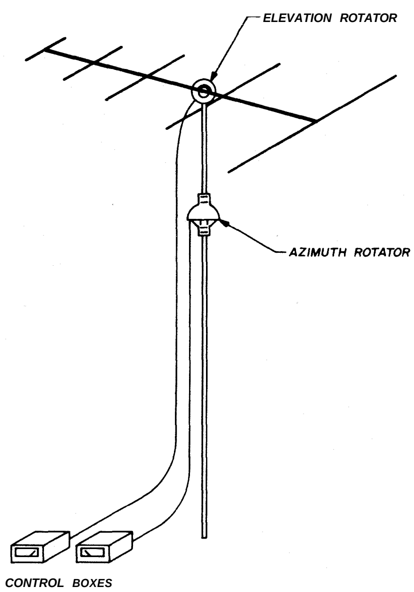


fig. 1. A typical satellite Az-El control system consists of a pair of rotators operated by separate control boxes. Usually the person carrying out the communications must also operate the antenna controls.

As mentioned earlier, OSCAR 7's slew rate can exceed the 6 degree/second speed of a typical rotator for less than ten seconds at the most on any given pass, But even after the slew rate decreases, it still takes a little time for the rotator to catch up. The worst-possible case is again the overhead pass,

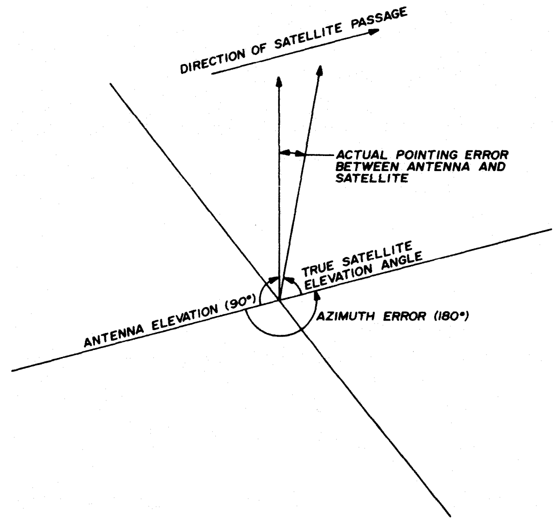
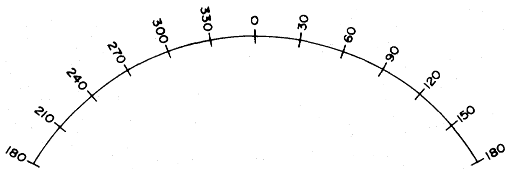


fig. 2. When the satellite is high overhead, or nearly so, the azimuth angle changes quite rapidly and the azimuth error may become quite large. However, with the antenna elevation at 90 degrees, the actual pointing error between antenna and satellite is fairly small.

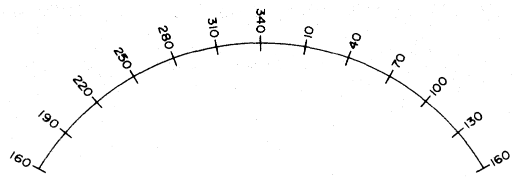
where the total duration of the lag is 30 seconds. (Appendix A gives details on how to compute the duration of the lag for other passes.)

Pointing error. Even though we can develop a temporary azimuth error of up to 180 degrees, the actual pointing error is not as bad as it might seem. For example, suppose we have an overhead pass and the elevation rotator hits the stops at 90 degrees, as shown in **fig. 2.** The azimuth suddenly swings 180 degrees, and the azimuth rotator direction is now in error by that amount. Even if the elevation control remained at 90 degrees and we waited 30 seconds for the azimuth rotator to come around, the satellite elevation angle would have changed by only about 8 degrees in that time. So the absolute pointing error between antenna and satellite true elevation angle would at most be about 8 degrees, *regardless of the magnitude of the azimuth error.* Once the azimuth rotator comes around to its proper alignment, the elevation rotator can be brought to bear on the satellite in less than 1-1/2 seconds (assuming a 6 degree/second turning rate).

With typical antenna beamwidths of 15 to 30 degrees at the half-power points, an 8-degree pointing error is inconsequential. Judicious operation of the



BEFORE MODIFICATION



AFTER MODIFICATION

fig. 3. Azimuth rotator dial face as normally supplied for the mechanical stop at 180 degrees and as modified when the stop is placed at 160 degrees.

elevation control can reduce the magnitude of that error even further.

reducing antenna pointing error

Everything we've said so far seems to indicate that the overall problem isn't nearly as great as we thought, so why even bother worrying about it? Well, first of all, the relatively small pointing error described above can be achieved *only* if the antenna is kept on track at all possible times. From a practical standpoint, in a manually operated control system, this simply isn't feasible. If you devote all your time to operating the antenna controls you have no time for communicating.

The end result is that the actual time lags and pointing errors are usually quite large. Nevertheless, a number of ways are available to improve the performance of such a system without putting an increased workload on the operator. Another good reason to attack this problem is that improvement in pointing accuracy allows the use of higher-gain antennas, which have inherently narrower beamwidths. This is especially beneficial when the antenna is under some form of automatic control.

Planning. A good rule in manually operated systems is to plan each pass ahead of time. Check to see how high the elevation angle will be to locate any potential trouble spots. During the pass, instead of playing a constant game of catch-up with the antenna, lead the satellite by half the interval between antenna corrections. For example, if you normally reposition the antenna every two minutes, then each time you move the antenna, set it to where the satellite will be one minute from that time. In this way the antenna will be a little ahead half the time and a little behind half the time, and the average pointing error will be reduced to *half* of what it would have been otherwise.

Mechanical considerations. Consider the azimuth at which the mechanical stop on the azimuth rotator should be placed. Since most rotators turn only 360 degrees, the ideal location for the stop would be at the azimuth corresponding to the point where the satellite first comes into view on a pass that will take it directly over the ground station. (For most locations, this would be an azimuth of about 162 degrees for ascending passes, or about 15 degrees for descending passes.) In this way there would never be any interference from the stop for the type of pass (ascending or descending) you selected for its location, and interference to the other type of pass would be no worse than if you did nothing a all.

It may seem that some rotators require the stop to be oriented to a particular direction (usually south). But this isn't true; you can orient the antenna to any direction you like on the mast. All that's required is that, once you've reset the antenna, you draw a new dial face for the position indicator. **Fig.** 3 illustrates an example where the regular dial face of an indicator requires that the stop be at 180 degrees and as modified for the stop at 160 degrees.

You might also try to obtain or modify a rotator that will turn more than 360 degrees. One that will turn 390 degrees with stops set at 180 degrees and 210 degrees, would work very nicely for both ascending and descending passes. As shown in **fig. 4,** if rotation starts from the 180-degree stop, the antenna can move clockwise a full turn through 180 degrees,

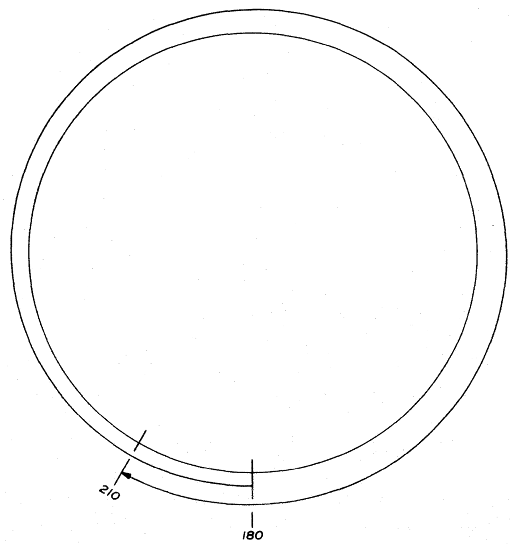


fig. 4. Turning pattern of rotator with 390 degree rotation

finally hitting the other stop at 210 degrees. These stops wouldn't have to be mechanical stops in the rotator but could be electrical stops in the control box or appropriate software if the system is under microprocessor control. Some sort of indicator (a pilot light, perhaps) could be used to let you know when the rotator is inside the extra 30 degrees, so you'll know which way to turn it if you're getting ready to track the satellite's initial azimuth as it comes over the horizon.

antenna elevation inversion system

Although most elevation control systems are set up with stops at 0 and 90 degrees, there's no law that says they have to be that way. And in most cases there is no physical reason that limits the equipment to this range. Remember the overhead pass? Suppose you don't touch the azimuth control, but when the antenna elevation reaches 90 degrees you let it keep right on going over on its back. All of a sudden you've eliminated a whole slew of problems. Instead of frantically chasing after the azimuth control, just ignore it, and every once in a while give the elevation control a slight touch up. Even when the pass isn't directly over head, much time and effort can be saved on high passes by inverting the antenna at the proper moment.

Antenna motion resulting from an elevation inversion system of this type is illustrated in **fig. 5.** Manufacturers of optical and electronic tracking equipment often incorporate this feature into their products, although they sometimes refer to inversion as "dumping."

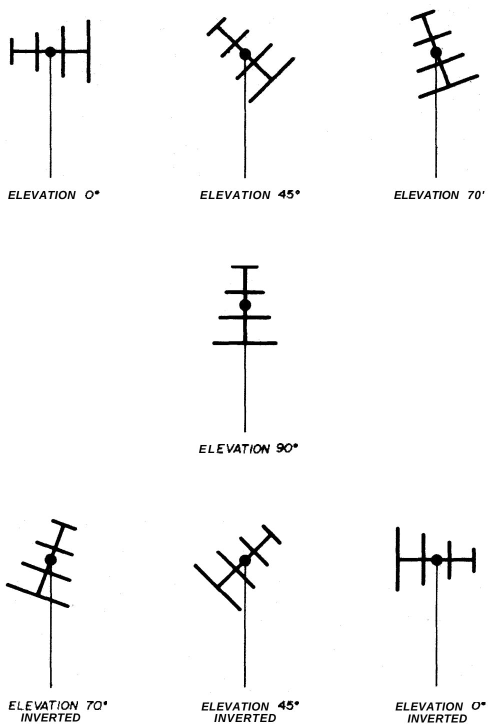


fig. 5. Antenna movement with an elevation inversion system. No azimuth change is made in these illustrations.

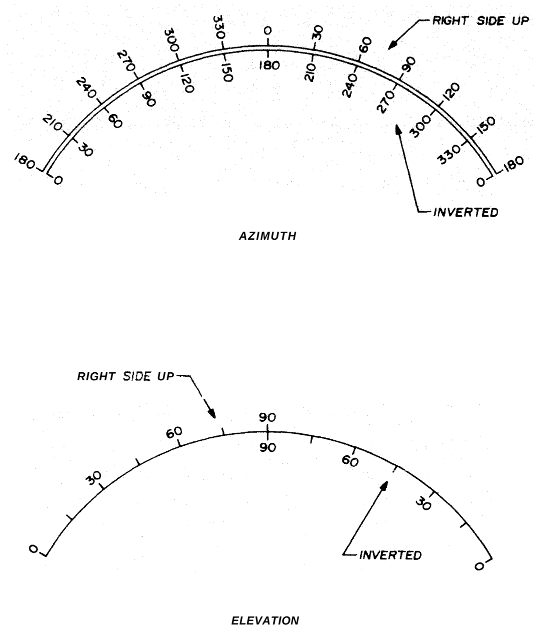


fig. 6. Azimuth and elevation dial faces for an antenna elevation inversion system. Numbers on the normal side of the dials are black; those for antenna inversion are red. When elevation dial is in the red, the corresponding azimuth red scale is used, and *vice versa.*

One thing you must do when using this type of system is to modify the azimuth and elevation control dial faces. Otherwise, when the antenna inverts, the azimuth dial wilt be 180 degrees off. It's normally a simple matter to extend the scale on the elevation control, and an example of how to label both faces is shown in **fig. 6.** The numbers on the normal (right-side up) side of the elevation dial are black and are red on the side that shows the antenna inverted. The azimuth control has its normal set of black numbers with an inner set of red numbers that are 180 degrees opposite in value to the black ones. When the elevation control pointer is in the black, you read from the black scale on the azimuth dial also. When the elevation dial is in the red, you read from the red scale on the azimuth control. It probably wouldn't hurt to have an indicator light illuminate when the antenna is inverted.

Renumbering dial scales isn't as hard as you might think. Dry transfer lettering works very nicely, and, if you want to be able to return the dials to their original condition, the new scales can be put on paper templates and attached with rubber cement. They can be peeled off later without damaging the original dial face.

systems under microprocessor control

Many of the ideas discussed so far, although especially suited for systems under manual control, are applicable to automatic control systems. Let's take a look at some factors that again affect both types of systems but are of special interest in systems under microprocessor control.

We already know that slew rates vary widely, from a small fraction of a degree per second up to a hundred degrees per second and higher. Most rotators, though, have only one speed, so you have to constantly turn it off and on to keep it on track. These start/stop operations put a lot of wear on the rotator, and the very nature of the procedure prevents the antenna from being exactly on the satellite at all times. Ideally, we'd like a rotator with a speed that varies according to the satellite slew rate, since this would produce a smooth and continuous motion with no pointing error.

Variable-speed rotators. Finding a variable-speed rotator suitable for your station would probably be no easy task, although you could build one from scratch with an appropriate motor. On the other hand, if your rotator is driven by a synchronous ac motor, it would be a simple matter to build a variable-frequency ac power supply that would in turn vary the motor speed. The microprocessor would select the correct frequency to produce the desired rotator speed. This, of course, would require calculating the slew rate in addition to the azimuth and elevation angles. The procedure for doing this is given in Appendix A.

Even the operator of a manually controlled system might find a variable-speed control handy, as it could reduce his workload to a certain extent. Commercially made speed controls, operated by a small joy stick and capable of handling motors of up to 15 watts, are available through a number of amateur telescope dealers. They are advertised as "drive correctors." You can save yourself some money, though, if you're willing to build you own.

It's even easier to vary the speed of rotators driven by dc motors, but whether the rotator is ac or dc, you'll have to invest some time in designing the interface circuitry between the control system and the microprocessor.

Microprocessor interface circuitry. Regardless of whether the rotator speed is variable or fixed, there will be times when the slew rate exceeds the rotator's capability. To reduce the complexity of the microprocessor software, the computer would calculate the slew rates for the pass ahead of time and store the information concerning the time period (if any) during which excessive slew rates will occur. Then, during actual tracking, the microprocessor would instruct the antenna to begin leading the satellite by a few degrees just before it gets to the bad area. The end result would be a worst-case pointing error of no more than 4 or 5 degrees, and then only for five or ten seconds.

There are, of course, many things the software should take into consideration; unfortunately we can't discuss all of them here. But whatever you do, make sure that the software knows where the rotator's mechanical stop is, and make sure it knows how to minimize interference from the stop.

operational procedure for a microprocessor-controlled tracking system

Now that we've discussed a number of the factors in developing a system, let's see how an interactive microprocessor-controlled tracking system might operate. The software would be designed for fully automated operation while at the same time incorporating interrupt capability to permit the operator to take control when desired.

The tracking program could be called up by typing in the word **TRACK.** Once activated, the program would accept additional commands. Entering NEXT **PASS** for example, would cause the computer to check the station's digital clock, compute the starting time of the next pass for each of the active satellites, and flash a message similar to the following on its display screen:

28 OCT 78 16:35 UTC

SATELLITE	MODE	NEXT PASS	TIME TILL START
Oscar 7	A	17:04	:29
Oscar 8	J	16:21	IP
Oscar 9	B	16:44	:09
RS2	A	17:51	1:16
RS4	A	16:50	:15

LENGTH	HIGHEST ELEVATION	DIR	PRIMARY COVERAGE
9	15	A	WEST
12	32	D	EAST
21	88	A	OMNI
4	6	D	NE
18	76	D	OMNI

which satellite?

Suppose, as in example above, you decide to work OSCAR 9, which has a Mode B ascending pass starting in nine minutes and lasting for 21 minutes. (Since it will be an overhead pass, the computer shows the coverage to be omnidirectional.) You would type in **OSCAR 9,** to which the computer would reply:

PREPARE TO TRACK?

Upon receiving a response of **YES,** the microprocessor would make calculations to determine if the azimuth slew rate will become excessive and would store the information regarding that part of the pass. It would then calculate the initial azimuth angle and swing the antenna to that point on the horizon (0 degrees elevation). As this operation is being carried out, the microprocessor would check the clock and display on the screen:

READY OSCAR 9 MODE B
T MINUS 8:43 AZ 163 EL 0

Following this, the computer would update the T time **(TIME TILL START)** every second as it counted down from 8 minutes, **43** seconds. You'd then be free until T minus 15 seconds, at which time the computer would begin emitting a persistent "beep-beep-beep" to alert you of the approaching satellite acquisition. At T minus zero the beeping would stop and the screen would flash:

TRACKING OSCAR 9 MODE B 21:00 REMAINING
AT T PLUS :00 AZ 163 EL 0

As the pass progresses the computer would count down the time remaining while counting up the T time and would continually update the azimuth and elevation numbers so that you'd know where the antenna is pointing at any given moment. A few minutes before the satellite reaches the point where the slew rate becomes excessive, the microprocessor would instruct the antenna to begin leading the satellite by a few degrees. The pointing error will therefore be kept to a minimum, and a few seconds later the antenna will be right back on target. When the pass ends, the screen flashes:

PASS COMPLETE AT T PLUS 21:W AZ 354 EL 0

At this point you can shut down the station if you're finished. Or, if you wish to try another pass, you can enter **NEXT PASS** to obtain an updated list of upcoming passes. Perhaps you're interested in what's available the next day. In that case you'd enter **ALL PASSES 29 OCT** to obtain a complete list. Or maybe you're interested only in Mode J for the next afternoon, so you'd type in **MODE J PASSES 29 OCT PM** to obtain a list of only the passes in which you're interested.

An interrupt capability would be incorporated into the software so you can stop in the middle of a pass if desired. For example, during the OSCAR 9 pass described above, suppose you communicate for ten minutes, then decide to try a different satellite. You'd type in **STOP,** which halts the tracking of OSCAR 9, followed by **NEXT PASS RS4,** to which the computer replies:

28 OCT 78 16:54 UTC

SATELLITE	MODE	NEXT PASS	TIME TILL START
RS4	A	16:50	IP

LENGTH	HIGHEST ELEVATION	DIR	PRIMARY COVERAGE
14	76	D	OMNI

PREPARE TO TRACK?

This tells you that the pass for the Soviet satellite RS4 is in progress (IP), having begun at 16:50, with 14 minutes remaining in the pass. Furthermore, it's a descending Mode A pass with a fairly high elevation angle and more or less omnidirectional coverage. In reply to the computer's question, you might type in **YES.**

The computer would immediately bring the antenna to bear on the satellite at its current position in the sky and begin to track RS4 as it flashed:

TRACKING RS4 MODE A 14:00 REMAINING AT
T PLUS 04:00 AZ 328 EL 11

The possibilities are unlimited, but the examples we have looked at should give you a good idea of the convenience that appropriately designed software can bring to a station that has a microprocessor-controlled satellite tracking system. You have a free hand in programming all the features that suit your situation.

summary

We began by discussing problems encountered in satellite tracking and described how some of them can cause serious pointing errors in both manually and automatically controlled systems if corrective measures are not taken. A number of approaches were covered for improving accuracy. A major concern is that, if at all possible, we reduce the workload on the operator (who would prefer to spend his time communicating rather than operating antenna controls), and we have presented some ways to do just that. Finally, we looked at some special problems associated with microprocessor-controlled systems and discussed some software features that should be included in the finished system.

We concluded by looking at an example of an operational procedure for a microprocessor-controlled tracking system. Here we discussed how the computer could provide detailed data on various satellites as well as controlling the tracking antenna in a convenient and accurate manner.

A number of hams have incorporated many of these features into their stations. The sky's the limit when you begin to work on your own system.

ham radio

diode noise source

for receiver noise measurements

Construction of a temperature-limited diode noise source for accurate automatic and manual noise-figure measurements to 500 MHz

The saturated temperature-limited thermionic diode has been used extensively for measurement of receiver noise figures in the high-frequency, vhf, and low uhf regions. Its characteristics are predictable and repeatable, and it may be used either in conjunction with an automatic noise-figure meter or with its own power supply and indicating meter for manual noise-figure measurements.

A recent article[1] described the use of diode noise sources with automatic noise-figure meters and indicated that many homebuilt sources could be used

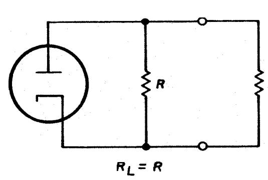


fig. 1. Basic circuit of a temperature-limited diode noise source.

with the Hewlett-Packard models 340A, W B , and 342A. The article also stated that most, if not all, of these noise sources were unsatisfactory because of their high VSWR at frequencies above 250 MHz or so. Described here is a noise source, using the Sylvania 5722 diode, which is patterned after the Hewlett-Packard Model 343A VHF Noise Source, and which appears to be comparable to that commercial unit at frequencies up to at least 450 MHz.

The noise source may also be used to make manual noise-figure measurements by either the twice-power or Y-factor method. Either technique requires a fixed plate supply and a variable filament supply, with an appropriate plate-current meter. In the past it has been normal practice to vary the diode filament voltage by means of a small variable auto-transformer or a power rheostat. Such control devices change voltage in discrete, albeit small, steps. This, coupled with voltage changes in the primary power source, make it difficult to establish and then to maintain the desired diode plate current. W6GXN described an improved power supply which minimized the problem.2 This article will present an updated version of his approach, using modern solid-state techniques.

diode noise source

To understand how a temperature-limited thermionic diode is used as a noise generator, we must start with the basic concepts of noise power. A resistance at a temperature other than absolute zero generates across its open-circuit terminals a voltage which is caused by the random motion of free electrons. This noise voltage, e_n, is infinitely broadbanded and defined by the equation

$$e_n^2/B = 4kTR \text{ volts}^2/\text{unit frequency bandwidth} \quad (1)$$

where

k = Boltzmann's constant 1.374×10^{-23} joule/°K
T = absolute temperature in °K
R = resistance in ohms
B = bandwidth, in Hz, of device under test

Since our treatment of noise deals with receivers or amplifiers of finite bandwidth, **eq. 1** is usually written as

$$e_n^2 = 4kTRB \quad (2)$$

When resistance R is connected to a matched load resistance, R_L (equal to R), maximum transfer of noise power will result. The noise power, P_n, dissipated in the load will be

$$P_n = \frac{e_n^2}{4R} = \frac{4kTRB}{4R} = kTB \quad (3)$$

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Eq. 3 defines the *available noise power* from resistance R.

Although its derivation is beyond the scope of this article, effective input noise temperature, T_e, is defined as

$$T_e = \frac{T_{ih} - YT_{ic}}{Y-1} \qquad \textbf{(4)}$$

where

T_{ih} = hot input noise temperature in $^\circ K$
T_{ic} = cold input noise temperature in $^\circ K$
Y = ratio of hot output noise power to cold output noise power

In the case of a temperature-limited thermionic diode (**fig. 1**), shot-noise current, i_n, can be determined from the equation

$$i_n^2 = 2qIB \qquad \textbf{(5)}$$

where

q = electronic charge, 1.600×10^{-19} coulomb
I = dc plate current in amperes
B = bandwidth, in Hz, of device under test

The total available noise power, P_n, from a diode noise source is the sum of the diode shot-noise power and the terminating resistor noise power. From **eqs.** 3 and 5,

$$P_n = \frac{R}{4}(2qIB) + kTB \qquad \textbf{(6)}$$

This equation can be factored and rearranged as

$$P_n = kB\left[(f)\left(\frac{IR}{2}\right) + T\right] \qquad \textbf{(7)}$$

Since

$$\frac{q}{k} = \frac{1.600 \times 10^{-19}}{1.374 \times 10^{-23}} = 11644.8 \qquad \textbf{(8)}$$

then

$$P_n = kB(5822IR + T) \qquad \textbf{(9)}$$

If, in **eq.** 3, the temperature is T_{ih}, then

$$kT_{ih}B = kB(5822IR + T) \qquad \textbf{(10)}$$

or

$$T_{ih} = 5822IR + T \qquad \textbf{(11)}$$

where T_{ih} is the noise temperature of the noise diode with its load resistance, R, at a temperature of T. Thus when I is zero, T_{ih} is equal to T.

The excess noise ratio, ENR, is defined as the ratio of the available noise power at temperature T in excess of that available at a standard temperature (T_o) to the available noise power at T_o, and is expressed as

$$ENR = \frac{kTB - kT_oB}{kT_oB} = \frac{T - T_o}{T_o} \qquad \textbf{(12)}$$

At a standard reference temperature of $290^\circ K$,

$$ENR = \frac{T-290}{290} = \frac{T}{290} - 1 \qquad \textbf{(13)}$$

$$\text{or} \quad ENR = \frac{T_{ih}}{290} - 1 = \frac{5822IR + T}{290} - I \qquad \textbf{(14)}$$

$$= 20.08IR + \frac{T}{290} - 1$$

Since the term $\frac{T}{290}$ is the noise power contribution of the load resistor, if a temperature of $290^\circ K$ is used for the load, **eq. 14** reduces to

$$ENR = 20.08IR \qquad \textbf{(15)}$$

If R equals 50 ohms and I is expressed in *milliamperes,*

$$ENR \doteq I \qquad \textbf{(16)}$$

$$\text{and} \quad ENR_{dB} \doteq 10 \log I \qquad \textbf{(17)}$$

Note that this commonly used equation is predicated on the temperature of the load resistance being 290°K (17°C or 62.6°F). In actual practice, the load resistance temperature may be as high as 310°K (37°C or 98.6°F). In this case, the excess noise ratio will be lower, yielding receiver noise-figure measurements which are higher than the true noise figure. The corrections for these errors are plotted in **fig. 2.**

A schematic diagram of the actual temperature-limited diode noise source is shown in **fig.** 3. You will note that the output circuit is considerably more complicated than those used in most of the previously published designs.[2-5] It is this output circuit, properly arranged physically, which makes this noise

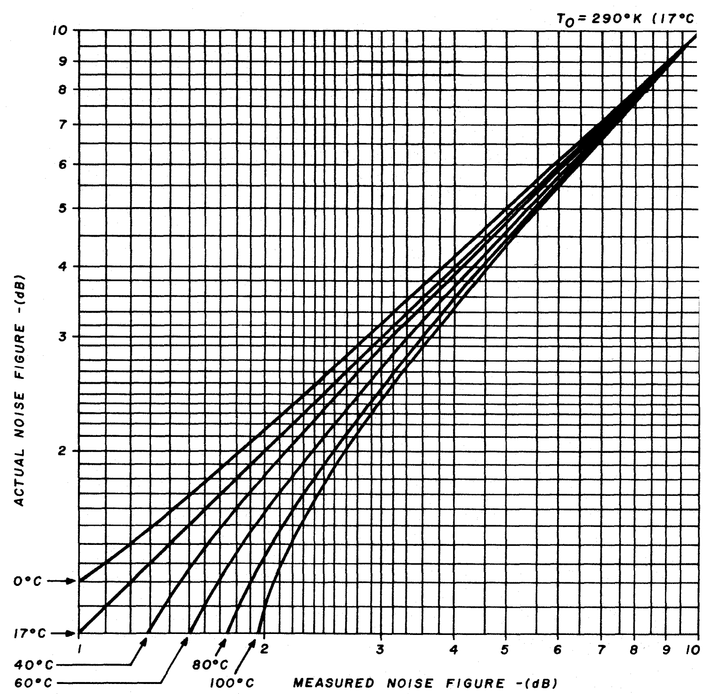

fig. 2. Temperature corrections for a temperature-limited diode noise source (courtesy Hewlett-Packard Company).

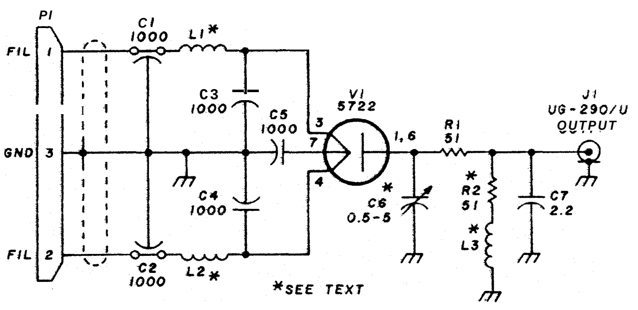

fig. 3. Schematic diagram of the diode noise source. Pin connections shown at P1 are those for a Cannon WK-5-22C-5/16 connector used to mate with a Hewlett-Packard 340B, 342A, or modified 340A Automatic Noise Figure Meter.

source comparable to the Hewlett-Packard model 343A.

The diode output circuit must theoretically satisfy two requirements: it must present a 50-ohm impedance (1:1 VSWR! to the receiver connected to the output connector, and it must present a load to the diode which results in a constant *ENR* over the usable frequency range. In practice, however, it appears that these two conditions cannot be satisfied concurrently, and, as in all designs, a compromise must be reached. The compromise in this case is to keep the VSWR as low as possible and to permit the *ENR* to change over a portion of the frequency range. The rationale for this is simply one of a known factor versus an unknown. If the VSWR is other than 1:1, the mismatch loss between the noise source and the receiver will be indeterminate. (Although noise power obeys ail power-transfer laws, noise is random in phase; therefore the loss is ambiguous rather than known.) On the other hand, it is possible to determine the *ENR*, although not to any precise degree of accuracy, so that measurements using a known value are possible.

An expanded schematic of the diode output circuit appears in **fig. 4.** The numbered components correspond to those shown in **fig.** 3, while those having letter subscripts represent the distributed reactive components, as follows. C_{pf} is the plate-to-filament capacitance of the tube; L_p is the series inductance of the tube structure and tube pins; L_s is the series inductance present in R1 and the plate pin connectors; and C_s is the shunt capacitance of the tube socket and/or C6.

If R1 and L3 are both replaced by shorts, and C7 is removed from the circuit, there will be a steep rise in the *ENR* to a resonant point, as shown on curve **A** of **fig. 5.** The VSWR at 432 MHz with this configuration will be about 3.5:1. K2PEY, in his circuit,[3] damped out this resonance by adding a 51-ohm series resistor (R1), resulting in curve B. However, the VSWR was still approximately 1.5:1 at 420 MHz according to his article, and about 1.8:1 at 432 MHz by measurement.

W8BBB modified the circuit further by adding L3 in series with R2. Since his article provided no figures on *ENR* or VSWR,[4] there is no basis for its comparision with the earlier configurations or with the one presented at this time.

Incorporating R1, L3, and C7 in the circuit results in curve **C** of **fig. 5**. *it* can be seen that the *ENR* has *a* rising characteristic with frequency, but the VSWR is less than 1.2:1 at 432 MHz. Thus, we have achieved a low VSWR, but at the cost of having an *ENR* which changes with frequency. As previously stated, however, we know the *ENR to* a fair degree of accuracy and we have minimized the indeterminate mismatch loss.

The noise source is enclosed in an 83 x 54 x 41 mm (3-1/4 x 2-1/8 x 1-518 inch) aluminum utility box. Two methods of mounting and making connections to the 5722 noise diode are presented, both of which have proved to be equally usable.

If suitable VSWR-measuring equipment is available (either a network analyzer or a slotted line and SWR indicator), the method shown in **fig. 6** is preferable, since it permits optimization of the noise-source VSWR by adjustment of capacitor C6. The tube is mounted by means of a 19-mm (314-inch) diameter wrap-around plastic clamp positioned so that the cable tube pins pass through a 16-mm (518-inch) diameter hole in the shield plate. The cable clamp is attached to the enclosure by means of a metal standoff post. Connections to the tube are made via contacts which have been removed from a 7- or 9-pin miniature tube socket and slipped over the appropriate tube pins.

If you are unable to measure the VSWR of the noise source, use a standard 7-pin, saddle-mount, mica-filled phenolic tube socket, from which the center ground post and the contacts for pins 2 and 5 have been removed; mount the socket on the shield plate. Do not use a black Bakelite or ceramic socket, or one which has a shield base. The added capacitance of the mica-filled phenolic socket is just about optimum at 432 MHz, so capacitor C6 can be omitted.

Other than the mounting and the elimination of C6, the socket method of construction follows that

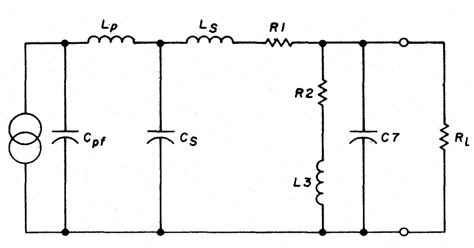

fig. 4. Expanded schematic diagram of the diode-source output circuit. Numbered components correspond to those in fig. 3. Components with letter subscripts are explained in the text.

shown in fig. *6.* In either case, mount the tube so that pins 3 and 4 are in a plane perpendicular to the bottom of the housing. To reduce the plate lead inductance, the contacts for pins 1 and 6 should be bent at right angles toward one another and soldered together. The usual vhf wiring techniques are required. Disc capacitors C3, C4, and C5 must be soldered directly between the tube pin contacts and the mounting plate, with lead lengths at an absolute minimum. Rf chokes L1 and L2 are each ten turns of no. 26 (0.4-mm) or no. 28 (0.3-mm) enamel-covered wire, close spaced and air wound to approximately 0.1 inch (2.5 mm) diameter.

Although R2 is shown as a nominal 51-ohm resistor, its optimum value is 50 ohms. It is suggested that a resistor as close as possible to 50 ohms (measured on a resistance bridge or accurate digital multimeter) be selected. Five per cent, 114-watt *carbon-film* resistors are recommended for both R1 and R2; the application of heat when soldering to the necessary short leads of ordinary composition resistors generally changes their values.

One lead of R2 is wound to form L3, a three-turn coil 2.5 mm (0.1 inch) in diameter and 5 mm (0.2 inch) long. Except for this lead, all other connections to the plate of V1 or to J1 must be made virtually leadless. C7, R21L3, and C6 (if used) are all grounded to a solder lug attached to one of the mounting screws for J1.

A shielded, two-wire cable, brought out of the housing through a neoprene grommet, is used to connect the noise source to either a Hewlett-Packard automatic noise-figure meter or to the noise-source power supply unit. If the noise source is to be used with a noise-figure meter, the appropriate pin connections to a Cannon WK-5-22C-5/16 connector are shown in **fig.** 3. Otherwise, any convenient pair of mating connectors can be used on the noise-source cable and the power supply. Obviously, the male of the pair should be attached to the cable so that high voltage is not exposed on the power-supply connector.

The 5722 diode generates a considerable amount of heat, which can affect the resistance values of R1 and R2. To minimize the heat within the housing, discard the U-shaped part of the utility box that mates with the half shown in **fig. 6.** In its place, use a similar piece made from perforated aluminum stock. This will retain the integrity of the shielding and will allow heat convection from within the enclosure.

If VSWR-measuring equipment is available and C6 has been included in the circuit, the VSWR of the noise source can be optimized. This can be done without applying filament or plate power to the diode. **Table 1** shows the VSWR of several configurations of the noise source, with and without a tube socket. The frequency at which C6 should be adjusted depends on your preference. Adjusting C6 for minimum VSWR at 432 MHz results in a VSWR of less than 1.2:1 over the entire usable range of the noise source. If the VSWR adjustment is made at 222 MHz, the VSWR increases to slightly less than 1.3:1 at 432

CURVE	R1	L3	C7	VSWR AT 432 MHz
A	0	0	∞	3.5:1
B	51	0	∞	1.8:1
C	51	≈10nH	2.2pF	< 1.2:1

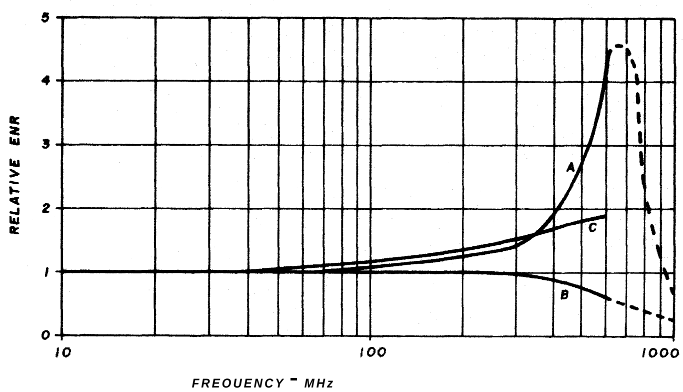

fig. 5. Relative *ENR* of a diode noise source for various configurations of the output circuit shown in fig. 4. Ideally, the *ENR* should be unity over the entire frequency range. Curves A and B have been calculated, thus the dashed portions above the maximum usable frequency of 600 MHz. Curve C is representative of the Hewlett-Packard model 343A and of the homebuilt noise source up to at least 450 MHz.

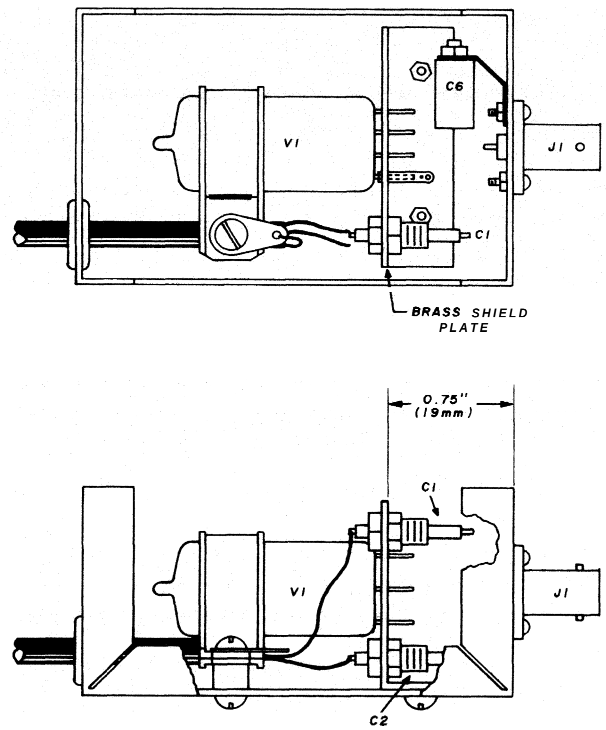

fig. 6. Physical layout of the noise source. The shield plate on which C1 and C2 are mounted is made of brass with a 16-mm (518-inch) diameter hole to clear the tube pins or to accommodate a 7-pin miniature tube socket (see text).

MHz and is only 1.1:1 at 145 MHz. Since noise-figure measurements at 432 MHz would appear to be of greater import than at other frequencies within the range of the noise source, it is recommended that the VSWR be optimized at that frequency. When so adjusted, the VSWR at the other amateur frequencies appears to be at least equal to that of the Hewlett-Packard model 343A.

For those amateurs interested in tweaking to the diode noise source for manual noise-figure measurements appears in **fig. 7.** It consists of an adjustable, regulated switching supply for the filament of the 5722 noise diode, supplied by transformer T1, and a simple half-wave rectified dc plate supply. High voltage is obtained from T2, a 12.6-volt transformer which has its low-voltage winding connected to the secondary of T1, thereby providing approximately 115 volts ac for the plate supply.

table 1. VSWR **measurements for several configurations of diode noise sources. See text and fig. 3 for description of configurations (second column).**

		VSWR						
unit	configuration	30 MHz	50 MHz	145 MHz	222 MHz	432 MHz	500 MHz	550 MHz
1	no socket, C6 and L3 optimized at 432 MHz	1.04	1.07	1.12	1.15	1.03	1.06	1.14
2	no socket, C6 and L3 optimized at 432 MHz	1.05	1.09	1.20	1.19	1.03	1.12	1.23
2	no socket, C6 optimized at 300 MHz	1.05	1.06	1.08	1.08	1.16	1.27	1.37
2	no socket, C6 optimized at 222 MHz	1.04	1.04	1.04	1.04	1.29	1.41	1.51
3	with socket, C6 omitted, C7 optimized at 432 MHz	1.04	1.06	1.09	1.14	1.05	1.15	1.25
4	with socket, C6 omitted, C7 optimized at 432 MHz	1.06	1.10	1.18	1.22	1.04	1.12	1.23
5	with socket, C6 omitted, no optimization	1.07	1.13	1.23	1.28	1.14	1.21	1.33
6	with socket, C6 omitted, C7 optimizedat432 MHz	1.05	1.08	1.13	1.17	1.09	1.23	1.34
	Hewlett-Packard 343A	1.04	1.05	1.08	1.06	1.09	1.14	1.17
	Hewlett-Packard 343A	1.02	1.03	1.05	1.06	1.12	1.18	1.23
	Hewlett-Packard 343A	1.03	1.05	1.07	1.08	1.17	1.16	1.13

ultimate, the inductance of L3 can be varied slightly by stretching or compressing the turns to minimize the VSWR. There will be some interaction between the inductance of L3 and the capacitance of C6, so that one will have to be readjusted if the other is changed. Alternatively, a trimmer capacitor can be substituted for C7 and its capacitance varied, in lieu of adjusting L3, to achieve the same results. Again, there will be interaction with the setting of C6, necessitating alternate adjustments of both capacitors.

It can be seen from the figures in **table 1** that even without VSWR-measuring equipment, careful construction should result in a noise source which is as good as, and probably better than, any which have been previously built (or described) by amateurs.

The final VSWR must be measured with both halves of the enclosure assembled. If the VSWR increases when the cover half is in position, it will be necessary to drill an access hole in it so that the trimmercapacitorcanbe adjusted with the cover in place.

power supply

The power supply circuit which is used with the

The object of the power supply is to establish a constant diode plate current by controlling the filament current and hence, the filament temperature. To accomplish this, op amp U2 is used to drive Q1, a Motorola MJE801 Darlington pair, which supplies current to the 5722 filament. A reference voltage is established at the inverting input of U2, derived from the regulated 5-volt output of U1 and set by R5, the ***diode current*** control. The diode plate-current return is through resistor R4 connected to the output of U1; a portion of the voltage drop across the resistor is applied to the noninverting input of the op amp by means of the voltage divider formed by resistors R7 and R8.

Once the diode plate current has been set by means of R5, any change in plate current will result in a change in voltage at the noninverting input of U2. Since the sensing circuit is in the negative return of the power supply, an increase in current will cause the noninverting input to become more negative. The resultant decrease in the output of the op amp will lower the base current of Q1, thereby decreasing the diode filament current to compensate for the increase in plate current. Conversely, if the diode plate current tries to fall, the noninverting input of U2 will become more positive. This will increase the output of U2, increasing the base current of Q1, and raising the diode filament current to negate the decrease in plate current.

The thermal inertia of the diode filament introduces a considerable time lag between the time that an error voltage is detected and the time that the filament current is compensated. Therefore the op amp, instead of operating in a linear mode, functions more like a comparator in that its output switches between a level equal to the voltage on pin 4 and about 1.5 volts less than the voltage on pin 8. The op amp output and the diode filament current are a series of rectangular pulses whose frequency varies between approximately 500 and 1000 Hz and whose duty cycle also varies; both the frequency and duty cycle depend on the setting of the *diode current* pot.

Zener diode CR6, connected between the base of Q1 and the diode filament return, has been incorporated to clamp the base voltage, and hence the diode filament current, to a safe value should there be any failure in the regulating circuit. Since the peak-to-peak amplitude of the voltage at the base of Q1 must be very close to 6 volts in order to obtain 20 to 25 mA diode plate current, a five per cent zener diode has been specified to preclude clamping at too low a voltage.

The positive side of the diode plate supply is returned to ground through a 1-milliamp meter and a selected shunt. Three usable ranges, of 3, 10, and 30 mA, have been incorporated; the maximum diode current is between 20 and 25 mA with the circuit shown. The values of the meter shunts (R12, R13, and R14) will depend on the internal resistance of the meter you actually use. R12 must be equal to the meter resistance divided by two, R13 to the meter resistance divided by nine, and R14 to the meter resistance divided by twenty-nine.

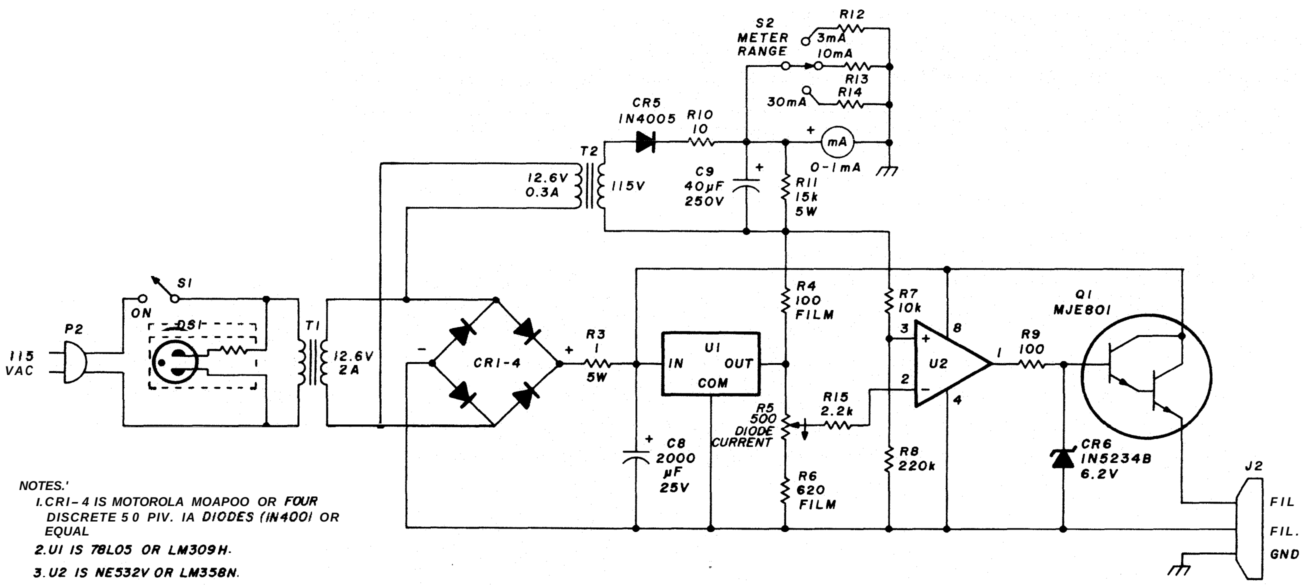

fig. 7. Schematic diagram of the noise-source power supply. The values of R12, R13, and R14 are discussed in the text. Film resistors may be deposited carbon or metal. The *meter range* switch must be a shorting type to protect the meter when the switch is rotated.

power supply construction

The power supply can be housed in any convenient enclosure and can be built in the same way as any conventional power supply. Only a few precautions need be observed. The negative terminal of plate-supply filter capacitor C9 should be returned directly to the transformer winding, and the negative side of the filament-supply filter capacitor C8 should be connected directly to the negative terminal of the bridge rectifier. Otherwise excessive ripple can appear in the outputs.

Transistor Q1 must be mounted on a heatsink which is insulated from chassis ground, or it must be insulated from a grounded heatsink by means of a mica or plastic insulator. If chassis-type construction is used, the chassis can serve as the heatsink. Otherwise, a heatsink having a radiating area of approximately 65 square cm (10 square inches) should be satisfactory. Be sure to apply a thin coating of silicone heatsink compound to the heatsink side of the transistor, and to the insulator if one is used.

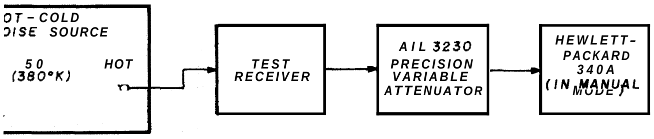

fig. 8. Test setup used to determine the test-receiver noise figure by means of a hot-cold noise source.

As mentioned previously in the discussion of the noise source, J2 can be any female connector which will mate with the noise-source cable connector. Inexpensive types, such as the Amphenol 91-MPM3L for the noise-source cable and Amphenol 78-PCG3 for the power supply, are more than adequate.

A word about the meter is in order. Since the accuracy of the noise-figure measurement is a function of the accuracy with which the diode current is measured, a high-quality meter should be used. The accuracy of shunt resistor R12 is equally, if not more, important. (R13 and R14 are of lesser importance, since the accuracy of noise-figure measurements on these ranges is generally not critical.) If possible, the accuracy and tracking of the 3-mA meter range should be checked against a milliammeter of known accuracy.

One way of trimming R12 to the proper value is to use a one per cent metal-film resistor whose resistance is the closest value higher than one-half the meter resistance. Then shunt the metal-film resistor with higher-value composition resistors (starting at about ten times the one per cent value) until the meter reading is correct.

After double checking the circuit, connect the noise source to the power supply. Set the *diode current* control for minimum current and set the *meter range* switch to its *30-mA* position. Energize the power supply and slowly increase the diode plate current by means of the *diode current* control. The current should increase smoothly from zero to a maximum of between 20 and 25 mA. If a plate current of at least 20 mA cannot be reached, it is likely that CR6 is limiting the base voltage of Q1. Reduce the plate current to zero, turn off the power supply, and disconnect one side of CR6. Then turn on the power supply again and increase the diode plate current to its maximum. If the plate current is now higher than could be obtained with CR6 connected, another zener diode should be used which has a slightly higher zener breakdown voltage. Alternatively, the existing zener voltage can be increased by about 0.6 volt by connecting an ordinary silicon diode in series with CR6; the diodes should be connected anode-to-anode.

If current limiting is not being caused by the zener diode, check the various voltages throughout the power supply. They should approximate the values listed below. Except for the output of U1, a 20 per cent variance in voltages can be expected because of component tolerances, especially in the transformers.

	voltage measurement at	
measure across	**1 mA plate current**	**20 mA plate current**
T1 secondary	13.3 Vac	12.4 Vac
T2 secondary	113 Vac	88 Vac
R3	0.9 Vdc	1.5 Vdc
C8	13 Vdc	10 Vdc
U1 output (to common)	5.0 Vdc	5.0 Vdc
CR6	6.2 Vp-p	6.2 Vp-p
V1 filament	4.4 vp-p	4.7 vp-p
R11	123 Vdc	72 Vdc

The voltages across CR6 and the filament of V1 must be measured using an oscilloscope because they consist of rectangular pulse trains. Use extreme care when making such measurements, since the scope ground connection will be elevated from the power-supply chassis ground by the plate-supply voltage.

The two characteristics of the noise source which required evaluation were the VSWR and the *ENR*. The former presented no problem, since suitable VSWR-measurement equipment was available. **Table 1** shows the VSWR of six of the homebuilt noise sources at various frequencies between 30 and 550 MHz. For comparison purposes, the measured VSWR of three Hewlett-Packard model 343A noise sources are also included. It can be seen that the comparison is favorable, especially at the amateur band frequencies of 432 MHz and below.

Measuring the excess noise ratio is another matter, however. Short of sending a unit to the National Bureau of Standards, or equipping a primary standards laboratory, there is no way to measure the noise output quantitatively. As derived earlier in this article, the theoretical *ENR* of a temperature-limited diode, when terminated by a resistive load at the standard reference temperature of 290°K, is nearly equal to 10 $\log I$ in a 50-ohm system, where I is the diode plate current in milliamperes.

While we might accept this expression as an abso-

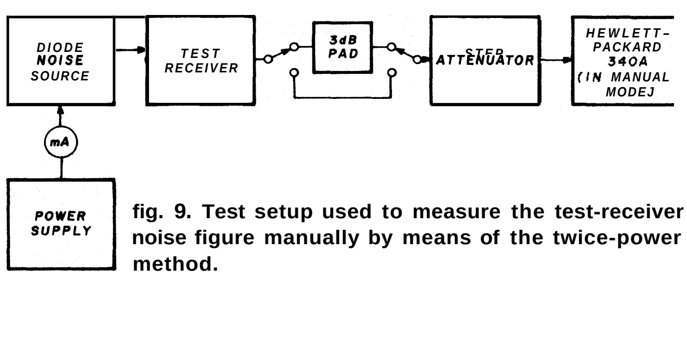

fig. 9. Test setup used to measure the test-receiver noise figure manually by means of the twice-power method.

lute value at frequencies below 30 MHz, we know from experience and from Hewlett-Packard's specifications for the 343A noise source that the *ENR* gradually increases above 30 MHz to the limit of the source's frequency range. (This frequency limitation results from the transit time of the electrons passing from cathode to anode becoming an appreciable part of the period at high frequencies, and from the series inductances shown in **fig. 4**.)

Since there was no method by which the actual noise output of the diode could be measured directly, we resorted to measurement by transfer. While this is a far from ideal technique, it was the only feasibie method, and provided us with usable data. The transfer method used was as follows. A hot-cold noise source and a precision variable attenuator were used in conjunction with a Hewlett-Packard noise-figure meter (in its manual mode as an indicator). The test setup is shown in **fig. 8**.

The noise source comprised two 50-ohm terminations, one immersed in liquid nitrogen (boiling point at 77.3°K) and one housed in a thermostatically controlled oven (temperature at 380°K). The VSWR of each termination was measured through the coaxial switch and found to be less than 1.02:1. The noise figures of four receivers, one each at 28, 144, 220, and 432 MHz, were determined by means of the test setup shown, taking into account the measured insertion loss of the coaxial switch (approximately 0.05 dB at 432 MHz, and insignificant at lower frequencies). These noise figures provided the reference for the transfer measurement.

The noise figures of the same receivers were then remeasured, using six different homebuilt noise sources and two Hewlett-Packard 343A noise sources. In each case, the noise figure was determined manually, by means of the twice-power method, and automatically, using a Hewlett-Packard model 340A Automatic Noise Figure Meter; the test setups for these measurements appear in **figs. 9** and **10.**

The results of these noise-figure measurements were analyzed and plotted as errors in the *ENR* of each noise source, compared with the *theoretical ENR* of a temperature-limited diode. The results are shown in **fig. 11,** along with the nominal and specified limits of the Hewlett-Packard model 343A *ENR*

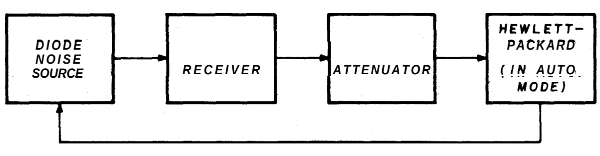

fig. 10. Test setup used to measure the test-receiver noise figure by means of an automatic noise-figure meter.

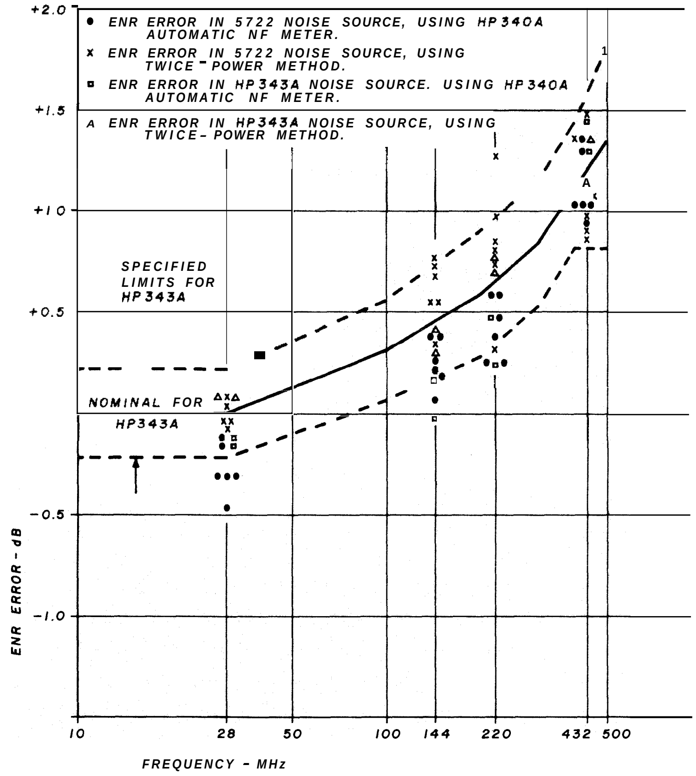

fig. 11. Errors in excess noise ratios of six homebuilt 5722 noise sources and two Hewlett-Packard model 343A noise sources, compared to theoretical noise power. Errors were determined by comparison with receiver noise figures measured using a hot-cold noise source. A positive error indicates output noise power in excess of the theoretical noise power; therefore a measured noise figure must be increased by the indicated error.

error. It must be reiterated that the *ENR* error is the deviation from the theoretical low-frequency *ENR,* and is a normal and expected deviation. Our tests were made to determine the magnitude of the deviation.

Analyzing the results of these measurements proved to be somewhat more difficult than performing the tests. At 28 MHz we expected the *ENRs* to be very close to the theoretical value. This proved to be the case for the manual measurements, but the automatic noise-figure measurements were generally on the high side, indicating a lower *ENR* than expected. The conclusion was that the error was in the automatic noise-figure meter, which is specified as having a possible error of ±0.5 dB. By way of comparison, the milliammeter shown in **fig. 9** was an accurate digital meter, the loss of the 3-dB pad had been checked at 30 MHz, and the noise-figure meter was used only as a reference indicator, so that any error in that instrument was eliminated by using a fixed meter reference reading. The trend of low readings in the automatic mode is continued at 144 and

220 MHz but not at 432 MHz, where both types of measurements yielded closely related results.

Although one of the homebuilt noise sources produced an *ENR* well outside of the expected range at 220 MHz (for some unexplainable reason), the results seem to indicate that a homebuilt noise source can be constructed and can be expected to provide an lett-Packard automatic noise-figure meters are calibrated, it follows that the ***ENR*** may be reduced to 5.2 dB by reducing the diode plate current. The required value of current plotted against frequency is shown in **fig. 12,** and is included in **table 2.** By setting the diode current to the value indicated for the frequency of measurement, no *ENR* correction need

table 2. Nominal corrections for temperature-limited diode noise sources at Amateur frequencies.

frequency (MHz)	ENR accuracy (dB)	*NF* correction (dB)	ENR (dB) at 3.31 mA	automatic noise-figure meter diode current (mA) to compensate for *NF* correction
28	±0.20	0	5.20	3.31
50	±0.23	+0.13	5.33	3.21
144	±0.28	+0.45	5.65	2.99
220	±0.30	+0.65	5.85	2.85
432	±0.38	+1.17	6.37	2.53

ENR within the range specified by Hewlett-Packard for their model 343A. No attempt was made to apply a temperature or mismatch-loss correction, since the intent was a comparison between the homebuilt noise source and the Hewlett-Packard 343A. Both types of noise sources were subjected to the same possible receiver mismatch and were operated under identical environmental conditions.

noise-figure measurements

As previously stated, the noise source can be used in conjunction with its own power supply to make noise-figure measurements, or it may be used with a Hewlett-Packard model 340B or 342A Automatic Noise Figure Meter. It may also be used with a modified Hewlett-Packard model 340A, as described in reference 1. The techniques employed in automatic noise-figure measurements are treated briefly in that article, and in detail in the Hewlett-Packard manuals covering such equipment.[6-9] The homebuilt noise source, when equipped with the appropriate power connector, can be considered as a direct replacement for the Hewlett-Packard model 343A at frequencies to at least 450 MHz.

Table 2 summarizes, for the amateur frequencies of interest, the nominal noise-figure correction which must be added to the measured noise figure because of the noise-source *ENR* increase with frequency, as shown in **fig. 11.** The table also includes the accuracy of the *ENR,* based on Hewlett-Packard's specifications for the model 343A noise source and the measured equivalence of the homebuilt versions. Thus, this is the uncertainty of the corrected noise figure due to *ENR* only, but does not include other errors discussed in reference 1.

Since the *ENR* is higher, at frequencies over 30 MHz, than the nominal 5.2-dB value for which Hewlett-Packard automatic noise-figure meters are calibrated, ... be applied to the noise-figure reading. Manual noise-figure measurements may also be made in conjunction with an automatic noise-figure meter, as described in the previously referenced manuals.

Manual measurements, using the power supply described in this article, can be made in one of two ways: the twice-power method and the Y-factor method. The former is the one most familiar to most amateurs, and will be covered first. **Fig. 13** shows four configurations, in order of preference, for the twice-power noise-figure measurement. In each diagram, the "receiver under test" means any receiver or portion thereof (such as a converter or mixer) which provides an output at either an intermediate or audio frequency. If **AGC** is incorporated in the receiver, the **AGC** should be disabled and the rf gain control

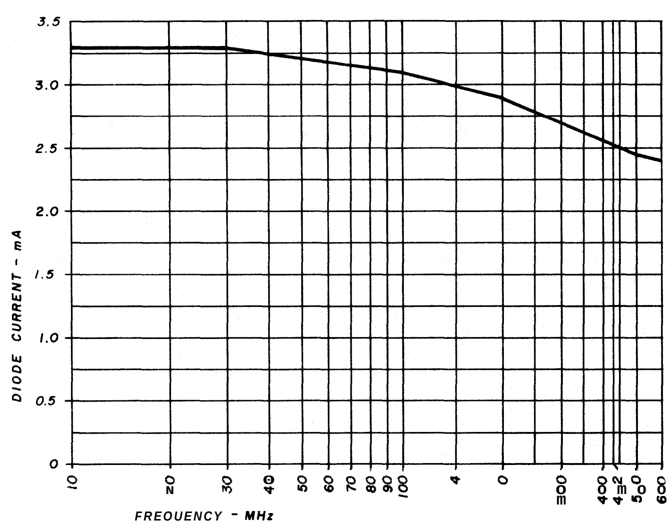

fig. 12. Diode current required by a temperature-limited diode noise source to maintain a nominally constant 5.2-dB excess noise ratio (curve by J. R. Reisert. W1JR).

set so that the receiver is not overloaded by the noise input. There should be a direct connection between the noise-source output connector and the loss pad, and between the pad and the receiver. This means no ***cables,*** and a minimum of adapters.

The use of a loss pad with a diode noise source is not absolutely essential, but can minimize several problems. First of all, most receivers do not present a 50-ohm input when optimized for a noise match. Because the rated *ENR* of a noise source is based on a 50-ohm load, there will be an indeterminate mismatch loss if the receiver VSWR is greater than 1.0:1. A 3 to 6 dB pad will not eliminate the mismatch ioss, but may reduce it somewhat.

A second reason for using a loss pad is to ensure a 50-ohm source impedance for the receiver, since the VSWR of the noise source is not a perfect 1.0:1. Any tendency of the receiver to "take off" when looking into an impedance other than 50 ohms will be reduced by the use of a pad.

In **fig. 13A**, the i-f output of the receiver under test is connected to a video or rf voltmeter (depending on the receiver output frequency), which is used only as an indicator. To terminate the 3-dB pad properly and present a constant load impedance to the receiver, the voltmeter must have a 50-ohm input impedance.*

With the noise source off and the 3-dB pad out of the circuit, a reference reading is established on the voltmeter. The 3-dB pad is then inserted between the receiver and the voltmeter, the noise source is turned on, and the noise-source plate current is adjusted until the same voltmeter reference is obtained. The uncorrected noise figure (in dB) of the receiver plus the loss pad between the noise source and the receiver is equal to 10 *log I,* where *I* is the diode plate current in milliamperes. The uncorrected receiver noise figure is determined by subtracting the attenuation of the loss pad from the calculated noise figure. The receiver noise figure must then be corrected for frequency by adding the noise-figure correction listed in **table** 2 or, at frequencies not listed in the table, the nominal Hewlett-Packard 343A error shown in **fig. 11.** Note that the noise-figure accuracy is limited by the degree of uncertainty in the noise-source *ENR,* as indicated in **table 2** and **fig. 11.** Other uncertainties in the measurement will be discussed later.

The advantages of the circuit shown in **fig. 13A** are twofold: the measurement is independent of both the voltmeter calibration and any nonlinearity in the receiver detector. **Fig. 13B** shows a similar setup, except that the 3-dB increase in noise power depends on the voltmeter calibration. in this arrangement, the voltmeter ideally should be a true rms type so its readings are proportional to the square root of the power. It need not have 50-ohm input impedance, but should be calibrated in dB, although the latter feature is not absolutely essential. An average-responding voltmeter can be used with little loss in accuracy, but a peak-responding type should be avoided if at all possible.†

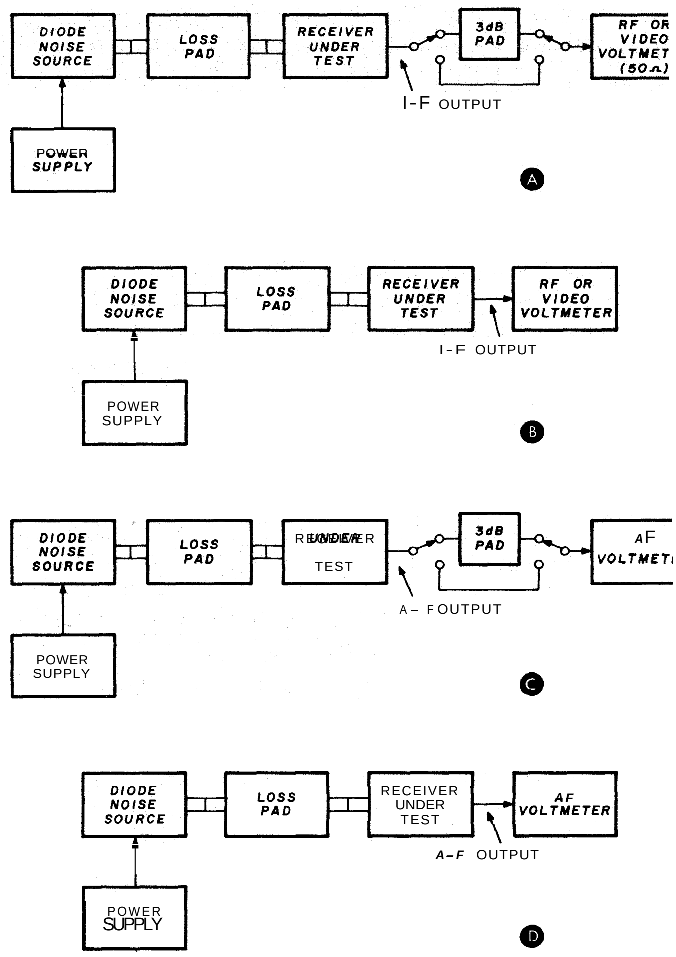

fig. 13. Four test configurations for measuring noise figure by the twice-power method, shown in order of preference. In A and C, the measurement is independent of the voltmeter calibration. In A and B, the measurement is independent of the linearity of the receiver detector.

*A high-impedance meter can be converted to 50 ohms by means of a 50-ohm, feed-through termination, such as the Heath SU-511-50, Tektronix 011-0049-01, Hewlett-Packard 10100C, or Systron-Donner 454.

†The terms "average-responding" and "peak-responding" refer to the voltmeter circuit, not the meter scale calibration. The Hewlett-Packard model 400D is a typical average-responding meter calibrated in rms volts. Voltmeters which employ rf probes, such as the Hewlett-Packard model 410B and the various Heath, Eico, and similar electronic voltmeters, are invariably peak-responding meters, with their meter scales also calibrated in rms volts. In all of these instruments, the rms meter calibration is based on a sinusoidal waveform, which is not valid for noise voltage.

When the arrangement of **fig. 13B** is used, a reference reading is established on the voltmeter with the noise source off. Then the noise source is turned on and the diode plate current is adjusted until the meter reading is exactly 3 dB greater than the reference reading. (If the voltmeter does not have a dB scale, the diode current should be adjusted until the meter reading is exactly 1.41 rimes the reference voltage.) The noise figure is determined as described for **fig. 13A.**

Figs. 13C and **13D** correspond to the test circuits just discussed except that the audio output of the receiver is measured by means of an audio-frequency voltmeter. Both of these circuit configurations depend on the linearity of the receiver detector; fortunately, most modern receivers use product detectors which are generally quite linear over a 3-dB range. The receiver beat-frequency oscillator must be on, of course, for the product detector to function.

Noise-figure measurements are made exactly as described for **figs. 13A** and **13B.** In **fig. 13C**, the impedance of the voltmeter need not be 50 ohms, but its impedance and that of the 3-dB pad must be the same. In some cases, it may not be convenient or even possible to disable the receiver AGC when making measurements in accordance with **figs. 13C** or **13D.** In view of the fact that the noise levels introduced are extremely low, the AGC in most, if not all, receivers will not be activated; so there is very little likelihood of any error occurring if the AGC cannot be disabled.

The less familiar Y-factor method of measuring noise figure requires a precision variable attenuator, as shown in **fig. 14.** The resolution of the attenuator should be at least 0.1 dB, and preferably 0.01 dB, and must be of known accuracy at the converter output frequency. With the noise source off, the precision attenuator is adjusted to obtain a convenient reference reading on the voltmeter. The noise-source plate current is then set to a value which corresponds

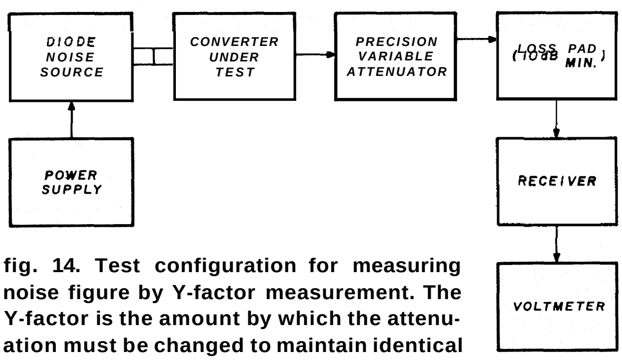

fig. 14. Test configuration for measuring noise figure by Y-factor measurement. The Y-factor is the amount by which the attenuation must be changed to maintain identical meter readings with the noise source off and on.

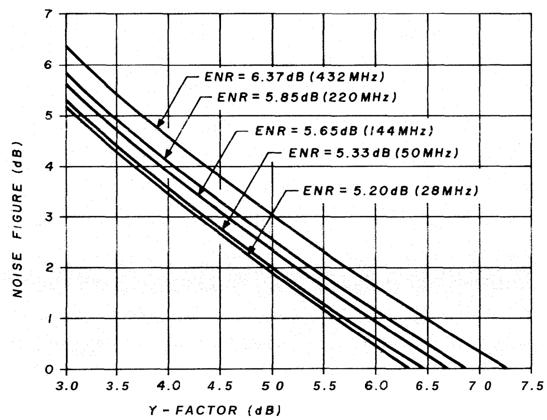

fig. 15. Noise figure plotted as a function of Y-factor for the nominal excess noise ratios of a temperature-limited diode operating at 3.31 mA plate current.

to a known ENR, *e.g.*, 3.31 mA, and the precision attenuator is readjusted to obtain the reference meter indication. The difference between the two attenuator settings, when converted from dB to the equivalent power ratio, is called the Y-factor, and is related to noise figure by the expression

$$NF = ENR - 10 \log (Y - 1) \qquad (18)$$

where both the noise figure (NF) and excess noise ratio (ENR) are in dB.* **Fig. 15** shows noise figure plotted against Y-factor (expressed in dB) for the nominal excess noise ratios obtained when the diode current is set to 3.31 mA. These curves obviously eliminate corrections because of differences in ENR at various frequencies, but are still subject to the uncertainty of the ENR tolerance.

There is a third method of manually measuring noise figure which is not recommended, but which should be mentioned because of its past appearance in some publications. It entails inserting a precision variable attenuator between the noise source and the receiver under test, then determining the noise figure from the ENR of the source and the calibration of the attenuator. This method is generally inaccurate because of variations in attenuator accuracy with frequency, and because the load impedance presented to the noise source by the attenuator-receiver combination may change as the attenuator setting is changed.

measurement errors

Some of the possible sources of error in noise-figure measurements are already apparent from the preceding discussion, especially that which is inherent in the noise source itself. However, there are addition-

*This relationship is derived in eqs. 1 through 9 of reference 1 al errors which must be considered and which are discussed in greater detail in reference 1. These errors comprise the following:

1. Noise-source accuracy, corrected for frequency

2. Noise-figure meter or power-supply milliammeter accuracy

3. Receiver image-response error

4. Temperature error

5. Mismatch error

If all errors are in dB. they accumulate additively. Therefore, the total measurement error will be the sum of the above. This is an imposing list, and could total well over 1.5 dB if all errors were of the same algebraic sign. However, many of these errors will cancel because of opposing signs, and generally the accuracy of commercial test equipment is better than the limits of its specifications. Nevertheless, these possibilities of error are very real and cannot be ignored except for ***comparative*** measurements using the same equipment at one particular time. And even then, the mismatch errors between the noise source and different receivers under test still exist.

acknowledgments

Bob Melvin, W6VSV, was responsible for both the noise-source analysis and the power-supply design, and must share credit for this article. Bryan Westfall, KGOJM, and Steve Mieth, K6YFK, provided equipment and invaluable assistance in making the hot-cold measurements used to evaluate the performance of the noise source. Appreciation is also due Cliff Buttschardt, W6HDO; Duke Moran, W6SPB; Paul Shuch, NGTX; and Bob Sutherland, W6PO; for the use of their commercial and homebuilt noise sources for the evaluation process.

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ham radio

ground currents measuring in 160-meter antenna systems

Several years of operating on 160 meters has verified one thing for sure: The antenna is the name of the game. All the store-bought boxes and smooth-turning dials are useless without that super skyhook. This article describes an instrument that will tell you the behavior of rf ground currents in your antenna system: where it's going, how much, where the current divides, and the location of any conducting element within the instrument's field.

top-band antennas

Few Amateurs are lucky enough to have room for even a simple dipole on 160 meters. Such an antenna requires a length of at least 76 meters (250 feet) and it won't work too well as a DX antenna unless it's 30 meters (100 feet) or more in the air.

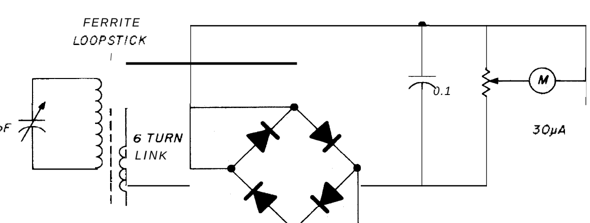
Loading a tower or short vertical antenna is often the best that space will allow. Many articles have been published on how to load this or that antenna on 160 meters and make it work. I've tried many and found that most have a common problem: *ground losses.* The shortest piece of wire that will resonate is %-wavelength long. That's 78 meters (257 feet) at 1824 kHz.

Shortened vertical antennas will work, but all verticals shorter than ½ wavelength must use the earth as a mirror image to supply the missing portion of the antenna. A typical example is the %-wavelength vertical that works against its mirror image to produce the patterns you see in the *Radio Amateur's Handbook.* The problem is usually found in the not-so-perfect ground around your station. In most cases this lossy ground becomes part of your antenna system. This is particularly true for short verticals and inverted L antennas. These antennas are electrically in series with the lossy ground. Working with such an antenna can be frustrating; often you can hear well enough, but much of your transmitter power ends up heating the worms.

Many of the better signals on 160 meters come from hams that have buried literally thousands of feet of copper in the ground to reduce these losses. The prospect of putting that much wire into the ground makes my back ache just thinking about it. Can't you hook onto your water pipes for some of that ground? How about the backyard fence? Would it help to drive a few ground stakes? All the mathematical gymnastics I could muster just didn't tell me where those ground currents actually went. What I needed was a device that would measure the rf current in the ground system and tell me where it was actually going. I finally built a device that does just that: it's called a magnetometer.

the magnetometer

The idea of a magnetic fieldstrength indicator is far from new. An article in the older *ARRL Antenna Handbooks* describes a less-sensitive model for higher frequencies. The 160-meter instrument described here uses a six-turn link and full-wave rectifier to make the impedance match to the meter more effective. Somehow the unit works better on 160 meters than on the higher-frequency bands. One reason is that it's harder to get a proper ground on 160 meters, and ground currents tend to be higher for longer distances through every possible path. By using the magnetometer you can tell the relative current and direction in *any* wire near your antenna or ground system — or in your antenna itself, for that matter. You can locate a buried ground or radial by sweeping



quantity	description
1	aluminum minibox, 102 x 128 x 77 mm (4 x 5 x 3 inches)
1	25-kilohm potentiometer
1	0-30 or 0-50 microammeter
4	germanium diodes (1N34 or equivalent)
1	102-mm (4-inch) broadcast ferrite loopstick (about 6.5 mm or ¼ inch — round works best)
1	365-pF variable
1	0.1 μF disc ceramic
1	terminal strip
2	rubber grommets

fig. 1. Schematic of the homebrew magnetometer.

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the magnetometer over the area and watching the meter deflection. You can compare relative currents in different radials or grounds.

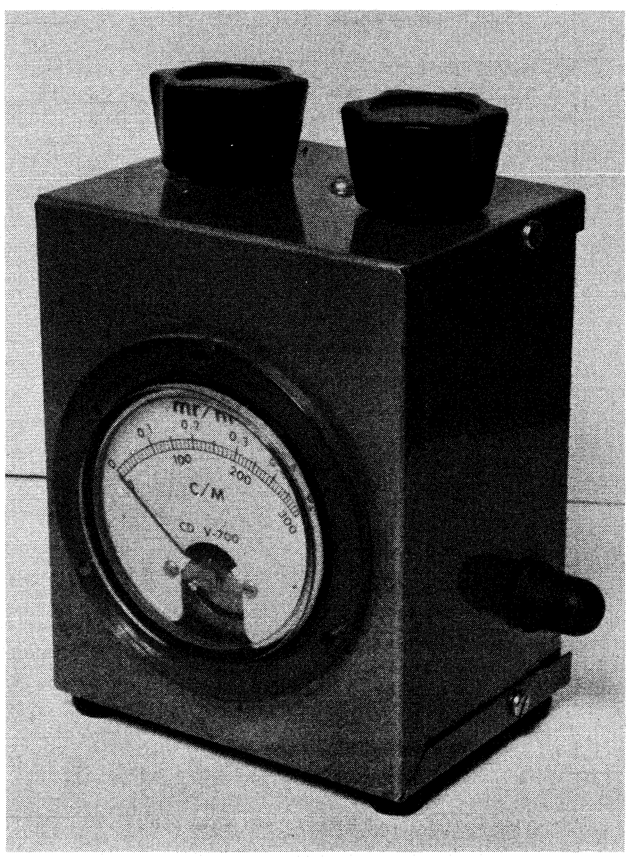
In my basement, where the rf current in the water pipes comes to a T intersection, I can measure how the current divides, if it does, and the relative currents in each section of pipe. In my case, the gas pipe was a better ground than the water pipe!

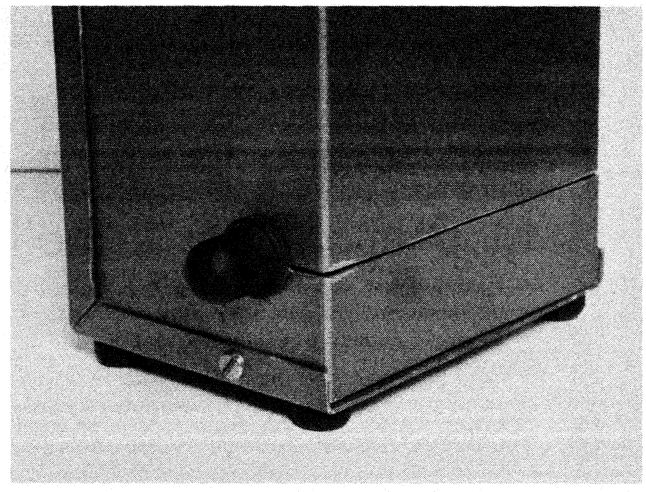
The magnetometer is not sensitive to the antenna's radiated field. It is shielded from this field by the aluminum box and will not act as a field-strength meter unless held perpendicular to a conductor carrying rf current. The slot must be aligned perpendicular to a conductor to allow the magnetic field to reach the loopstick.

It is this magnetic field that operates the device. Fifty watts of rf power into your antenna is usually enough to give a useful reading. (Do your testing during the daylight hours when 160 meters is quiet.) A few minutes with the magnetometer will tell you exactly where your rf energy is flowing.

Unless your antenna is textbook perfect, and few are, you're sure to get a few surprises. I checked WØNFL's station in this manner. Jim was using an inverted L antenna and loaded it through a series-tuned capacitor. Although he had several ground stakes and many buried ground wires, some rf still

Front view of the magnetometer. Small, inexpensive, and easy to build, this device traces one of the most common problems on 160 meters: ground losses.





Rear view of the magnetometer. The slot must be aligned perpendicular to a conductor to allow the magnetic field to reach the loopstick (see fig. 1).

flowed into the station. Ail pipes and the furnace air duct in the basement were bonded together and tied to the ground system.

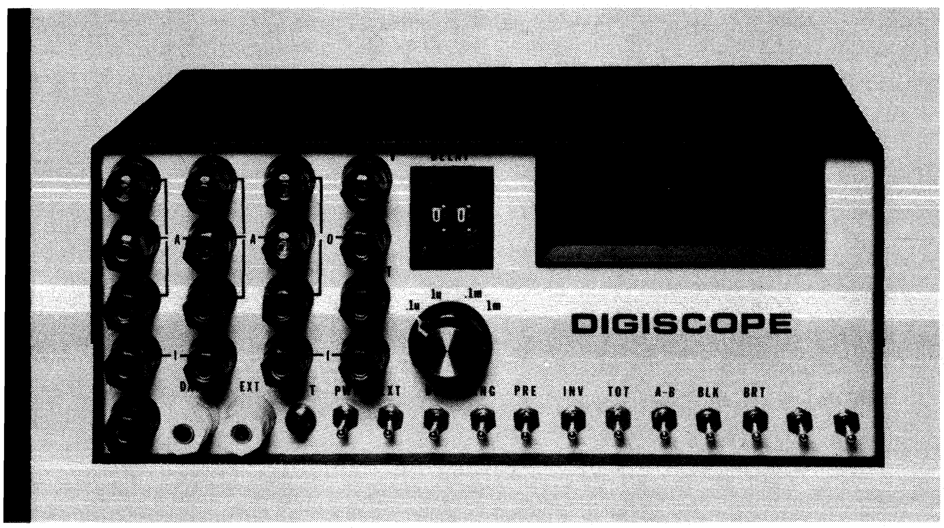
I still remember Jim's surprise when we found that he was loading his kitchen sink. It turned out that the soil pipe provided the best rf ground, and the only path to that was through the S-trap under the kitchen sink. Boy was it hot!

I've used my magnetometer for about four years and have had nothing but good luck and fun with it. I sent a sketch to a couple of my 160-meter cronies, W9GDW and NØBD. Both report that it works well for them too.

construction

The best part about the magnetometer is that it is inexpensive, requires no power supply, and most hams can build it in one evening, often from parts in the junk box. Parts placement isn't critical, and I'm sure that any reasonable approximation of the schematic (**fig. 1**) will work. Use the most sensitive meter you can find and germanium diodes for best results. The loopstick I used was made to operate in the broadcast band with a 365-pF capacitor. I removed about ten turns from the coil to obtain resonance in the 160-meter band. A grid dipper works well to check tuning range. The loopstick is mounted 25.5 mm (1 inch) from the bottom of the minibox and about 19 mm (¾ inch) in from the back. Drill a 12.5-mm (½-inch) hole in each side of the box at these points. Use a hacksaw to cut a slot from the back of the box down to the holes. I used rubber grommets to mount the loopstick. The pickup winding is a six-turn loop of hookup wire wound over the center of the loopstick winding. A few more holes and there you have it: another gadget that no serious 160-meter antenna nut can live without.

ham radio



the digiscope

Eliminate a lot of expensive test equipment by using the digiscope, a self-contained test instrument for TTL circuitry

If you experiment with or repair digital circuitry, this article is for you. Even if you don't, consider the fact that most new equipment contains at least some logic circuitry. Whether you're a digital expert or merely contemplating an upgrade of your trouble-shooting skills, you'll find the digiscope very useful.

To understand my claim, look at a few digital basics. Logic circuitry uses only two levels of voltage, and at times a dc voltmeter is all you need to find a problem or debug an experiment. Unfortunately, voltage levels in many circuits don't stand still, and rapid level changes make the voltmeter almost useless. Professional technicians depend on the oscilloscope to see what is happening in such circuits. The cost of a good oscilloscope may exceed the cost of the device you're trying to check. That's where the digiscope fits in. The objective is to provide oscilloscope function at multimeter prices.

You might be interested in how the digiscope evolved. Last year, several of our club members decided to built the frequency counter described by Jim Pollock.[1] Naturally, some of the counters (including mine) didn't work on the first try. I had worked with digital circuitry previously and enjoyed helping with a few of the counters using an oscilloscope for the repairs. As I used the scope, I began to realize that, for these types of circuits, oscilloscope functions could be approximated using inexpensive TTL devices. The result is that many more people could work with complex circuits without making a major investment in test equipment.

oscilloscope functions

To see how I have approximated an oscilloscope function, take a brief look at the digiscope features.

1. A level indicator serves as a basic voltmeter by indicating a 0 or 1 level in the circuit under test.

2. Pulse duration from 0.1 microsecond to 999 milliseconds can be measured. A variation of this feature allows the measurement of time between the trailing edge of a pulse and the leading edge of the next.

3. The length of time between the leading edges of two pulses at different circuit points can be measured.

4. You can defer the measurement of a pulse until after the occurrence of a pulse at another circuit point. This is similar to an oscilloscope's external sync. In addition, this measurement can be delayed for up to 99 milliseconds after the occurrence of the sync pulse, with a delay resolution of 0.1 millisecond. This approximates the oscilloscope's delayed sweep.

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5. Auxiliary gates are available on the front panel, including a latch to check for pulse occurrence or coincidence between two or three pulses.

6. An event counter, inherent in the design, allows counting up to 999 events with overflow indication. Since some readers may not be familiar with the equivalent oscilloscope functions, I will include examples of digiscope use in circuit explanations.

clock and display circuits

Before looking at uses for the digiscope, I'll examine the clock and display circuits, since they are the key to understanding basic digiscope functions. At the heart of the display circuitry is the 74143 TTL IC which combines the functions of the 7490 and 7447, serving as a decade counter as well as a decoder for the displays. Three of the counters are used, permitting up to 999 counts with a latched overflow indicator. Input to the first 74143 is from the clock, with options of 0.1 microsecond, 1 microsecond, 0.1 millisecond, and 1 millisecond.

The pulse to be measured acts as a gate for the counter, allowing clock pulses to be counted during the time the pulse being measured remains at a logic 1 level. Initially, the counter is reset from the front panel. The next pulse to arrive allows counting. After the first ends, further counting is inhibited so that the display may be read. Input clock pulse options allow a measured range from 100 nanoseconds to 999 milliseconds. Resolution varies from 0.1 microsecond to 1 millisecond. A decimal point is displayed when necessary to indicate either 0.1 microsecond or 0.1 millisecond resolution.

The clock circuit and divider chain is almost identical to that used in the WB2DFA 50-MHz counter[1] referred to earlier. The only change is that a 10-MHz crystal is used to obtain 0.1-microsecond measurement resolution. Although oscillators of this type are usually recommended for use up to 1 MHz, the 10-MHz crystal has worked well. No starting or stability problems have been observed. Counting range is changed by selecting the required frequency via S1A.

As an example of operation, consider the pulses shown in **fig. 1.** Suppose you would like to examine the pulses of **fig. 1B**. To do this, connect the data-in test lead to the appropriate point, and set the range switch to 0.1 millisecond. Push the reset button to enable the scope. During the pulse time, pin 13 of U1 (see **fig. 2**) is a 1 for its duration, or 10 ms. During this 10 ms, one hundred pulses from pin 12 of U4 are passed to the counter/display circuits. At the end of the 10 ms pulse time, the display is then held with 100 displayed. SIB selects the necessary decimal point line causing the display to read 10.0 milliseconds, the length of the pulse. So that I may view the measurement, subsequent pulses are blocked, until the reset is pressed, so that the display may be read.

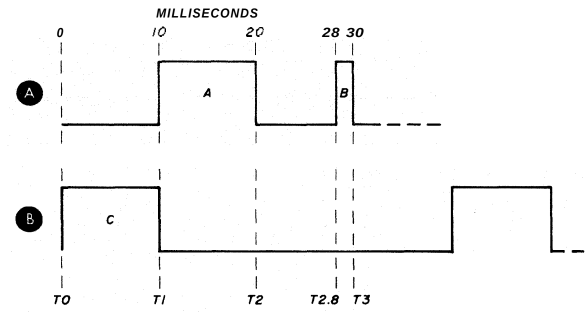


fig. 1. Timing diagram of the two pulses used for explanation of the digiscope's circuits. The time scale is 2 ms per division.

Having measured the pulse width of **fig. 1B**, you might be interested in the length of time between two consecutive pulses. To determine this, simply change S10 to the invert position and press reset. **Fig. 1B** would now appear as a train of 30-ms pulses with 10-ms separation. Using the same circuitry, the scope would now indicate a pulse length of 30.0 milliseconds. Other factors, such as rise and fall times as well as the existence of jitter, have not been determined, however repeated measurements may help determine the extent of any jitter.

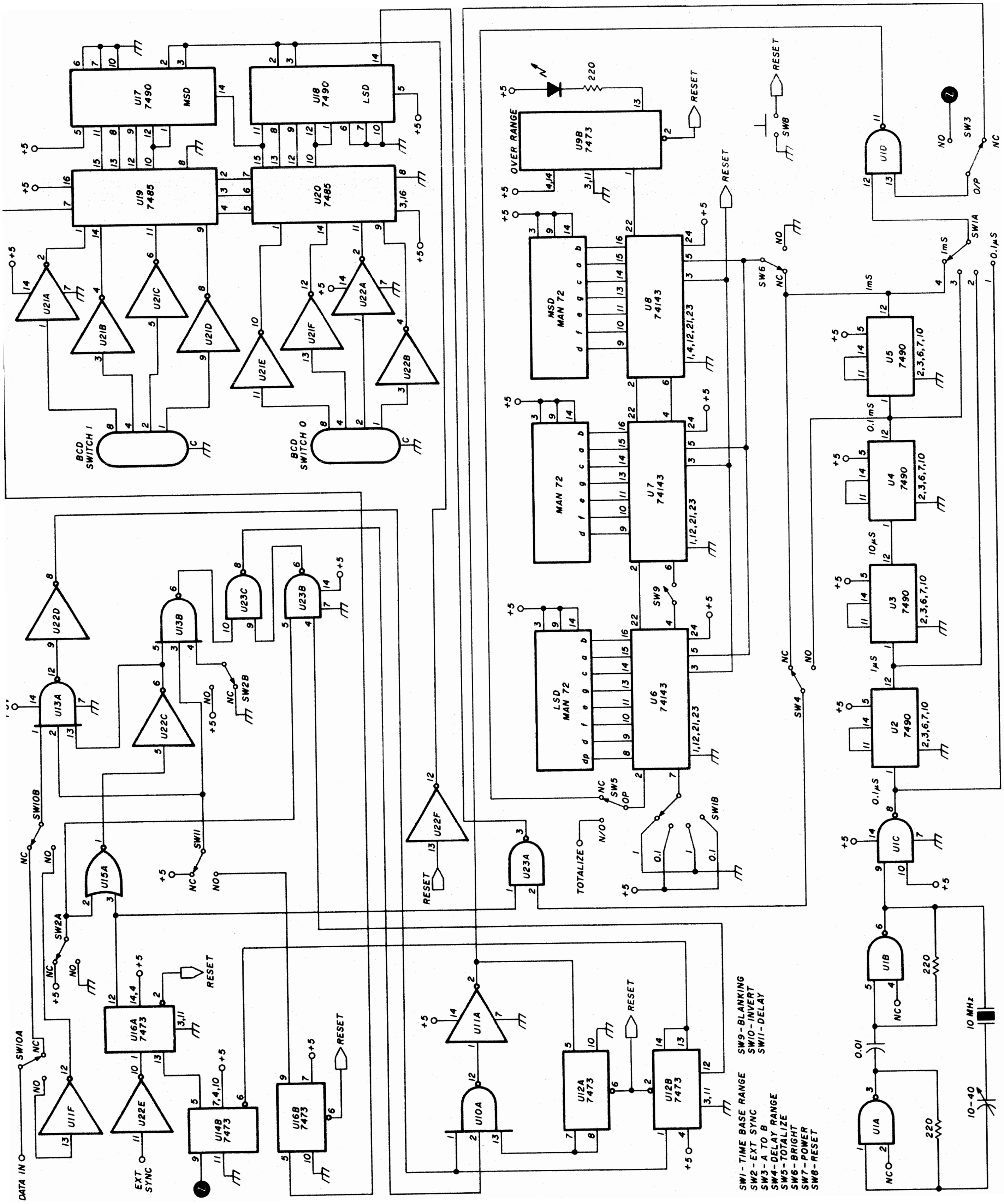
one-shot/swallow circuitry

Recall I said that after the measured pulse ends, further pulses would be ignored. This is accomplished with U12A. The trailing edge of the measured pulse causes U12A to be set, pulling pin 13 of U10 low which prevents subsequent pulses from passing through U10A. This condition exists until the reset switch is pressed.

An interesting problem could arise when the scope is reset. Suppose that the instant after the reset line is grounded you're halfway through a pulse. An erroneous measurement would be obtained. To prevent this, the scope discards the first pulse after reset, ensuring that a partial pulse is not used. After reset, pin 2 of U10 is low, preventing pulses from passing to the counters. The first pulse, whether partial or complete, is blocked with its subsequent trailing edge used to set U12B. Indirectly, this causes pin 2 of U10 to go high. The second pulse is now successfully passed through U10A to the counters. By swallowing the first pulse, I ensure that only complete pulses are measured.

external sync

Examining pulses of uniform length is a fairly easy task. Unfortunately, the pulses may be of varying lengths, as illustrated in **fig. 1A**. By using the gating


SW1-TIME BASE RANGE
SW2-EXT SYNC
SW3-A TO B
SW4-DELAY RANGE
SW5-TOTALIZE
SW6-BRIGHT
SW7-POWER
SW8-RESET
SW9-BLANKING
SW10-INVERT
SW11-DELAY

method I have described, repeated measurements would randomly show pulses of 10 milliseconds and 2 milliseconds. In our simple case, merely finding both pulses may be sufficient. A more precise method is to look for a pulse only at the time it should occur.

As an example, suppose I want to measure the first pulse of **fig. 1A**. Note that this occurs after the pulse of **fig. 1B**. By connecting the external sync input to the appropriate spot and placing the norm/external sync switch in the external sync position, U16A is set by pulse C. The next pulse to arrive at data-in is pulse A, the one I want to measure. In the external sync mode, the pulse-swallowing circuitry is disabled by U13B. Since I know exactly when to expect pulse A, partial pulses are not a problem.

delayed sync

To illustrate the delayed sync logic, assume that you want to look at only pulse B in **fig. 1A**. To avoid random measurement, use the external sync feature. If pulse C began between T1 and T2.7, the external sync feature would be sufficient. However, what is needed is a way to use pulse C as the sync but delay the input until after pulse A.

This cabability is provided by the delayed sync circuitry. With the digiscope set up as in the last example, place S11 in delay position. Note that in the delay mode, setting U16A is still not sufficient for U13A to pass a signal; U16B must now also be set. This will occur at T0 plus some amount of delay selected on the front panel. In this example, the delay switches are set for 20 ms. Thus, at T2, U16B will set, enabling pulse B to pass through U13A to the counter. In this example, pulse B could be selected with a delay ranging from 11 to 27 milliseconds.

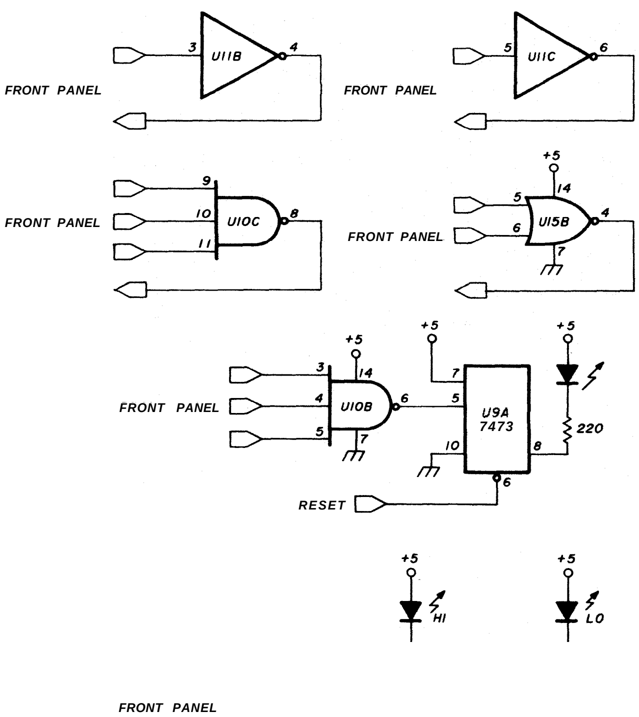


fig. 2. At left, schematic diagram of the digiscope. The RESET switch simultaneously controls all reset functions. Above, the external inputs and conditioning circuits.

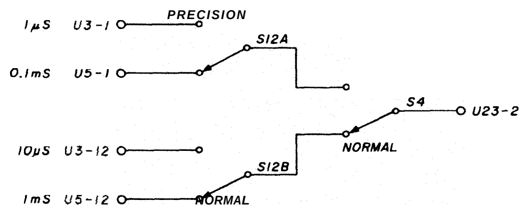


fig. 3. Optional circuit to enable the operator to select smaller increments of delay time.

The delayed sync logic allows me to enable measurement at a precise instant after the occurence of the external sync. This delay can range from 0.1 to 99 ms. S4 selects delay increments of 0.1 or 1 ms. Counters U17 and U18 are responsible for delay timing and are held at zero after the reset is pressed. When the external sync pulse arrives at U16A, U17 and U18 are allowed to begin accumulating delay pulses. The desired delay is set into the thumbwheel switches. These switches present a BCD format to U19 and U20, where they are compared with the count in U17 and U18. When the count exceeds the amount set in the switches, a negative pulse is passed to U16B. This results in the set of U16B and the enabling of data-in.

The precision of the sync delay circuit is very good, since it is based on the crystal-controlled clock. Accuracy however, is dependent on the resolution inherent in the delay range you have selected. Actually, the delay begins not at arrival of the external sync, but at the next clock pulse after external sync. Thus delay error could be from zero up to one unit of resolution. Any error always results in early expiration of delay. Suppose I select a 0.1-ms delay increment and set a delay of 4.6 ms in the thumbwheels. Actual delay will vary between 4.5 and 4.6 ms. For greater resolution, I could simply use smaller units of time in incrementing the delay counters. A circuit illustrating optional resolution of 1 or 10 microseconds is illustrated in **fig. 3**. It should be recognized however, that an increase in resolution reduces the total delay available.

Occasionally it may be necessary to measure the time between two pulses at different pins. As an example, I might wish to measure the time between the rise of pulse C and the rise of pulse A. To do this I would connect the pulse A line to the data-in jack, while the other point would be connected to the external sync input and S3 placed in the A-B position. S2 should be placed in the external sync position, S10 in the invert position, and the reset button pressed. When pulse C arrives, it will set U16A, which will set U14B. The Q output of U14B enables the counter input and timing begins. When pulse A arrives at data-in, U12B is set. The $\overline{Q}$ output of U12B resets U14B, ending the counting cycle. With the range switch set to 1 ms, 10 ms will now be displayed.

accessory circuits

To assist in examining unusual situations, some accessory circuits are provided. These are simply logic gates with inputs and outputs extended to the front panel. U11D and U11E provide a basic logic tester. Connecting the tester input to another logic circuit will cause the appropriate front panel LED to indicate a 0 or 1 condition. Note that a 1 will be indicated if connected to an open circuit. Both 0 and 1 will be lit if connected to a pulsing circuit or one with a faulty voltage level between 0 and 1. Each LED is mounted in a 14-pin socket next to the counter display sockets. LED function is indicated by transparent lettering on a negative film placed between the bezel and the LED. The overrange and test latch indicators are also located in this socket.

Also available on the front panel are AND, OR, and inverter circuits. These can be used to combine pins of a circuit for making tests not otherwise possible. An example would be measurement of the overlap of pulses A and C. I could connect the lines of **fig. 1** to the front panel connections of U10. The output could be inverted using U11B. The output of U11B is then connected to data-in. If overlap exists, it can now be measured to the nearest 100 ns.

If I simply wanted to detect the existence of overlap, a test latch circuit is available. It is composed of U10B and U9A. In this example, I could connect to any two inputs of U10B. Overlap would be indicated by the lighting of the test latch display at the output of U9A. This latch is reset with the reset switch.

A totalize function is provided by S5. With this switch in the TOTALIZE position, front panel input is provided to the counters. External events, which are available as TTL pulses, can then be accumulated. I suggest using an accessory inverter to shield the more expensive counter chip from possible excessive voltage levels. Accumulation can be reset at any time with the front panel reset switch.

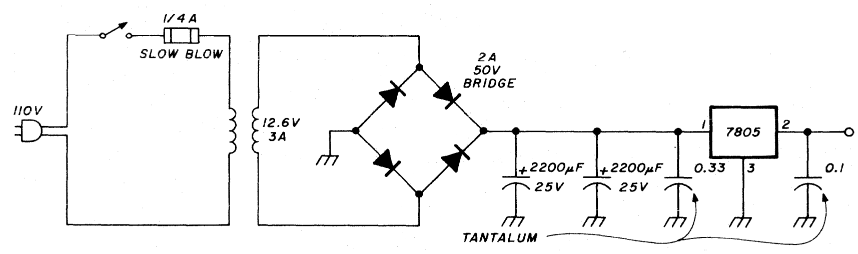


fig. 4. Schematic diagram of a suggested power supply for use with the digiscope.

construction

While building the digiscope I found it difficult to stop adding features. If you build this project, I think you'll find it a good idea to leave room for additions. For example, I think a good addition would be selectable CMOS inputs for working on CMOS circuits. To facilitate this, I left some blank IC positions on my circuit board. While a circuit board is suggested, you could wire wrap this project; my original effort was done on plugboard and didn't seem to suffer too much from all the stray wire.* One note on the 10-MHz crystal is that not any crystal will work well in this circuit. I achieved the best results with a crystal obtained from Jan Crystals. A suggested power supply is illustrated in **fig. 4.**

conclusions

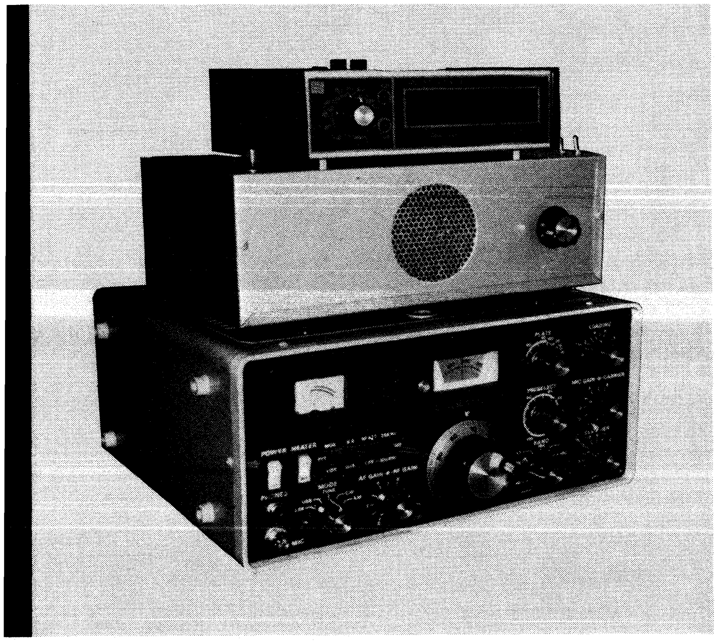
The digiscope is an exciting approach to digital trouble shooting. It can give a way of working on complex circuits without expensive test equipment. If you are interested in digital circuits but don't know how to start, I suggest obtaining the TTL *Cookbook.* I know you'll find working with logic a lot easier than you expected. Whether you're new to TTL or a veteran in logic circuits, I think the digiscope will be a valuable addition to your workbench.

reference

1. Jim Pollock, WB2DFA, "Six-Digit 50-MHz Frequency Counter," ***ham radio,*** January, 1976, page 18.

ham radio

*A drilled and plated set of circuit boards with additional instructions are available for $19.00 postpaid from RTC Electronics, Box 2514, Lincoln, Nebraska 68502.



talking digital readout for amateur transceivers

Adapting the talking calculator's synthesized voice to digital displays for the visually handicapped

During a ragchew with Ted Albrecht, W9IFJ, the subject of the talking calculator was brought up. This machine is manufactured by Telesensory Systems in Palo Alto, California, for the visually handicapped. We agreed that it would be most interesting if the talking calculator's synthesized voice were available separately. Perhaps it could be interfaced with digital-readout equipment in general and Ted's in particular.

*The S2A Speech Synthesize Module is available from Telesensory Systems, Inc., P.O. Box 10099, Palo Alto, California 94304. Contact Mr. Wladimir N. Walko, Speech Products Manager, (415) 493-2626. Unit price to hobbyists is $95.00.

Happily, upon contacting Telesensory in Palo Alto, we found that the voice synthesizer module was available separately to hobbyists, and its numeric vocabulary was in standard TTL BCD code that already existed at the input of the 7447s in Ted's DD-1 Digital Display Unit.

Ted's purchase of the English speaking S2A module from Telesensory Systems* represented the committment necessary to get the project moving. Design of an interface between it and the DD-1 is the project this article describes. Much of the design should be credited to Tim Blank, WB9GYU, who is more knowledgeable of digital logic design than I, and who provided the assistance in debugging the assembled unit.

intefacing considerations

Interfacing the voice with the six-digit readout involves changing the parallel (all-at-once) visual dispiay to a seriai or sequential form, so that the digits from left to right are said in order with a minimum of delay between words. It would be possible to clock across at regular time intervals if allowance for the longest word were made, but much wasted time would occur, so the clock approach was abandoned. Since each digit can be 0 to 9, or anything in between, four data lines of binary for each are required. To read the digits in sequence then requires that the first four digit data lines of the speech synthesizer module be sequentially connected to the four digit data lines coming from each digit of the visual display. This adds up to twenty-four data lines between the DD-1 and the interface board!

design

The most reasonable approach was to use four eight-position multiplexers, one for each data line. They could be stepped by a binary address of three bits, from 0 through 7. The device that provides this address to the four 74151 multiplexers is a 74193, a most remarkable MSI chip (**fig. 1**). The 74193 can provide a four-line address with a capability of 0 through 15, which would seem to be a problem. However, the 74193 can be reset after any given number of steps very readily. In fact, it can be told to skip a preset number of the initial steps in its output address. Its fan-out is about twenty, indicating no trouble in driving the multiplexers, so it seemed a reasonable basis for an interface. Since the 74193 could be told which addresses to provide, considerable imagination can be used in this aspect of the design.

I decided to use the eight-position capability of the

By Ray Brandt, N9KV, 824 North Marion Avenue, Janesville, Wisconsin 53545

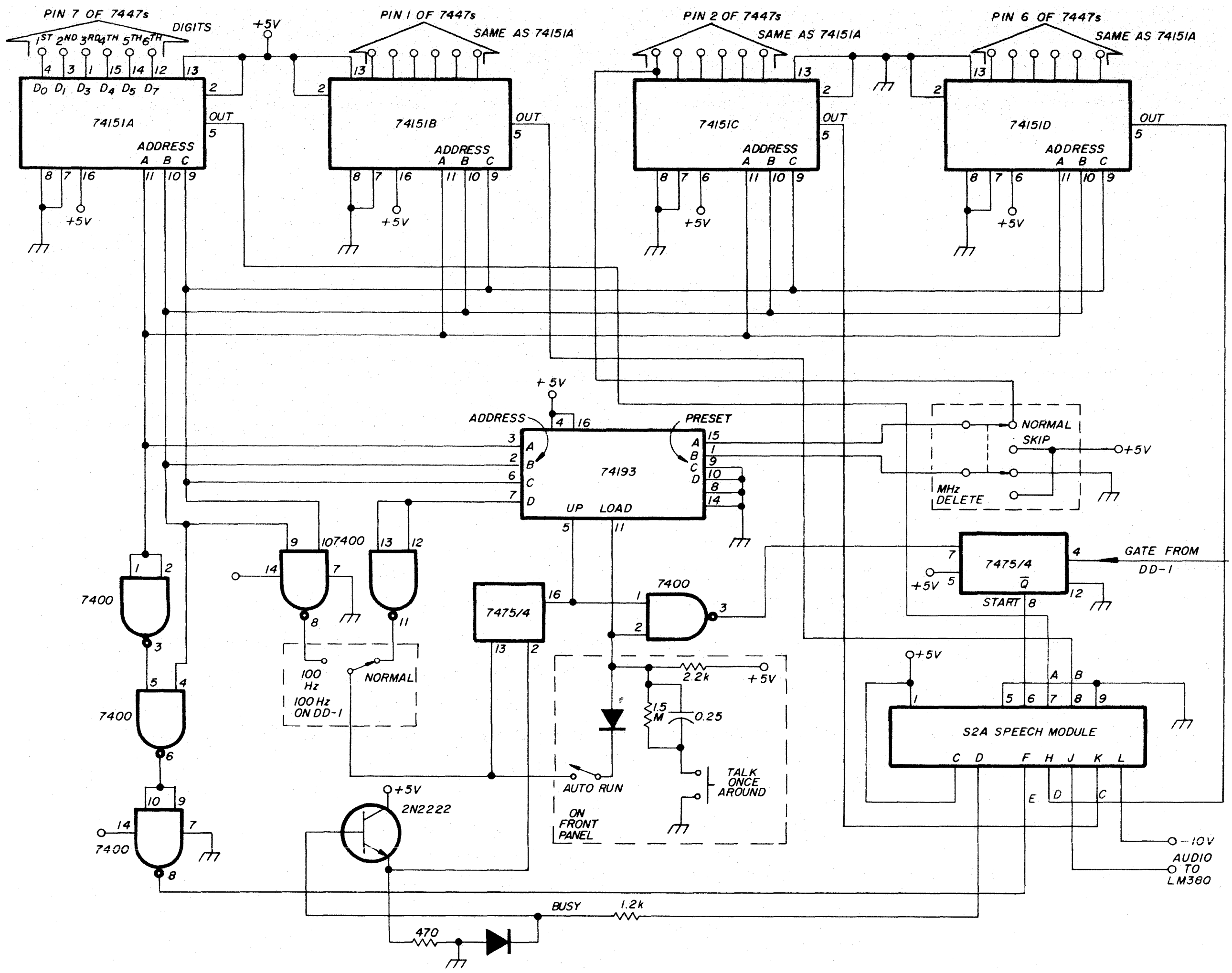


fig. 1. Schematic of the talking readout. The 74193 provides the address to the four 74151 multiplexers. The inputs to the 74151s come from the designated pins of the 7447s in the frequency display.

multiplexers, using the two spare positions to have the voice say POINT between megahertz and kilohertz, and between kilohertz and hertz, provided by the readout unit. Looking at the S2A vocabulary **(table 1)**, it's necessary for the **A,** B, and E data lines to be high, so the **A** and B data lines were hard-wired in the 2 and 6 data positions. An additional circuit was designed to bring up the E data line in these positions only.

A dual NAND gate will produce a 0 only when both inputs are a 1. Studying the address truth table **(table 2)** for what is unique about these two positions of address, it's seen that B is, of course, a 1 for both *2* and 6 and adds to **C** for the 6 position. But looking again, B is uniquely present on only 2 and 6 if odd numbers involving an **A** can be ignored. This suggests the approach. Invert the **A** and we have a 1 only for even numbers – use it for one gate input with the B data line on the other gate input.

Now we have a 0 out of this gate on only the 2 and 6 positions. Since we needed a 1, another section of the 7400 provides the inverter between it and the E data line of the S2A.

Taming the 74193. Several conditions must be met for the 74193 to step along through its addresses in sequence, reset at the end of the sequence or sentence, and for its selected data to be presented to the speech module only:

1. When it is valid data

2. When the module is not busy saying the previous word.

Consider first the data provided by the digital display unit (see **table 3**). The count periods are a precise, crystal-controlled, ten milliseconds each followed by 50-millisecond periods of stored display data, alternating about seventeen times a second between the two. We must accept data only during the storage time, or the readings will appear random, having been taken on the "fly" during a count. The waveform that defines these two periods is present

table 1. S2A Speech Synthesizer vocabulary.

data lines up	speech	data lines up	speech
none	OH	C,D	PERCENT
A	ONE	B,C,D	LOW
B	TWO	A,B,C,D	OVER
A,B	THREE	E	ROOT
C	FOUR	A,E	EM
A,C	FIVE	B,E	TIMES
B,C	SIX	A,B,E	POINT
A,B,C	SEVEN	C,E	OVERFLOW
D	EIGHT	A,C,E	MINUS
A,D	NINE	B,C,E	PLUS
B,D	TIMES*MINUS	A,B,C,E	CLEAR
A,B,D	EQUALS	D,E	SWAP

on pin 6 of the 7400 at the right edge of the DD-1 board – the second chip from the back. It is in the logic 1 state during count, 0 during read. We want a signal that is up during the storage period. Since there was an unused section of this 7400, I made an inverter of it by connecting its pin 6 to 9 and 10, bringing out the desired signal to the interface board from pin 8.

Interconnection hints. The grubby details of interconnecting with the DD-1 deserve comment. The picture looking into the top of the unit shows a small terminal board mounted at front right above the 7447s. The circuit was built from a small rectangular piece of circuit board stock, cross-hatched with a hacksaw to provide rows and columns for the data lines. The ventilating holes to the front and right of the power transformer were enlarged to accommodate the twenty-eight interconnecting leads, leaving out the bottom of the accessory unit. I removed the mounting feet and replaced them with spacers, which fasten the DD-1 to the interface unit.

The 100-Hz pushbutton of the DD-1 has a spare spdt section; the three leads added to it and the blanking output previously described are included in the harness.

DD-1 mods

To facilitate connection to the 7447 inputs and provide the blanking output, it's necessary to remove the DD-1 from its case – not a trivial job. Take out the power transformer mounting screws, all screws mounting the circuit board, and remove the input connector, the knob, and bandswitch mounting. Push in the LEDs and the panel should lift out from the back. Make short connections to the added terminal board; the harness may be added after reassembly. A magnifying lamp and small iron and solder are essential. If you decide to drill any holes in the circuit board, for gosh sakes hold the board up to the light first so you won't drill through conductors that may be on the other side.

Modification of the DD-1 is the most miserable part of the project, but finding the patient well after all this surgery should buoy you up for the remaining pitfalls awaiting you.

interfacing the speech module

Now consider interfacing the speech synthesis module. The ROM of this unit is permanently stored with digital data which, when swept out by its internal clock (roughly 12 kHz), produces an audio waveform imitating the twenty-four words of its vocabulary. When properly filtered to reject the dominant clock frequency and amplified, acceptable speech is reproduced by the speaker. The complete program of the S2A is shown in **table 1.** (The unused words may come in handy when debugging the unit.) **A** is the least-significant bit; its pin is 7. B appears on 8, **C** on K, D on H, and E on pin F of its twenty-pin edge connector.

If you're eager to try the speech board before finishing the project, connect all the power leads and indicated ground leads, then tie all data leads either up or down (but don't leave them floating or you'll have problems). Pulsing the start line up with a clip lead, then back to ground, should produce the word you've coded into the speech board by the data line hookup you've chosen.

The **BUSY** line goes to almost –10 when saying words, then goes back to +5 on completion. An emitter follower, with its emitter returned to ground, converts this to the TTL voltage swing required by the interface.

event sequence

Two latches are used in the control loop to guarantee the proper sequence of events. The left-hand latch normally passes on the information that the last word is complete. It then steps the address into the next position. However, when the **D** address line goes high for address **8,** the inverter opens the latch, no more steps are permitted, and the sentence is terminated. If the 100-Hz switch on the DD-1 is not depressed, the lower gate operates at address ***6,*** cutting the sentence two words short.

***table 2*. Address truth table.**

multiplexer address position	multiplexer address lines A	B	C	D
0	0	0	0	0
1	1	0	0	0
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

The right-hand latch has a logic 0 entering it at completion of a word, after the address has been stepped, which it passes along and inverts the next time a valid read period occurs from the DD-1. The latch Q output is inverted internally and serves as a start pulse to the speech module, telling it to **GO** while the data is valid.

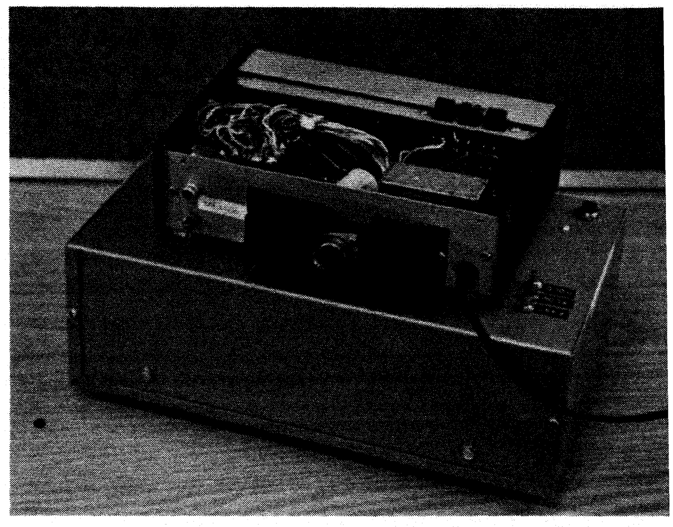
As I've indicated, the unit normally stops talking at the end of a sentence; this is its initial condition with all address lines high. The pushbutton pulls down the 74193 **LOAD** terminal, resetting it to the 0 address, so it runs through the permitted sequence to address **8** and stops.

Another feature of the 74151 involves the use of its preset capability. If all four preset lines are in the 0 logic state, a down pulse on the **LOAD** terminal will cause it to go to the 0 address. If the **A** preset line is tied high, the reset will go to address **1.** If both **A** and B preset lines are tied high, while **C** and **D** are at ground, the unit will reset to address 3.

This design permitted two useful features to be incorporated. When the DD-1 is presenting 160, 80, or 40 meters, only one digit is needed to enumerate the megahertz. Spectronics connects all four data lines for the first digit high on these bands to cause the digit to blank out. But this presents an **A,** B, **C, D** to the S2A and it will say **OVER** in this first-digit location. Since the DD-l reads only a 1 or 2 when it reads anything in the first-digit position, only the **A** and B data lines need be connected to the multiplexers. Connecting the **C and/or D** data lines to the **A** preset line will cause the program to skip that digit entirely, which will shorten the beginning of the sentence by one word. The designers of the 74193 must have had this in mind, wow!

Rather than connecting directly to the preset line, however, **I** ran the **C** and **D** data lines through one section of a dpdt toggle switch so that both the **A** and B preset lines can be tied high in the other switch position (see **fig. 1**). The other switch position will delete the **ONE FOUR POINT** from the start of the 20-meter reading when desired; call it megahertz Delete, **I** guess. Shortening the sentence to just the three-kilohertz numerals takes only 318 as long to say, permitting more rapid checking while tuning around the band with the transceiver. This can, of course, be switched in and out.

While playing around with the unit at this stage of development, **I** noticed that connecting a diode with its anode on the 74193 **LOAD** terminal and cathode on the sentence-terminator clock input of the left latch permitted the unit to **AUTO-RUN,** repeating however short or long a sentence were chosen, ad nauseam. It apparently provides a sufficient down pulse to the **LOAD** terminal to reset at the completion of each sentence. Really slick for an afterthought!



Back view with the cover of the DD-1 removed. The harness and circuit board are visible at the left. The harness enters the talking readout unit through an enlarged hole adjacent to the power transformer in the DD-1.

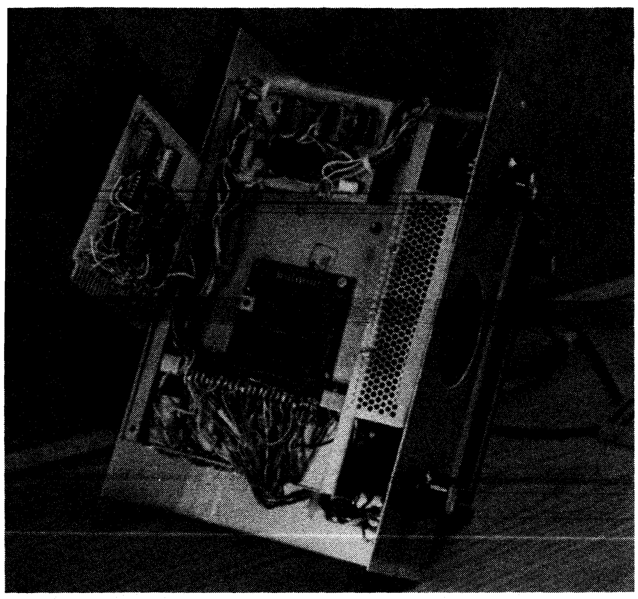
The interface unit was built on a Radio Shack project board, about 102 mm (4 inches) square with a **44**-pin edge connector printed on one edge. There are three rows where up to nine device sockets can be mounted as well as printed tabs permitting two or three connections to each chip terminal. Keeping the data lines in groups of four, along side of the connector, makes tracing easier; the extra one or two can be at the end that didn't quite make it.

table 3. Relationship between the address. displayed frequency, and output of speech module.

DD-1 display	1	4	•	2	5	0	•	3
speech	ONE	FOUR	POINT	TWO	FIVE	OH	POINT	THREE
address position	0	1	2	3	4	5	6	7

Make a sketch and assign each 74151 a data-line letter: **A, B, C,** or **D.** One of each four data lines for a given digit will go to the same number pin of each 74151. The first digit lines go to pin 4, second to pin 3, third to pin 1, fourth to pin 15, fifth to pin 14, and sixth to pin 12. Pins 2 and 13 of the **A** and B multiplexers go to +5 volts; the same pins of the **C** and D multiplexers go to ground as a part of the fixed decimal **POINT** at these address locations.

Wiring can be completed, on the device-side of the board, **I** recommend about no. 24 (0.5-mm) stranded



Internal view of the talking readout chassis. The speech synthesis module is in the center of the chassis, directly behind the speaker. The interface board has temporarily been removed from its connector. At the top of the chassis is the clock board which has been added to incorporate a talking clock feature. This unit will be described in a subsequent issue of ham radio.

wire as about the largest practical size to work with. The picture tells the story; it's a real birds nest! Use the bus down the center row of devices for ground; the other two for +5 volts. The only reason for signals getting where they shouldn't is solder bridges, *and* if they don't get there at all it's poor connections. After installing all the devices, a couple of point-to-point checkouts with the ohmmeter are worthwhile. It boggles my mind to imagine this as an etched board — perhaps someone will accept the challenge!

power supply

The power supply **(fig. 2)** is conventional and requires little comment. The +5 volt supply uses a 6.3-volt, I-amp transformer. I slipped in a half dozen extra turns over the existing secondary winding without tearing down the transformer so that its output is close to 7 volts, providing adequate voltage for good regulation. The total load at 5 volts is about 300 mA. A small 300-mA, 12-volt transformer from Radio Shack is adequate for the light load of the negative supply.

The case of the 7805 may be bolted to the chassis; the 7815 can be insulated with a mica or fishpaper shim. To realize a −10 volts, the positive terminal is connected to the +5 volt output; the negative common will then be 10 volts below ground. The full 15 volts of the negative supply is used by the LM380 audio amplifier, so the speaker should be returned to the −10 volt bus.

some afterthoughts

There are other ways of building this unit. Perhaps mounting the four multiplexers on a small board internal to the DD-1 would make things easier and minimize the interconnection problem. This approach shoulg permit a unit no longer than the DD-1 for the balance of the circuit. I hope this is only the first article on the subject. Good luck on your talking second operator!

Oh yes — should you desire to interface a visual readout with only the seven-segment connections to the LEDs available for data, National Semiconductor makes the 86L25, which will convert seven-segment data back to the BCD mode. One would be required for each digit, of course.

ham radio

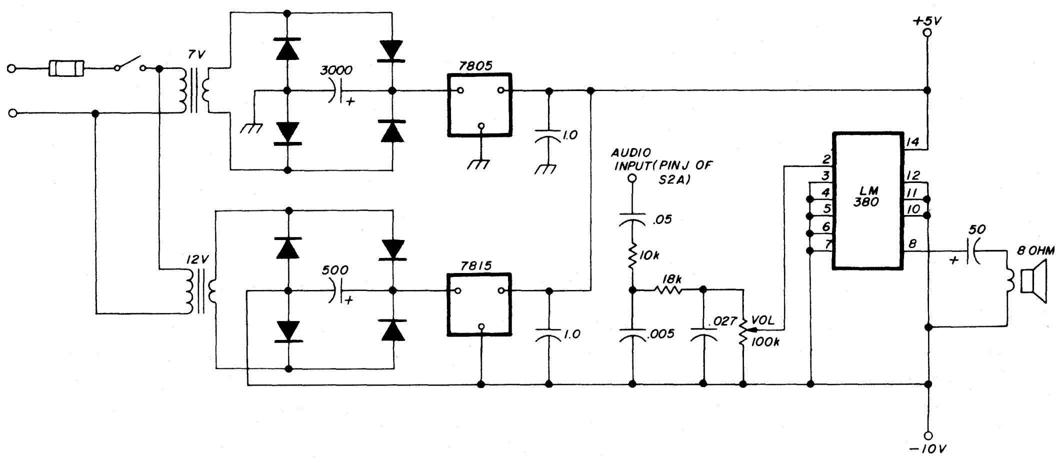


fig. 2. Power-supply schematic. Circuit is conventional. Parts are available from Radio Shack and other popular suppliers.

an introduction to packet radio

An interesting idea from our Canadian colleagues: computer time sharing over vhf links on the vhf bands

If ever two hobbies were made for each other, they must be microcomputers and Amateur Radio. More and more hams seem to have taken the plunge and are starting up small computer systems. A logical outcome of this marriage of activities is that sooner or later the ham will wonder if he can't use the radio to send his computer information to his friends on the air. It can be done, and this article describes one of the best ways of doing it.

time sharing

Most computer users have been exposed to the time-shared computer. The computer is very fast, and the users require and generate information very slowly. If each user is connected to the computer by a separate line **(fig. 1)** or a radio link **(fig. 2)**, then all the computer has to do is to check each line periodically and divide the processing time among the users. Typically it might take 20 seconds to type in a line on your terminal and only a few milliseconds at most for the computer to process that data line. Time sharing permits each user to think he has the whole computer to himself. It also permits the computer to act as an intelligent clearing-house for programs in which the users interact with each other.

The next step is to look at the lines, or radio links, themselves. The line that took 20 seconds to type might contain 64 eight-bit ASCII characters, or 512 bits of information — an average rate of 512/20, or about 26 bits per second.

It's easy to send 2400 bits per second through a normal voice channel and it's theoretically possible to use much higher rates. So, if we were to store each line locally then send it in a short burst, it could be sent in about 0.2 second! More detailed estimates and calculations show that several hundred users could be accommodated on one voice channel.

Also, because only one frequency (or two for full duplex) is used, newcomers can join the system easily by getting onto that frequency without having to wait for new channels to be assigned at the computer. Now we have a time-shared radio link working with a time-shared computer **(fig. 3)**. As we'll see, the time-shared radio link is useful by itself, without the central computer. Its major advantage is a huge saving of spectrum space (not to mention knowing what frequency to look on to find your friends).

packets

Each of the short bursts mentioned above is called a ***packet*** and most contain, in addition to the data, the identification of both sending and receiving stations and some form of error checking, so that the recipient will know if the information is correct. If it is correct, the recipient sends an acknowledgment (ACK) to the sending station (a fully automatic process). Full details of the packet format appear below.

Transmission of radio packets by Amateurs is now legal on several vhf and uhf bands in Canada. U.S.

By Ian Hodgson, VE2BEN, 296 Malcolm Circle, Dorval, Quebec, Canada H9S 1T7

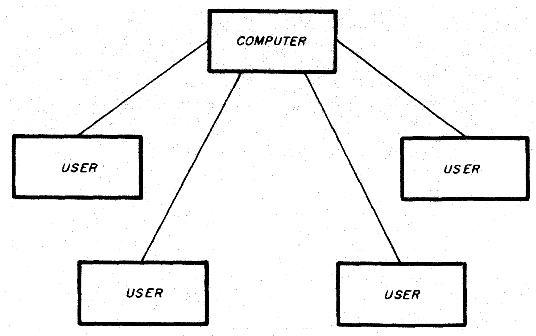


fig. 1. A wired time-sharing system requires separate connections for each user. It permits communications between users as well as computing functions.

Amateurs may want to apply pressure to be permitted to send packets also.

networks

Although two stations can send packets back and forth, the method doesn't really come into its own until many users share a packet repeater (called a ***node).*** Many nodes can be linked (possibly by a geostationary satellite — this system may actually be working in a couple of years) to form a network **(fig. 4**). Each packet contains its own address information, so the network can automatically forward the information to its destination.

The first radio packet network in use was set up by the University of Hawaii and is called the Aloha net. It links terminals on several islands to a central computer on about 400 MHz.

the node

The repeaters in a network may take several forms. They may be simple rf repeaters used to extend range, as we now use on two meters. This scheme, however, doesn't fully exploit the advantages of packets.

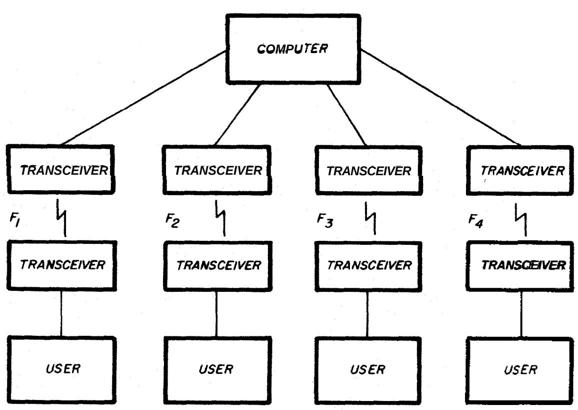


fig. 2. Radio equivalent of a wired system. A different frequency and a separate transceiver at the computer are required for each user.

A better repeater is the ***store and forward repeater,*** which receives packets from one or more users, processes them, and forwards them to a central node, which would otherwise be too far away to access. If major cities had intelligent nodes, store and forward repeaters would make them accessible to users in outlying areas.

Major nodes would likely be computer controlled to perform error checking, acknowledging, and routing of packets between users. Ultimately, the node might be connected to a central computer with more power than the individual users could afford to have at home. You could have access to utility programs, a ham community bulletin board, repeater council data, a swap program, or even play multi-user space war instead of rag chewing.

what else can it do?

Any information that can be put into digital form can be sent through packets. This information includes RTTY, computer data, and even voice and TV pictures. Of course, data rates must be much higher for the last two modes. The packet system could be as easy, fast, and reliable for talking to hams a continent away as for talking across town on the repeater.

how about details?

Let's take a closer look at the details of the process. The packet format hasn't yet been finally decided upon, but here's what we've been experimenting with (each byte is eight bits).

bytes	contents
1	packet initiation byte (hex A7)
1	start of header (SOH, hex 01)
6	destination call sign
2	destination node
6	originator's call sign
2	originator's node
1	service message flag byte (set to hex 41, A, for ACK)
1	spare
1	end of header (EOH, hex 04)
2	CRC16 (this is a type of error check) for header
(Everything above is the header)	
1	start of text (STX, hex 02)
64	DATA (could be ASCII, binary or EBCDIC)
1	end of text (ETX, hex 03)
2	CRC16 for data
91	bytes total

Note that all packets must be the same length if the system is to operate efficiently (if you don't understand why, let's just say that it can be shown mathematically, and, even though I can do the figures, I don't really understand either).

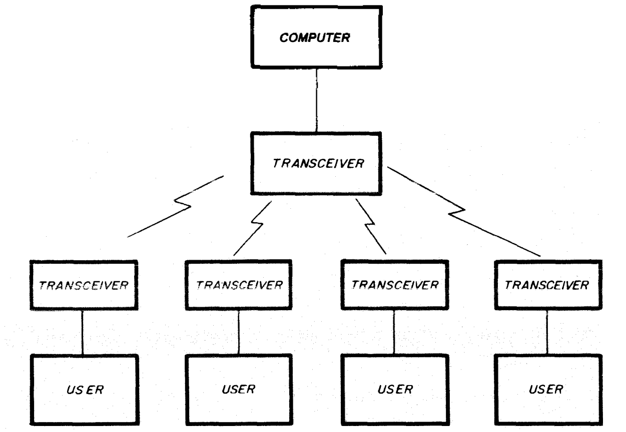


fig. 3. A packet network needs only one frequency and one main transceiver. In addition to computing power, the network offers flexible communications between users, a fact that Amateurs can put to good use.

An ACK is simply the header from the received message with the originate and destination addresses interchanged and with the service byte set to indicate that it is an ACK. No data is included in an ACK.

interference and ACKS

It will undoubtedly come to pass that two users will, sooner or later, send their packets at the same time, thus causing total or partial loss of both. We can deal with this problem by using what's called a CSMA POSACK system. This acronym stands for Carrier Sense Multiple Access Positive Acknowledgment. Here's how it works:

You finish typing your line of data and hit "return," or whatever, to send the packet. Your system checks with your receiver, and if no signal is being received, the system immediately sends the packet. It then waits for a time of one packet length

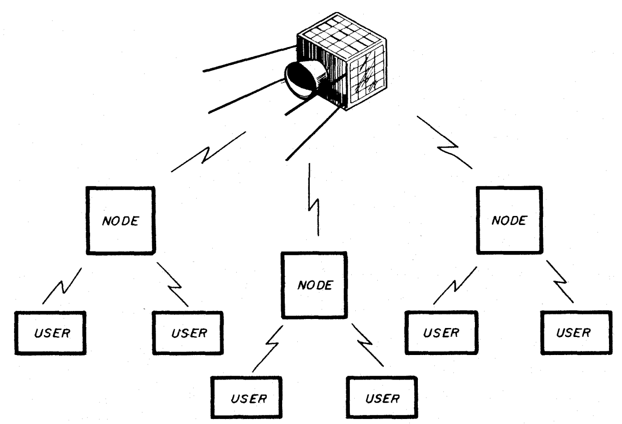


fig. 4. A packet network allows time-shared communications between users and a variety of nodes. Nodes in various cities may be connected together by uhf links or perhaps by satellite. Not as far out as it seems, this scheme may be in operation within two years.

for the ACK. If not received, the system waits an additional, random, number of packet lengths and tries again. The process is repeated three times. If still no ACK is received, the system returns a message to the sender, telling him that the transmission was unsuccessful.

Why do we wait a random delay? If there were no random delay, and two stations sent their initial packets at the same time, then all three tries would collide. This way, they don't.

When receiving packets, your system intercepts all packets regardless of their address. The system checks the destination of every packet, and if it's for you, the error check (CRC 16) is performed. If this is OK, the system immediately sends an ACK. Note that

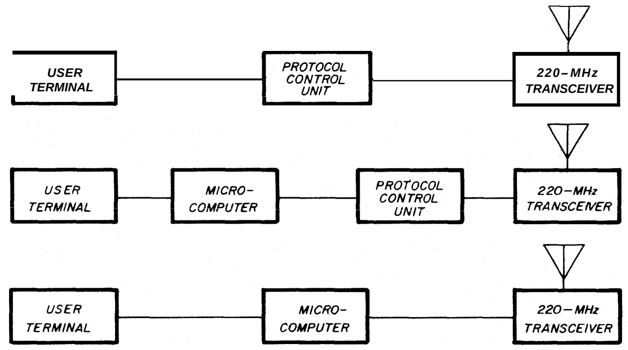


fig. 5. Three possible packet radio setups. The protocol control unit may be used by itself to implement the packet protocol, or it may be used in conjunction with a microcomputer to allow computer information exchange between users. Alternatively, the microcomputer may be programmed to take care of the protocol itself, with some reduction in its availability for other purposes.

ACKs have priority on the system, as all transmitters wait one packet length after each packet sent so the ACKs can get through. Also if a signal is present on the initial receiver check, the packet transmission from your station is delayed long enough so that the other station can be acknowledged.

hardware

Sounds complex? Well, it is. But don't forget how inexpensive complexity is becoming; witness the pocket calculator, which for less than $10 is more complex than everything else in your house put together. Here's what you really need:

a. A vhf (220-MHz in Canada) transceiver (fm will do)

b. A computer, Protocol Control Unit, or both (see **fig. 3**)

A single-board microcomputer can form the Protocol Control Unit (PCU) to implement all the above details (these details are called the packet ***protocol).*** It would require only about 2k of memory and could probably be built for less than $100. We are working on this too. In due course, these should be available commercially, but for now let's build our own. Full details of the final protocol will be published when available.

what frequencies can I use?

In Canada certain portions of some vhf bands have been set aside for packets:

frequency (MHz)	transmission mode
220.1-220.5	shared packet and other modes
220.5-221.0	shared wideband packet and other modes
221.O-223.0	packet only
223.0-223.5	shared wideband packet and other modes
433.0-434.0	packet only (wideband, 100 kHz, for repeater links)
24,000-24,010	shared packet and other modes

Obviously the bulk of the activity is expected to be on 220 MHz. If you've looked at the low prices on rigs for that band, it seems like an excellent idea.

how it all began

In April, 1978, the Department of Communications (DOC)in Canada announced its intention of changing the regulations to permit only "Packet Radio" (cries of, "What the devil is that?") on 220-225 MHz. Amidst the horrified screams of many Amateurs, a few of us decided to have a closer look. We liked what we saw. With the close cooperation of the DOC, who modified their original proposal accordingly, we started experimenting with packets. The group doing most of the work is based in Montreal.

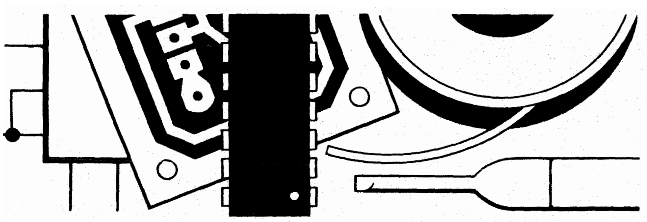
One more thing to note. The Amateur Radio community is the group that will develop this system on a low-cost, widely distributed basis. Our work will no doubt be closely watched by commercial interests, so let's earn our privileges as hams and contribute once more to the advancement of the state of the art.

acknowledgments

I'd like to thank Bob Rouleau, VE2PY, for enthusiastically supporting the packet concept and keeping the rest of us going; Paul Laflamme for his work on the protocol and programming and for helping to prepare this article. Special thanks to Dr. John de Mercado, Director General of the Telecommunications Regulations Branch of the DOC, for giving us the initial kick by changing the regulations and for his continued support and close cooperation. I wish ***all*** government departments worked this way!

ham radio

the **weekender**



biquad bandpass filter for CW use

The biquad bandpass filter has many advantages when used as a CW filter. First, the biquad can be adjusted to control both center frequency and Q. Second, the filter is very stable and will not break out into ear-shattering oscillations. Third, the filter can be built by hams like you and me. That is to say, the components are available at Radio Shack, the tolerance of none of the components is critical, and no special test gear is needed.

circuit description

Fig. 1A shows the traditional circuit, which requires the plus and minus supplies. The advantage of this configuration over the single-voltage circuit is the capability to couple directly to the input and output of the filter. In addition, the op amp has better gain characteristics with the higher supply voltage.

The single supply circuit is very similar (see **fig. 1B**). Most references recommend providing a low-impedance dc return for the input circuit. For this reason, I avoided for a long time even trying the single-voltage supply. But my experiments with the μA741 showed that the circuit does a remarkably good job. By providing half the supply voltage to the positive input of the op amp, all of the outputs are operating above ground. This is the obvious reason for needing the capacitive coupling on the output and input. The μA741s will drive headphones directly, but for general use, an amplifier speaker is recommended.

By James M. Rohler, NØDE, 1967 Bristol Drive, Bettendorf, Iowa 52722

After a short period of experimenting, you will find the adjustable features useful. The Q is especially interesting, since it can be adjusted too high, causing ringing. Under some conditions, the narrower band-

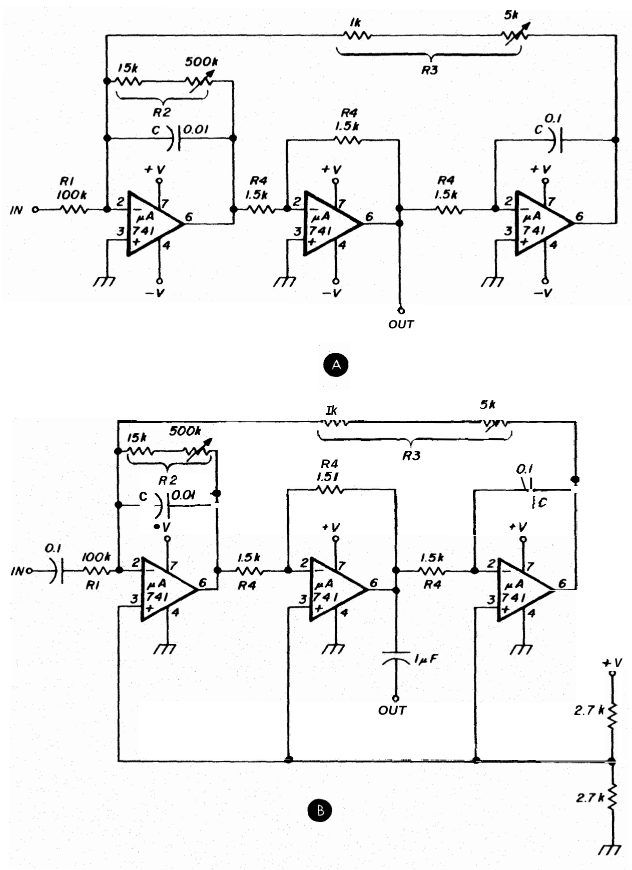


fig. 1. Schematic diagrams of the biquad active filter for dual supply (A) and single supply (B).

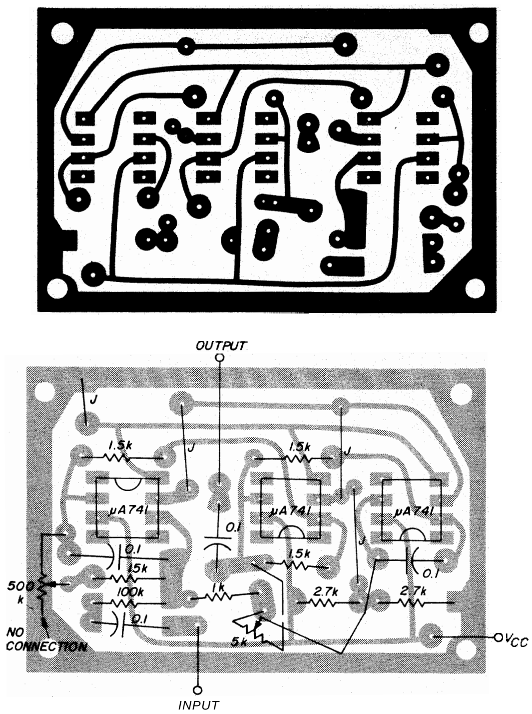


fig. 2. Foil pattern for the active bandpass filter (above). The parts placement diagram is shown (below).

width and ringing is a fair tradeoff for eliminating the QRM.

components and construction

Fig. 2 shows the foil pattern and parts placement diagram. If potentiometers are not desired, fixed resistor values can be determined from the following equations:

$$f_o = \frac{1}{2\pi C \sqrt{R3R4}} \qquad (1)$$

$$B = \frac{1}{2\pi CR2} \qquad (2)$$

$$G = R2/R1 \qquad (3)$$

where f_o is the center frequency in Hertz,
B is the bandwidth in Hertz,
G is the gain of the circuit.
The Q is the reciprocal of the bandwidth times the center frequency.

For single-supply operation, jumpers are used to connect the junction of the bias resistors to pin 3 of each IC. In addition, pin 4 of each IC is connected to ground. To use a dual supply, the pin 3s are connected to ground, and pin 4s to the minus supply. Also, in this case, the input and output capacitors and bias resistors can be eliminated.

ham radio

gallon-size dummy load

Homebrew dummy load with low SWR that includes rf-voltage monitorina provisions

In a previous article,' **I** described a home-built dummy load based on intelligent overload of a bank of carbon resistors good for 5-second tuneup of a 1-kW transmitter (50 per cent loaded at tuneup). The dummy load served me well for many years and, in fact, is still in use. But an incident occurred recently during a TVI test of a new linear that emphasized that a rating of 500 watts for 5 seconds doesn't mean a full kilowatt continuously! A new set of resistors got the load back to normal, but the incident caused some time out to build the dummy load described here, which is capable of running tests at the 2-kW input level.

review of rating techniques

The following applies to Allen-Bradley 2-watt resistors. Other sizes, or resistors of other manufacturers, can be used, but the rating coefficients must be obtained from the manufacturer.

A 2-watt, A-B resistor of the 20 per cent series is rated for a continuous load of 2 watts for 100,000 hours when the resistor is mounted with 25.4-mm (1-inch) leads and has a body temperature of 100C (212F). This occurs when the ambient temperature is 50C (122F). The life rating increases by a factor of ten for a 50C (122F) reduction in temperature. The allowable load increases by 40 per cent for a 10:1 reduction in life. For short-term loads the resistors are rated at 44 watt-seconds for the same mounting and ambient temperature, *i.e.*, 44 watts for 1 second, and so on.

The required life for a test dummy load is very short; a few hundred hours in many years. Also, good cooling can be provided by mounting resistors with metal fins touching the resistor body and immersing the resistor body and fins in oil. Even after considerable testing, the resistor body temperature need not exceed 50C (122F).

The shorter life requirement allows the power input to be increased by a factor of 1.44 and the lower temperature by a factor of 1.4. The rating now becomes 10.6 watts continuous, or 47 watts for 5 seconds. It's somewhat over 28.5 watts for 10 seconds and well over 20 watts for 20 seconds. (Ambient temperature isn't a major factor for very short loads, but resistance change is.)

At high power there's another factor to consider. The maximum voltage rating of a 2-watt resistor is 500 volts, which will affect the mounting method.

choice of design values

Since some work is done at full 2-kW input, a dummy load that could accept this power for short test periods seemed desirable. A reasonable assumption for maximum linear efficiency is 60 per cent, so the desired rating was 1200 watts for 5 seconds, which would allow 300 watts continuously.

The voltage rating must be reduced by the peak-to-average ratio of the applied signal. Also, for high frequency power use some dielectric heating of the resistor body will occur. With some allowance for this, the rf voltage across a resistor should not exceed 250 volts rms.

At this voltage the minimum resistance for a peak dissipation of 47 watts is about 1325 ohms. A total of 26 resistors in parallel would be needed for a 50-ohm load. However, since this is not a standard value, some adjustment is necessary.

The local surplus emporium had no 2-watt resistors close to the 1200-ohm value, but there was a large bin of 470-ohm units at a very good price. A quick calculation showed that eighteen of these in parallel would give a resistance of 25 ohms, with two such banks in series giving the desired 50 ohms. Alternatively, two banks of fifteen resistors each

By R. P. Haviland, W4MB, 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

would give 60 ohms, nearly the mean between 50 and 75 ohms. The last combination was chosen for construction.

dummy-load circuit

While the dummy load is usually used with a wattmeter, an independent power check is sometimes useful. A built-in rf voltmeter can give this measurement. Because of the high voltage present, a voltage divider must be included. The circuit is shown in **fig. 1**.

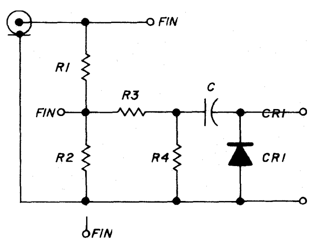
With the resistors in hand, the mounting and oil immersion problems must be worked out. For very short periods of operation, a 0.946-liter (l-quart) can would be suitable. However, the temperature rise is rapid, and more oil is desirable, so 3.8 liters (1 gallon) allows reasonable test periods.

construction

It's easier to show the construction than to describe it. **Fig.** 2 shows the load assembly. Three fins provide the connections for the two banks. One outer fin soldered to the can case provides most of the assembly support and the connection to the outer coax lead. The other outer fin is drilled to fit over the center conductor of the coax receptable, providing the remainder of the support. (Note that the three fins should be the same size to keep the voltage equal across the banks.) The voltage-monitoring components are mounted on a terminal strip with tip jacks used for the output.

If possible, use hermetically sealed connectors. Transformer oil, the preferred cooling medium, tends to migrate, and unsealed connections will always be oily. Even light mineral oil will migrate. Hydraulic brake fluid, the other possible coolant, also migrates. If you must use standard connectors, be generous with silicone rubber sealant.

Performance of the unit is good. The SWR is about 1.2:1, with a 50-ohm wattmeter, to above 30 MHz. However, at higher frequencies the uncompensated reactances affect performance. The apparent resistance on 144 MHz is appreciably reduced, and the SWR is undesirably high. While the unit is usable at vhf, it's basically a high-frequency design.



R1,R2 EACH 470 OHM 2 WATT, ALLEN BRADLEY RESISTORS IN PARALLEL
R3,R4 1k 1 WATT
C 0.001 μF CERAMIC DISC
CR1 GERMANIUM DIODE

fig. 1. Dummy-load schematic, including voltage-monitoring provisions.

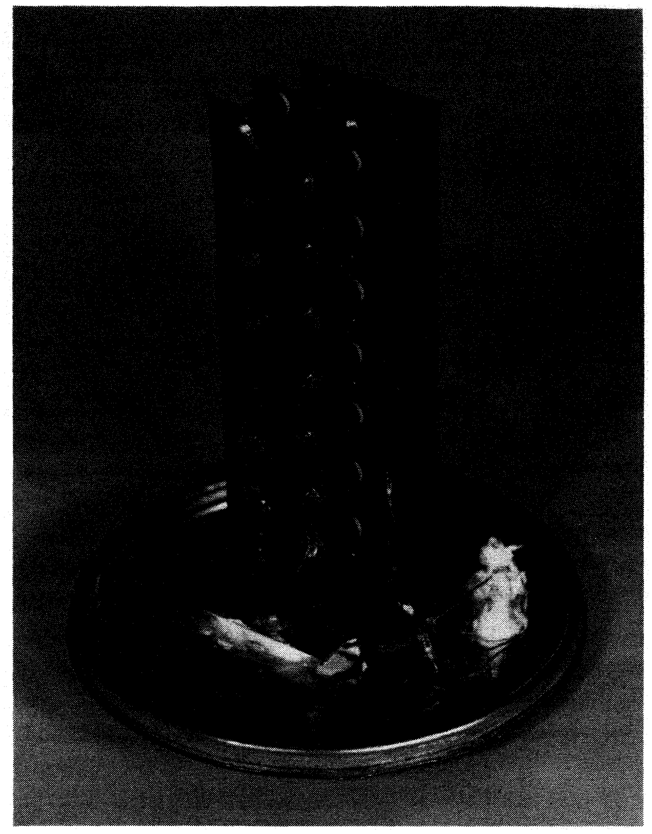


fig. 2. Construction of the gallon-size dummy load. Fins are made from copper flashing. The 2-watt resistor leads project through holes in the copper, are bent over, and then are soldered. The fins should touch the resistor body. Resistor spacing should be two body diameters or more. The fins should clear the can sides by 12.5 mm (½ inch) or more.

variations

Many variations of the basic design are possible. Sixty resistors would handle a kilowatt transmitter for long periods. Exact 50- or 75-ohm loads can be worked out for SWRs around 1.05:1. Special resistances are possible, such as 8 ohms for audio amplifier tests. Stray reactance can be tuned out for vhf operation.

The unit shown has been operated many times with a linear amplifier running at 1-kW input for periods of 10 minutes or so. No measurable change in resistance values was found. The unit has paid its way: the linear is now free of TVI thanks to the testing time possible.

reference

1. Robert P. Haviland, W4MB, "Superfluous Signals," *ham radio,* March, 1976, pages 40-43.

ham radio

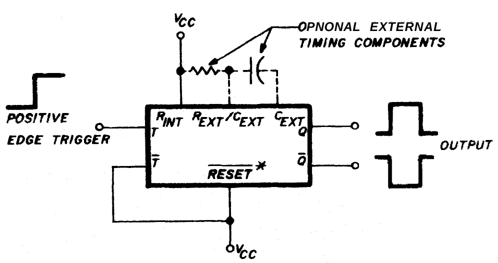
ITAL TECHNIQUES

multivibrators and analog input interfacing

Previous parts of this series have concentrated on only one multivibrator, the bistable, or fiip-fiop. This part will examine the monostable multivibrator, or one-shot, and look into ways of converting analog signals to digital levels.

A one-shot will create a single pulse of controllable width in response to a trigger. The trigger can be either a positive-going or negative-going state transition. There are two basic types: conventional and retriggerable. The latter will automatically reset and start the pulse again if a trigger arrives before the pulse is complete. Most of the former will inhibit any retriggering if the pulse has already started.

A one-shot symbol is shown in **fig. 1.** It has **Q** outputs and direct set and clear inputs as in a flip-flop.



* RESET (CLEAR) AND SET (PRESET) MAY BE EITHER ACTIVE-LOW OR ACTIVE-HIGH DEPENDING ON DEVICE

fig. 1. One-shot multivibrator symbology with optional external components.

Some devices allow a choice of positive or negative triggering. As in the flip-flop, unused control and input lines must be tied high or low depending on function.

Output pulse width depends on an RC time constant, either internal or external. The device is partly analog in operation, and a full description is found in texts.[1] Each one-shot IC is specified in width from a simple formula or a graph showing width versus RC combinations. Most all one-shots have internal resistors and capacitors that can be changed by external components. Most confusion in RC selection comes about by the three pins marked R_{int}, C_{ext}, and R_{ext}/C_{ext}. Connection rules are:

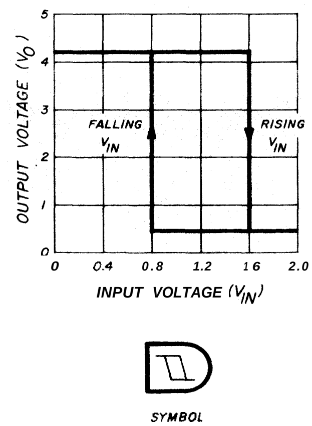


fig. 2. Schmitt trigger gate hysteresis curve and symbol.

Internal C only:	C_{ext} pin left open
External C:	Capacitor between C_{ext} and R_{ext}/C_{ext} pins
Internal R only:	Connect R_{int} pin to V_{cc}
External R:	R_{int} pins left open, resistor between V_{cc} and R_{ext}/C_{ext} pin

Internal capacitance and resistance are specified for each device. External components should be mounted as close as possible to the package; R and C pins are very susceptible to noise pickup, even with TTL.

External resistance may be variable for adjustment. The trimming potentiometer must not be wirewound for short pulses; winding inductance will change the time constant.

astable multivibrators

These are free-running oscillators with digital level outputs. Their use is generally restricted to digital VCOs (voltage-controlled oscillators). Most are of the emitter-coupled variety for stability." Details can be found in application notes or reference 1.

Most of today's timing is obtained from quartz crystal or LC oscillators, with or without dividers for lower frequencies. Such oscillators don't have the fast rise and fall times required for digital circuits. TTL

By Leonard Anderson, 10048 Lanark Street, Sun **Valley,** California 91352

"Not to be confused with ECL, or emitter-coupled-logic devices will not work properly with transition times of about 5 microseconds or more; internal circuitry will actually oscillate while a transition is made from logic 0 maximum to logic 1 minimum. CMOS is a bit more tolerant. There are two solutions.

the Schmitt trigger

A Schmitt trigger is a high-gain amplifier with feedback to give hysteresis. Hysteresis is a nonlinearity between input and output and, in digital form, helps to discriminate signal and noise when feeding digital inputs.

Some gates and inverters incorporate Schrnitt trigger circuits at each input. So do a few one-shots. Typical input versus output is shown in **fig. 2** for a TTL gate. The hysteresis symbol is used in gates and inverters having Schmitt inputs.

A rising input voltage will cause the output to "snap" from high to low when it crosses the 1.6-volt threshold. Internal feedback allows a very slow threshold crossing to change the output very rapidly. Schmitt inverters and gates are very good for interfacing an analog signal within input maximum voltage swing to conventional digital circuitry.

Other uses are given in **fig.** 3. Circuits of **figs. 3A** and **3B** are useful for automatically resetting a digital circuit during power-on. **Fig.** 3C may be used to debounce a single-throw switch, but with some caution. Values are for TTL, and closure time constant is much shorter than opening time constant. It will work well with CMOS, where resistor values can be equal and much higher.

comparators

Comparators are high-gain, wideband operational amplifiers with output clamps to restrict output voltage swings to digital levels. A typical application is shown in **fig. 4.**

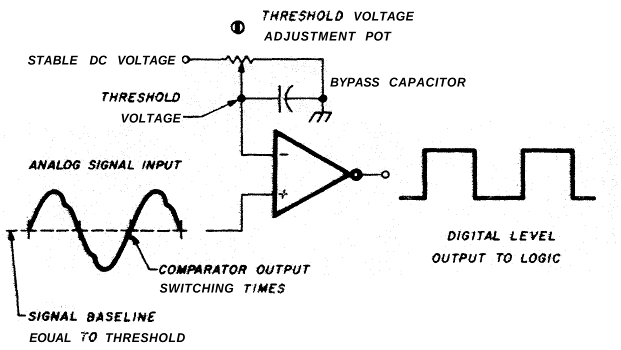


fig. 4. Typical comparator circuit.

Comparators usually need two more supply voltages. This requirement is offset by the ability to threshold signals with a dc baseline way off digital voltage limits. Input voltage thresholds and baseline shifting are the same as for lower-frequency operational amplifiers.

The main characteristic in choosing comparators is slew rate, or response time. Slew rate is rate of change of output voltage per unit time compared with input voltage. Slew rate should be as high as possible. The newer rating is response time and is found on data sheets as time-related graphs of output voltage change for different overdrives.

Overdrive occurs when the input voltage exceeds threshold voltage. Most high-speed comparators have nearly the same output voltage transition with any overdrive; the major change is a slight delay with small overdrives.

Packages such as the Motorola MC3430 have four separate comparators. The MC3430 needs only two voltages, +5 volts and −5 volts, and also has a common strobe control line. All four outputs are TTL-compatible and are three-state. When the strobe is high all outputs are in a high-impedance state, which allows wiring several outputs in parallel. Feedback may also be added for hysteresis.2

The Harris HA-4905 is also a quad comparator with an interesting supply connection. Two package pins are used for the comparator section and two more are used for the output circuits. One or two supplies can be used for the comparator with a differential of 5-15 volts. The output-circuit supply may be set to equal the digital-circuit supply for signal compatibility.

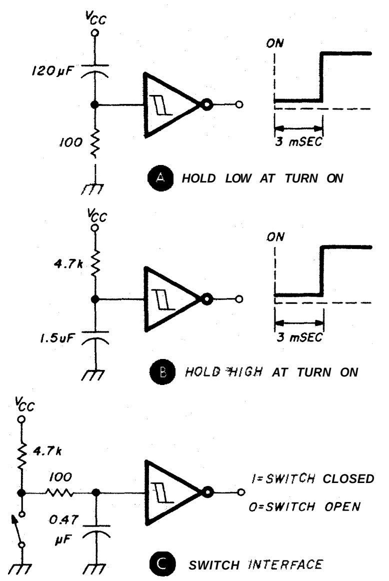


fig. 3. Schmitt trigger applications with inverters.

references

1. Jacob Millman and Herbert Taub, "Pulse, Digital, and Switching Waveforms," Chapter 11, McGraw-Hill Book Company, 1965.
2. Motorola Semiconductor Data Library, Volume 6, Series B, Motorola Semiconductor Products, Inc., 1976, page 6-23.

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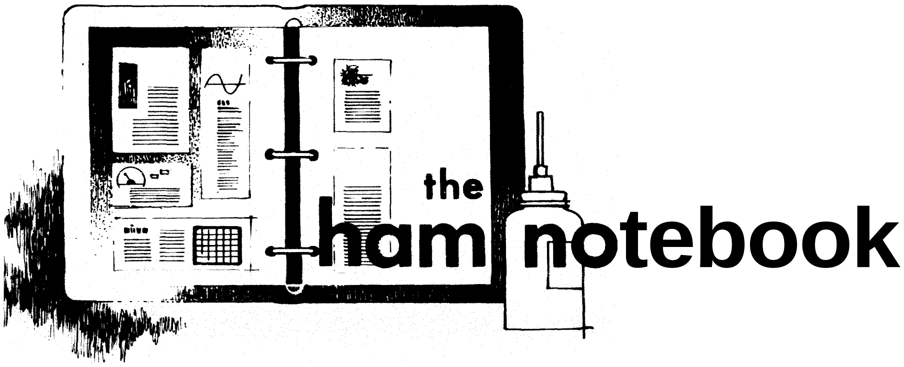


ıcreased break-in delay range for the Heathkit HW-8

The Heathkit HW-8 transceiver reak-in delay time constant is such ıat, even with relatively fast keying nd maximum time delay set in, the g returns to the receive mode after very word. This is fine for rapid reak-in, but most CW operation is ot rapid break-in, and all that's ac-omplished is unnecessary relay wear nd operator fatigue. The fix is sim-ıle: Change C92 from 10 μF to 50 μF, from 10 μF to 50 μF, the transmitter and sidetone monitor oscillator keying develop tails because of the coupling back through the solid-state devices. This side effect may be eliminated in the transmitter keying circuit by lifting the end of R64 nearest the back of the printed circuit board, where wire "W" connects, soldering the anode lead of a IN4148 silicon diode (Radio Shack 276-1122) in the

In addition, receiver muting is delayed excessively by the increased size of C92. This may be corrected by placing a 4700-ohm resistor in parallel with C43. Remove C43 from the circuit board, solder it across a 4700-ohm resistor, and place the pair where C43 was originally installed. (Refer to **fig. 1**.) All resistors are ¼ watt.

After making the modifications,

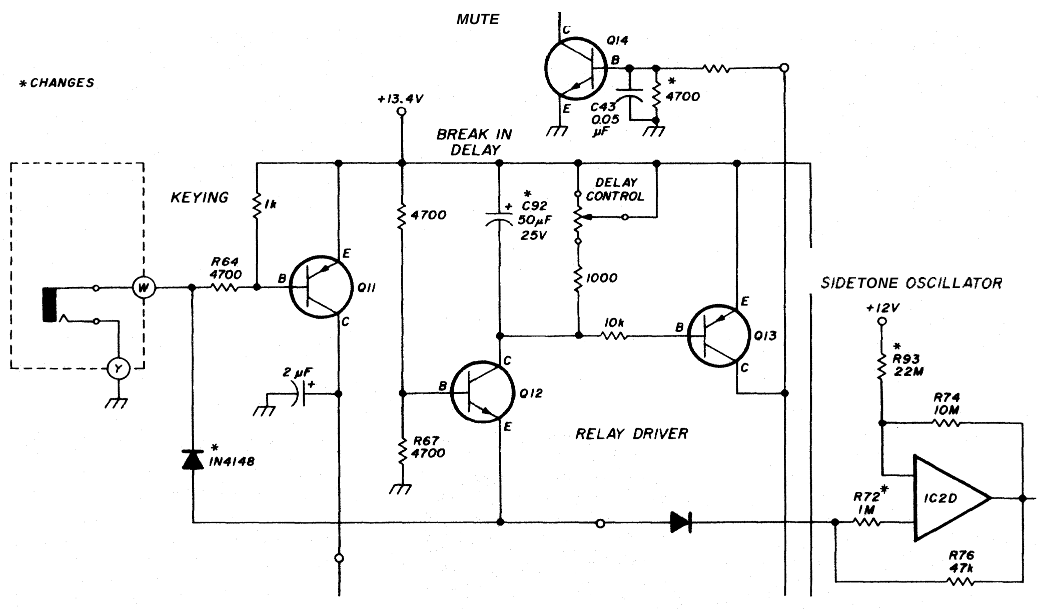


fig. 1. Changes to the Heathkit HW-8 for increased break-in delay range.

25 volts; R72 from 3.3 megohms to 1.0 megohms; R73 from 10 megohms to 22 megohms; parallel C43 with $700 ohms; and add a diode in series with R64 (see **fig. 1**).

The time constant is controlled by C92. However, when C92 is increased empty hole, and soldering the cathode lead of the diode to the unconnected end of R64. Reconnect wire "W" to the junction of the diode and R64, above the board. In the sidetone oscillator circuit change R72 and R73 as indicated above.

the DELAY CONTROL potentiometer will provide a wide range of time constants. Each operator should adjust the DELAY CONTROL to suit his particular keying speed, with high speed requiring the least delay time.

John Abbott, K6YB

Collins 516F-2 high-voltage regulation

Shortly after acquiring my Collins S-line equipment, I became aware that the CW waveform as viewed on the SB-610 monitor scope left something to be desired. A definite trough appeared in the waveform, which is characteristic of a high-voltage power supply with insufficient regulation (fig. **2).** The as-built, 10-μF filter is sufficient for SSB operation. However, during key-down CW operation, with the amplifiers running full-bore and exhibiting a widely varying load, there's a need for increased capacitance in the filter section. I found that a minimum of 25-μF of filter capacitance provided the smooth waveform shown in fig. 3.

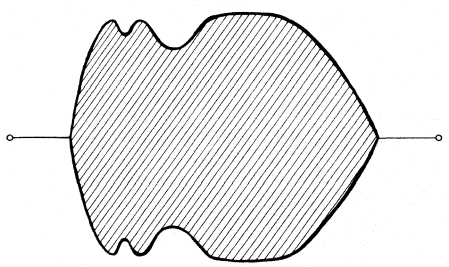


fig. 2. Monitor scope display of Collins S-Line CW waveform showing insufficient high voltage supply regulation.

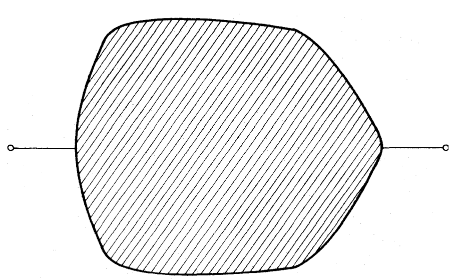


fig. 3. Monitor scope display of Collins S-Line CW waveform showing good high voltage supply regulation, which is obtainable in practice.

With under-chassis space being at a premium, three 80-μF caps (Sprague TVA 1716) were used to replace the originals. These capacitors will provide over 2½ times the former capacitance and will fit into the chassis-mounted clamps without protruding below the chassis bottom. If other types or sizes are used, be sure sufficient chassis clearance remains.

Before removing the original caps, observe the physical wiring arrangement. Place the new capacitors in the same direction as those to be removed. The original capacitors are manufactured with terminals instead of wire leads, but no difficulty should be encountered if the full-length leads of the new units are used and only the negative lead of the ground-end capacitor (C4) and positive lead of the high-voltage bus capacitor (C2) are trimmed after the connections have been made. Be sure to insulate the negative/positive interconnecting leads of the three capacitors. They are ***not*** at ground potential, and they pass near the capacitor mounting brackets. See fig. **4.**

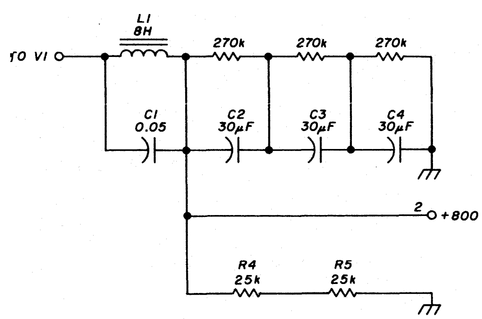


fig. 4. Modified Collins 516-F2 supply. High-voltage filter caps C2, C3. and C4 are replaced with 80-μF caps.

After installation, make certain that no short circuits exist. When the supply is reconnected to the 32-S ( ), ensure that the bias pot adjustment is correct.

Paul **Pagel, NIFB**

loss of torque in HAM-M rotators

Several years after I purchased my first HAM-M rotator, it started to slow down. At first I thought the slowdown was because of the extra antennas I'd added, so I increased the voltage with another transformer in series with the internal transformer. This helped for a while, but the rotator gradually slowed down again until it took five minutes to rotate 360 degrees!

After several inquiries and investigations, I found the cause. It seems that the electrolyte in the motor capacitor, C2, a 120-140 μF, 50-Vac capacitor, dries out in time. I installed a substitute capacitor, and the antennas almost took off. I removed the extra transformer. Even then, with a heavy antenna load, the 1-rpm speed was fully restored.

The motor-start capacitor is easily changed. It's conveniently located in the control box. The as-built Cornell Dubilier capacitor part number is 51172-10, priced at $2.50. A higher quality (but possibly larger in size) replacement is probably available from your local electrical or motor supply house. The capacitance should be at least 120 μF.

I've experienced this problem several times. In one case, it happened after only one year on a new HAM-M-II.

Joe Reisert. **W1JR**

phono plug wiring

Soldering the braid of coaxial cable, such as RG-58/U, to the shell of a phonoplug can be a messy and sometimes frustrating job. Excessive heat may be applied, resulting in the center-conductor insulation's being damaged, causing either a shorted plug or the possibility of future intermittent trouble.

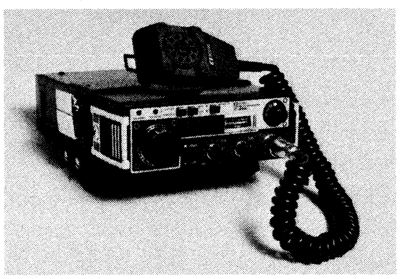
A neat and virtually short-proof connection may be made using heat-shrink tubing. Instead of soldering the coaxial braid to the shell of the plug, a 2.5-cm (l-inch) length of 9.5-mm (318-inch) diameter heat-shrink tubing is slipped down the coax and over the braid until even with the end of the shell. The heat of a match finishes the job quickly, neatly, and permanently.

Paul **Pagel, NlFB**

NEW products

For literature on any of the new products, use our *Check-Off* service on page 110.

IC-280 mobile transceiver



The versatility of a microprocessor is exemplified in the introduction of the ICOM IC-280 fm mobile radio for 2 meters. Referred to as the "remotable" radio, the IC-280 actually comes assembled for immediate operation as one box. However, the same radio may be operated as separate units by removing the head and connecting the optional remote cable to each unit. You can then mount the central head in a small place where almost no other radio will fit.

"Remotability" is not the only reason to have an IC-280. The microprocessor covers all 4 MHz of the 2-meter band, plus some at both ends, in 15- or 5-kHz steps which are selected by the user or the microprocessor. In addition, there are three memory channels to store any frequency which can be programmed on the dial.

The modular 10-watt output stage has plenty of power to drive the most popular amplifiers to full output. The continuous display of frequency in transmit, receive, or memory position makes the IC-280 the easiest-to-use fm radio, and the best-performing fm radio that ICOM has designed to date. All ICOM dealers should have them in stock and on display now. See one today at your authorized ICOM dealer. or contact ICOM East, Inc., 3331 Towerwood Drive, Dallas, Texas 75234; or ICOM West, Inc., 13256 Northrup Way, Suite 3, Bellevue, Washington 98005.

multiple-output lab power supply

A new lab power supply, capable of functioning as three separate power supplies and featuring an exclusive automatic tracking circuit, has been introduced by the B&K Precision product group of Dynascan Corporation.

The new Model 1650 offers a 5-volt dc, 5-amp output, and two separate 25-volt dc outputs at 0.5 amp. The exclusive automatic tracking circuit allows the B output to "track" voltage changes of the A supply.

Designed for use with solid-state equipment where both linear and digital circuitry may be encountered, the unit's three outputs are completely isolated, offering the user full versatility to connect the outputs in series or parallel.

The valuable tracking circuit allows proportional control of the B supply when the A supply output is varied. When the TRACK mode is selected, if the B control is set at 100 per cent, the voltage level selected with the A control will appear at both the A and B outputs. If the B control is set to a 50 per cent position, the B output will be 50 per cent of the voltage at the A output. Tracking is controlled by means of a pulse-width-modulated control signal, which is coupled through an opto-isolator. This unique BBK Precision design permits complete electrical isolation of both supplies in the tracking mode.

The tracking feature of the 1650 allows it to be used as a substitute power source for breadboard and prototype circuits and other equipment. For test purposes it can provide single or simultaneously varying voltages to observe operating effects on the circuit under test. The tracking feature can also provide positive and negative voltages for operational amplifier circuits and offers a convenient means of evaluating the performance of differential amplifiers with voltage changes.

The 1650 features automatic current limiting and short-circuit protection on all ranges and outputs. IC control circuits maintain the highest reliability and stability. All power outputs use color-coded, heavy-duty, six-way binding posts.

The BBK Precision Model 1650 multiple output power supply is a money-saving alternative to separate supplies. It's available for immediate delivery at local BBK Precision distributors for $275.00. For additional information, contact BBK Precision Sales Department, 6460 West Cortland Street, Chicago, Illinois 60635.

DE-130 electronic keyer

The new DE-130 Digital Electronic Keyer, by Dynamic Electronics, Inc., is designed to provide all the features required of a high-quality keyer at a minimum cost without additional accessories. For example, an ac power

supply is included, eliminating the need for batteries or an add-on supply. The cost of an external paddle assembly is not necessary since an electronic paddle called a "touch key" is included. The electronic

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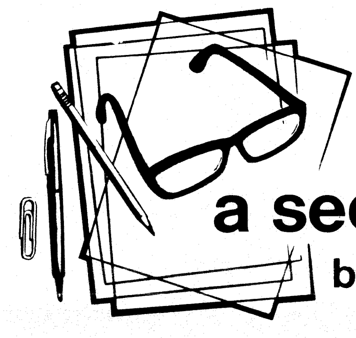

a second look

by Jim Fisk

This is the time of the year when many amateurs are working on their antenna systems for the coming DX season. If you're considering installing a new tower, however, there are some potential legal problems you should consider, even before you dig the hole for the base and pour the concrete. Mervyn Hecht, Attorney-at-Law and a Trustee of the Personal Communications Foundation (PCF), discussed four of the possible problems in a recent PCF bulletin:

Error in calculating the property line. "This can come about in two ways. First, the property line may not be where you think it is, especially on hillside properties. Imagine how expensive it will be to move your misplaced tower if your neighbor discovers it is on his property and will not let you keep it there! If there are no property line survey marks you can rely on, and the tower is to be positioned anywhere near a property line, have that line surveyed before you dig the hole for the base.

Secondly, don't forget that the antenna will be wider than the tower. If the tower is right next to a property line, the antenna will protrude into your neighbor's "air space." If that happens, your neighbor has the right to make you move the tower.

Blocking the neighbor's view. This problem seems to crop up primarily on hillside properties. It may seem reasonable for the valley dweller with a hillside near his house to place his tower on the hillside, above the surrounding hills, but to a person who lives on the top of the ridge an antenna sticking up at the edge of his yard — so he has to look between the director and the reflector to see the setting sun — can be very frustrating. The legal aspects of blocking the view (or sunlight) are now in a state of change, but the trend is toward recognition by courts of these rights, and away from the absolute property rights characteristic of earlier times.

The radio operator must recognize the potential problem and try to position his tower and antenna where it will not interfere with any often-used view nor block the sunlight. If there is some problem in avoiding this result, consider an alternative such as a) a motorized or hand-cranked tower so the antenna can be lowered when not in use; b) a smaller sized antenna; c) meeting with the potentially offended neighbor to obtain the neighbor's permission to erect the tower on some less offending spot owned by the neighbor.

Interference with underground or property line easements. Many property titles are legally "burdened" by deeds to telephone companies, electric companies, cable television operators, and other utilities which give these services various rights. Usually these rights are to install (either under or over the ground) various cables and pipes, and often to enter onto the property to replace, service, and check these installations. These easements are often so broad that although you own the property — and pay the property taxes — you have given up the use of these (usually five-foot wide) strips of land. If you install anything which blocks the utility company's rights, or prevents them from exercising the rights granted, you may be required to move your tower. Even if the utility is not using the easement now, it may in the future (perhaps a few weeks after you install the tower), or the utility may just be run by difficult people who are intent on enforcing their rights.

Causing damage to the neighbor's structure. There are three general ways I have seen this happen. First, mechanical drilling, such as with a jack hammer, which can cause shock waves to nearby structures. Second, digging a hole may result in loss of lateral support which can cause unexpected land movement resulting in damage to nearby structures. By far the most common major problems I have seen resulting from property line excavation, however, are related to water drainage. Particular care should be taken not to change any drainage pattern because, during a rain storm, the slightest change can cause thousands of pounds of water to accumulate in unexpected places."

As if this is not enough to think about, Attorney Hecht further notes that although he has "not even mentioned deed restrictions, height limitations, airport clearance and lighting regulations, city permits, convenants running with the land, or neighbors running after you with a shotgun . . ." he does not wish to discourage radio amateurs from installing a tower. Just be aware that if you are going to dig a hole for a tower base, don't dig near a property line unless you take special care to avoid the special problems that can arise.

Jim Fisk, **W1HR**
editor-in-chief

comments

lightning protection

Dear HR:

Many thanks for publishing K9MM's excellent article, "Lightning Protection," in the December issue. It is hoped that Amateurs will heed the warning.

Author Becker suggests grounding wooden poles by placing a grounding conductor on the pole — this is quite effective, but it may act as another antenna, radiate minute ground electrical noises, and reradiating some transmitting power. It is suggested that the ground wire be cut into 10-foot (3-meter) lengths with each end bent 180° on a half-inch radius (except the top spike) and stapled to the pole with each loop separated by 2 mm (1/16 inch) from its neighbor. This spacing is adequate for Amateur use.

It is further suggested that Amateurs become acquainted with the provisions of the National Electrical Code (NFPA no. 70) published by the National Fire Protection Association; it lists specific requirements for station protection. This document is available at most public libraries and is quoted briefly in part on page 645 of the 1978 *ARRL Handbook.* Compliance with the provisions of the NFPA Code may avoid insurance adjuster hassles if lightning plays havoc with your property.

Marchal H. Caldwell, Sr., W6RTK
Sacramento, California

lightning protection

Dear HR:

I have recently become a professional associate in the Lightning Protection Institute, a group of professionals, installers, and equipment manufacturers who have joined together to promote lightning protection and the safe design and installation of lightning protection systems. For this reason, and because I am a Radio Amateur with several wire antennas, I found the article in the December issue of *ham radio* particularly informative, accurate, and up to date as it discussed lightning theory, protection against direct strikes, and protection against surge and transient high voltages.

As is usually the case, the article was well written and documented, and, in my opinion, serves as the best source of lightning protection information that I have seen to date for the Radio Amateur. Unless I miss my guess, this is a better treatise by far than that found in the ARRL *Radio Amateur's Handbook.* Quite frankly, I feel it is so good that those editors ought to consider lifting the article and using it in the *Handbook* in toto.

Gerald B. Curtis, WB2FBL
Westrnont, New Jersey 08108

OSCAR 10-meter downlink

Dear HR:

Now that the 10-meter band is practically fully open for worldwide direct communications, perhaps it's time to curtail Oscar 10-meter downlinks. The reason for my position is that 100 kHz is a lot to take out of the usable hf spectrum; now that the Soviets have orbited their two vehicles this has increased the "forbidden" territory to 200 kHz (29.3 to 29.5 MHz).

The pass time over any given area may be good for only 15 to 20 minutes of acquisition, but since direct skip prevails over such a wide area, in all directions, this means that the effective interference chances are multiplied by the number of passes and orbit time.

There are several other factors which seem to make satellites with 10-meter output unacceptable, one being the large number of Civil Emergency Preparedness stations which have been on these frequencies since 1945, another being the many low-power stations which have been squeezed out of the lower part of the 10-meter band.

I have suggested to AMSAT that if they lead the way now by deleting the 10-meter downlink, the Soviets may do likewise on their next venture. AMSAT is to be congratulated for serving 10 meters during the sunspot time, but the time to discontinue is now!

Samuel H. Beverage, W1MGP
North Haven, Maine

Dear HR:

My Digital Display, which appeared in the March, 1979, issue of *ham radio,* occasionally will reset to 999.9 instead of 000.0. This problem is easily eliminated with the following simple circuit changes:

1. Lift pin 3 of U6 from ground and reconnect it to pins 2 and 8 of U6.

2. Disconnect the line between pins 6, 7 of U12, and pin 10 of U9. Reconnect pins 6 and 7 of U12 to pins 2, 3, and 8 of U6.

This ensures U12 will always begin a count cycle in the same state.

Frank C. Getz, Jr., N3FG
Media, Pennsylvania

display SSTV pictures on a fast-scan TV

Using an integrated circuit CRT controller plus software to provide the interface between a computer memory and a fast-scan TV

The missing link for all my SSTV projects1 was the interface from computer memory to a normal fast-scan TV set. Although commercial units are available, their cost tends to be high and designs complex. In this article I will present a hardware design that is simple and can be easily constructed for less than one-hundred dollars.

Product goals. Prior to starting the project I set a few goals for the interface:

1. The hardware design should be simple, and make use of a minimum of components.
2. The hardware should be reproducible, and flexible enough to allow for future expansion.
3. The software should be modular and use as much relative addressing code as possible and run in ROM.

Obviously, to accomplish a task like this took careful planning and much thought. The item which made the whole project possible was the new family of Large Scale Integration (LSI) chips called CRT controllers.

Specifications. The hardware and software package in this article will accomplish the following:

hardware

1. Display a digitized slow-scan TV picture located in RAM on a normal TV set with 128 pixels/line and 16 gray levels, expandable to 256 pixels/line.
2. Allow for transmission of medium-scan Amateur television2 in any format.
3. Provide the flexibility to enhance the digitized picture by simple hardware program commands which include interlaced or noninterlaced video and fast-scan picture zooming.

software

1. Receive or transmit Amateur Radio SSTV with 128 or 256 pixels per line and sixteen gray levels.
2. Zoom by the use of software to transmit on SSTV or display on fast-scan TV any one of five quadrants of a digitized picture.
3. Receive quarter-framed SSTV pictures, and display them on fast-scan TV or transmit the pictures on a composite single-framed picture.

In order to accomplish these feats, you must first have some means of digitizing the picture and interfacing the computer with the Amateur Radio receiver. Fig. 1 is a block diagram of my entire computer configuration. The detailed design of my SSTV analog interface board is contained in my previous articles.[1]

general background

Fig. 2 is a block diagram of the computer video-interface card which is used to display the fast-scan TV. To help you understand the function of this card, I'll discuss how a microprocessor functions in this type of application.

A microprocessor is generally a complex logic element which moves data to and from memory or ports external to the system. In this process, the data could be altered in any manner. In my application, data is first moved from a port which contains an analog-to-digital converter attached to an SSTV demodulator. The data is then formatted in the microprocessor memory in such a way that if it is accessed in a serial manner, converted to an analog

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signal, and mixed with sync pulses, you could display the information as a picture on a television set. This movement of data is accomplished by a series of events called instructions. The art of creating these instructions is called computer programming.

If the microprocessor were fast enough, this process could be accomplished by software with a small amount of hardware. To date, none of the commonly available low-cost microprocessor chips are fast enough. Obviously, if you would like to display pictures from memory, some sort of hardware device must be constructed. As a result of this requirement, I designed the following video-interface card.

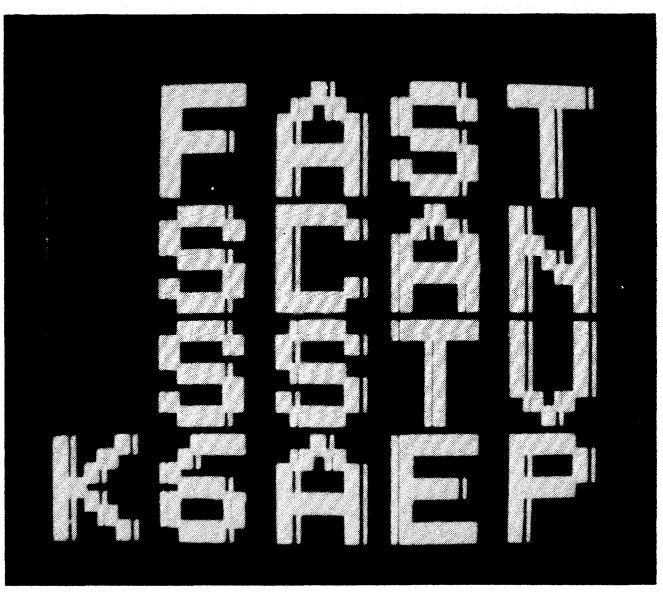
CRT controller block

At this time, four manufacturers have developed LSI CRT controller chips. They are basically complex devices which replace the function of approximately forty ICs and combine their functions into a single package. These chips are designed to interface with digital computers and are used mainly in communications terminals.

Each controller typically contains a number of functional units. **Fig.** 3 is a block diagram of the MC6845 CRT controller. These devices are used to address memory as a data refresh buffer, serialize the data from RAM, and mix it at the correct time with TV sync pulses. Each manufacturer's chip design has different features. Some of these features range from built-in character generators to full-color graphics.

Since an SSTV fast-scan display application consists of addressing large blocks of RAM (up to 16k), only one CRT controller met this requirement, the Motorola MC6845.[3,4] The MC6845 is unique, because the chip has more versatility than do any of

SSTV characters as seen on fast-scan TV.



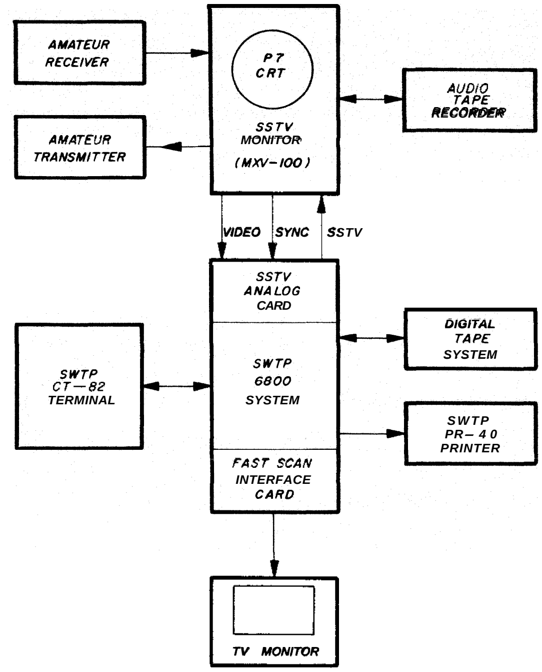


fig. 1. Block diagram of the computer and SSTV systems.

the other controllers. This IC contains eighteen programmable registers which control the vertical1 horizontal timings and refresh RAM address, number of scan elements, and lines per picture.

The unique feature of the chip is that the software controls how the chip performs. For example, calculations of the various controller chip-register values were made by fine tuning with an oscilloscope and a TV monitor to achieve optimum results. This feature is similar to changing variable capacitors or potentiometers in older hardware technologies. Software is much faster and more reliable than the older techniques. With software you can achieve textbook waveforms with a little experimentation.

memory accessing

Another subject which should be briefly discussed is direct-memory accessing (DMA). DMA is a technique for reading or writing to or from memory at a much faster rate than allowed by the microprocessor software. The three DMA techniques that can be implemented in microprocessors are:

1. Halting the processor.

2. Cycle stealing.

3. Multiplexing the CPU and DMA.

Each technique has its own advantages. When planning the project, two DMA methods were investigated as possibilities in this application, halting the CPU and multiplexing. The cycle-stealing method would not be fast enough for the refreshing or accessing of a video display system.

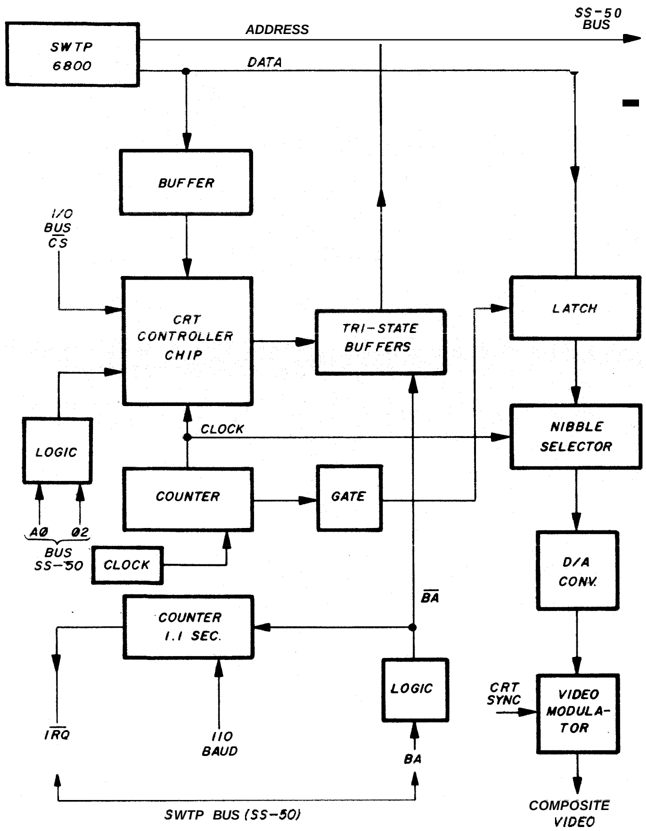


fig. 2. Block diagram of the video-interface card.

Multiplexing had a big advantage since the CPU and display could share the picture RAM area at the same time. This multiplexed-type of operation is easily implemented in the 6800 microprocessor since both phase 1 and 2 cycles have the same durations. In this configuration the CRT controller chip would drive the CPU. The CPU would operate as normal, addressing RAM at a phase 2 cycle time. The display would access memory during phase 1 cycle time. This method was considered the most desirable, however the standard SWTPC 6800 MP-A CPU card derives its baud rate from the CPU clock. Therefore, this DMA approach was not chosen. If it were, the baud rates derived from the CRT controller clock would be nonstandard.

The halting of the CPU was the method I chose for my card design. Additionally, halting of the CPU is the easy type of design to develop. All that has to be done to halt the CPU is to ground the halt line on the bus (SS-50). In my application, I decided to be a little more elegant than just grounding the halt pin to start the operation. The 6800, like many of the more popular microprocessors, has an instruction called wait (WAI). When the microprocessor executes this command, it waits or does nothing until it receives an interrupt (or signal) from another source. This is the method I chose to halt the CPU. While in wait state, my special card takes control of the bus and provides the signals to RAM to sequentially take the pixels (picture elements) and transfer them to the TV monitor. When the microprocessor receives the interrupt, the interface card is disengaged and normal microprocessor operation resumes. The SSTV hardware interface to the software is an analog-to-digital (A/D) converter and a digital-to-analog converter (D/A). The A/D is used to receive the SSTV pictures, and the D/A is used to transmit SSTV. Both units are connected to an MC6820 PIA parallel port. Fig. 4 describes the 6820 bits assigned by the software and used by the analog card.

video-card theory

Fig. 5 contains a schematic diagram of the video-interface card. Some of the major components have been previously discussed, with the details to follow.

Initialization. Before the whole process can start, the CRT controller chip must be initialized. This initialization tells the controller chip the type of video which will be displayed and where the picture is in the microprocessor memory. To accomplish this, registers in the controller chip must first be loaded with data by software, with hardware selecting the chip at the correct time.

Chip selection in the SWTP 6800 system is conveniently located on the system mother board. I found it convenient to run a wire from the I/O selection line on the small socket to the $\overline{CS}$ line on the CRT controller chip. This left only the chip ENABLE control decoding which had to be installed on the card. Since data is valid only during phase 2 of the 6800 clock cycle, only two signals had to be ANDed, W/$\overline{R}$ and phase 2. The least significant bit of the address line (AØ) was connected to the chip RS input. Thus when A0 is true, the register number can be loaded; when off, the data is loaded into the controller chip.

One important point should be noted when debugging this circuit: It does nothing without initialization. When initialized, you can observe the various output lines changing.

Buffering. Since the card shares the same bus as the microprocessor, all lines must be buffered with either tri-state or low-power Schottky ICs (74LS). Tri-state devices have three output states, ground, high, and floating. When floating, these chips exhibit a high impedance which is almost an open circuit. When not in use, the video card floats on the bus.

When a WAI instruction is executed by the microprocessor, the bus available (BA) line rises and the CPU card floats. I found that the SWTP MP-A CPU card has a bug and requires a slight modification to properly DMA the bus. Apparently Southwest designed the card for DMA; however, they allowed one buffer to be in the high state. You can correct this problem by cutting the lead on U12 between pins 4 and 5. Leave the connection on pin 5, but connect a wire from pin 4 to pin 2 of U12. Since many manufacturers produce 6800 CPU cards, I suggest that you consult your schematics to ensure that your card will tri-state the bus when the 6800 BA line goes positive.

One line that my video card does not control during DMA is R/$\overline{W}$. Since the video card is always reading RAM during refresh, I let this line float positive. The main reason for doing this is that I ran out of buffer modules on the three tri-state buffers (U16, U17, and U18). A more desirable state would be to condition this line positive.

Memory addressing. The memory addressing of the video card is the simplest part of the whole operation. All of the difficult work is accomplished by the CRT controller chip. When initialized, the memory is addressed at one half the pixel rate (650 ns). The appropriate sync pulses are also outputted from the chip and mixed with the video.

Data flow. The most important part of the entire video card is the flow of RAM data to the TV set. Since the CRT controller chip does the difficult task of addressing memory, all that has to be done is to serialize the information. When RAM is addressed, the data is presented to two latches (U5 and U6) where each holds four bits of a byte (nibble). The data is latched at the end of an addressing cycle, with both latches feeding a multiplexer (U7). First the lower nibble then the upper nibble are fed through the multiplexer. Since each pixel is a nibble, and a byte contains two pixels, the data rate is twice the memory-accessing rate. This allows the use of 450 ns memory to refresh a picture with 128 pixels per line. If you wish to refresh a larger buffer with more pixels per line (256), you must make the following changes.

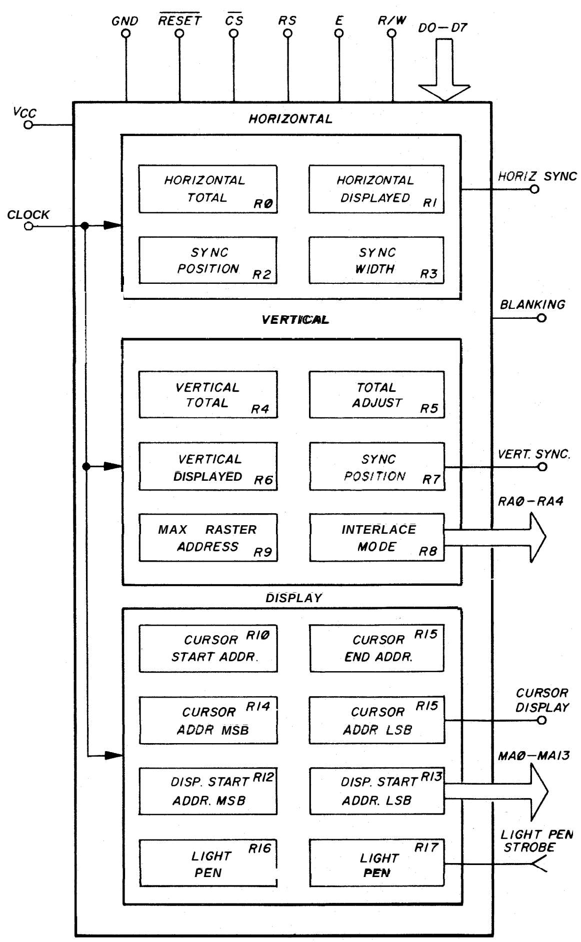


fig. 3. Diagram of the functions internal to the MC6845 CRT-controller IC.

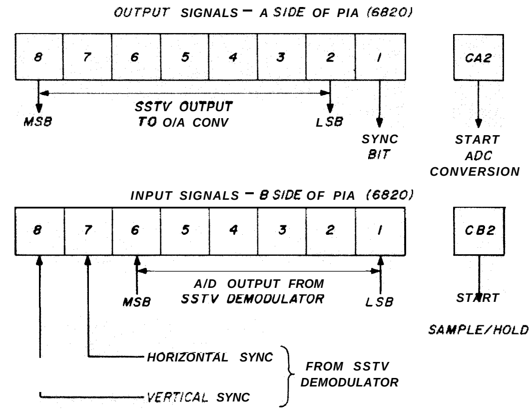


fig. 4. Format of the digital signals into and out of the 6820 PIA.

1. Change the crystal frequency to twice the rate, 12.2888 MHz for 60-video standards.
2. Re-initialize the CRT controller chip for 128 characters per line and a memory start address of 0000,
3. Load a picture in RAM from address 0000 to 3FFF (16k bytes).
4. Place 16k of static 250-ns access time RAM at locations 0000 to 3FFF.

These steps will produce a fast-scan television picture with twice the resolution of comparable commercial units.

video modulator

The video modulator was designed for its simplicity and low cost. At first I considered using a digital-to-analog converter for the modulator. This method, on the surface, appeared to be a good approach. However, the component count and cost were considerably higher than I expected. **Fig. 6** is a plot of the excellent linearity of this modulator. The four 1 per cent resistors could be replaced with 5 per cent values by a selection process. Only one resistor had to be fine-tuned in the modulator circuit, the 3.3 kilohm resistor which controls the 70-to-30 per cent

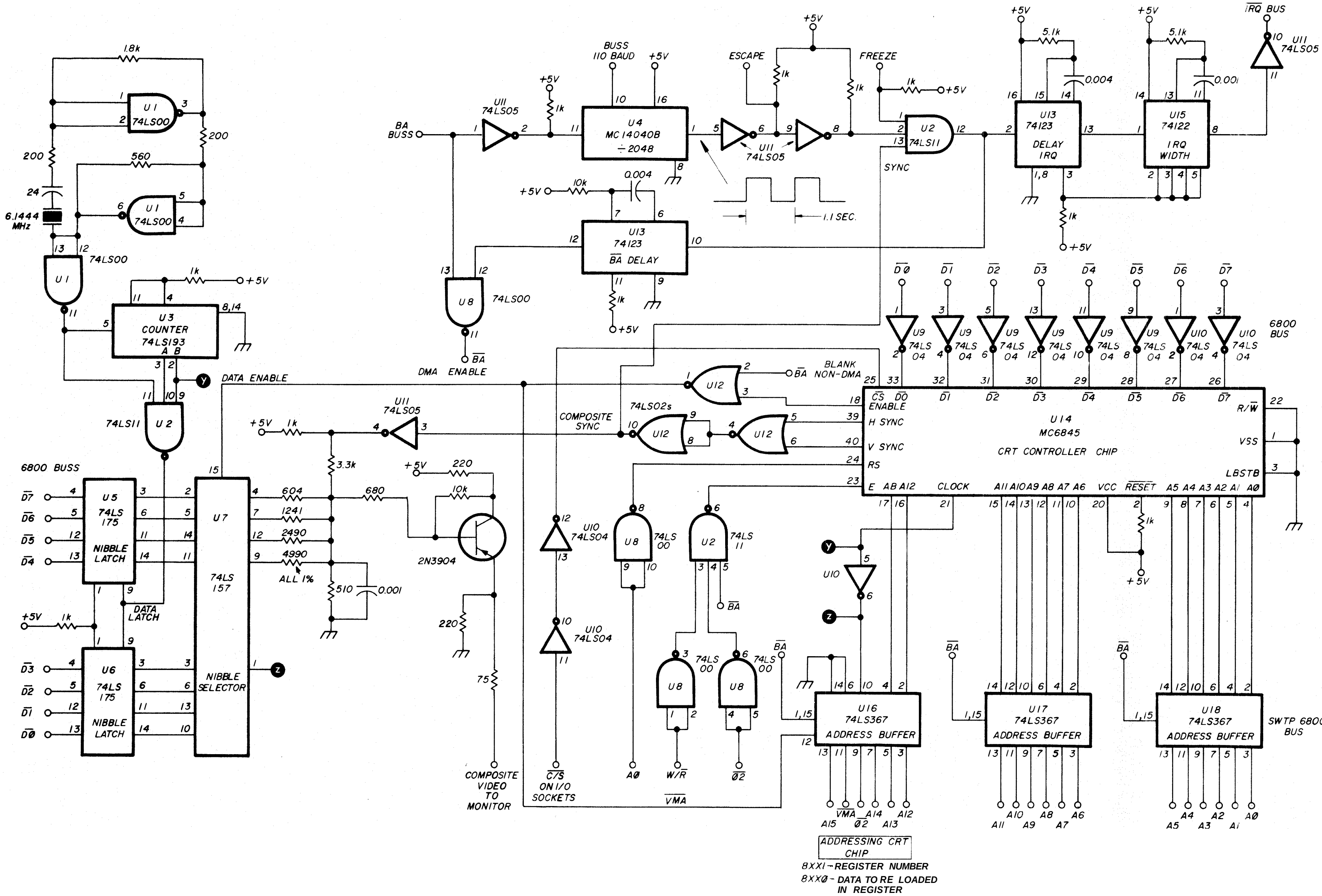


fig. 5. Schematic diagram of the video-interface card.

relationship between video and sync. If the sync is too low or high, change the 3.3-kilohm resistor in series with U11. The sync portion of the video should be 30 per cent of the total swing of the video signal.

counters and timers

The main counter on the card is very simple. The clock signal is generated by U1 and is divided down by U3. The only tricky part is the nibble latch signal derived from U2. Since the data is valid only at the end of the addressing cycle, it must be latched as close to the fail of the address as possible. Calculations show that if 250-ns memory were refreshed, 256 pixels/line could be refreshed and latched with this scheme. If the latch is marginal, inverters could be placed in series with the line for additional delay.

The timer is used to return from a wait condition of the CPU. I derived this return signal by counting down the 110-baud rate on the SS-50 bus by 2048. This method gave me an interrupt every 1.1 seconds. I serviced this interrupt by software, with the interrupt generated by grounding $\overline{IRQ}$ on the bus line for approximately 2 μs (U15). The visible effect on the

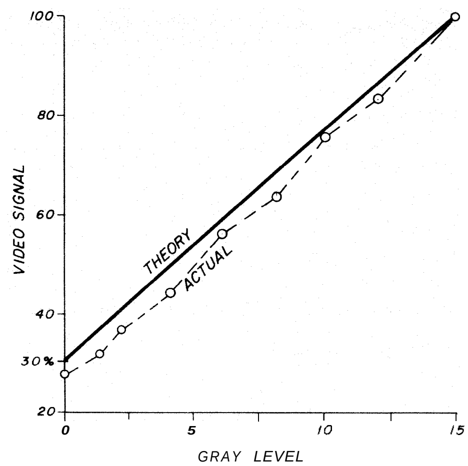


fig. 6. Diagram of the linearity of the fast-scan modulator in percentage of video signal to gray level.

issuing the interrupt. This time could be reduced since the capacitor used in the single shot was the only value I had in my junkbox when the card was developed. Two external control signals were placed on the I/O connector of the card. These signals are

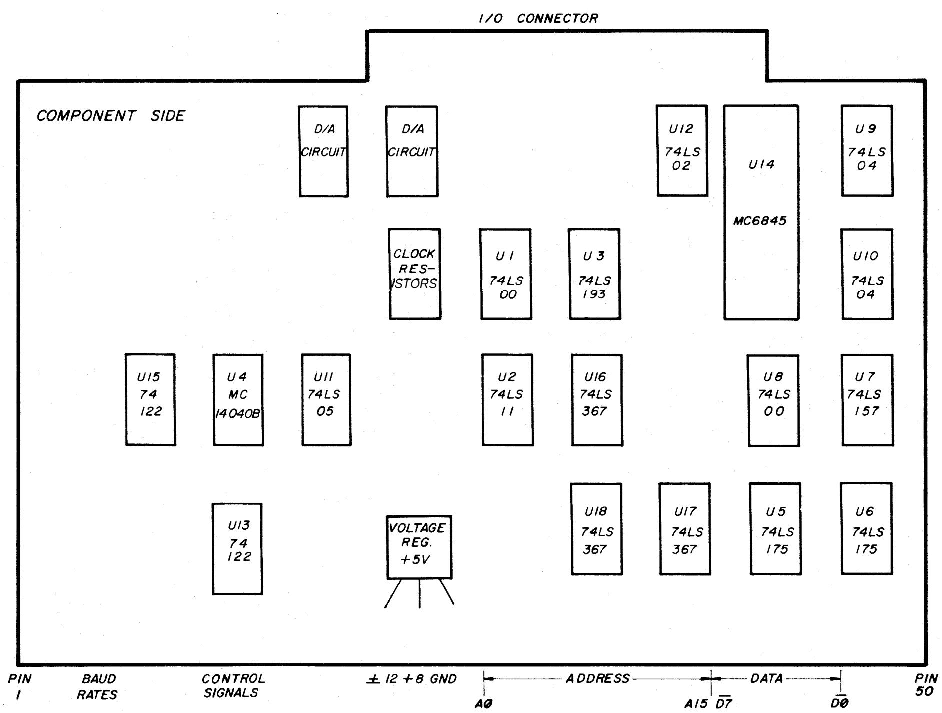


fig. 7. Layout of the video-interface circuitry on the prototype board.

fast-scan TV is one missing scan line every 1.1 seconds.

U13 is used to drop the DMA at the correct time and issue the interrupt. The BA delay side of U13 causes the video card to float 8 μs prior to and after

freeze and escape. Freeze is used to display a continuous picture on the TV screen by grounding the central line with a switch. Escape can be used to return to the CPU from DMA by grounding this line with a pushbutton. The push button will cause an interrupt.

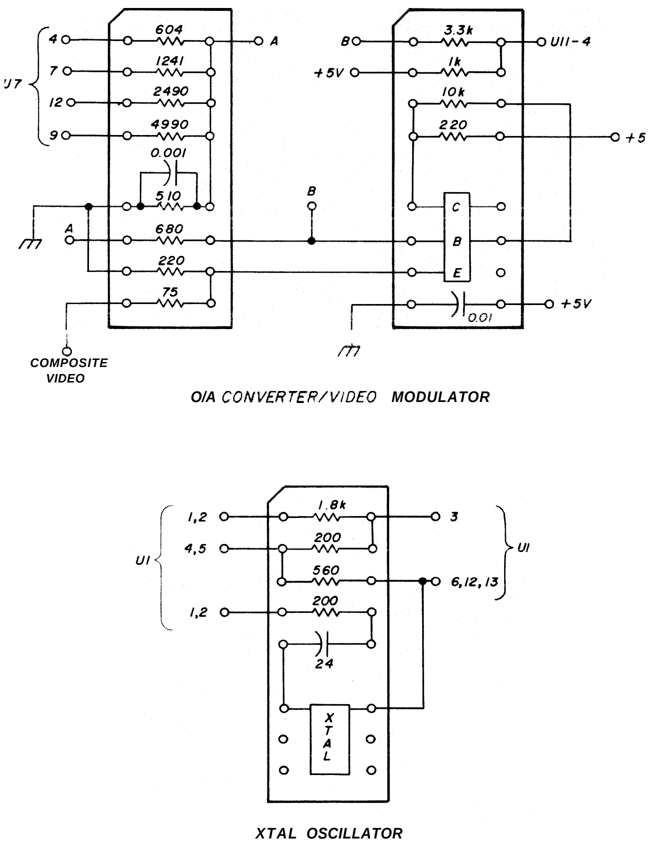


fig. 8. Suggested component mounting layout for use with DIP plugs.

video-card construction

My card was solder wired on an SS-50 bus prototype card. All signals to and from the card were routed through the connector at the top of the card. The power (8 volts) was obtained from the SS-50 bus, and regulated to 5 volts by a three-terminal regulator on the card.

Fig. *7* is the layout of the card. I found that the component layout was not critical due to the low frequencies involved. Three sockets were used for the components on the modulator and the clock circuit. The use of dip plugs make a convenient means of mounting discrete components. If you wish to experiment with different clock frequencies, new crystal plugs could be exchanged quickly. Fig. **8** is a layout of the dip sockets I used.

The only component which may be difficult to obtain is the MC6845. This component is currently available by mail order from Jade Electronics for $29.95. The crystal and other chips can be obtained from the same source. The crystal frequency is a standard microprocessor frequency. A scattering of 0.01-μF capacitors should be placed between the 5-volt line and ground. These capacitors are not shown on the schematic since they are a function of the card layout. The card requires no adjustments. The software handles all initial adjustments."

I decided to revise my software from the previous articles of this project for three reasons:

1. To allow the programming code to reside on EPROM with slight changes. No self-modifying code was used.
2. To simplify software operation by providing routines that have proven effective in the majority of Amateur Radio SSTV contacts.
3. To add new features to the software.

Fig. **9** contains a memory map of the software. The software was written in a top-down manner with all of the major routines or subroutines callable from other programs. This means that if the software were placed on EPROM, the routines could be called like macros and used as a basis of a high-level programming package. Since the software demands the use of a small amount of RAM, twenty-two bytes were reserved in the A000 region which resides physically on the CPU card. The routines were made as versatile as possible by placing all delay constants and some limited code in this RAM region.

The RAM constants are initialized during execution of the program, and can be modified at any time to produce new effects in the reception or display of SSTV signals. **Table 1** further defines the RAM constants.

CRT controller-chip software

The CRT controller chip contains eighteen registers which can be programmed to produce almost any type of video. The chip has quite a large amount of flexibility in horizontal and vertical timings. However, the crystal frequency selected must be approximately correct. I selected the CRT controller crystal by the following calculations.

The first step in crystal selection is to determine the over and under scan limits of your monitor. I determined that I could lock on video with a horizontal picture display time between 34 and 46 ps. Since the most desirable condition for my monitor (Sanyo VM4092) is to display a picture with a slight underscan, I found that 42 μs was optimum. The next step was to calculate the pixel time. I chose to display 128 pixels per line, and each pixel consisted of 64 bytes. Dividing 42 by 64, the pixel time was 656 ns. Converting this time to frequency and multiplying by four, which is the counter-divide ratio, 6.095 MHz was found to be optimum. This value was close to an off-the-shelf commercial frequency of 6.1444 MHz (651 ns).

The next task was to program the eighteen registers. The specification sheet for the MC6845 provides

*A copy of the source code is available by sending a self-addressed, stamped envelope to *ham radio,* Greenville, New Hampshire 03048.

more detail on the constant selection. I'll provide some of the logic on how I selected my constants, since I found the specifications somewhat confusing.

Horizontal-total register **(R0)**. The horizontal total register is the television horizontal frequency divided by the clock –

$$63.5\ \mu s / 651\ ns = 96$$

Bytes to be displayed **(R1)**. For 128 pixels per line use 64, and for 256 pixels per line use 128.

Horizontal-sync frequency **(R2)**. This register moves the horizontal-sync position. The effect on the TV set is to move the centering of the picture right or left. A value of 77 was found to be optimum.

Horizontal-sync width **(R3)**. The pulse width should be 4-5 ps, which is a value of 7 (4.55 μs).

Vertical-total register **(R4, R5)**. These two registers determine the vertical frequency. These values were determined experimentally by changing R4 (coarse) and R5 (fine tune) ending at 127, 10.

Vertical-displayed rows **(R6)**. This constant determines the number of character rows that will be displayed. Since an SSTV picture displayed on fast scan

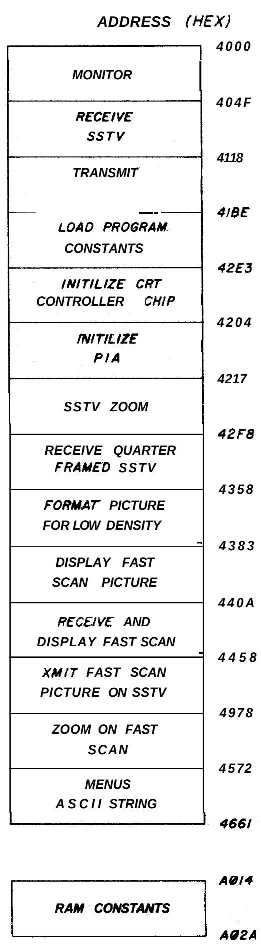


fig. 9. Program memory map for the SSTV routines.

```
                         *
                         * CRT CHIP
                         * INITILIZATION
                         * SOFTWARE
                         *
                         * 1/31/79
                         * C. W. ABRAMS
                         *
5000                             ORG    $5000
5000 5F                  CRT     CLRB
5001 CE 50 13                    LDX    #CRT2
5004 F7 80 3D            CRT1    STAB   $803D
5007 A6 00                       LDAA   X
5009 B7 80 3C                    STAA   $803C
500C 08                          INX
500D 5C                          INCB
500E C1 10                       CMPB   #$10
5010 26 F2                       BNE    CRT1
5012 39                          RTS
                         *
                         * CONSTANTS
                         * FOR INIT
                         *
                         *
5013 60                  CRT2    FCB    96,64,77,
7
5014 40 4D
5016 07
5017 7F                          FCB    127,10,12
0,120
5018 0A 78
501A 78
501B 01                          FCB    1,1,0,0
501C 01 00
501E 00
501F 12                          FCB    $12,$40
5020 40
                                 END
          NO ERROR(S) DETECTED
```

fig. 10. CRT-controller IC initialization software.

must have the correct aspect ratio, a value of 120 was selected. This format causes eight lines not to be displayed on the fast-scan screen.

Vertical-sync position **(R7)**. This constant was chosen by trial and error. A value of 120 produced optimum results.

Interlace **(R8)**. This constant can select three types of interlaced video: normal, interlaced sync, or interlaced sync and video. An interlaced picture produced the best video. Therefore, a constant of 1 was selected.

Scan-line register **(R9)**. This register is used to tell the controller the number of scan lines per character.

Fast-scan picture of a girl's face.

Since a pixel line has only one scan line, a value of 1 was programmed.

Refresh-buffer address (R12 and R13). These two registers determine the starting address for the refreshing of the fast-scan video picture. Since 120 lines are displayed, the top four and bottom four SSTV lines were truncated. Therefore, a refresh start address 0240 (hex) was chosen. This centers the SSTV pictures on the fast-scan TV display.

Fig. 10 is a source listing of a routine to load the CRT-controller registers.

A CRT controller gives you a flexibility never before attainable in a video display system. One example of this flexibility is the newly proposed medium-scan TV. When the medium-scan format is standardized, a special crystal dip socket could be constructed and new constants placed in the software. When this is done, the composite video could be connected directly to an Amateur Radio transmitter.

With a little imagination, numerous tricks can be played with the CRT-controller chip to enhance the pictures for display purposes.

software

To provide flow charts for this entire package would be an enormous task. Therefore. I've selected

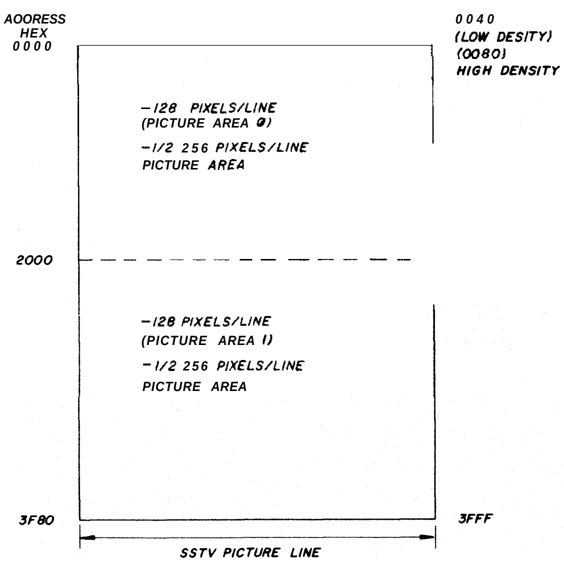


fig. 11. Diagram of the RAM space required for the SSTV pictures.

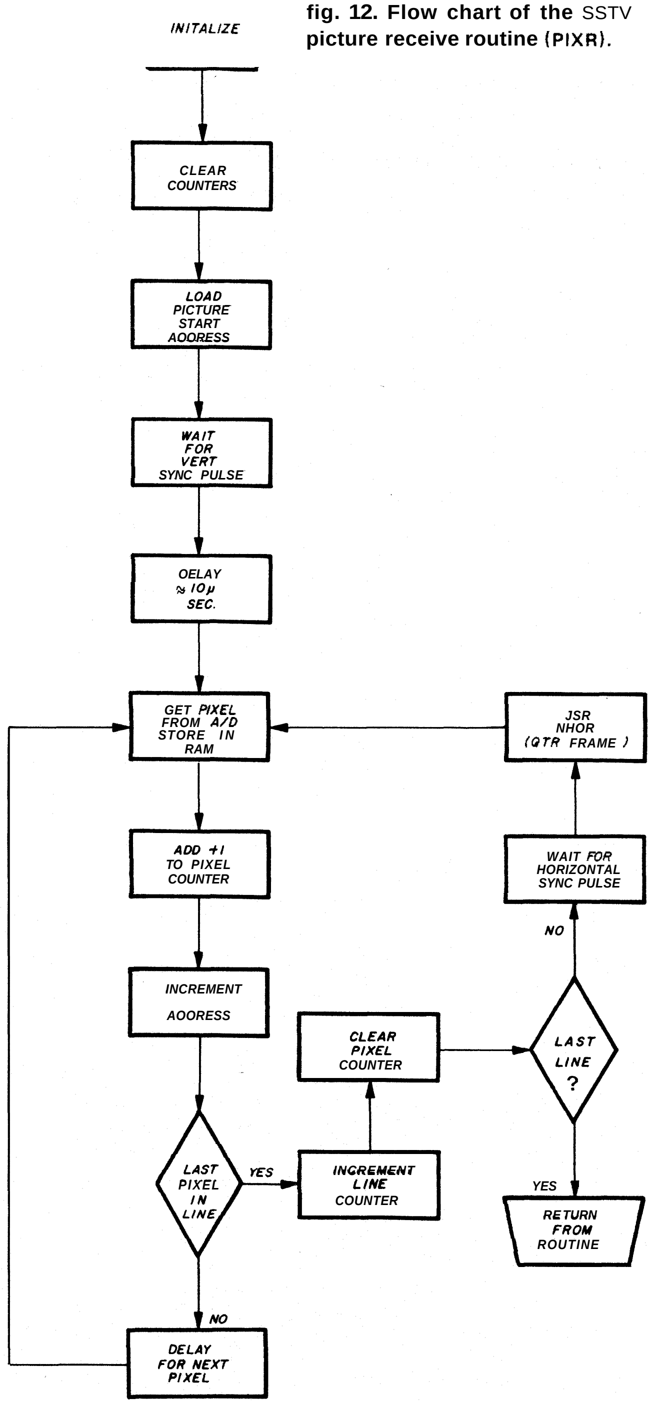


fig. 12. Flow chart of the SSTV picture receive routine (PIXR).

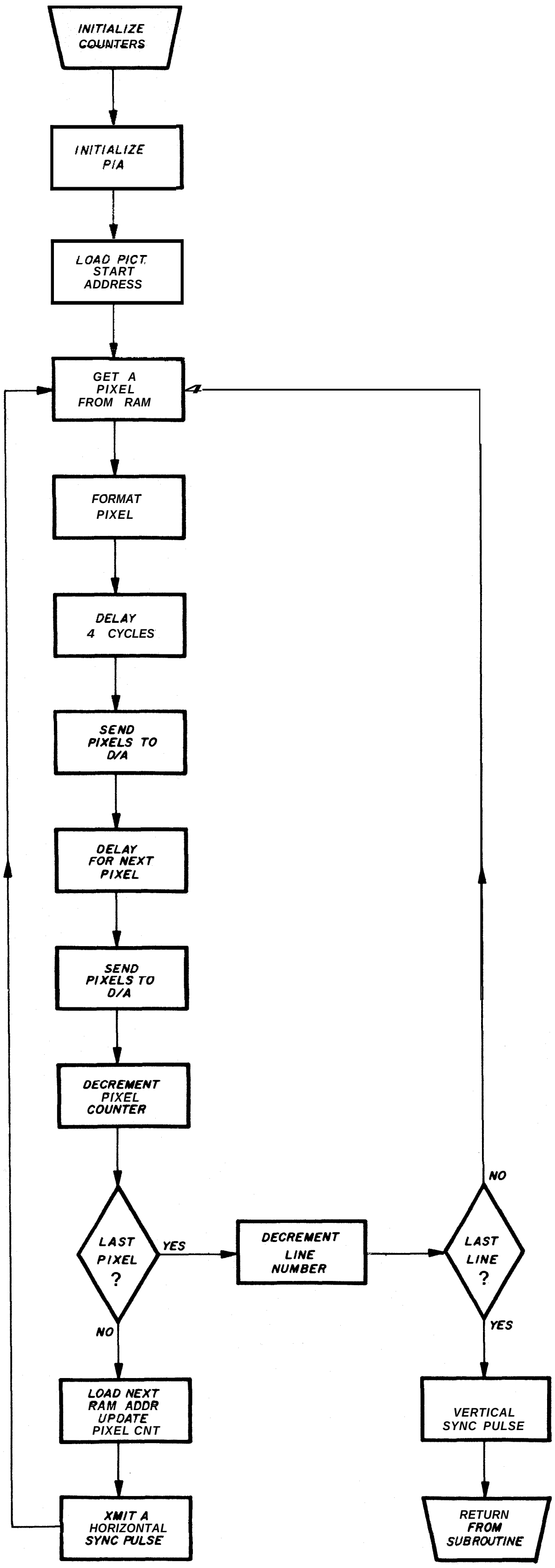


fig. 13. Flow chart for transmitting an **SSTV** picture **(TRANS1)**.

portions of critical routines to flow chart which represent how the package performs.

picture area

Fig. 11 provides a pictorial view of the digitized picture in RAM. The picture consists of a 16k-block of RAM. This 16k region could be divided up into two portions for low-density TV, or a single high-density area. I decided to use both densities in the software package. The low-density television area was designated as a primary and a secondary region. The primary region was used for receiving an SSTV picture with 128 pixels per line, and for displaying on low-density fast scan. The secondary region was used for picture enhancements and a second low-density picture-storage area. If you are somewhat confused, don't worry, I'm sure you will understand as I discuss the routines in more detail.

Main-line routine (START). The program is started by executing the instruction at location 4000 (hex). The first routine executed is START, which places a menu on the screen to display the program options. A routine selection is made by hitting a single key. The ESC key is used to jump to an undefined program. The ESC jump address assembled into this program is a location immediately after the ASCII table. This was done for future program expansions. Three levels of messages were programmed. The highest level is the START routine, used for the reception or transmission of SSTV with 256 pixels/line.

The next level is FAST which is used for displaying fast-scan TV or transmitting SSTV with 128 pixels/line. The lowest level of messages are in the various routines. This menu scheme allows the calling of routines from other programs with the entry messages displayed.

table 1. RAM constants used to describe the displayed picture.

label	location	bytes	description
XSAV	A014	2	temporary index register store
XSAV1	A016	2	temporary index register store
PIXC	A018	1	pixel counter
LINE	A019	1	line counter
CNT2	A01A	1	general counter storage
CNT1	A01B	1	general counter storage
RPIXC	A01C	1	receive pixels
RLINE	A01D	1	receive lines per picture
RSTAT	A01E	2	picture start address
RECVC	A020	1	receive delay constant
TPICT	A021	1	transmit pixels per line
TLIN	A022	1	transmit lines per picture
TRX	A023	1	transmit delay constant
NHOR	A026	4	program modifications used for quarter framing, also used for temporary byte store

Receive **SSTV (RECV1)**. This routine is a general-purpose SSTV reception program. Table **1** lists the four constants (starting with R) which must be initialized for the routine to receive pictures from the A/D converter. This routine calls six other subroutines which store the picture in RAM (STORE), wait for vertical- and horizontal-sync pulses (VERT, HORIZ), and get pixels (GETA) from the A/D converter. The constant NHOR is used for quarter framing. This will be explained later. This constant is initialized with a RTS (39 Hex) which means the call immediately returns to the calling JSR. Fig. **12** is a flow chart of the PIXR routine which receives a SSTV picture.

Transmit **SSTV** (XMIT). This routine is a general-purpose transmission routine. Like RECV1, the routine can be used as a universal transmit routine. The three constants in the RAM region, starting with T, control the type of SSTV transmitted. This routine is set up to transmit 60-Hz SSTV. For those in 50-Hz countries, you may wish to redefine the TRX transmit delay constant. The XMIT routine is the main line, and four other subroutines are called; TRANS1, SVERT, SHORIZ, and DEL5. The XMIT routine MENU allows for the transmission of up to nine pictures to be transmitted. Fig. **13** is a flow chart of TRANS1, which is used to transmit an SSTV picture.

Initialization routine **(INIT** and LOAD). Two routines are used to initialize the system prior to execution of the routine, INIT and LOAD. INIT is a simple subroutine which initializes the PIA for transmission or reception of SSTV. The port assigned by the software is 8010 (Hex). The LOAD routine initializes the receive and transmit program constants, picture start address, and monitor jump address in the first part of the subroutine. In the second part of the routine, the CRT-controller registers are loaded with the constants previously described. The initialization routines set the SSTV picture formats for 256 pixels/line and the CRT controller is set to display 128 pixels per line.

SSTV zoom **(ZOOM)**. ZOOM is used to enlarge a 256 pixel per line picture by two times and transmit it on SSTV. Five locations can be selected, the four picture corners and the picture center. The original picture is not destroyed in the process. The enlargement process is quite simple. Each pixel and line is retransmitted twice, the net result is a picture enlargement. The TRZ routine performs this operation, and is callable from another program. To use this routine, load RSTAT with the address of the upper left-hand corner position of the picture, then call the routine with a JSR instruction. Fig. **14** is a flow chart of this routine. The ZOOM main-line routine selects the program con-

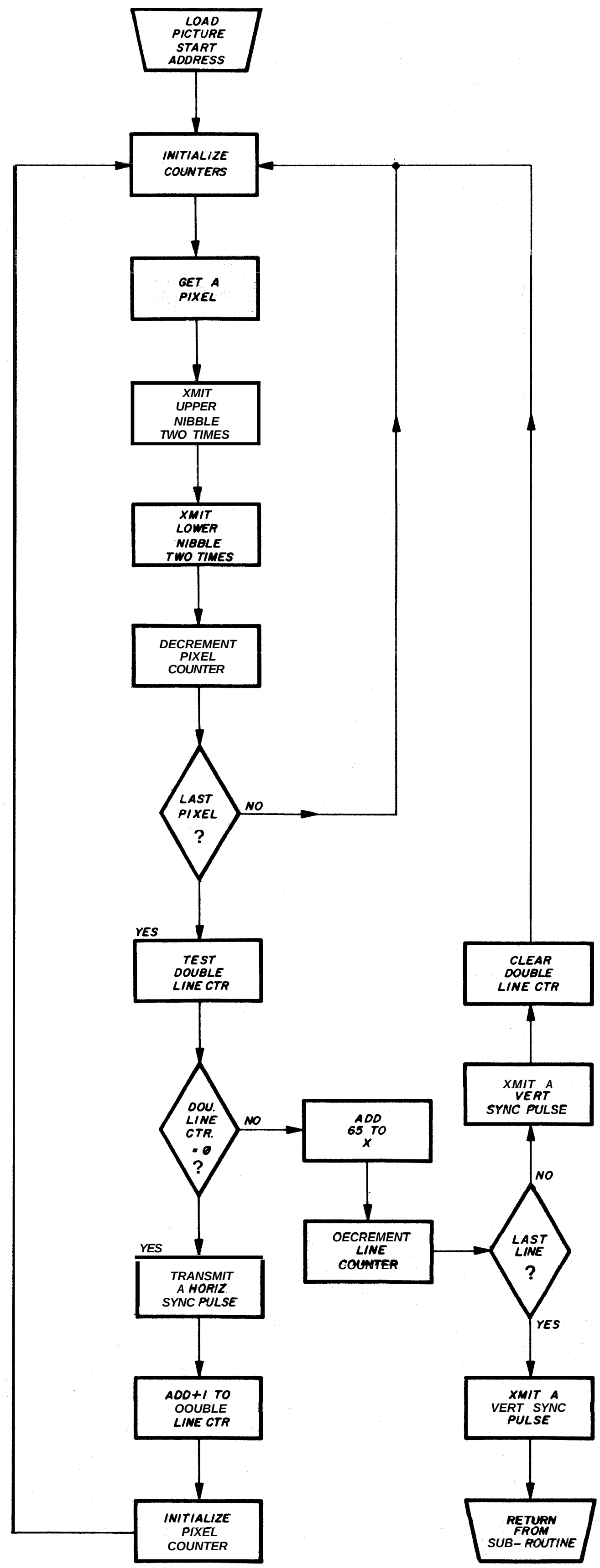


fig. 14. Flow chart of the zoom routing on the SSTV picture (TRZ).

stants and the number of loops through the program by displaying the routine's menu on the terminals screen.

Fast-scan format (**FMT**). This is the last of the 256 pixelslline routines. This routine takes a high-density picture and compresses it **128 pixels/line** (low density). The process is accomplished by simple averaging of pixels. Fig. 15 is a flow chart of this routine.

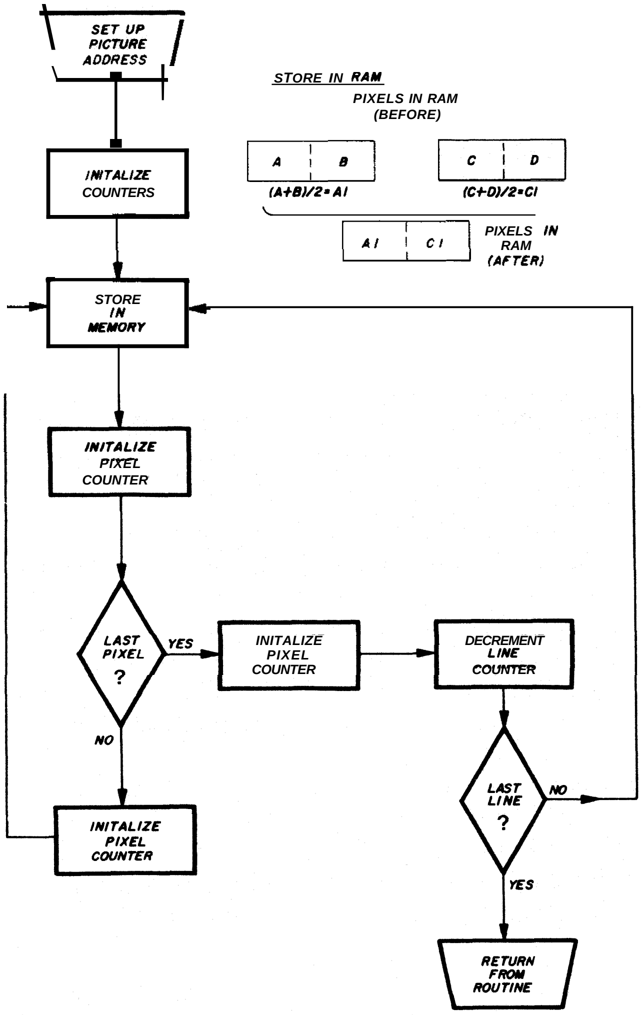
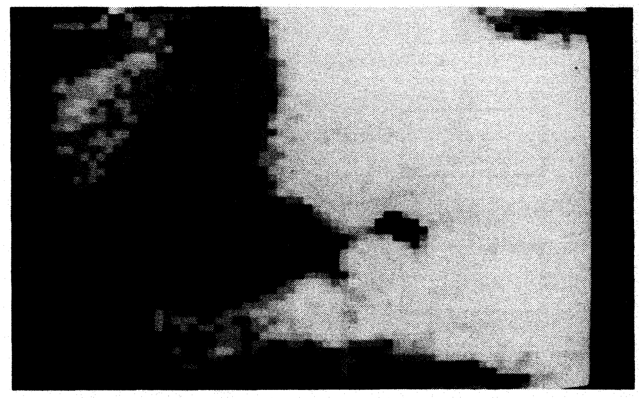


fig. 15. Flow chart of the SSTV format routine (FMT).

This process produces some interesting effects when a high-density picture is compressed and displayed on high-density SSTV. The result is the original picture duplicated two times on the upper half of the screen. The compression of high-density pictures to low density by this algorithm process produces some artifacts. These artifacts can be seen in the photographs. I concluded that this problem was due to the simple algorithm I used. Since the artifacts occur mainly with black and white edges, I decided to live with this condition.

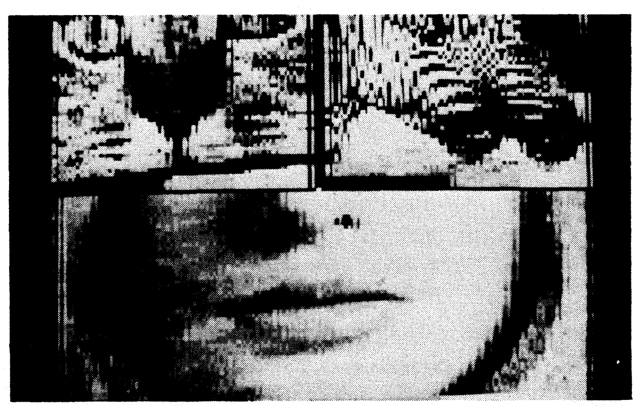


Fast-scan picture with the zoom routine.

Quarter-framing (**QTR1**). This routine is used to receive four different SSTV pictures and place each picture in a different location in a composite picture. The routine, when executed, asks which quarter frame is to receive a picture. The options are 0 through 4, which equates to upper left and right and lower left and right. The addresses of the various locations are listed in TABLE, which is stored in RSTAT when selected. Since a quarter-framed picture has **128** pixels per line and 64 lines, RPICX and RLINE are changed to 64. Additionally, every other line is received, the RTS instruction at NHOR is modified to jump to GETP1. This causes the PIXR routine to add 64 to X and wait for a second horizontal sync pulse prior to placing a new SSTV picture line in RAM. This routine assumes that the picture received is the same frequency as the last picture received, *i.e.*, 50 or 60 Hz.

Fast-scan **TV** routines (FAST). The main-line routine for all fast-scan options is FAST. All pictures displayed or transmitted in these routines have a density of **128** pixels. In these routines two picture areas are used. These areas are identified as areas 0 and 1. Each routine uses three areas in a slightly different manner. I'll discuss their use later. In order to return from FAST to START any other key, except **1** through

Girl's face with quarter frames.



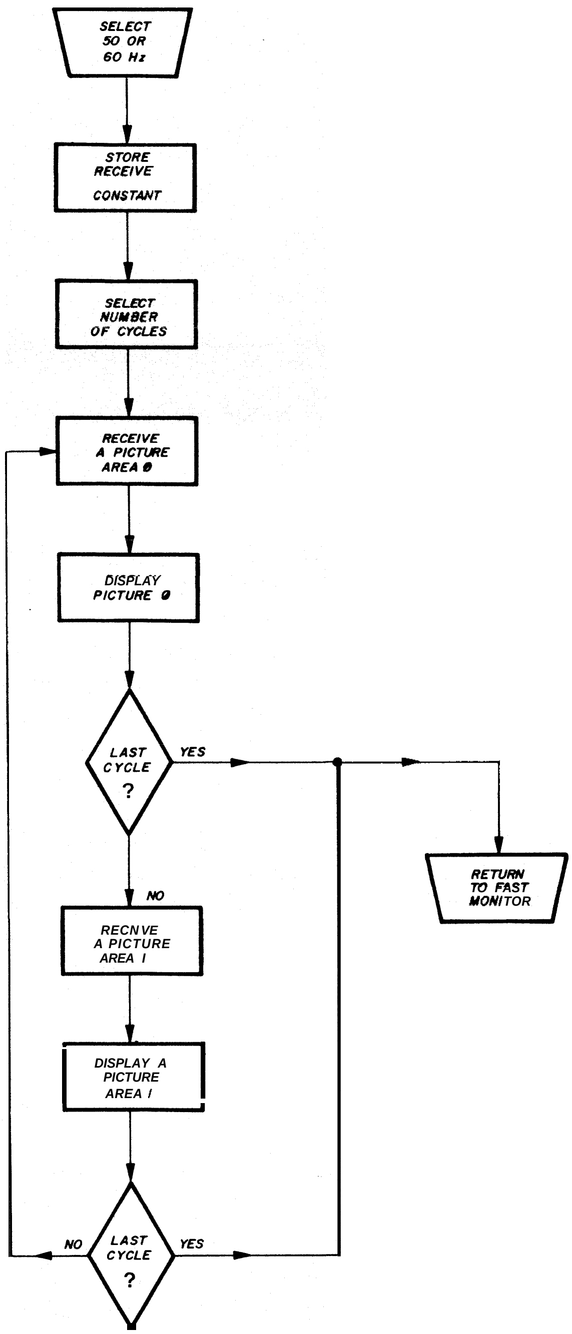


fig. 16. Flow chart of the receive then display on fast scan routine (FRECV).

5, can be pressed. All routine options selected by FAST will return to this routine after they are completed.

Display fast scan **(FAST1, DISP).** This subroutine controls all fast-scan displaying. When the main line program jumps to FAST1, the 256-pixel picture is formatted to 128 pixels and then displayed. If the jump is to **FAST4,** the 128-pixel picture is displayed. The routine places a menu on the terminal which asks for the number of cycles. A cycle is the number of times the DISP subroutine is executed. The DISP subroutine displays a picture for 7 interrupt cycles, or approximately 7.7 seconds/cycle. The response to this message should be 1 to F (15). The DISP subroutine operates in the following manner. Prior to displaying a picture three initialization steps are performed:

1. The B accumulator is loaded with the number of cycles to be displayed.
2. The interrupt service routine is loaded into the IRQ vector address in the monitor (AØØØ).
3. Address AØØE is cleared. This step is required if you use an RT-68MX monitor, as I do.

The interrupt mask is then cleared by a CLI and the display process is executed by a WAI command. The WAI causes the BA signal on the SS-50 bus to go positive, which tri-states the CPU card for DMA. After 1.1 seconds, an interrupt is generated by the hardware interface card and the RTI instruction (DISP3) is executed. This instruction restores all registers including the PC counter. The next instruction after WAI is executed and the process is repeated six more times.

Fast-scan **receive/display (FRECV).** This routine is used to receive and then display an SSTV picture. When FRECV is executed, a message asks if 50- or 60-Hz SSTV is to be displayed. The next message asks for the number of cycles. The response should be 1 to F (15). The number of cycles is the number of SSTV pictures you wish to receive and display on fast scan. The displayed picture will remain on the screen just long enough to allow reception of the next picture.

When two or more cycles are selected, the picture is first loaded into the secondary picture area 0. The next picture is loaded into picture area 1. This process allows you to store two low-density pictures in RAM. A flow chart of the FREC routine is shown in fig. **16.**

Transmitting a low-density **SSTV** picture **(FXMIT).** This routine allows the displaying of low-density pictures on SSTV. The routine is an example of how to call the XMIT subroutine. Prior to calling XMIT, the number of pixels/line and display constants are changed. When XMIT is called, a low-density picture is transmitted on SSTV. The FXMIT routine displays a message which prompts the operator to display picture 0 or 1.

Fast-scan zoom **(FZOOM).** This routine demonstrates the flexibility of the CRT-controller chip. The routine is used to examine on fast-scan TV an SSTV picture in RAM. The picture is examined by magnifying (zooming in on) the picture by two times. The zoom is accomplished by moving the SSTV picture in area 0 to area 1. In this movement process, each pixel is doubled along with doubling the lines. The CRT-controller chip is then re-initialized for the picture 1 area by CRT3, and is then displayed on fast scan as a magnified picture. **Fig. 17** is a flow chart of this routine.

This completes the description of the operation of the software package. A few systems considerations will be discussed in this section. The programming package assumes that a MIKBUG-like monitor is used. All MIKBUG calls are contained in the equate table at the beginning of the source code. The software assumes that two memory-mapped I/O ports

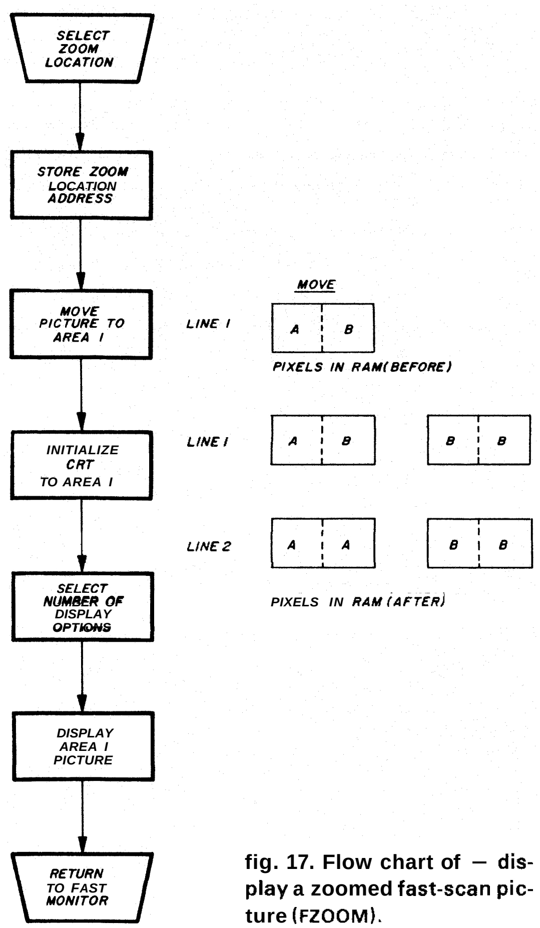


fig. 17. Flow chart of — display a zoomed fast-scan picture (FZOOM).

are used for the SSTV analog-card (8Ø1Ø) and the CRT-controller card (8Ø3C). The 803X address was used because my system has a second mother board. All of the CRT-controller program I/O calls are located in the routine with CRT 1 through 3 labels. Don't forget that the CRT-controller-chip's least-significant bit (LSB) of the address identifies either a register number or register data, *i.e.*, LSB = 0 (data), LSB = 1 (register number).

Delay loops are also used for timings. The timing loops are used to receive and transmit SSTV pictures through the hardware interface. Delay-loop constants in the software were selected and based upon the SWTP MP-A CPU clock frequency of 1.7971 MHz. If you use another CPU card, the constants will have to be altered.

EPROM program relocating

If you wish to relocate the software to reside on EPROMS, the task is quite simple. I assembled this program on a boundary which can be easily relocated. I tried to use a minimum amount of absolute code in my assembly. To relocate the program, scan the source listing for JMP or JSR instructions which address the 4XXX RAM region. All that has to be done is change the 4XXX hex to another digit, *i.e.*, CXXX. For example, I have relocated the software to run in 2708 EPROMS at location CXXX by this technique. Since the package is greater than Ik, two 2708 EPROMS are required.

summary

The displaying of gray-level pictures by use of microprocessors opens many new areas for the home hobbyist. The software techniques discussed in this article could be easily adapted for receiving1 displaying weather satellite pictures by simply changing program constants. Only the hardware demodulator interface has to be modified. Potential applications are almost endless.

Some may find my SSTV character-generator picture of interest. I created this picture by software which is coresident with the fast-scan program. I do not intend to publish this program in this form.

acknowledgments

The interfacing of new devices like the CRT-controller chip was quite a challenge. Although Motorola did a fine job on their spec sheet, numerous questions arose. Without the help of Bruce Kinney (W6TED), I would have labored numerous extra hours.

If you wish to obtain a printed-circuit board for my fast-scan card, it can be obtained from Geoff Chapman, VK2AIT,* for 25 dollars, Australian, which includes shipping. If you wish to write me for more information, please include an SASE for my reply.

*G. N. Chapman, VK2AIT, 70 Cliff Road, Epping, NSW, 2121, Australia.

references

1. Clayton W. Abrams, KGAEP, "SSTV Meets SWTPC," *73*, November, 1978. page 168, December, 1978, page 152.
2. Dr. Don Miller, W9NTP, "Medium-Scan Television System," *A5 Magazine,* November, 1978.
3. *MC6845 Data Sheet,* Motorola Semiconductor Products, Phoenix, Arizona, 1977.
4. Jack Fister, "Programming CRT Controllers," *Electronics Engineering Times,* June 29, 1978.

ham radio

uhf local-oscillator chain

for the purist

Design and construction of an LO system that offers excellent spectral purity, stability, and calibration tolerance with output between 380 and 540 MHz

For many experimenters a major stumbling block toward building a high-performance uhf or microwave station seems to be the local-oscillator chain. This is especially true in EME applications, where any degree of success demands spectral purity affording at least 60 dB spurious rejection, calibration tolerance to within a few hundred Hz, and frequency stability of a few tens of Hz over the temperature extremes encountered in the station. These stringent requirements, along with the extensive test-equipment required to verify them, seem to put the whole business of LO design and construction into the category of "more art than science."

background

Certainly the most artistic uhf LO developed in recent years is the circuit by Joe Reisert, W1JR, originally published in his now-defunct *W6FZJ 432 MHz EME Newsletter.* The circuit has since been presented in *ham radio*[1] and duplicated by hundreds of uhf enthusiasts with a high degree of success. I used Joe's circuit in my original 1296-MHz transceive converter[2] and was entirely pleased with the results. The design offers exceptional spectral purity (fig. 1), good thermal stability, and adequate calibration tolerance (all as functions of the crystal used, of course).

Joe's circuit was designed to be built in three-dimensional space above an unetched PC board, which was used merely as a ground plane. I developed printed-circuit artwork for this LO some time ago to improve its repeatability by ensuring proper component layout. During the PC-artwork development, it seemed reasonable to replace a number of discrete components with their microstripline equivalents, thus reducing component count and cost. This task completed, I trimmed from the circuit every nonessential component in further attempts at cost reduction. When I realized that my circuit bore little resemblence to Joe's original design I threw caution to the wind, abandoned his original framework altogether, and ended up with a completely new uhf LO.

spectral purity

The result is shown schematically in fig. 2. I call it "Mr. Clean" in recognition of the excellent spectral purity achieved (see figs. 3 and 4). The circuit can be

By H. Paul Shuch, N6TX, Microcomm, 14908 Sandy Lane, San Jose, California 95124

built for a 5-mW output at any desired frequency between 380 and 540 MHz, thus serving well as an LO for 432-MHz converters, 1296- or 2304-MHz converters (if followed by an appropriate ×3 or ×4 multiplier), or a weak-signal source. The circuit offers spurious rejection of more than 40 dB, a calibration tolerance of ±10 ppm, and temperature stability on the order of ±0.3 ppm/°C over the range of −10 to +60C. To date I've built more than 50 copies of this circuit, all with equal performance. The design has also been successfully duplicated with little difficulty by W6OAL, KØJHI, and WA6TLX.

The importance of spectral purity in a uhf LO cannot be overstressed, especially when the output frequency is multiplied into the microwave region. **Fig. 5** is an example of the LO output of a popular European 1296-MHz receiving converter. Compare this figure with the LO of my 1296-MHz system **(fig. 6)**.

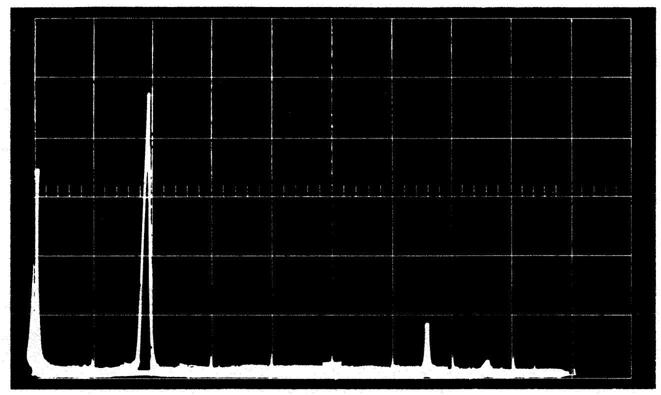


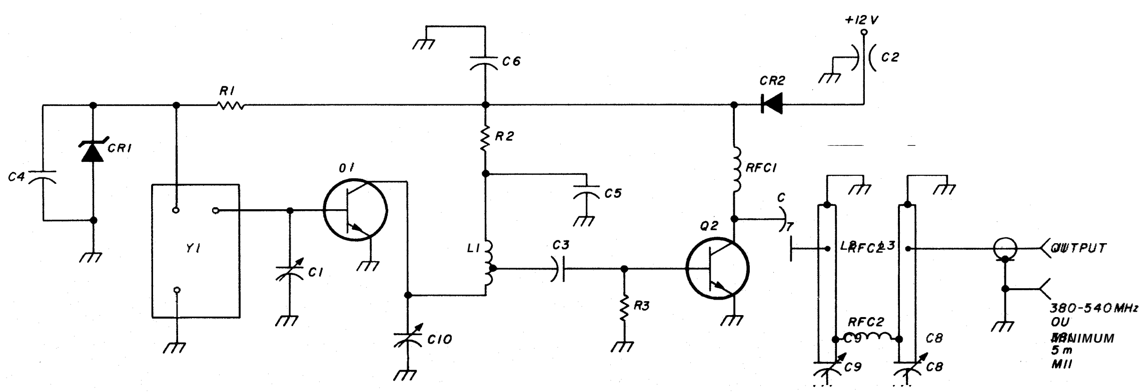
fig. 1. Dc-to-2 **GHz** spectral display of the Reisert local oscillator (reference **1)** shows its excellent spurious-response rejection. Vertical scale is **10 dB/division**; horizontal scale is 200 **MHz/division**; resolution is **1** MHz.

oscillator circuit

The primary requirements for a usable local oscillator, as mentioned previously, are stability and spectral purity. Frequency stability is generally achieved by using an oscillator circuit that draws minimum current through the crystal (thus minimizing crystal heating). This in turn dictates operating the basic oscillator at an *extremely low output power level* and making up the necessary gain in the following buffer or multiplier stages.

I learned from Joe Reisert some time ago that it's important to let the crystal oscillate at its natural resonant frequency. That is, when plugging a crystal into the oscillator circuit *for which it was cut,* you'll achieve the greatest stability by letting the crystal oscillate wherever it wants to. Any attempt to VXO or "rubber" the crystal's oscillation frequency to achieve a desired dial calibration will result in a net degradation of local-oscillator stability. This is especially true when the crystal frequency will subsequently be multiplied to the ultimate output frequency.

Since frequency pulling of the oscillator is to be avoided in the interest of stability, great precision is required in the crystal frequency calibration, with reespect to the particular oscillator circuit used, if the i-f calibration is to bear any relationship to the operating frequency. Crystal manufacturers can generally optimize custom-ground crystals for operation



C1	8-24 pF ceramic trimmer
C2	1000-pF feedthrough
C3-C6	0.01-μF miniature ceramic discap
C7	33-pF chip cap (ATC 100B or equivalent)
C8-C10	1-9 pF piston trimmer (Triko 203-09M or equivalent)
CR1	9.1-V zener (1N757A or equivalent)
CR2	1N3600 (or equivalent general-purpose silicon diode)
J1	Output receptacle (E. F. Johnson 142-0298-001 or equivalent)
L1	4t 1-mm (no. 18 AWG) tinned, 17.8 mm (0.7 inch.) diameter x 5 mm (0.2 inch) long. Tap l t up from C5 end
L2, 3	Microstripline inductors (see PC artwork)
Q1, 2	2N5179
R1, 2	180 ohms 5% ¼ w carbon composition
R3	330 ohms 5% ¼ w carbon composition
Y1	International Crystal OE-5 oscillator module. Frequency between 95-135 MHz ($f_{out}/4$)
RFC1,2	0.33 pH miniature molded choke

fig. 2. Schematic of the Microcomm Model **LO-70** uhf local-oscillator chain, which evolved from the design by Joe Reisert, **W1JR.**

in a particular circuit, provided the schematic is supplied. Unfortunately, when ordering high-precision crystals from two different reputable manufacturers for use in Reisert's oscillator circuit, I found calibration errors on the order of 10 kHz at 432 MHz — certainly beyond my expectations for a $30 crystal! And since you can be certain the crystal manufacturer certainly didn't build up Reisert's circuit and check the crystal for proper operation in it, perhaps it's better to use an oscillator circuit with which the crystal manufacturer has some experience.

I decided to use an oscillator circuit of the crystal manufacturer's choosing. In so doing I found that

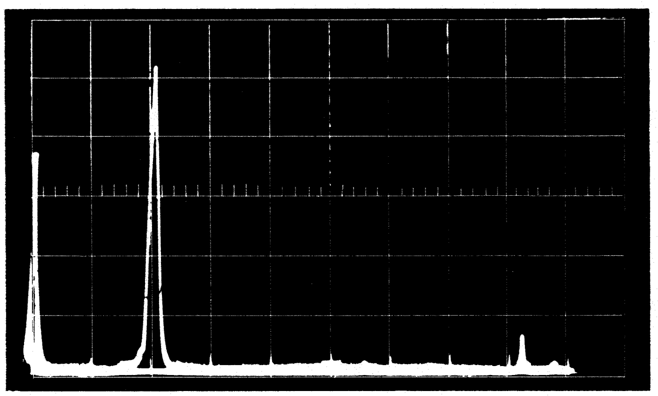


fig. 3. Spectral response of the local oscillator presented in this article compares favorably with the Reisert circuit. The analyzer settings are as in fig. 1.

crystals from widely separated production runs all fell well within the manufacturer's calibration tolerance limits of ±0.001 per cent, as well as the claimed thermal stability specifications of ±0.002 per cent from −10 to +60°C. The crystal *and* oscillator circuit (on a PC board, inside a can for shielding) are available as a preassembled module from International Crystal Manufacturing Company as their Model OE-5. Specifications and the oscillator schematic are shown in **fig. 7.** This assembly costs around $20 in single quantities, supplied at your selected operating frequency in the 100-MHz range. Since it's no more expensive than a crystal of equivalent specifications ordered separately, why bother to build your own oscillator circuit?*

Output power from the OE-5 oscillator module is low, on the order of 1/2 mW, which certainly holds down crystal heating. Spectral purity is enhanced by starting with the highest crystal frequency practical (in this case, around 100 MHz), and performing low-order integer multiplication in active stages whose bias current is optimized to the favored conduction

*Lead time for this oscillator module runs typically six to eight weeks, so order well in advance.

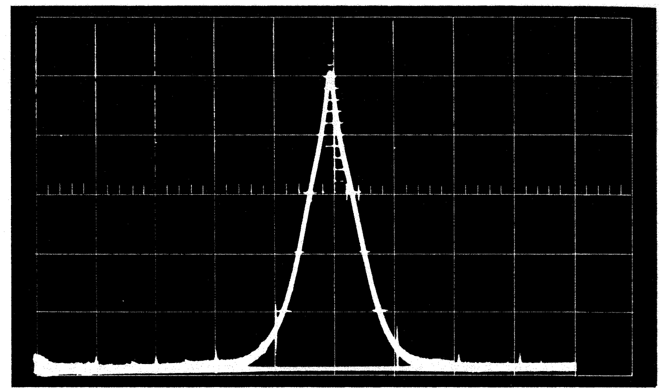


fig. 4. Spectral display of the Microcomm LO-70 LO-chain (operating frequency ±100 MHz) shows the absence of close-in spurious components. Vertical scale is 10 dB/division; horizontal scale is 20 MHz/division; resolution is 1 MHz.

angle for the multiplication desired. This oscillator, like Reisert's, uses two common-emitter bipolar doublers operating Class C. Each multiplier stage affords about 5 dB gain, so the output power from the LO chain is on the order of 5 mW (7 dBm).

Note that, from the OE-5 circuit schematic in **fig. 7,** the oscillator output is taken from a link in a parallel-resonant circuit. This coupling link provides a dc bias return for the base of first doubler transistor, Q1, as seen in the LO chain schematic, **fig. 2.** Trimmer capacitor C1 is used to resonate the OE-5 module output coupling link, which provides a double-tuned interstage and greatly improved spectral purify. However, as with all double-tuned circuits, these two tanks are somewhat interactive, so during tuneup it may be necessary to readjust the OE-5 trimmer capacitor along with C1.

circuit description of the LO chain

The 200-MHz signal from Q1 is applied through

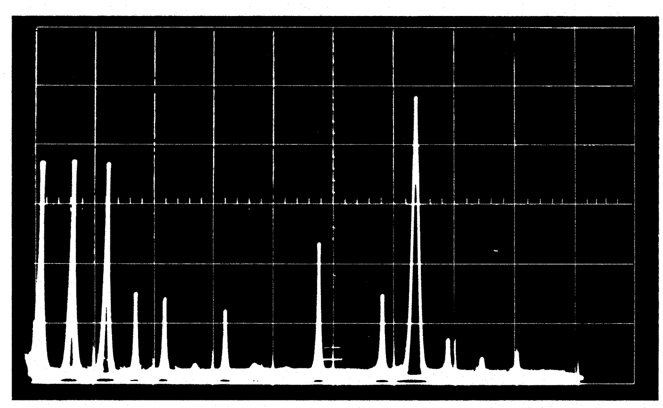


fig. 5. The importance of spectral purity in a uhf local oscillator chain is compounded when the output frequency is multiplied into the microwave region. This is the output of the local oscillator used in a popular European 1296-MHz receive converter. The effect of inadequate filtering at all stages is evident.

tuned circuit L1, C10, to the base of the second doubler, Q2. Nothing exotic here, but Q2's collector feeds a rather unusual output-filtering arrangement, which is largely responsible for the spectral purity of this LO. Basically, microstripline inductors L2 and L3, with trimmer capacitors C8 and C9, form two parallel-resonant circuits. RFC2 inductively top-couples them for a standard two-pole bandpass.

There are really more filtering elements here than meet the eye. For example, Q2's collector feed choke, RFC1, and coupling capacitor, C7, form a single L-section high-pass filter, which keeps any 200-MHz component from Q2's base from entering the output filter. Additionally, C8, C9, and RFC2 form a pi network lowpass filter which supresses harmonics from Q2. Thus the entire output circuitry consists of one lowpass filter, one high-pass filter, and two bandpass sections – all ensuring that Mr. Clean lives up to its name.

construction

All components including the OE-5 oscillator module mount on a 63.5 x 76 mm (2.5 x 3 inch) PC board. PC-board artwork is provided in **fig. 8.** Microstripline dimensions are a function of the material used, so be certain to etch this board on 1.6 mm (0.0625 inch) thick fiberglass-epoxy PC laminate, double-clad with l-ounce copper (0.035 mm or 0.0014 inch thick). One side of the board should remain unetched to serve as a ground plane.

Drill the board as in **fig. 9.** Note that to avoid short circuiting the OE-5 power and output pins as well as the rf output at J1, it's necessary to remove ground plane metal from around the three holes marked countersink in **fig. 9.** A 3.25-mm, (0.125-inch) twist drill does an adequate job. Be careful not to drill too far through the board!

Note that in **fig. 2** the ends of the output filter

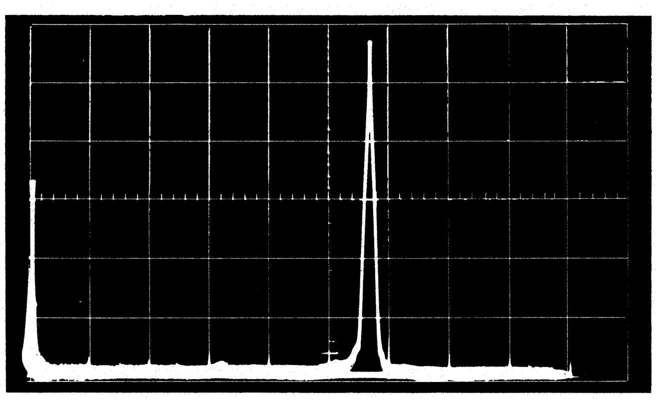


fig. 6. The output spectrum of the author's 1296-MHz system contrasts sharply with that of fig. 5. This clean display results from driving the uhf LO described in this article into a well-filtered multiplier circuit. Spurious rejection over the dc-to-2 GHz region approaches 60 dB.

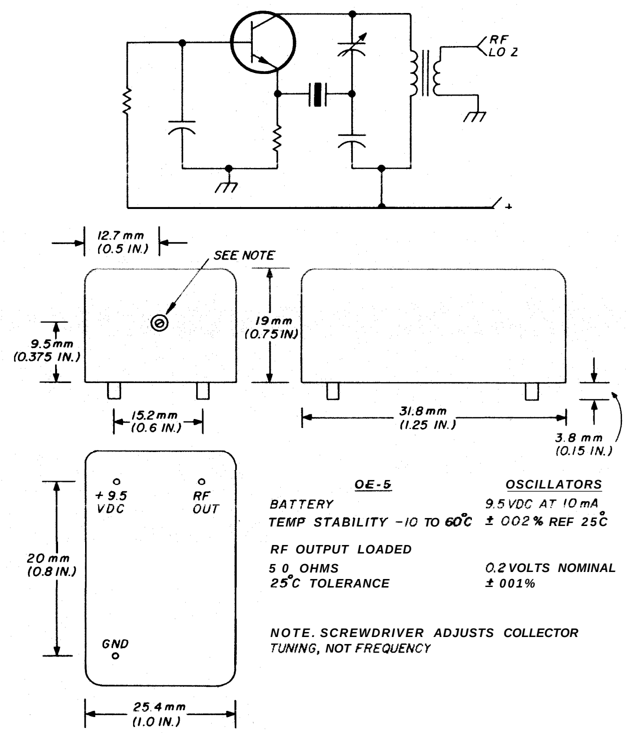


fig. 7. Schematic, dimensions, and specifications on the International Crystal OE-5 oscillator module.

microstripline inductors (L2 and L3) are grounded through the board. This can be accomplished by installing small 1-mm (0.04-inch) OD eyelets through the board at two locations indicated in **fig. 9.** These eyelets can be set then soldered to ***both*** sides of the board to ensure a reliable ground. For those who prefer not to prepare their own board, an etched, drilled, and plated board, with eyelets in place, is available from the author.*

Component layout on the printed circuit board is shown in **fig. 10. I** recommend mounting all components ***except*** the OE-5 oscillator module; save that for last. Note that R3, C8, C9, C10, and the emitter and case leads of Q1 all ground to a large ground plane area on the stripline side of the board as well as the unetched ground plane side. Be sure to solder these components at *both* sides.

When mounting the OE-5 module to the ground plane side of the main board, there's a slight chance that circuit traces on the OE-5 board might short circuit to the ground plane. Make a thin spacer from sheet acetate or Teflon the size of the OE-5 board, with holes drilled for the three pins. When installing

*An etched, drilled, and plated board for this local oscillator, complete with grounding eyelets for L2 and L3, is available for $6.50 postpaid in the U.S. and Canada ($7.00 elsewhere) from Microcomm, 14908 Sandy Lane, San Jose, California 95124. A completely assembled LO chain, adjusted to your selected operating frequency between 380-540 MHz, is also available at nominal cost. Send a stamped, self-addressed envelope to Microcomm for details. Amateurs indicate callsign on correspondence.

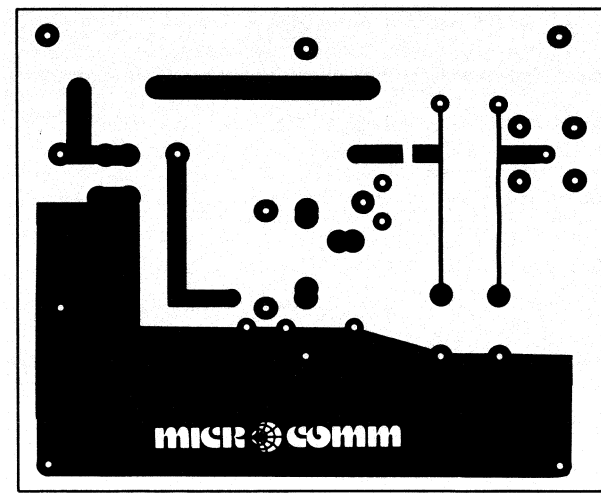


fig. 8. Full-size PC-board layout for the uhf local oscillator. Substrate material should be 1.6 mm (0.0625 inch) thick fiberglass-epoxy PC board, double-clad with 1 ounce copper (microstripline side shown).

the OE-5, grasp the three pins firmly with needlenose pliers. Gently ease each pin, one at a time, into the main board. Do not use force.

tune up

There are at least three different techniques for tuning this local oscillator chain. I hope you can keep them straight, because in this section I'll tell you a) how *not* to tune an LO, b) how I tune my LOs, and c) how to tune yours.

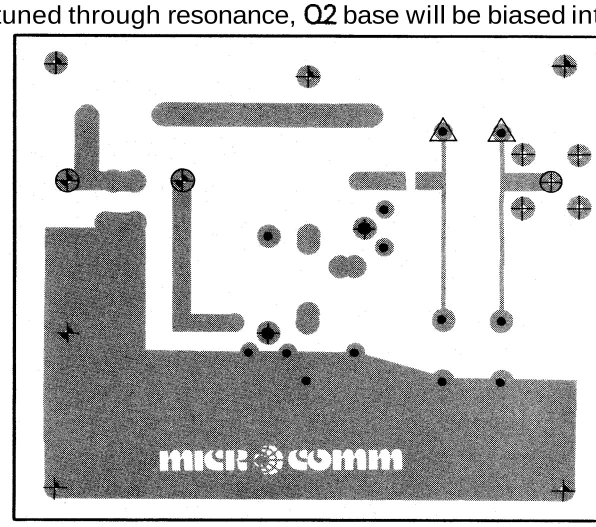
Avoid like the plague the "maximum smoke" technique. It's *not* possible to successfully tune this circuit, Reisert's circuit, or anyone else's LO circuit, for maximum indicated output power alone. Fig. **11** illustrates quite graphically that if you tune for maximum power it's likely to be distributed over a maximum number of frequencies. I can't overemphasize the importance of tuning up uhf LO chains using proper test equipment together with a systematic procedure for minimizing spurious spectral responses. I'm a firm believer in the use of spectral analysis and wouldn't dream of tuning up one of my own LO chains without the use of a microwave spectrum analyzer. Take another look at fig. 3 and compare it with fig. **11.** You can see the dramatic effect of tuning each stage of the LO chain for maximum spectral purity rather than maximum output. And the test equipment needn't put you into hock forever. Even a simple homebrew spectrum analyzer3 will allow you to achieve spectacular purity.

But you probably don't have a spectrum analyzer and cringe at the thought of having to build one before you can tune the LO you've just finished constructing, right? There's another way; I call it the "poor man's spectrum analyzer." You'll need a vhf cavity wavemeter (a grid-dip oscillator in the wavemeter mode will do), some sort of a relative power-indicating device (the one I told you not to use in method a above), and a bandpass filter tuned to the approximate LO output frequency. (There are some constraints surrounding the selection of the proper filter, which I'll cover later.) You'll also need some sort of resistive attenuator or pad, 3 to 10 dB, with a 50-ohm impedance and a volt-ohmmeter.

Preliminary steps. First connect the pad to the LO output connector, the power meter to the other end of the pad, and a +12 Vdc supply to feed-through capacitor C2. Caution: Do *not* exceed 12 volts, as this is the V_{ceo} (maximum collector-to-emitter potential) of the 2N5179s used as the multiplier transistors! In fact, series diode CR2 does provide some protection, and I have not had any transistor failures operating at 13.5 volts from a fully charged car battery — but why take chances?

With power applied, tune C1 and the OE-5's trimmer cap until the OE-5 oscillates, as indicated by an abrupt increase in supply current. With the OE-5 oscillating, set up the grid dipper in the wavemeter mode, tune it to the crystal frequency, and sniff around the microstripline going to the base of Q1 for some rf. Once you've found it, disconnect V_{cc}, then reconnect it. Did the oscillator start? If not, try retuning the OE-5 trimmer and C1 slightly until the oscillator starts reliably each time.

Now tune the first multiplier. As L1 and C10 are tuned through resonance, Q2 base will be biased into



Key
- △ **install eyelet here (see text)**
- • **no. 60, 1 mm (0.04 inch) (11 places)**
- + **no. 56, 1.2 mm (0.046 inch) (5 places)**
- + **no. 49, 1.85 mm (0.073 inch) (3 places)**
- + **no. 27, 3.7 mm (0.144 inch) (5 places)**
- + **no. 12, 4.8 mm (0.189 inch) (2 places)**
- ○ **countersink ground plane side here (see text)**

fig. 9. Drilling template for local oscillator PC board (viewed from microstripline side).

conduction and supply current will increase by 5-10 mA (Q2's collector current). Try it. The only problem is that tank L1/C10 may resonate at more frequencies than the desired $F_{xtal} \times 2$. So tune the dipmeter (still in wavemeter mode) to the second harmonic of the crystal frequency, couple it *loosely* to L1, and tune for an indication of rf. Once you've found it, repeak C1 and C10 for the combined occurrence of *maximum* rf and *maximum* supply current; then check to make sure the oscillator still starts up each time power is applied. If not, retweak the oscillator trimmer *slightly* until it does. At this point you should start to see some indication of rf at output connector J1. Remember that your relative-power indicator can't distinguish between frequencies, but

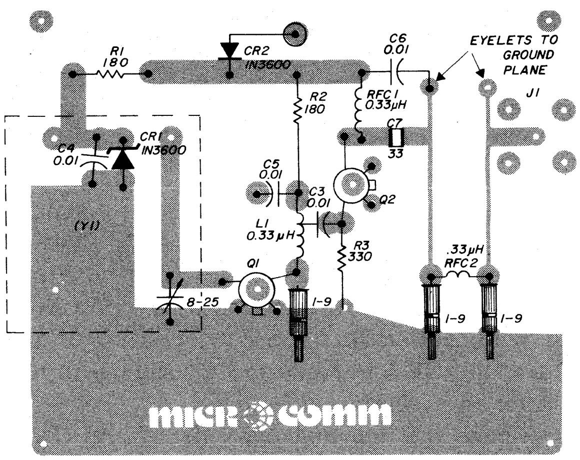


fig. **10.** Parts layout, Microcomm model LO-70 uhf **local** oscillator (**microstripline** side).

just for starters, tune C8 and C9 for maximum output. At this point the output spectrum (if you could see it) would probably appear as in **fig. 11**, but don't worry about it.

Final adjustments. Now that you've completed the preliminary tune up you're ready to clean up your act. Insert the bandpass filter (tuned to the desired output frequency) between the pad and the power indicator, and *carefully* retweak all trimmers for maximum output power. The adjustments will interact, so go back and do it again. Check to make sure the oscillator still starts each time you apply power. When you're finished remove the filter and pad, and measure output power at J1. It should be on the order of 5-10 mW, and the spectrum should appear as in **fig. 3**.

But don't count on it. The first time I tried this procedure, I ended up with almost as much output at 1200 MHz as I had on 400 MHz. This didn't make much sense to me, as I was using a high-Q, quarter-

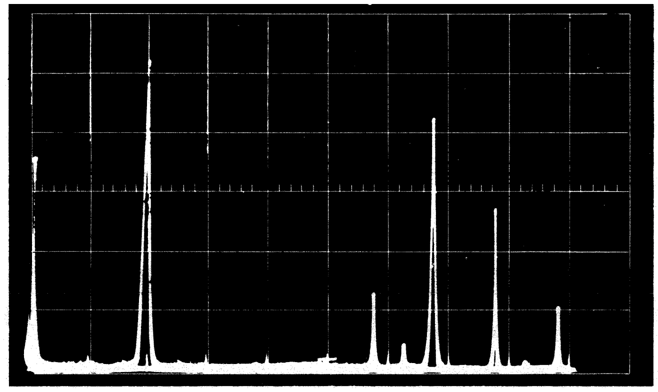


fig. 11. This spectral display, with measurement conditions as in figs. 1 and 3, illustrates the importance of tuning the local-oscillator chain using a spectrum analyzer. This is what results if the LO is tuned for maximum output as indicated on a power meter. Note that the worst spurious component is down by only 10 dB.

wavelength, trough-line filter in aligning the LO, and I knew it had sharp skirts! Just for good measure I swept the filter, and the problem became immediately evident (see **fig. 12**). A quarter-wave transmission line, shorted at one end, makes a dandy resonator. Unfortunately, so does a three-quarter-wavelength transmission line. The filter I chose exhibited a passband at the third harmonic of the LO frequency, and by tuning for maximum signal through the filter, I was actually optimizing the spurious output! I mention this because quite a few bandpass filters exhibit multiple resonances. The halfwave slab resonator described in the ARRL *VHF Manual and Handbook* does, so it would not yield a spurious-free output if used as a tune-up aid

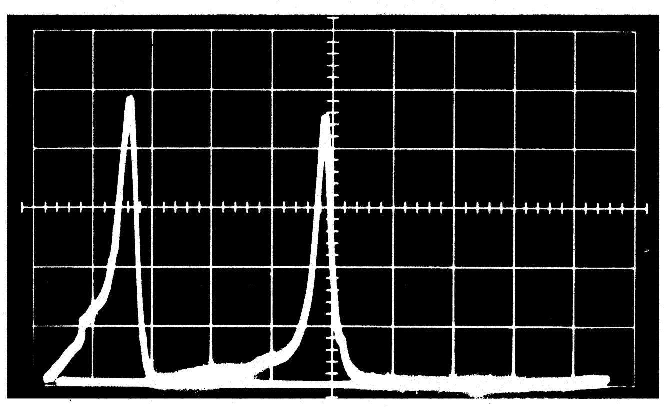


fig. 12. Swept-frequency response of the quarter-wave, trough-line resonator first used as a tune-up aid for this LO, as described in the text. Vertical scale is 5 **dB/cm**, with the second major division from the top of the screen **representing 0** dB insertion loss. Horizontal scale is 250 **MHz/cm**, yielding a dc-2.5 **GHz** display. Note that, while the insertion **loss** at 400 MHz (the desired frequency) is less than 1 dB, the insertion loss at the third harmonic **(1200 MHz)** is on the order of only 3 **dB**.

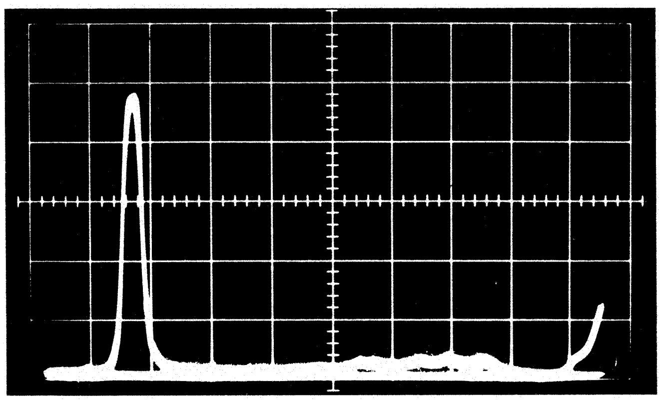


fig. 13. Swept-frequency response of the Microcomm model PB-70 filter ultimately used as a tuning aid for aligning this LO, as described in the text. Vertical and horizontal scales are as in fig. 12. Note absence of spurious responses out to 2.5 GHz, as well as the passband insertion loss of less than 1 dB.

for this LO. Best bet is to use either an interdigital filter or helical resonator, or a multipole design whose interstage coupling is designed to supress higher-order modes. The Microcomm model BP-70 is one such filter (see **fig. 13**), as is the Spectrum International model PSf432. Also useful are the surplus military filters of the F-197/U variety, which have recently surfaced at numerous ham auctions and flea markets around the country.

Given a single-response bandpass filter tuned to the approximate operating frequency of the LO, it's possible to tune this oscillator circuit to a degree of spectral purity rivaling that achieved on a laboratory microwave spectrum analyzer.

conclusion

I've presented a uhf local-oscillator chain that offers stability, calibration tolerance, and spectral purity on a par with Joe Reisert's very fine circuit, but with fewer components and easy assembly on a PC board. I am currently using this LO in my 432-MHz receive converter, driving multipliers in my 1296- and 2304-MHz converters, in my 1296-MHz hand-held transceiver, and in an S-band satellite ground station design I am producing commercially. I find the circuit easy to assemble and extremely reliable. I hope other uhf and microwave experimenters find it useful.

references

1. Joe Reisert, W1JAA, "What's Wrong with Amateur vhf/uhf Receivers — and What You Can Do to Improve Them," *ham radio,* March 1976, page 45.
2. H. Paul Shuch, WA6UAM, "Easy-to-Build ssb Transceiver for 1296 MHz," *ham radio,* September 1974, page 8.
3. H. Paul Shuch, WA6UAM, "Low-Cost Microwave Spectrum Analyzer," *ham radio,* August 1977, page 54.

ham radio

linear amplifier design

The designer of a linear amplifier should be concerned with the proper potentials required to make the power tube operate in a linear manner. The word *linear* implies that the output signal of the amplifier is an amplified replica of the input signal. There's no such thing as a perfect linear amplifier, and the designer's problem is to make the practical amplifier (*i.e.*, the amplifier that can be built) as linear as possible.

When a linear amplifier is driven by a complex signal, such as the human voice, nonlinearity results in intermodulation distortion. This unpleasant form of distortion creates a broad, raspy signal that throws annoying "buckshot" into adjacent channels. Proper design and operation of a linear amplifier reduces this distortion to a minimum.

amplifier circuit and mode

There's a lot of confusion with regard to the so-called "grounded-grid" amplifier. Rf power amplifiers are classified according to circuitry and mode of operation. The two classifications should not be confused with one another. For Amateur service, the two most popular circuits are the grid-driven circuit and the cathode-driven circuit. As shown in **fig. 1**, the circuits are remarkably similar, the most obvious difference being the placement of the ground point in relation to the input and output circuits.

The mode of operation refers to the dynamic operating characteristics of the tube (class AB_1, class B, or class C). Characteristics of the classes are given in reference material listed at the end of this article. For linear service, the power tube amplifier is commonly run in either class AB_1 or class B service. Thus, modern equipment may have an intermix of circuitry and mode — the cathode-driven amplifier may be operated in a class AB_1 mode, for example, or the grid-driven amplifier may be operated in the class B mode.

So far, I've not discussed the popular grounded-grid amplifier. This is a sloppy term which usually refers to a cathode-driven amplifier, working in the class B mode. "Grounded grid" implies cathode drive, but in such a circuit the grid may not necessarily be at dc ground potential, especially with respect to screen voltage (see **fig. 2**). Rf ground and dc ground are not always the same in a linear amplifier, and most circuit engineers shudder at the use of the term.

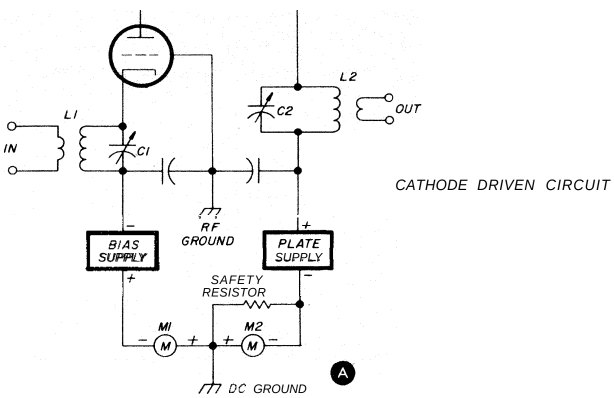
amplifier plate circuit

While this series of articles concerns itself with linear, cathode-driven-amplifier design, the remarks about the plate circuit apply equally well to grid-driven amplifiers. It is desirable to operate any linear amplifier with a very minimum of intermodulation distortion, with high-plate efficiency, and with high power gain. The latter is especially important, as it affords maximum power output with a given amount of drive power. The class B mode of operation meets these requirements.

Shown in **fig.** 3 is a graphical representation of a class B amplifier, showing the operating cycle of the tube. This is the portion of the electrical cycle over which the tube grid is driven positive (approaching $+e$) with respect to the cathode (or the cathode driven negative with respect to the grid). When the grid potential is highly negative with respect to the cathode (approaching $-e$), the tube is cut off and is inoperative. In the class B amplifier, the operating cycle is about one-half the electrical cycle, or approximately 180 degrees. The transfer curve plot shown indicates that the tube delivers power only over one-half of the electrical cycle and is cut-off during the other half of the cycle. Does this mean that the output signal consists of half-sine waves as shown, and is therefore highly distorted? Not at all.

The amplifier plate circuit (often called the tank circuit) saves the day, since the energy storage ability (Q) of the circuit balances the energy between the halves of the cycle, much as the flywheel stores energy during the operating cycles of a gasoline engine. The plate circuit must, therefore, be designed to have sufficient Q or energy storage, for good operation. A Q value of 12 is commonly used for linear amplifier service, as it provides ample energy storage and at the same time provides reasonable reduction of harmonics generated in the amplifier.

By William I. Orr, W6SAI, 48 Campbell Lane, Menlo Park, California 94025



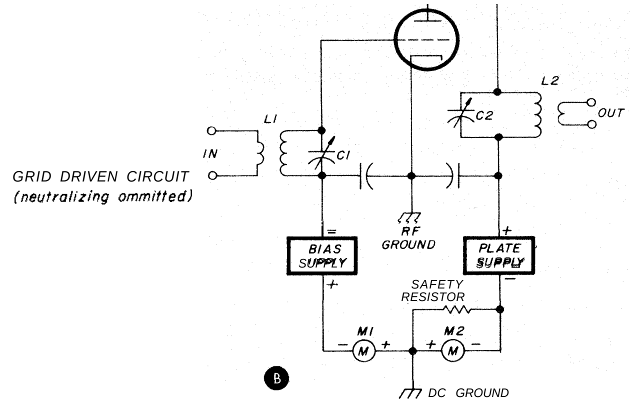


fig. 1. A comparison between grid-driven and cathode-driven amplifiers. Rf and dc circuits have been simplified for clarity. In both cases, the **grid**- and plate-current meters are placed in the ground return circuits to remove any dangerous voltage from the meter movement. This, however, places the plate supply above dc ground by virtue of the voltage across the plate meter. If the meter coil should open, the negative lead of the supply rises to the value of the plate voltage. As a safety factor, a wirewound resistor is usually placed across the plate meter, and often the grid meter. The circuit configuration determines the difference between **cathode**- and grid-driven service. The applied voltages determine the mode of operation.

A rigorous design of the plate circuit calls for manipulation of the plate voltage and current to determine the operating parameters of the tube. The results of these tedious calculations can be summed up in simple formulas that provide the designer with circuit data in everyday terms.

A network is required that matches the plate load impedance of the power tube to the characteristic impedance of the transmission line, while at the same time maintaining a Q value of 12. The popular pi network can do the job. The plate load impedance (Z_L) for a class B rf amplifier can be closely approximated by:

$$\text{load impedance (ohms)} = \frac{\text{plate voltage}}{2 \times \text{peak dc plate current (amperes)}}$$

As an example, a pi network is to be used to match a pair of 3-500Z tubes to a 50-ohm transmission line. The tubes operate with 2500 volts plate potential with a peak dc plate current of 800 mA (0.8 amp) for a PEP input of 2 kW.

$$\text{load impedance} = \frac{2500}{2 \times 0.8} = 1560 \text{ ohms}$$

Thus, the pi network plate circuit has to match a load impedance of 1560 ohms to a 50-ohm termination.

designing the plate circuit network

The approximate values of the pi network can be determined from three simple graphs. The plate inductance from **fig. 4,** the tuning capacitance (C1) from **fig. 5**, and the loading capacitance (C2) from **fig. 6.** The graphs are entered at the x axis and read up until the sloping line denoting a particular Amateur band is intersected. The value of the component is then read horizontally off the y axis. For example, the required inductance for a plate load of 1560 ohms for the 15 meter band is about one microhenry — as close as the graph can be read. Note that capacitor C1 is commonly referred to as the tuning capacitor and C2 the loading capacitor.

The graph for C2 tells us that the pi network cannot cope with impedance transformation values much greater than 100-to-1 at this value of Q. Note how the curves bunch together and "fall-off the graph" at plate impedances much higher than 5000 ohms.

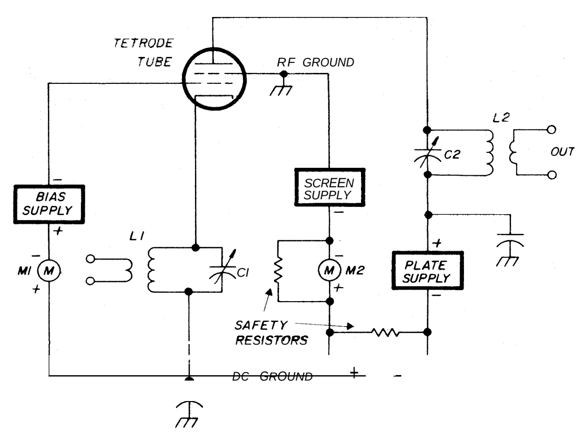


fig. 2. Diagram of the so-called "grounded-grid" amplifier. The grid and screen elements are bypassed to ground as far as rf is concerned, but each element has normal operating voltages applied and are "above ground" as far as dc is concerned. Metering is inserted in the supply return leads to dc ground. Rf ground is placed at the positive screen voltage level. This eliminates the screen bypass capacitor, a tricky component that often causes circuit instability at the higher frequencies.

A more accurate, computer-derived summary of pi network values is given in table **1.** Note that, for a given plate impedance, when the operating frequency is doubled the capacitance and inductance values are halved. (Fifteen- and forty-meter constants are related by a factor of three as 21 MHz is the third harmonic of 7 MHz.)

coil winding

Winding plate coil L1 to a given value of inductance takes an inductance meter, or a degree of expertise and a dip-meter. A simple formula for calculating inductance when the coil dimensions are known is:

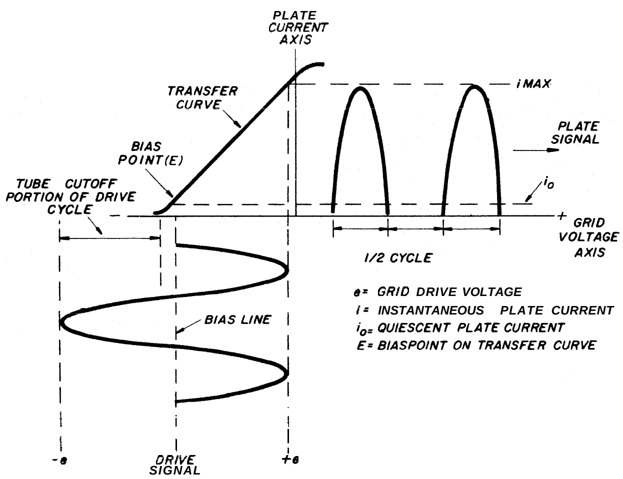


fig. 3. Transfer curve and operating cycle for a class B amplifier. The transfer curve is determined by a static test of the tube where plate current is plotted against grid bias. Once the transfer curve is established, the operating cycle may be determined. The sine wave drive signal **(e)** is drawn about the bias line, determining both the zero-signal plate current **(i_o)** and the peak plate current **(i_{max})**. Note that when the grid driving signal swings negative, no plate current is drawn and the tube is cut-off for one-half cycle. Pulses of plate current only appear when the drive signal is positive with respect to the bias voltage. Thus, the output waveform of a class B rf amplifier consists of a series of half-cycles, much in the manner of a half-wave rectifier. The distorted waveform is restored to a sine wave by the plate tank circuit which, by virtue of its *Q,* or flywheel effect, stores energy on the active half of the cycle and releases it on the inactive half. Circuit engineers, working from a transfer curve, can determine actual dc operating potentials for a linear amplifier.

$$Inductance\ (\mu H) = \frac{R^2N^2}{9R + 10S}$$

where R is the radius of the coil in inches
S is the length of the coil winding in inches
N is the number of turns

These calculations have been simplified in the ARRL type-A "Lightning Calculator," which is a simple slide rule providing direct read-out of the coil dimensions if the inductance is known. It takes the hard work out of designing coils.

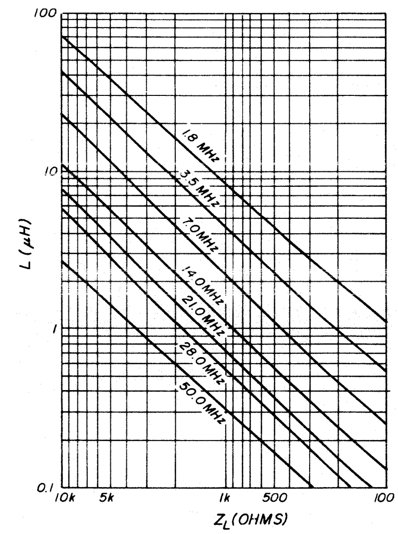


fig. 4. Plot of the plate inductance vs. plate load impedance for the high frequency Amateur bands (Q = 12).

Once the plate circuit has been designed and built, it is a good idea to "breadboard" it up and check it out with a dip-meter before the connections are finally soldered. Coil taps may have to be moved a bit to compensate for capacitance of the components to the chassis and adjacent parts.

amplifier-cathode circuit

The cathode-input circuit provides an impedance match between the 50-ohm coaxial output circuit of the driver/exciter and the input impedance of the cathode-driven amplifier (see table **2).** The input im-

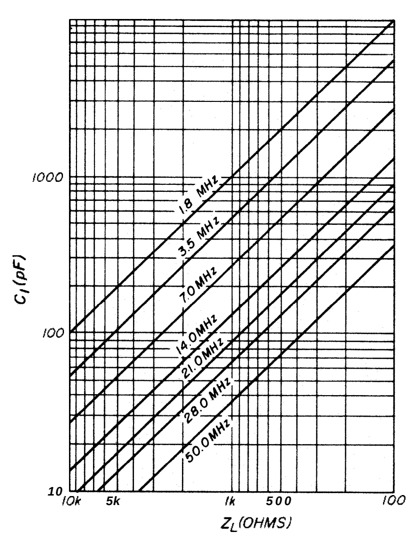


fig. 5. Plot of the tuning capacitance **(C1)** vs. plate load impedance **(Q = 12).**

pedance (Z_t) of a cathode-driven tube is related to the ratio of the peak cathode signal voltage to the peak cathode current (sum of grid and plate currents), and is commonly given in the tube data sheet. For the 3-500Z at 2500 volts, it is about 110 ohms. And for two tubes in parallel, it is about 55 ohms, but *only* over the operating cycle.

It is tempting to jump to the conclusion that if the amplifier input impedance is about 55 ohms and the coaxial line impedance driving it is 50 ohms, that no cathode impedance matching circuit is required. In fact, many commercially manufactured amplifiers leave it out for economy's sake. This omission is poor engineering practice, as the circuit *Q* is required in the cathode circuit as well as in the plate circuit. Omission of the cathode-tuned circuit can lead to distortion of the driving signal, increased intermodulation distortion, reduced amplifier efficiency, and driver loading problems. A circuit *Q* of 2 is adequate, and a simple rule of thumb is that the network circuit capacitances at resonance should be about 20 pF per meter of wavelength for one-to-one impedance transformation.

practical amplifier circuit

Armed with the information discussed so far, it is possible to draw up a schematic for a cathode driven, 2-kW PEP linear amplifier using two 3-500Z tubes in parallel (see **fig. 7**). This is a true "grounded-grid" circuit, as the grids are at both dc and rf ground potential.

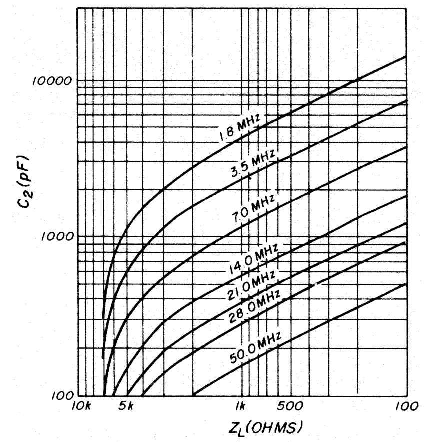


fig. 6. Plot of the loading capacitance (C2) vs. plate load impedance (Q = 12).

Note that plate and grid currents are measured in the cathode return circuit. This requires the amplifier plate power supply to "float" a little above ground potential in order to insert a meter in the negative lead to measure plate current. This removes the lethal plate voltage from the meter. The grid meter is out of the critical rf ground return path, which simplifies the metering circuit. A filament voltmeter is included. Filament voltage should be held to within

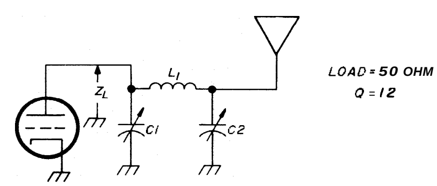


table 1. Computer-derived values for a pi network having a *Q* of 12 and working into a 50-ohm load. Values for C1 include the output capacitance of the tubes. These values are taken from a computer program derived by Bob Sutherland, W6PO.

component		Z_L plate load impedance (ohms)							
	band	**1000**	**1500**	**2000**	**2500**	**3000**	**3500**	**4000**	**5000**
C1	160	1060	690	531	430	354	309	265	212
	80	546	364	273	220	182	159	136	109
	40	273	182	136	110	91	80	68	55
	20	136	91	68	55	45	40	34	27
	15	91	61	45	37	30	26	23	18
	10	68	45	34	30	23	20	17	14
C2	160	4421	3487	2865	2440	2105	1849	1594	1186
	80	2274	1784	1473	1263	1082	951	820	610
	40	1137	892	737	632	541	475	410	305
	20	568	446	368	316	271	237	205	153
	15	379	297	246	211	180	158	137	102
	10	284	223	184	158	135	118	102	76
L1	160	8.84	13.26	16.61	20.10	24.13	27.80	31.47	38.63
	80	4.55	6.57	8.54	10.90	12.41	14.29	16.18	19.87
	40	2.27	3.28	4.27	5.50	6.20	7.15	8.09	9.93
	20	1.14	1.64	2.14	2.70	3.10	3.57	4.05	4.97
	15	0.76	1.09	1.42	1.82	2.07	2.38	2.70	3.31
	10	0.57	0.82	1.07	1.36	1.55	1.78	2.02	2.48

±5 per cent of 5 volts, and it is prudent to monitor this voltage when expensive tubes are used. A plate voltmeter may be included in the amplifier, but it is easier to place it in the power supply.

Amplifier standby plate current is reduced by means of a 10-kilohm, 25-wait cathode resistor which is shorted out by the VOX relay of the exciter, causing the tubes to operate at the proper resting plate current when the amplifier is on the air. A zener diode is placed in series with the cathode dc return path to reduce the quiescent plate current during amplifier operation.

A 50-ohm wirewound resistor from the negative side of the plate supply to ground makes certain that the negative supply terminal does not rise to the value of the plate voltage if the positive side of the supply is accidentally shorted to ground.

Two reverse-connected diodes are shunted across the safety resistor to limit any transient surges under a shorted condition which might cause wiring insulation breakdown. In addition, the diodes protect the meters from transient currents. A resistor across the zener diode provides a constant load for it and prevents cathode voltage from soaring if the zener safety fuse opens.

Note that a 10-ohm, 50-watt wirewound resistor is placed in series with the B-plus lead to the plate rf choke. This resistor serves as a vhf choke to suppress harmonic currents in the power lead and also protects the tube and associated circuitry in case of a flash-over in the tube or plate circuit. The tremendous amount of energy stored in the power supply is instantaneously "dumped" into the amplifier when a

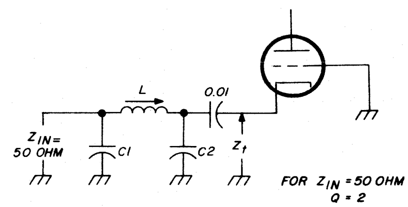


table 2. The pi-network circuit for a cathode-driven amplifier. This chart provides approximate values for the components of the cathode circuit. Capacitors should be 1-kV silver mica or equivalent. The inductor can be wound on a slug-tuned form. Value of C2 should take into account the cathode-grid capacitance of the tube which appears in parallel with C2 (information is from a computer program by W6PO).

cathode Z_t (Ω)	band	C1(pF)	C2(pF)	L(μH)	cathode Z_t(Ω)	band	C1(pF)	C2(pF)	L1(μH)
	160	3300	4100	2.50		160	3300	2870	3.81
	80	1700	2120	1.34		80	1700	1540	2.05
20	40	900	1120	0.68	75	40	900	770	1.03
	20	440	560	0.33		20	440	380	0.51
	15	300	370	0.22		15	300	250	0.34
	10	220	275	0.16		10	220	180	0.25
	160	3300	3900	2.84		160	3300	2520	4.20
	80	1700	2100	1.52		80	1700	1350	2.26
30	40	900	1050	0.77	100	40	900	680	1.14
	20	440	520	0.38		20	440	330	0.56
	15	300	350	0.25		15	300	220	0.38
	10	220	258	0.19		10	220	160	0.28
	160	3300	3360	3.01		160	3300	2100	4.81
	80	1700	1800	1.62		80	1700	1130	2.59
40	40	900	910	0.82	150	40	900	570	1.30
	20	440	450	0.40		20	440	280	0.66
	15	300	300	0.27		15	300	180	0.43
	10	220	220	0.20		10	220	138	0.32
	160	3300	3300	3.33		160	3300	1800	5.32
	80	1700	1700	1.79		80	1700	980	2.86
50	40	900	900	0.90	200	40	900	490	1.44
	20	440	440	0.45		20	440	245	0.71
	15	300	300	0.30		15	300	164	0.48
	10	220	220	0.22		10	220	120	0.35
	160	3300	3100	3.53		160	3300	1640	5.78
	80	1700	1670	1.90		80	1700	880	3.11
60	40	900	840	0.96	250	40	900	440	1.57
	20	440	417	0.47		20	440	220	0.78
	15	300	275	0.32		15	300	140	0.52
	10	220	205	0.23		10	220	100	0.38

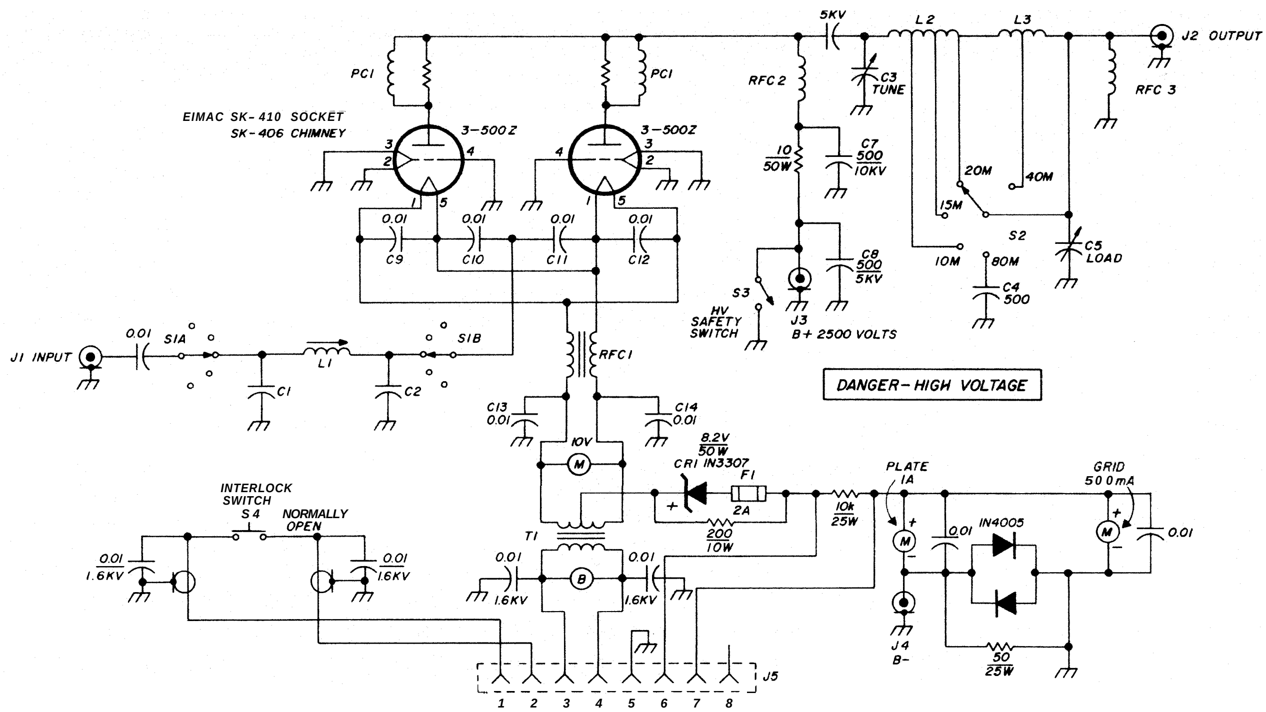


fig. 7. Schematic diagram of **the 3-500Z** linear amplifier.

C3	250 pF, 4.5 kV plate spacing — Johnson 154-16
C4	500 pF, 4.5 kV
C5	1000 pF, 500 volt plate spacing
C6	0.001 μF, 5 kV — Centralab 858S-1000
C7, C8	500 pF, 10 kV TV-type "door knob"
C9-C14	0.01 μF, 500 volt mica capacitor. Ceramic disc is a suitable substitute if rated 1 kV.
PC 1	Three 100-ohm, 2-watt resistors in parallel
PC 2	Three turns of no. 14 AWG (1.6 mm) wound with 12.5-mm (0.5-inch) diameter and 19-mm (0.75-inch) length connected in parallel with the resistors. The coil may be wound around one of the resistors.
RFC 1	50 μH; 14 bifilar turns of no. 10 AWG (2.6 mm) enameled wire wound on ferrite core 12.5 cm (5 inches) long and 12.5 cm (0.5 inch) in diameter (Indiana General CF-503 or equivalent).
RFC 2	100 pH, 1 ampere dc; 112 turns no. 26 AWG (0.4 mm) spacewound wire diameter on 2.5 cm (1 inch) ceramic form 15 cm (6 inches) long (Centralab X-3022H insulator). Series resonant at 24.5 MHz with terminals shorted (B&W 800).
RFC 3	2.5 mH, 100 mA
T1	5 volts at 30 amps (Chicago-Standard P-4648)
Blower	13 cu. ft./min. Use a no. 3 impeller at 3100 rpm (Ripley 8472, Dayton 1C-180, or Redmond AK-2H-01AX)

flash-over occurs, and much of this destructive energy is dissipated in the resistor.

Many modern-generation Amateurs have never worked with equipment operating at voltages higher than 12 volts. This amplifier, with the high-voltage plate supply, is positively lethal and the operator can be killed if his hands are inside the unit when the high voltage is on. It is ***imperative,*** therefore, that safety switches be incorporated in the amplifier design. It is poor engineering practice to leave these devices out! S4 is a normally open, pushbutton device that is closed only when the lid is placed on the amplifier enclosure. S3 is a shorting switch that shorts the high voltage to ground when the lid is removed. Construction of this special switch will be covered in a future article. ***Always remember — high voltage kills!*** Take necessary precautions.

Although not shown on the schematic, it is a good idea to use a filament transformer having a primary winding tapped for 105, 115, and 125 volts. This provides a plus or minus ten per cent adjustment from a normal line voltage of 115 volts. If a closer filament adjustment is desirable, the transformer can be run on the 105 volt tap with a rheostat in series with the primary winding to place the filament voltage "on the nose."

The plus and minus leads to the high voltage supply should be run through high-voltage connectors and high-voltage cable. Test prod wire having a 10-kV breakdown is satisfactory. As an alternative, RG-58/U coaxial cable can be used for high-voltage leads along with PL-259 plugs and reducers and SO-239 receptacles. The shield of the coaxial line is grounded by the connectors. **ham** radio

short Beverage for 40 meters

Discussion of a short Beverage antenna for 40 meters with particular emphasis on the matching transformer and termination

Basically, my problem is one of geography. Living in a moderately rare DX location, I have become weary of pile-ups and the quick signal report exchanges. The insipid hello, goodbye, PSE QSL routine fails to satisfy the rag chewer that I am, with the result that I now tend to avoid the higher frequency bands and seek refuge lower in the spectrum. The 40-meter band offers attractive rag chewing possibilities, but everything about my location militates against a 40-meter pipeline to the folks back home.

For a starter, the band assignment in this part of the Pacific is only from 7 to 7.1 MHz. This narrow band is cluttered with Asian BC stations, and from about 7.03 to 7.1 MHz there is one continuous roar of JA ssb signals. To copy any W/K signals above 7.03 is well nigh impossible without a highly directive antenna. The Asian signals so totally overwhelm the receiver as to completely bury the much weaker W/Ks arriving from over 9600 km (6000 miles) away. After many frustrating attempts at rag chewing while listening on my vertical, I was convinced that, without some highly directive receiving antenna, it was a losing proposition.

Extensive research and meditation on this dilemma brought me to the conclusion that some sort of Beverage antenna offered the only hope in my circumstances. For me, multi-element phased or parasitic arrays on 7 MHz were out of the question, but a patch of jungle behind the house offered possibilities for a Beverage-type long wire. At this point, I must acknowledge my debt to others for supplying me with the three basic premises upon which my project was founded.

First, a simple low-to-the-ground, properly terminated long wire can achieve astonishing rejection of signals from unwanted directions.[1] Second, such a wire will exhibit maximum front-to-back ratio if it is an odd number of quarter wavelengths long.[2] And third, although most publications show a simple 600-ohm resistor for termination, no simple resistor alone will ever give optimum termination. A little inductance will always be needed in series with the resistor.[3] (Apparently the intrinsic insulator and end capacitance of such a wire causes a slight mismatch which must be inductively cancelled out.)

After an hour of tramping about in the jungle, taking repeated compass sightings, I finally located a group of four trees that lay in a perfectly straight line, on the exact bearing needed to beam W/K. A coconut palm and a breadfruit tree about 58 meters (190

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feet) apart provided the end supports, with two trees in between providing intermediate support points along the span.

There is no simple formula by which to determine the height of the wire above ground. I wanted it low enough that I might perform all adjustments while be used in place of the slug; however, one precaution should be observed. The primary and secondary windings should be placed on opposite sides of the circle, with a tight electrostatic shield between them. The object is to prevent proximity coupling direct from the end of the hot wire over to the coax input.

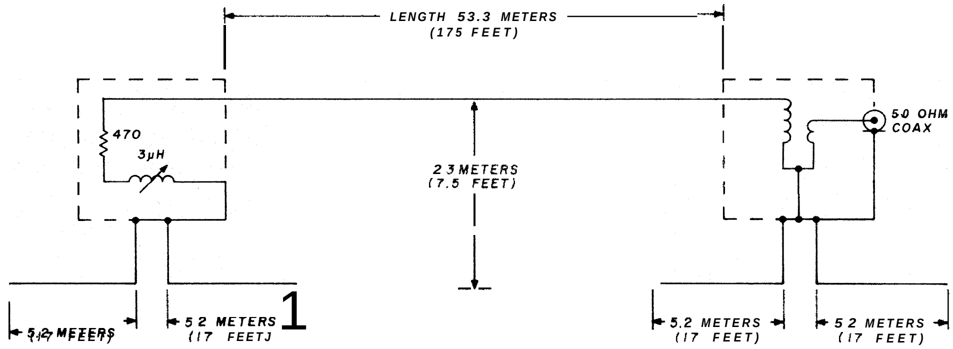

fig. 1. Diagram of the Beverage in use at the author's station. Each of the four ground wires is buried about 15 cm (6 inches) deep, on opposite sides and perpendicular to the Beverage wire. All wire used was number 10 AWG (2.6 cm), with the antenna made from Copperweld.

standing on an ordinary kitchen stool, and yet high enough to be well above the hands of any pedestrian traffic through the woods. A height of 2.3 meters (7.5 feet) satisfied both requirements. A piece of number 10 AWG (2.6-mm) copperweld was cut to a length of 53.3 meters (175 feet), which is 1.25 wavelengths at 7 MHz. With a block and tackle it was stretched taut as a fiddle string, so that with the two intermediate support points, it hangs straight as an arrow. The coupling and termination enclosures

For best unwanted signal rejection, all coupling must occur through the core material.

To compute the characteristic impedance of my wire, I used the standard single wire transmission line formula $Z_o = 138 \log_{10} \frac{2h}{p}$ where h is the height of the wire above ground and p is the radius of the wire measured in the same units. For my case, the com-

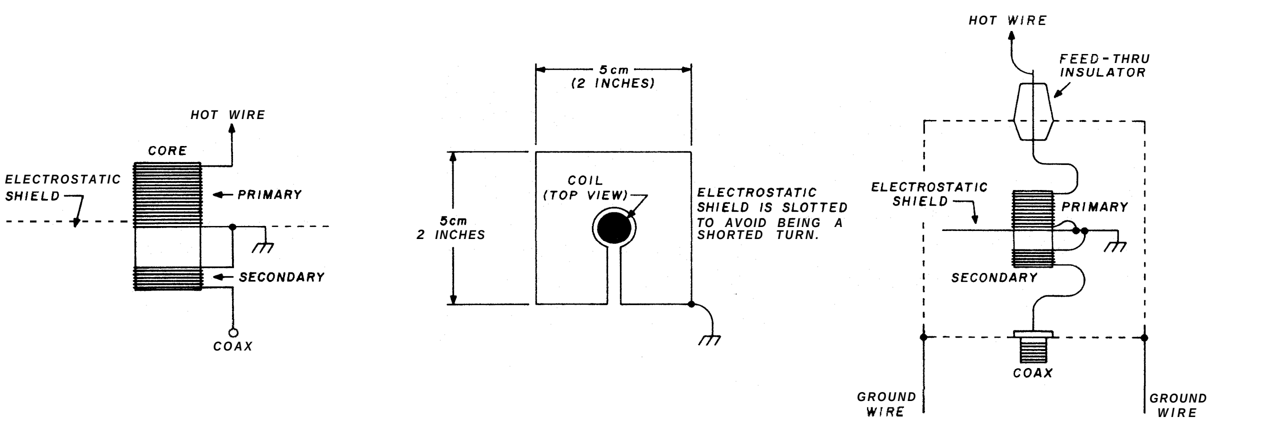

fig. 2. Diagrams of the matching transformer. The coil slug is made of powdered iron, 2.5 cm (1 inch) long and 1.3 cm (0.5 inch) in diameter. The primary is 30 close-spaced turns of number 26 AWG (0.4 mm) enameled wire, while the secondary is nine turns of number 26 AWG (0.4 mm) enameled, also close spaced. The two windings are separated by approximately 6.5 mm (¼ inch). The electrostatic shield is approximately 5 cm (2 inches) square and is made from copper foil; it is slotted to avoid being a shorted turn.

were hung at wire level, and the ground system installed as shown in **fig. 1.** Any type of minibox enclosure may be used as long as it is all metal for total shielding and weather proofing.

The coil detail is shown in **fig. 2.** A surplus powdered-iron slug of unknown pedigree was used here because I had nothing else. A toroid could well

puted Z_o is 489 ohms. Reasoning that the series coil would add a few ohms of rf resistance to the lumped resistor, I chose the next smaller resistor value, 470 ohms. A commercial slug-tuned coil was used for the inductance. After all adjustments were complete, the inductance actually in use was 3 μH, or an X_L of 132 ohms at 7 MHz.

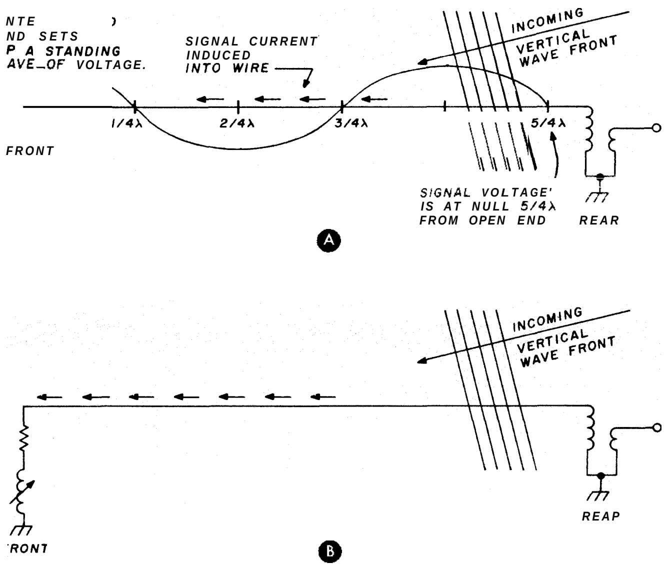

fig. 3. Diagram (A) shows the standing wave on the unterminated wire caused by an incoming signal from the rear. The signal voltage across the transformer is at a null 5/4λ from the open end. When the wire is perfectly terminated (B), the signals from the rear of the antenna are absorbed by the termination.

The Beverage antenna is located such that it is broadside to the transmitting antenna (a base-fed vertical dipole), and located 38 meters (125 feet) away. The feedline is 50 meters (165 feet) of RG-58A/U coax. Such orientation ensures minimal coupling between the two. Under key-down conditions, with 500-watts input to my trusty old 813, the measured rf voltage at the receiver input is only 0.15 volts rms.

Before step-by-step adjustments are described, a little discussion of the back rejection theory will be helpful. As shown in **fig. 3A,** incoming signals propagating along the wire from the back side will induce a signal current into the wire. If the wire is unterminated at the front end, a standing wave of voltage will be set up with a maximum at the open end. Reflections traveling backwards along the wire for an odd number of quarter wavelengths will create a voltage null where the coupling coil is located at the rear end. Because there is almost no voltage acting on the relatively high impedance of the coil, very little signal will be coupled through to the receiver.

This can be clearly demonstrated by simply removing the coil from the circuit and connecting the hot lead of the coax directly to the rear end of the wire. Under these conditions, a standing wave of current will appear in the wire with a maximum at the rear end, coupling very well into the low impedance of the coax. With the front end of the wire unterminated, take an S-meter reading on some sky wave signal arriving from the back side. This establishes the ability of the wire to receive a certain signal level, apart from any termination or phasing attenuation. Once an average level is noted, reinsert the coupling transformer. On my antenna, the rear-side signal dropped by 18 dB when the coil was reinserted, showing that some amount of front-to-back ratio is achieved through the phase relationship of the odd quarter wavelength. The drop in signal noticed is not related to coil losses when the coil is reinserted. A similar comparison was made with the signal from a Kuala Lumpur station broadcasting on 6.03 MHz. At this frequency, the wire is about one wavelength long, and it showed no front-to-back change at all when the coil was reinserted into the circuit. Coil losses with iron-core coupling are so insignificant that they do not show up in the meter readings.

After my termination was reconnected and carefully adjusted for minimum back pick-up, the front-to-back ratio was increased by another 15 dB, giving a total front-to-back ratio of 33 dB for both effects working together. Referring to **fig. 3B,** if the front end termination is a perfect match for the characteristic impedance of the wire, the induced rear side signal current is almost totally absorbed in the termination, and there is no reflection to speak of going back to the coupling end.

In my location there was no possibility of enlisting the help of a "local" 40-meter station to provide a rear side signal for tuning purposes. The closest dry land off my back is the island of Borneo, 4000 km (2500 miles) away! I selected a station in Kuching, Sarawak, (broadcasting on 7.16 MHz) to be my reference for all front-to-back adjustments. To get accurate measurements on a signal from that far away is difficult but not impossible. Taking readings on a vertically polarized local signal can be misleading. The Beverage does respond to vertical polarization, but it depends on the slight forward tilt of the incoming wave front to produce the small horizontal vector actually coupling into the wire. If a local vertical signal is used, the wavefront is so square to the ground that coupling into the Beverage is much less than a low-angle sky wave arrival would provide. Comparisons with a front-side vertically polarized signal from KG6RJ, only 3.2 km (2 miles) away, showed the Beverage about 12 dB less responsive than it would be to low-angle sky waves from the same direction.

An aged Hammarlund HQ180 receiver S-meter was used for all readings. The bandwidth was set to the narrowest position so the S-meter would respond only to the carrier of the station concerned and not to the buckshot of BC stations or QRN. The rf gain must be set to maximum for all readings. Since all readings are a comparison of the vertical reference (transmitting) antenna vs the Beverage, the first step is to establish a loss figure for the Beverage. Because it presents a very small "capture area" to incoming wave fronts, and being very long and close to lossy ground, the Beverage is very inefficient, compared with the vertical. In order to derive its intrinsic loss, the termination must be disconnected and the coupling coil temporarily removed from the circuit. The center lead of the coax is jumpered to the end of the Beverage wire for the first comparison.

The Kuching signal was then tuned in using the vertical antenna. The HQ-180 antenna trimmer was detuned to where the Kuching signal just peaked to the 20 dB over S9 mark. This is to avoid taking any readings at the compressed top end of the S-meter scaie where the calibrations are very inexact. The meter is watched for about one full minute to get the feel of the QSB and to make sure the needle never swings beyond the 20 dB mark. Then the vertical coax is disconnected from the receiver and the Beverage connected.

With the Beverage antenna in use, the meter is again watched for about one full minute, noting the very highest swings of the needle. With mine, the peaks were an average of 20 dB lower, showing the basic Beverage wire alone has a receiving loss of 20 dB compared with the vertical.

Next, the coupling coil and termination were reattached. The same comparison procedure was followed; first tuning in the signal on the vertical, adjusting the antenna trimmer until the signal peaks 20 dB over S9; then attaching the Beverage antenna and noting the drop in S-meter readings. The QSB is accentuated on the Beverage because it is very selective to polarity, responding best to vertically polarized incoming wave fronts.

The procedure is repeated with different settings of the termination coil, changing it in increments of about two turns of the slug between trials. As the optimum point is approached, the S-meter responses on the Beverage will go lower and lower. When I got my termination to the optimum setting, the S-meter dropped dramatically from the 20 dB over S9 mark on the vertical to only S3.5 on the Beverage — counting downward from the 20 over 9 mark, allowing 6 dB per unit, a decrease of 53 dB.

Note that 53 dB is ***not*** the front-to-back ratio of the Beverage working alone! That is merely the difference between the two antennas. In the first test described, it was already established that the Beverage had an efficiency of 20 dB less than the vertical. Subtracting this constant from the difference readings between the two gives an absolute front-to-back ratio of 33 dB for the Beverage alone.

It must be remembered that S-meters are notoriously unreliable as regards absolute decibel calibration. While useful for noting changes in a given signal, they must be regarded with suspicion when seeking to establish accurate decibel levels. Listening experience with this antenna would seem to indicate the actual suppression is not quite as dramatic as these figures indicate. The Hammarlund HQ180 exhibits different sensitivity on each band, so, obviously, if the meter were accurate on one band, it would be lying on all the others. At 7 MHz, it seems to be a bit generous with its dB read-out. Nevertheless, the suppression of this antenna must be called excellent, more than worth the very small investment required for materials. It is hard to imagine more suppression per dollar than the Beverage offers.

Taking hundreds of readings by the same comparison method, I found that when the Beverage was terminated at its best front-to-back rejection, the front-to-side rejection was also best. A strong Chinese broadcast station (7.025 MHz) from somewhere in the rear quadrant was reduced 30 dB by the pattern of the Beverage alone. Japanese ssb signals arrive from a 30-degree-wide sector and do not all show the same front-to-side rejection ratio. The average values were from 36 to 40 dB front-to-side ratio for the Beverage alone (in the sector of 60 to 90 degrees relative in the pattern).

The proof of the pudding is always in the eating, and this antenna has certainly proven itself with my goals in view. The Beverage does not completely eliminate the undesired Asian signals, but it does knock them down far enough that distant signals, which would have been completely overwhelmed with the vertical, can now be heard. The only disadvantage noted is that because of its polarity selectivity, the Beverage antenna shows magnified QSB effects. That is a small price to pay for the rejection in unwanted directions. Rag chewing with W/K generals on 40-meter CW is now commonplace and pleasurable, whereas before it was a grim struggle if possible at all.

In conclusion, let me add that, due to my unique location, all of these rejection figures were derived on signals arriving from 2400 to 9500 km (1500 to 6000 miles) away, which implies low-angle arrival. I am unable to specify just how the rejection figures would work out for high-angle signals. Perhaps someone situated in the center of the United States could perform further experiments to add high-angle rejection data to what I have already established for DX-only responses. Any takers?

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2. The ***ARRL Antenna Book***, 13th edition, American Radio Relay League, Newington, Connecticut, 1976, page 172.
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ham radio

matchbox plus two

Improvements for your Johnson Matchbox antenna tuner — coax-to-coax tuning plus antenna switching

The Johnson Matchbox antenna tuner has been around for a long time. The Matchbox was manufactured in two versions, one for 275 watts and one for 1 kilowatt. The circuit used excellent quality components and was conservatively rated. This article describes modifications that can be made to the Matchbox that eliminate the need for disconnecting and connecting coax and wire feedlines and add the convenience of changing antennas with the turn of a switch.

features

The 275-watt Matchbox uses Johnson Series 154 capacitors, which are rated at 3000 Vac. The coil is wound with no. 12 (2.1-mm) wire. It is to be noted that transmatch articles in *QST* and the ARRL *Handbook* specify Series 154 capacitors for use at 2 kW.

The Johnson Matchbox will match the 52-ohm output of a transmitter into loads ranging from 25-1200 ohms for balanced transmission lines to 25-3000 ohms for unbalanced lines. The tuned circuit provides at least 15 dB harmonic attenuation.[1] Most Amateur transmitters and amplifiers for the high frequency bands use pi networks. The nominal design load impedance for this network is 52 ohms, with a maximum VSWR of 2:1[2,3,4,5] Maximum efficiency is obtained when the amplifier is loaded to its rated input and the load impedance is within design limits.

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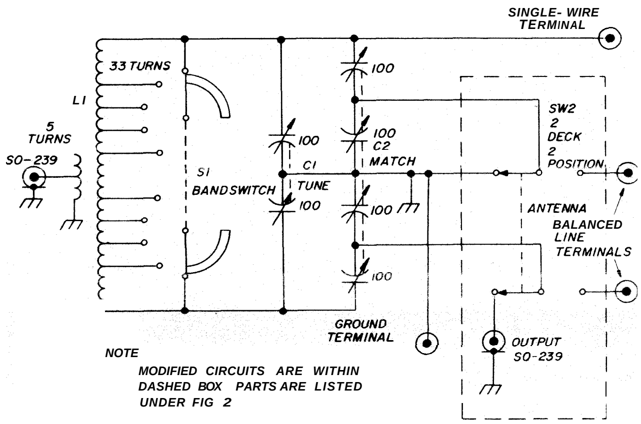


	Johnson part no.	description
C1	154-505-4	capacitor, 100ED30, 3kV peak, 0 19-cm (0.075-inch) spacing, 15-plateslsection, dual 100110 pF
C2	169-25	capacitor, 100EDA30, 3kV peak, 0.19-cm (0.075-inch) spacing, 15-plates/section, dual differential, 100110 pF
SW1	22.884	bandswitch
L1	23.1041	inductor, 33 turns no. 12 (2.1 mm), spaced 0.13 cm (0.05 inch), airwound on spreaders at 90°. Five turns in center are double spaced. Diameter 6.35 cm (2.5 inches). Winding length 12 cm (4.75 inches). Taps at 8.8. 12.7, 14.6, and 15.5 turns from each end. Coupling coil: 5 turns 2.6 mm (no. 10) double spaced, wound around center of inductor

fig. 1. Schematic of the Johnson Matchbox showing modifications (dashed box). The original parts in the Matchbox are shown for reference. These parts descriptions were omitted in early instruction manuals.

typical antenna mismatch at band edges

My triband beam, the CL-33, is assembled to favor the phone frequencies. The VSWR at the band edges is:

frequency MHz	vswr
14.00	4.0
14.35	1.8
21.00	1.5
21.45	2.0
28.00	3.0
29.70	4.7

Tuning coax with the Matchbox and experimentally selecting a line length will produce a 1:1 VSWR at the transmitter output on all frequencies within the antenna and Matchbox ranges.

test connections

The rear-panel photograph shows the simple hookup I used to experiment with coax-to-coax tuning before modifying. Remove the screw from the top center of the rear panel. Temporarily mount a SO-239, by one corner, into the screw hole. Run a lead from the center terminal of the SO-239 to the balanced-line terminal, directly below. Ground the other balanced-line terminal. Connect the end of the coax line from your beam to the temporary SO-239. Run the transmitter output through a VSWR bridge to the Matchbox input. Use the tuning procedure you used for balanced lines.

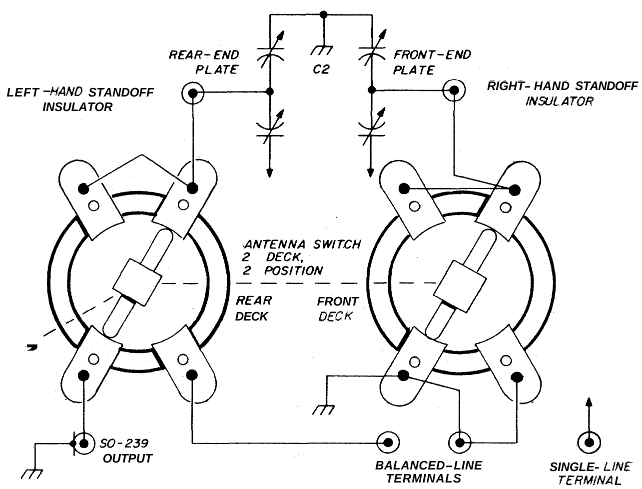
modifications

The schematic of fig. 1 shows the Johnson Matchbox as built with the modified circuits in a dashed box. The parts list in fig. 1 shows the values and descriptions of the parts used in the original Matchbox. (Early instruction manuals for the Matchbox did not include this information.) The parts needed to make the modification are shown in fig. 2. Proceed as follows:

1. Bypass or remove the relay. Run a line directly from the input SO-239 to the braid from the coupling coil.

2. Remove the tuning capacitor C1 by removing the nuts on the stators. Remove the support posts and slip C1 away from the solder lugs on the connecting leads. Do not distort the leads.

3. Remove the leads attached to the front and rear end plates of matching capacitor, C2. Remove the



quantity	description
1	antenna switch, Fair Radio cat. no. ML7464910-G1/407 or equivalent
2	standoff insulators, H.H. Smith 9502 or equivalent
1	SO-239 coaxial receptacle, Amphenol 83-1R
1	flexibleshaft coupling, Fair Radio cat. no. COUN007
1	switch extension shaft, brass or aluminum, 16.5 x 0.64 cm (6.5 x 1/4 inches)
1	bar knob with brass insert
1	switch mounting plate (refer to fig. 3 and text)

fig. 2. Graphic display of all connections for the modifications, from C2 through new SW2, to the output terminals.

other ends of these same leads from the balanced-line output insulators. The leads will be used to connect C2 to the standoff insulators. Do not remove the single-wire output lead.

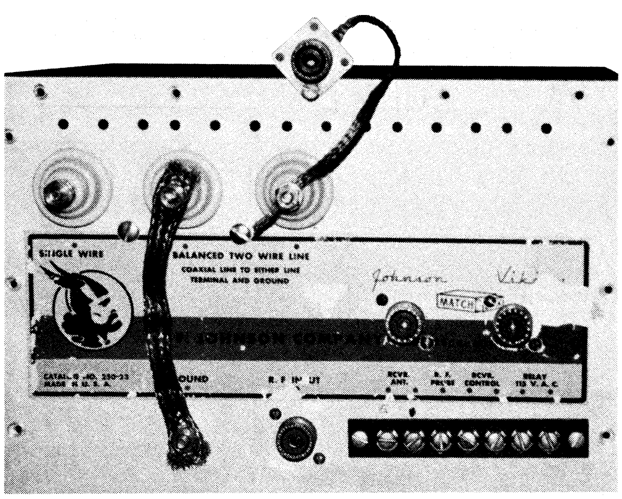
4. Install standoff insulators on each side of the center output insulator. See photograph of wiring connections.

5. Punch or drill a 1.27-cm (112-inch) hole for a SO-239 connector into the rear panel directly behind and below C1. Drill two mounting holes and install the connector from the inside. (You may want to add more connectors if your switch has more than two positions.)

6. Use **fig.** 3 as a template. Cut a 6.35 by 10.16 cm (2.5 by 4 inch) switch mounting plate from 0.17-cm (0.065-inch) aluminum sheet. Use a sharp center-punch and mark the centers for all holes, right through the template. All screw holes are for 6-32 (M3.5) screws. (I drilled mine for 8-32 [M8] to allow for minor errors.) The ML7464910-G11407 switch has a 1.27-cm (112-inch) shaft bushing. Centralab Ham Switch 2551 has a 0.95-cm (318-inch) bushing. The 2551 has six positions. The rear shaft bushing nut on C1 requires a 1.59-cm (518-inch) hole for clearance.

7. Assemble the switch plate and switch. Attach this assembly to the rear end plate of C1. Use 6-32 by 114-inch (M3.5 by 0.635 cm) brass screws. Longer screws will damage the C1 insulator. Install C1 in its original position. Attach connections from the band switch to the stators. Check the rear stator plate lead. It must clear the switch plate.

8. Drill a 0.635-cm (114-inch) hole in the cabinet front panel 6 cm (2-3/8 inches) above the C1 shaft hole.



In this view the rear panel holds an extra SO-239 which has been installed to experiment with coax-to-coax tuning.

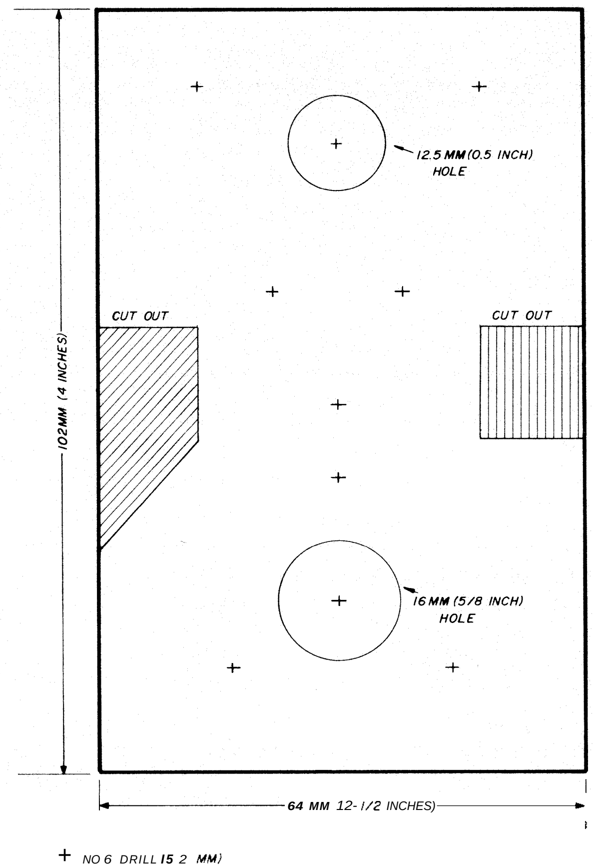


fig. 3. Switch-plate drilling template. Punch or drill through the drawing into the blank switch plate.

To complete wiring refer to the schematics, drawings, and photographs.

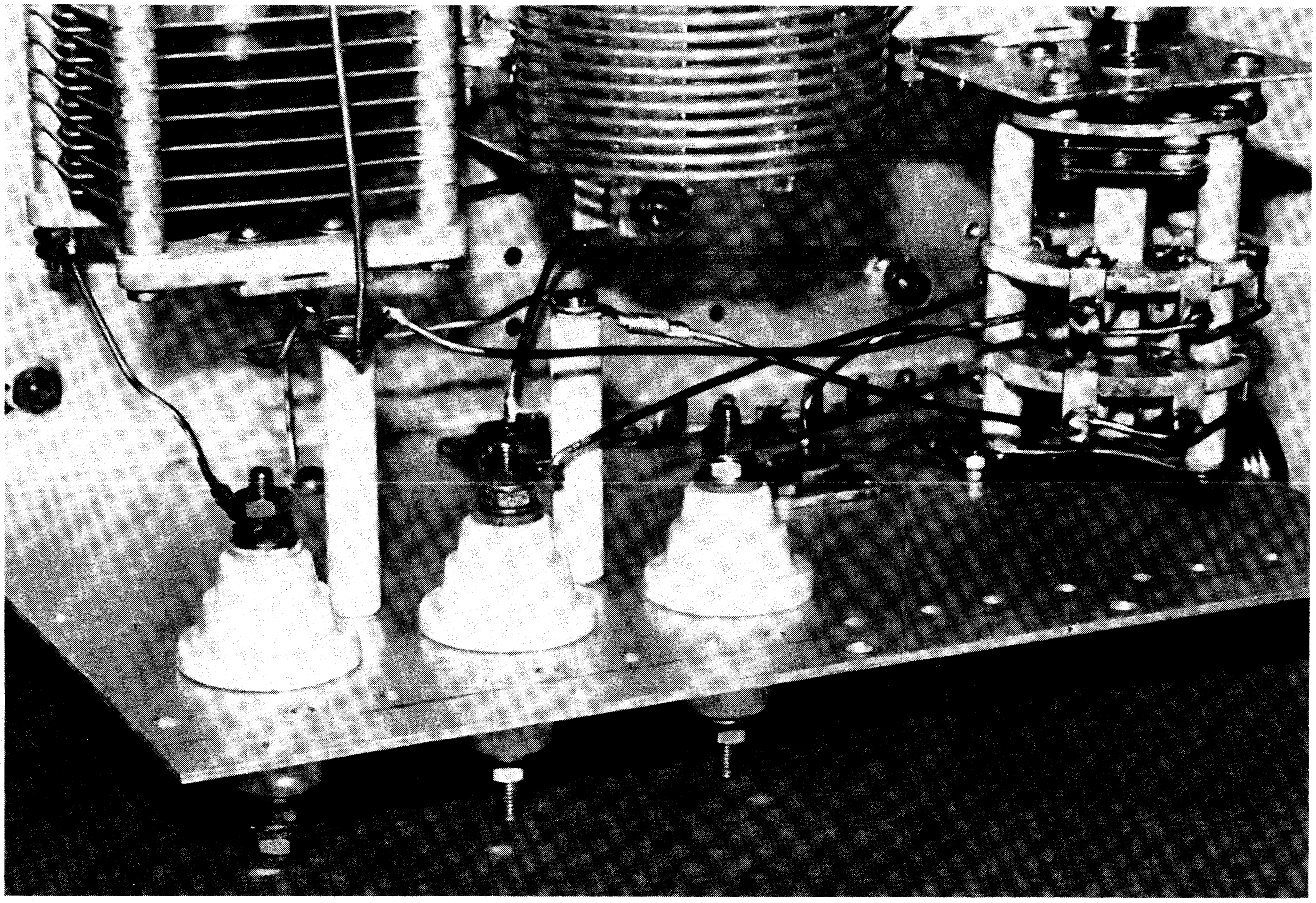
Before replacing the cover, run through these checks:

Temporarily install the dials and switch knobs. *Remember C1 and **C2** are hot! Use the insulated coupling on **C2.** Keep your hands clear. Even at low power, you can be burned!*

Run tests on 20, 15, and 10 meters. If the impedance at your transmitter is outside the Matchbox range on any band change the feedline length.6

feedline tests

My feedline length was determined by inserting a 180-degree 14.0 MHz line between my wall entrance panel and the Matchbox. The impedance at the input of this 7-meter (23-foot) line was outside the Matchbox range on 20 meters but satisfactory on 15 and 10 meters. I reduced the line length by making 0.3-meter (I-foot) cuts to 0.9 meter (3 feet). I recorded readings for each length, dial readings, and measured VSWR. A review shows that at a 5.8-meter (19-foot) line length, the system was in tolerance at the edges of all bands. I made a new 5.8-meter (19-foot) line and verified the previous tests.



All wiring connections from C2 to the new selector switch are visible in this inside view of the modified Matchbox. This photo also shows the two new standoff insulators which are used for the junction points.

wrap-up

Replace the cover. Record dial readings at each 100 kHz on 20 and 15 meters and at each 500 kHz on 10 meters. Using a dummy load I can tune and load the transmitter, set the Matchbox to the recorded dial settings, select the antenna, and start transmitting without a touch-up.

During one of the many tests conducted with this Matchbox, I found that I could match the CL-33 antenna on 7155 kHz. W4TBU at Henderson, Kentucky, was worked from Winter Haven using my triband CL-33 beam. The report was S2. I haven't tried loading the two-meter beam.

some additional suggestions

A number of switches are satisfactory for this modification. The ML7464910-G1/407 from Fair Radio Sales* is excellent and inexpensive. Centralab's 2551 Ham Switch is a good commercial unit. It has two decks and six positions. Contacts are rated at 9 amperes. It has 2000-Vac insulation.7

"Fair Radio Sales, Post Office Box 1105, Lima, Ohio 45802

The kilowatt Matchbox can be modified for coax tuning. Switch ML7464910-G1/407 will work in the kilowatt model. Catalog no. ML7762999-G1/397 switch requires more space, 7.62 cm diameter by 12.7 cm long (3 x 5 inches). It has two decks and three positions. Contacts and insulation are more than ample for the legal Amateur power limit.

Centralab switch JV9033 will also work in the kilowatt Matchbox. This switch has two poles, eight positions, 17-ampere contacts, and 3000-Vac insulation.7

references

1. Operation *and Instruction* Manual, Johnson Viking Matchbox.
2. Mack Seybold, W2RYI, "Components for Pi-Coupled Amplifiers," RCA Ham Tips, RCA Tube Division.
3. Drake T-4X Instruction Manual.
4. Heath SB-200 Instruction Manual.
5. Earl W. Whyman, W2HB, "Importance of Standing-Wave Ratios," ham radio, July 1973, pages 26-33.
6. "Feeder-to-Transmitter Matching Networks," The ARRL Antenna Book, American Radio Relay League, Newington, Connecticut, 13th edition, page 181
7. Centralab Industrial Distributor Catalog, Series 201.

ham radio

test-equipment mainframe

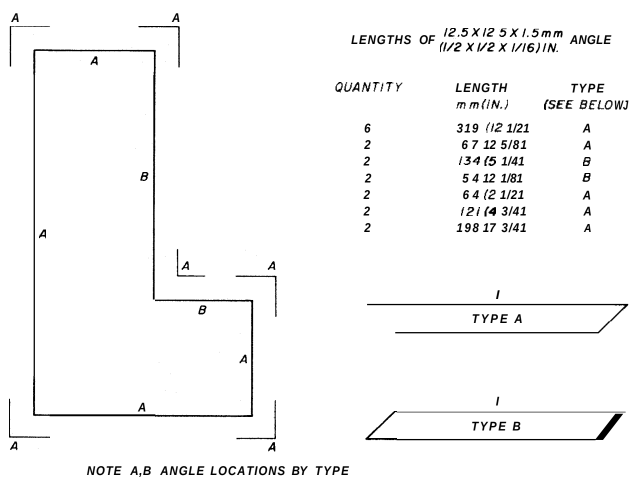
Construction ideas for those wanting convenience and a professional appearance for test equipment

Over the years I've built many pieces of test equipment, at first without paying much attention to size or appearance. Size became important because of extended operation aboard a sailboat and led to a series of units packaged in standard 51 x 77 x 128 mm (2 x 3 x 5 inch) miniboxes. As these units were developed, I realized that good appearance wasn't really difficult. (See reference for a description of many of these designs and for hints on obtaining good appearance.)

These units served me well but had some disadvantages. The main one was in assembly of a test set up, which involved chasing down the right unit or units, getting batteries together, and interconnecting the units to secure the desired signals. Storage was another problem, as well as keeping batteries on hand.

The appearance of the Tektronics 500-series plug-in units and mainframe led me to adopt some of their ideas and to the development of the mainframe and plug-in system described here — very handy and a great time saver.

The idea of the mainframe can be seen in the photo of the first (or prototype) version. At the bottom a chassis contains power supplies and control circuits, some interconnect circuits, and a series of switches. Receptacles at the top of the chassis are spaced to accommodate four miniboxes of the same



QUANTITY	LENGTH mm (IN.)	TYPE (SEE BELOW)
6	319 (12 1/21	A
2	6 7 12 5/81	A
2	134 (5 1/41	B
2	5 4 12 1/81	B
2	6 4 (2 1/21	A
2	121 (4 3/41	A
2	198 17 3/41	A

fig. 1. Dimensions of angle stock needed for the first version of rack. Location of the pieces are shown.

size 51 x 77 x 128 mm (2 x 3 x 5 inch) for individual instruments. A receptacle on the front panel allows use of another instrument on an extension cable or makes the power supplies and interconnect points available externally.

The chassis used in the prototype are special boxes, having an L-shaped cross section. This can be seen more clearly in the photo of the second version,

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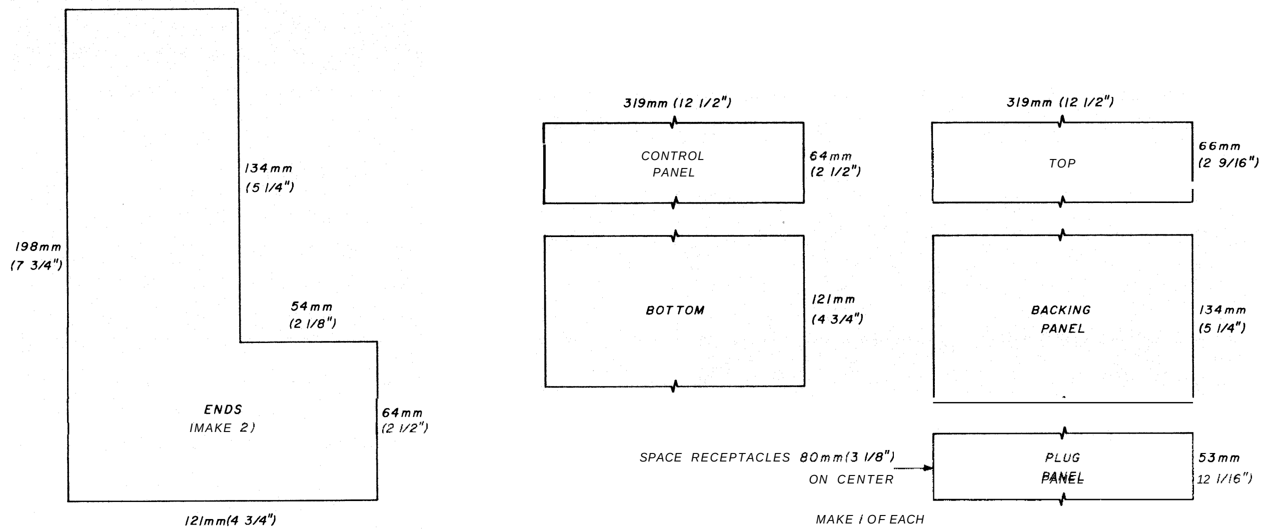


fig. 2. Dimensions of sheet metal for first version of the rack. Hardware "hobby" aluminum sheet is satisfactory.

which also shows the receptacles (Jones plugs) that accept the instrument packages. This version uses a different method of interconnect (tip jacks) instead of switches feeding a pair of buses.

mainframe construction

Two methods of construction were used. The four-unit frame is designed for hacksaw and tin-snip fabrication. The five-unit frame is designed for forming with a sheet-metal brake. With care and a little time, the bends of the second type can be made with a vise. The pieces may be fastened with rivets or self-tapping screws.

Dimensions of the four-unit rack are shown in **figs. 1** and **2** for the sheet metal and angle pieces respectively. The do-it-yourself stock available in hardware stores is easiest to work, although the rack could just as well have been made from sheet iron and iron angle.

Dimensions for the five-unit rack are shown in **fig. 3**. This unit can also be built from thin do-it-yourself aluminum stock. It will be more rugged if at least 1.3-

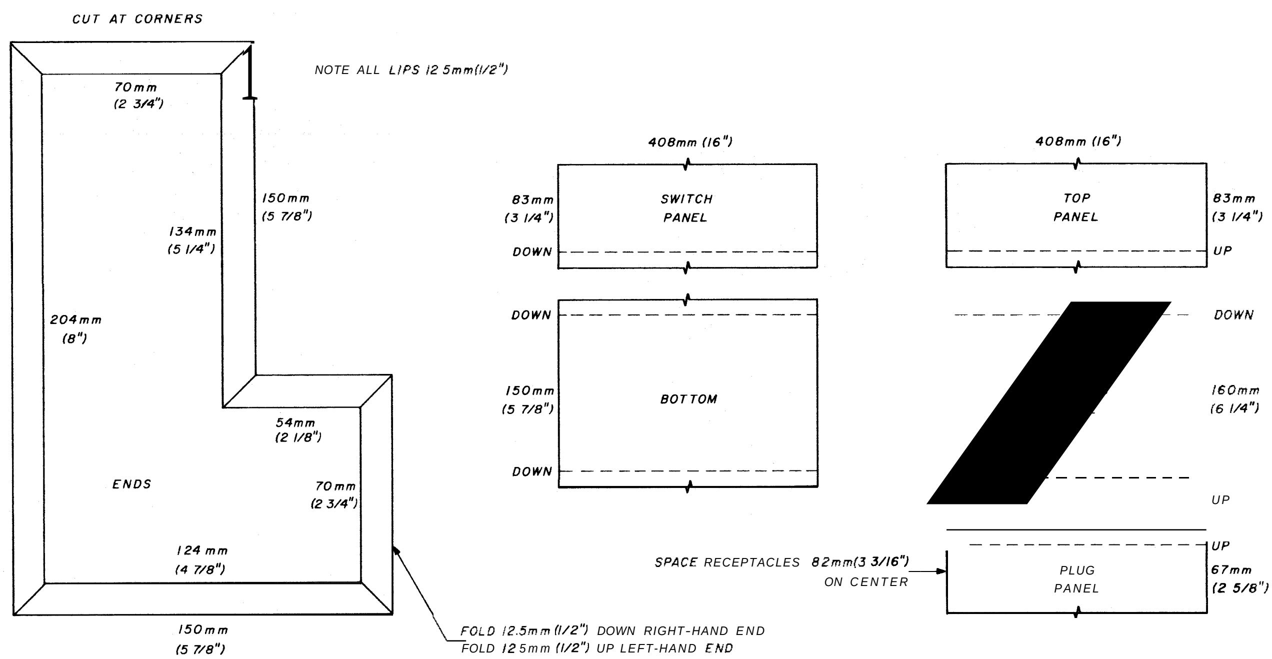
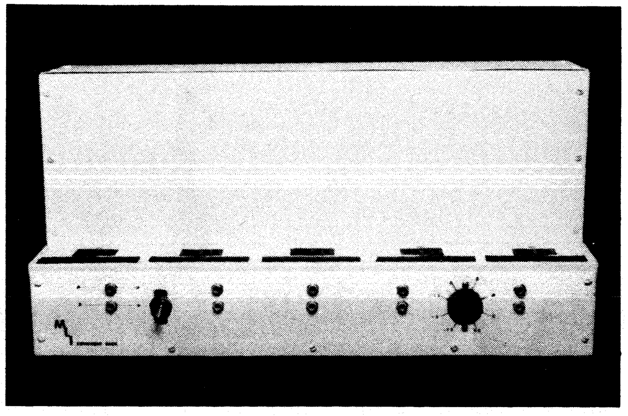


fig. 3. Sheet-metal dimensions for the second version of the rack. Bend on the dotted lines to 90 degrees. If prepainted sheet stock is used, watch the direction of bends carefully.



First version of the home-built equipment rack with home-built plug-ins. Instruments are described in reference 1. This unit can be built with tin snips and hacksaw.

mm (0.05-inch) stock is used. Aluminum sold in rolls as mobile-home skirting is satisfactory, as is sheet iron.

I've decided on a flat white finish for all small instruments, with black lettering. The mainframe was finished the same way. Steps for obtaining a good appearance are as follows:

1. Complete the instrument and get it working.

2. Remove the case and clean it with steel wool and soap. Polish any scratches; smooth all corners.

3. Spray with desired color paint, using several thin coats.

4. Add lettering, using a pressure transfer kit. Watch location with respect to dials and binding posts. If any are crooked, remove and start over.

5. With lettering complete, warm the case to about 92C (200F), which improves letter adhesion.

6. Finish the instrument with a *thin* coat of transparent spray, such as Krylon spray varnish.

7. Reassemble.

With a little practice, a professional-looking appearance can be obtained.

power supplies

Each of the units shown contains two independent power supplies. One provides 112 volts regulated. This is used mainly for instruments based on op amps but is also for general use. A 12.6-volt ac line is also brought out of this supply. The second unit provides +5 volts, primarily for TTL logic.

Circuits for the supplies used in the prototypes are shown in **fig. 4A.** These are IC regulated circuits. The original instrument design was based on ± 9 volts rather than ± 12 volts. If desired, provisions for these (or other) voltages can be made. Alternative connections for this requirement are also shown **(fig. 4B)**.

unit interconnections

In addition to the power leads, three bus leads are provided for unit interconnection. The interconnection schematic of the four-unit rack is shown in **fig. 5.** One bus, C, is common to all units. It is usually used for a sync signal. The other two can be switch selected as shown.

In the five-unit rack, the common bus is retained, but the unit leads A and B are brought to tip jacks, which gives more flexibility but is slightly less convenient.

The front panel also has an 8-pin receptacle (a tube socket was used in the prototype). This receptacle can be used with an extension cable to power another instrument or to power experimental equipment. It's convenient to make up several cables, a long and short one with instrument receptacles, and another pair with pigtail leads. A supply of "short preventers," made from alligator clips covered with transparent vinyl tubing, is a further convenience.

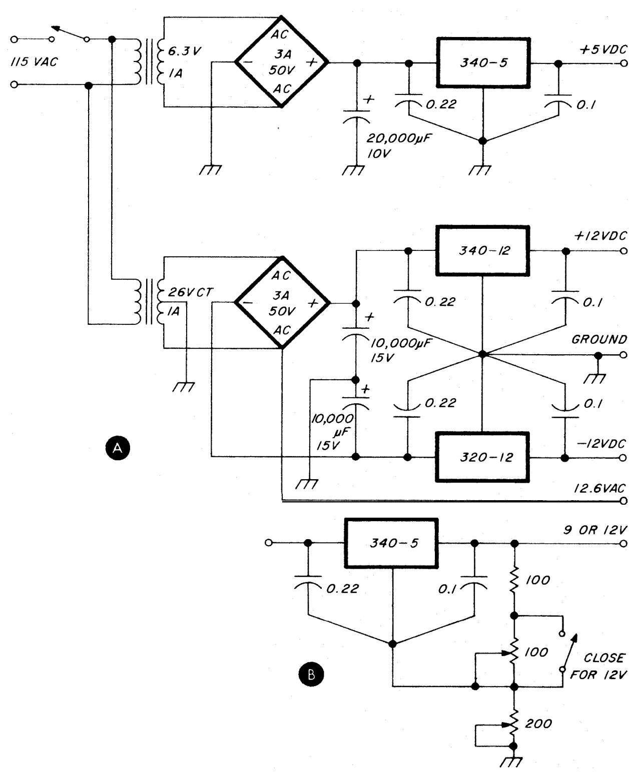


fig. 4. Circuit for power supply used in the racks, (A). The 320 regulator case must be insulated from the ground. (B) shows an alternative connection for +9, +12 volts. The negative line would be similar.

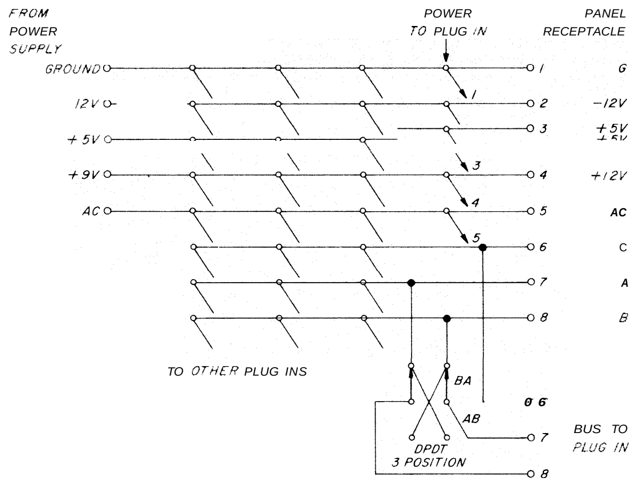


fig. 5. Schematic of interconnections for the first version of rack. For the second version, the leads from pins 7 and 8 of the receptacles end on tip jacks to give greater flexibility in interconnection.

The nominal 51 x 77 x 128 mm (2 x 3 x 5 inch) instrument cases are usually 54 x 77 x 131 mm (2-1/8 x 3 x 5-1/8 inches) and vary from one manufacturer to another. Rack spacing is laid out with this in mind.

The drilling and cutout pattern for the end of the instrument case is shown in **fig. 6.** This is for 8-pin Jones plugs. The alignment of the plug-in with the mainframe affects appearance, so the plug cutout is oversize to allow for adjustment.

A first trial at the rack used tube sockets for receptacles. These were undesirable, giving poor alignment and requiring too much insertion/removal force. The Jones plugs are much better.

battery operation

Most of the instruments used with these racks have low power drain, so battery operation for portable use is possible. A convenient way of obtaining this is to build up a case with a receptacle on each end. Eight A-cells will last quite a while as the 12-volt supply. Four D-cells with one diode in series will do reasonably well for the 5-volt supply, or a lantern battery can be used. If much portable operation is planned, use nickel-cadmiums instead of regular dry batteries.

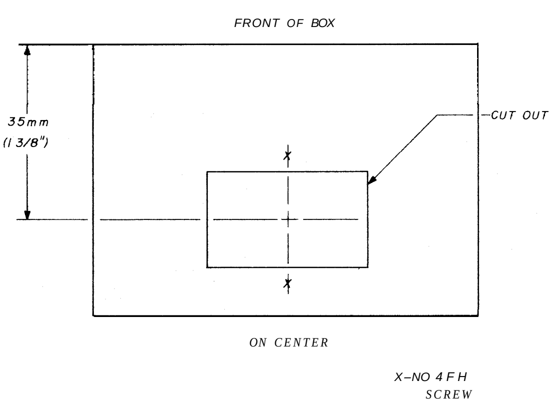
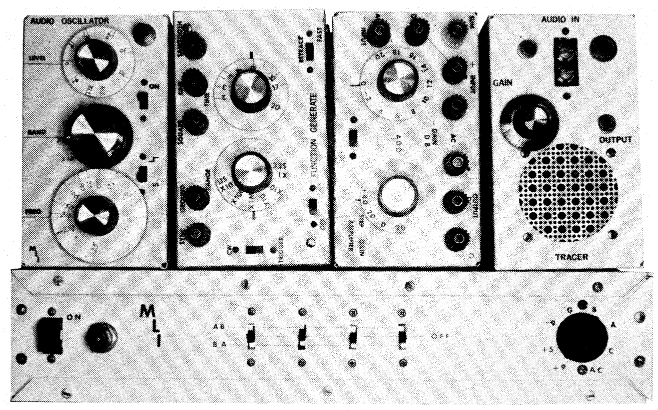


fig. 6. Location of the plug cutout on the end of the minibox. Location of the plug should be adjusted if necessary, since alignment affects appearance when the unit is plugged into the rack.

instruments

The instruments shown in the prototype setup are, from left,

Sine-wavelsquare-wave audio oscillator, 20 Hz-20 kHz.



Second version of the equipment rack designed for forming from sheet metal. Strips of rubber tape in front of the sockets stabilize the plugged-in instruments.

Function generator, square-wave, triangle, pulse and sine-wave, period, 20 microseconds – 20 seconds.

Summing step-gain amplifier, –40 to +40 dB gain.

Signal tracer or general-purpose amplifier

These instruments and others are fully described in reference 1.

Two of the instruments shown in the prototype photo have been refinished and relettered to fit the plug-in format. The other two have been modified for the connector plug but have not been refinished.

Additional instruments have been designed and built. I hope to describe these in a later article.

reference

1. R. P. Haviland, "Build-It Book of Miniature Test and Measurement Instruments," TAB, (No. 792), Blue Ridge Summit, Pennsylvania, 17214, 1976.

ham radio

scaling antenna elements

Many high-performance dipole and Yagi-Uda antenna systems have been developed over the years, but scaling them for use at other frequencies can produce disappointing results when the elements' length and diameter cannot, from a practical standpoint, be scaled directly. Fig. **1** was developed a number of years ago for element lengths near a *half* wavelength.' The advent of the modern hand calculator has turned nuisance calculations into a challenge.

using fig. 1

As seen in fig. **1,** the relative wave velocity on any element is a function of that element's length-to-diameter ratio, with:

$$\frac{l_n}{d_n} = r_n \qquad (1)$$

where l_n = the length of the nth element
$d,$ = the diameter of the nth element
r_n = the length-to-diameter ratio

Once the relative wave velocity, v_{rn}, has been determined from fig. **1,** the free-space wavelength and element length are related by:

$$l_o = \frac{l_n}{v_{rn}} \text{ or } l_o = \frac{d_n r_n}{v_{rn}} \qquad (2)$$

where l_n = the length of the nth element
l_o = the free-space wavelength

For changes in an element's diameter, the new and old lengths can be equated by:

$$l_o = \frac{l_n}{v_{rn}} = \frac{L_n}{V_{rn}} = \frac{D_n R_n}{V_{rn}} \qquad (3)$$

where $L,$ = the new length
$D,$ = the new diameter
V_{rn} = the new relative velocity
$R,$ = the new length-to-diameter ratio

practical examples

Length for **a** given diameter. How long should a 318-inch diameter rod be when it is a one-half wavelength radiator at 150 MHz? From the standard wavelength formula, a free-space, half-wavelength radiator is 39.3429 inches long.* Rearranging eq. 2 yields:

$$\frac{l_o}{d_n} = \frac{r_n}{v_{rn}} \text{ with } \frac{r_n}{v_{rn}} = \frac{39.3429}{0.375} = 104.9$$

By moving along fig. **1,** you will find a point on the curve where the length-to-diameter ratio divided by the relative velocity equals 104.9, $r_n = 96.3$ and $v_{rn} = 0.918$.

From eq. **1,** the rod length then becomes

$$\begin{aligned} l_n &= d_n \cdot r_n \\ &= (0.375)(96.3) \\ &= 36.12 \text{ inches} \end{aligned}$$

Change of diameter. Assume that one-half wavelength element is 391-118 inches long, and that its diameter of 1.5 inches should be increased to 2.0 inches for added strength. What should the new length be? By eq. **1**:

$$r_n = \frac{l_n}{d_n} = \frac{391.125}{1.5} = 260.75$$

From fig. **1,** the relative velocity, v_{rn}, is about 0.9411, and substituting into a rearranged eq. 3 produces:

$$\frac{l_n}{v_{rn} D_n} = \frac{R_n}{V_{,,}} = \frac{391.125}{0.9411(2.0)} = 207.802$$

Move along the curve on fig. **1** and find the new length-to-diameter ratio divided by the new relative velocity factor which gives the above ratio, or:

$$\frac{R_n}{V_{rn}} = 207.802 = \frac{194.3}{0.9351}$$

The new length-to-diameter ratio is 194.3, and the new relative velocity factor is 0.9351.
From this, the new element length becomes

$$\begin{aligned} L, &= R_n D_n = 194.3(2.0) \\ &= 388.63 \text{ inches} \end{aligned}$$

Change of frequency. Assume that a 146-MHz director has a diameter of 318 inch and a length of 35.0 inches, and is to be used at 14.2 MHz with a new diameter of 2.0 inches. What is the new length?

"Contrary to normal ham *radio* style, the examples in this article do not, for two reasons, include metric conversions. First, the element sizes are common in the U.S.. used here as examples rather than for absolute conversion to the metric system. Second, the added complexity of metric conversions would tend to hinder understanding of this article and its formulas.

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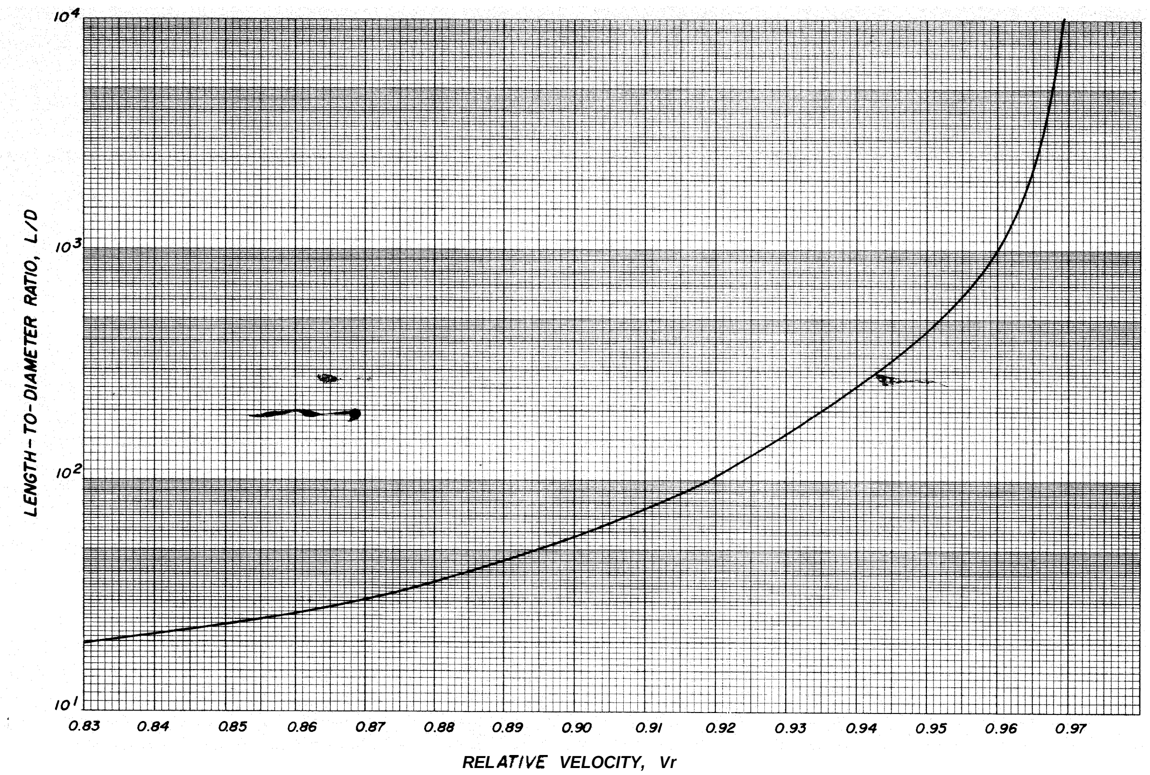

fig. 1. Graph of the cylindrical wire relative velocity vs. an element's length-to-diameter ratio.' This excludes the small spacing effects in series-fed antennas.

First:

$$\frac{l_n}{d_n} = \frac{35.0}{0.375} = 93.333$$

From **fig. !,** the relative velocity factor is about 0.9168, and the free space wavelength is:

$$l_o = \frac{35.0}{0.9168} = 38.1763 \text{ inches}$$
$$\text{or } 0.4722\lambda \text{ at } 146 \text{ MHz.}$$

From this, the 0.4722 λ at 14.2 MHz is 392.5162 inches (L_o.)

Therefore:

$$\frac{L_o}{D_n} = \frac{392.5162}{2.0} = 196.2581 = \frac{R_n}{V_{rn}}$$

Move along the curve on **fig. 1,** and find that:

$$\frac{R_n}{V_{rn}} = 196.2581 = \frac{184.0}{0.9375}$$

The new element length, L_n, becomes

$$L_n = R_nD_n = 184.0(2.0)$$
$$= 368 \text{ inches}$$

summary

As the above examples show, **fig. 1** can be used to solve a number of element scaling problems in a short period of time when a hand calculator is available. Moving from left to right on the figure increases (quite rapidly) the R_n/V_{rn} ratio, and solving for L, in terms of R_nD_n ***and*** $V_{rn}L_o$ (where L, $= L_n/V_{rn}$) is a good check (as well as refinement) of the ratio obtained from the **fig. 1** curve.

I have used this procedure many times to scale beam elements with excellent results. For example, I scaled one high-performance, free-space, 3-element Yagi-Uda array directly from the 2-meter Amateur band to the 40-meter Amateur band, and the maximum gain frequency shift on the 40-meter band was only 10 kHz!

It appears that good results can be obtained when the element lengths are within ±20 per cent of free-space, one-half wavelength, but the error is also a direct function of the number of antenna elements. I have not found this error to be significant in ***array*** scaling.

Fig. 1 assumes that the elements are either shunt-excited or are parasitic. When driven elements are series-fed, a small gap capacitance exists across the end of the feed line which is in parallel with the element self-series impedance. This affects the driven element impedance more than it does the driven element gain. When the driven series-fed element is essentially one-half wavelength, good scaling results occur when the gap is omitted in scaling this element.

'For greater accuracy, a full-size copy of the author's original graph is available by sending a self-addressed, stamped envelope to ***ham radio,*** Greenville, New Hampshire 03048.

reference

1. S. A. Schelkunoff and H. T. Friis, ***Antennas: Theory and Practice,*** John Wiley & Sons, Inc., Chapter 13; 1952.

ham radio

predicting close encounters:

Oscar 7 and Oscar 8

This article presents a fast, simple, and accurate method for predicting close encounters between Oscar 7 and 8

In early 1975, Amateur Radio operators added another first to their long list of communications accomplishments when two earth stations communicated via a path involving a direct satellite-satellite link.[1] Each station transmitted to AMSAT-Oscar 7 on 432 MHz; the signals were then relayed to AMSAT-Oscar 6 on 146 MHz, and back down to the ground on 29 MHz. Never before, in any radio service, had two satellites been directly interlinked to support communications between two ground stations.

Communication via the ESSE (earth-satellite-satellite-earth) path, using Amateur spacecraft, has been possible only when the satellites involved were relatively close to each other. Of course, it's also necessary that the transponder frequencies be suitable. Close-approach periods, lasting approximately three weeks, occurred about every six months for the AMSAT-Oscar 6 and AMSAT-Oscar 7. I still clearly remember the reference orbit (first orbit of the GMT day) on Wednesday, February 5, 1975, during the first close-approach period. AMSAT (and the new RS) satellites are reserved for special experiments on Wednesdays, with a very interesting test scheduled for this particular day. The distance between the two spacecraft was less than 1200 km on the reference orbit, and Amateurs with 432-MHz transmit capabilities were being encouraged to try for interlinking QSOs. Strict cooperation was needed if the tests were to succeed; anyone transmitting to Oscar 6 on 146 MHz might desensitize the transponder and con-

By **Martin Davidoff, Ph.D., K2UBC,** Mathematics Department, Catonsville Community College, Catonsville, Maryland 21228

fuse stations receiving on 29 MHz. The results are history — the tests were a huge success. Cooperation was excellent; rapid fading, which many feared might be a serious problem, was minimal, and dozens of QSOs were made. Rag-chewing quality signals were heard on the 29-MHz downlink from W2GN, W8DX, VE2BYG (VE3SAT), K3JTE (W3PK), and many others uplinking on 432 MHz.

AMSAT later received written reports of completed ESSE QSOs from fifty-five Amateurs in twelve countries during the January/February, 1975, close-approach period. Contacts were reported for satellite separation distances ranging up to 2000 km, and reception of the Oscar 7 mode B beacon repeated by the Oscar 6 transponder was reported for satellite separation distances ranging up to 7000 km. ESSE tests involving these two spacecraft continued during periods of close approach until mid 1977, when Oscar 6 ceased operation.

The transponder frequencies for Oscar 8 where chosen so that ESSE tests could resume. Both mode A and mode J are suitable. **Fig. 1** illustrates the links and transponder frequencies involved in ESSE communications using Oscar 7 and Oscar 8. Almost immediately following the launch of Oscar 8 (March 5, 1978), it became apparent that ESSE communications could take place when sensitivity measurements by K1HTV and W6CG of the Oscar 8 transponders (modes A and J) showed that, as long as the satellite agc wasn't being activated, good return signals could be obtained over line-of-sight paths of 3000 km by ground stations using as little as 100 milliwatts to a dipole. These expectations were accidentally confirmed a few days later when a number of Amateurs using the mode B transponder on Oscar 7 were inadvertently repeated by the mode A transponder on Oscar 8. To date, I'm not aware of any successful interlinking experiments using mode J.

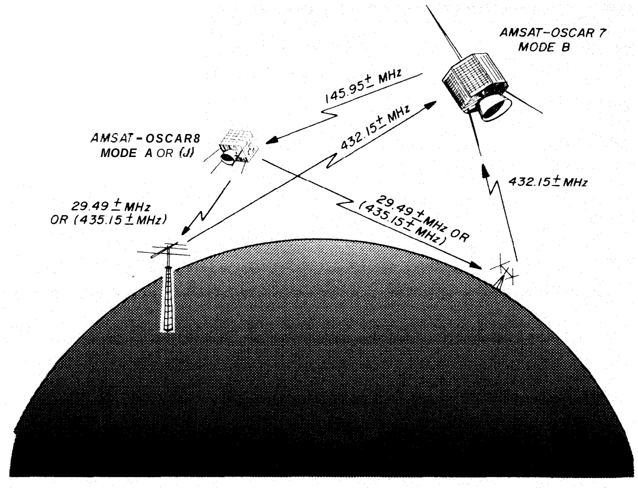

fig. 1. Diagram of the paths and frequencies involved when two earth stations communicate via a direct link between Oscar 7 and Oscar 8.

> Oscar 7, now more than four years old, is beginning to show signs of age. The unscheduled mode jumping and erratic behavior of the mode-B transponder and beacon observed in recent months are likely to continue. With a little luck, ground station cooperation (use minimal uplink power), and careful management by the command stations, Oscar 7 may continue to operate for many more years.
>
> Operating modes for Oscar 7 are not specified in the 1979 WGPAJ Orbit Calendar because instructions to the ground command stations are being formulated on a short lead-time basis to help prolong the satellite's life (battery voltage and internal temperatures are being carefully monitored), and because the spacecraft often jumps modes of its own accord. Users may operate through whichever transponder is on, except, of course, on UTC Wednesdays when Oscar 7 and Oscar 8 are reserved for special authorized experiments, or if AMSAT announces a spacecraft emergency. The WGPAJ Orbit Calendar can be relied upon to accurately provide Oscar 7's position, but it's up to the user to determine the operating mode. Atmospheric absorption at 10 meters (resulting from high solar activity), and almost total loss of the mode-B beacon make this a bit more involved than it first appears. Often, the only way to determine if the mode-B transponder is on is to transmit in the uplink passband, Mode B itself appears to have two distinct operating states, a normal state where it works as well as when first launched, and a degraded state where sensitivity and power output are way down. If nothing is heard on mode A or mode B, the transponder is probably in the recharge mode.

the problem

With Oscar 6 and Oscar 7 in similar orbits, close-encounter periods lasted a few weeks and were repeated on approximately a six-month schedule. The situation with Oscar 7 and Oscar 8 is quite different. Close encounters now last only part of an orbit, but they occur almost once a day. A given ground station is in range of only a small percentage of such close encounters. Since satellite separation distance is a critical factor in ESSE communications, stations seriously interested in experimenting with this mode need an accurate method of determining when close encounters will occur and what the separation distance will be. The goal of this article is to present a fast, simple method for obtaining this information. It involves the "close-encounter curves" shown in **fig.** 2. With these curves, Amateurs can select the best orbits (and sections of orbits) for linking experiments. The mathematical basis of close-encounter curves is described in the appendix.

close-encounter curves

The vertical axis of the close-encounter curve **(fig.** 2) represents the separation distance between AMSAT-Oscar 7 and AMSAT-Oscar 8. This information is presented as a function of the elapsed time (in minutes) since the last Oscar 7 ascending node. Eight curves are shown in **fig.** 2. Each one is labeled by a parameter τ, which indicates the difference (in minutes) between the Oscar 7 and Oscar 8 ascending nodes. Our convention is to use positive values for τ when Oscar 8 crosses the equator *after* Oscar 7. For example, if an Oscar 7 ascending node occurs at 01412, and an Oscar 8 ascending node occurs at 01442, the curve labeled $\tau = 3$ applies; if both ascending nodes occur at the same time, the curve labeled $\tau = 0$ is used. Fractions of a minute should be rounded off. The following series of questions and answers best illustrate how the curves are interpreted.

Q. Under what conditions will Oscar 7 and Oscar 8 pass closest to one another?

A. Fig. 2 shows that the intersatellite distance approaches a minimum value on orbits when the Oscar 8 ascending node occurs about four minutes after the Oscar 7 ascending node. The point of closest approach will occur about 27 minutes after the Oscar 7 ascending node. You can also see that the separation distance is never less than 550 km, the difference in altitudes of the two spacecraft.

Q. How can a ground station at 40 degrees north latitude pick the optimum orbits for intersatellite communications?

A. A station at 40 degrees north latitude has access to Oscar 7 only until 23 minutes after the ascending node (on an overhead pass). The station cannot access the satellite at the point of intersatellite closest approach. Looking at **fig.** 2 you can see that, beginning at 15 minutes after the Oscar 7 ascending node, the intersatellite distance becomes less than 1700 km when $\tau = 2$, 3, or 4 minutes (Oscar 8 ascending nodes occurring 2, 3, or 4 minutes after the Oscar 7 node). The time slot between 15 and 23 minutes after the Oscar 7 node is the best window. We've arbitrarily chosen 1700 km as our cutoff point because signals will be down 10 dB relative to 550 km (absolute closest approach) due to $1/r^2$ path losses.

Each ground station using close-encounter curves will find it convenient to shade in the area during which access to Oscar 7 is not possible. For example, a station at 40 degrees north latitude would shade in the region between 23 minutes and 35 minutes on **fig.** 2.

Q. For my ground station at 40 degrees north latitude, I see that the time period between 15 and 23 minutes after an Oscar 7 ascending node is optimal for intersatellite communications if Oscar 8 has an ascending node 2 to 4 minutes later. How is this information used to select specific orbits?

A. Just read down the Oscar 7 and Oscar 8 time columns in the W6PAJ orbit calendar until you locate an Oscar 8 node occurring 2 to 4 minutes after an Oscar 7 node. When you find one, use your usual tracking aid (Satellabe, Oscarlocator, *etc.)* to determine if *both* satellites are within range during any portion of the time slot (15 to 23 minutes after the Oscar 7 node).

Q. Should a station at 40 degrees north latitude concentrate on morning descending orbits or evening ascending orbits for intersatellite communications experiments?

A. It's hard to say. **Fig.** 2 shows that, during 1979, the evening orbits provide shorter intersatellite distances. Other factors, however, such as the normally lower transponder loading early in the day, might lead to better results on morning orbits. For morning passes, a station at 40 degrees north latitude would select orbits with $\tau = 4$, ***5,*** *or* 6 and concentrate on times between about 35 and 40 minutes past the Oscar 7 ascending node.

Q. Over what period of time can **fig.** 2 be used?

A. If the relative orientation of the Oscar 7 and Oscar 8 orbital planes remained constant, **fig. 2** could be used indefinitely. Because there is a slight drift in the relative orientation of these planes, a graph like the one shown in **fig.** 2 must be drawn for a specific date. In this case, the predicted positions of the orbital planes for July 1, 1979, were used. The drift, however, is so slow that **fig.** 2 will provide reasonably good results (within one minute or 200 km) for all of 1979.

Q. How often can one expect to find Oscar 7 and Oscar 8 in a position suitable for interlinking tests?

A. A close approach, while both satellites are in range of your ground station, will occur about once every seventeen days for evening (local time) orbits. For morning orbits, the figure is also about once in seventeen days. On the average then, if morning and evening passes are considered, any ground station will have a good shot at interlinking about once every eight or nine days.

These answers were derived as follows. The probability of a close encounter occurring on a given south-north satellite pass is equal to the probability of τ being 2, 3, or 4 minutes. With three suitable one minute time slots out of a 103-minute period, the probability of a close encounter occurring on a specific orbit is 31103. On the average, there are two

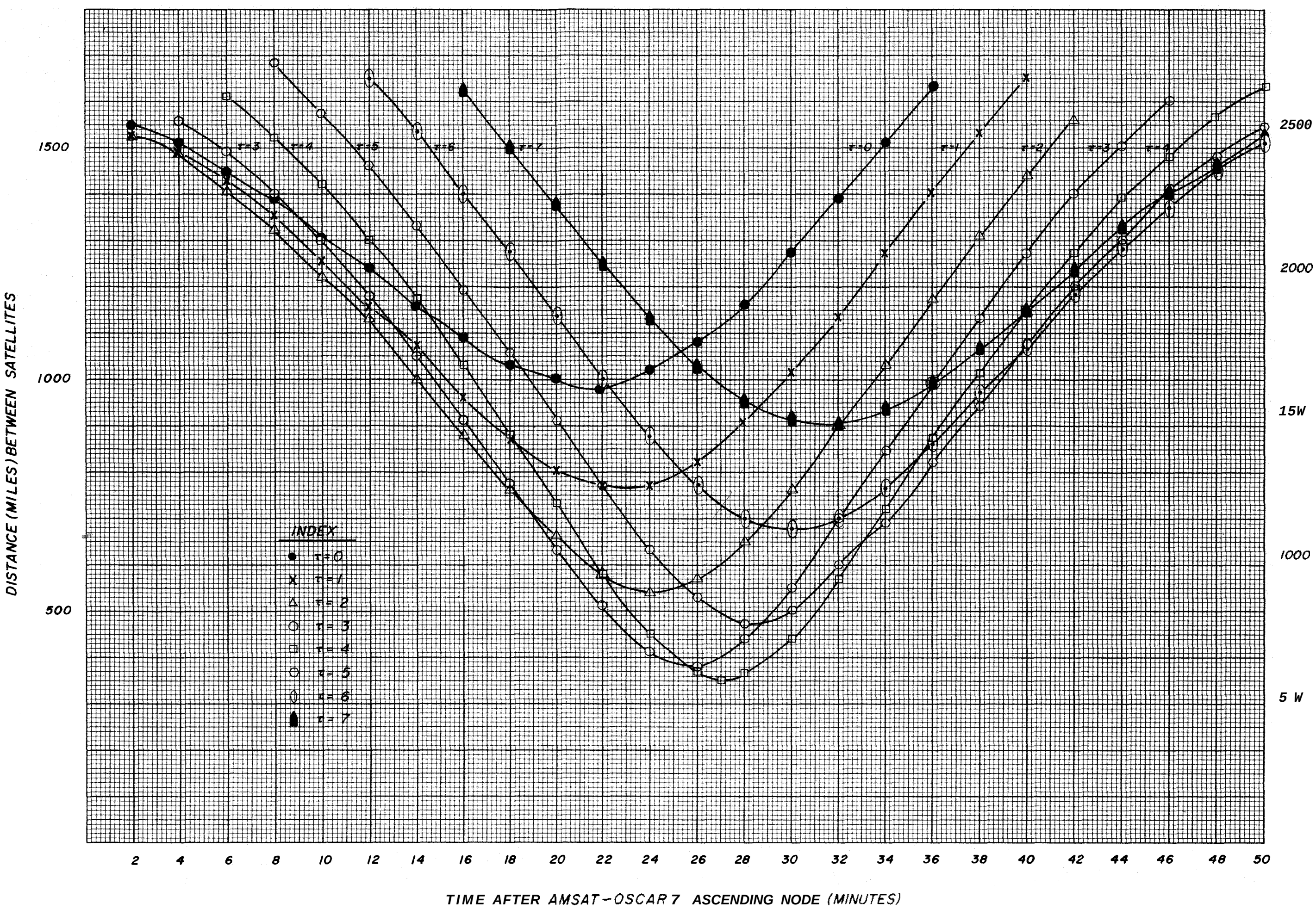

fig. 2. Close-encounter curves for Oscar 7 and Oscar 8, useful for 1979.

close in Oscar 8 south-north passes each day, so the probability of a close encounter is 2(3/103) per day, which translates into once every seventeen days. The analysis for north-south orbits is similar, except for the fact that the three desirable time slots for τ are 4, 5, and 6 minutes.

Keep in mind that Oscar 7 must be in mode B if communications are to be possible during a close encounter. Concern for Oscar 7's health and longevity has forced AMSAT to begin selecting modes on a day-by-day basis, so you just have to take a chance that it will be in mode B.

Q. Is Doppler shift a serious problem?

A. Doppler shifts are not a serious obstacle. The real problems are spurious responses (birdies) and desensitization in the receiving system, Finding your own downlink signal will always involve some searching, and you'll quickly realize the value of good filtering on the transmitter and receiver (to prevent birdies and desensitization) and thorough testing of your ground station when the satellites are not in range (to learn the location and characteristics of any remaining birdies).

The following considerations should help you reduce the amount of searching you must do for your downlink. The total Doppler shift consists of contributions from each of the three links involved. The 70-cm link(s) produce the largest effect. For a mode-B/mode-A linkup, the total Doppler should be less than ±6 kHz. A rough, but close, guess of the value can be obtained by just considering the position of Oscar 7 (mode-B transponder) and using the frequency offset regularly seen for a mode-B QSO. With a mode-B/mode-J linkup, the Doppler can be up to ±12 kHz. Once again, the value can be estimated by concentrating on the 70-cm links. If both satellites are moving either toward you or away from you, the Doppler shifts will tend to cancel. During optimal-access periods, one of these two cases usually exists, so searching for your downlink can be confined to a ±6 kHz window.

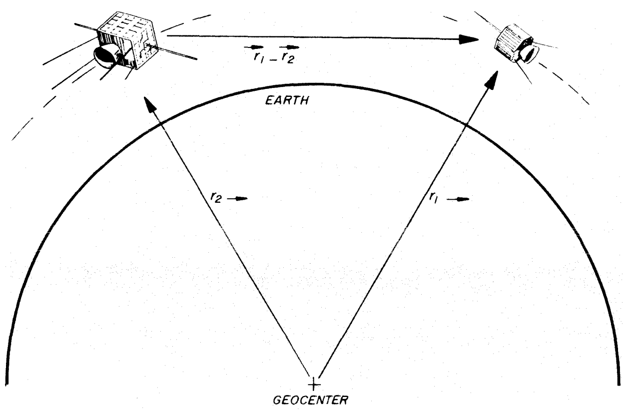

fig. 3. The position vectors for two satellites. This diagram represents Oscar 7 as satellite **1** and Oscar 8 as satellite 2.

Q. I've been using the close-encounter curves to listen to the Oscar 7 mode-B beacon through Oscar 8 mode A, and signal strength correlates reasonably well with the curves but not completely. Why?

A. Distance is only one aspect of Oscar 7/Oscar 8 radio-link performance. Another important factor is relative orientation of the spacecraft antennas. This can be accurately modeled if a computer is available. The close-encounter curves would enable us to save a great deal of computer time by restricting path-loss studies involving antenna patterns to a small portion of a limited number of orbits. Ground stations must also take into account transponder loading and the performance of the satellite-ground communications link if they hope to explain all observations of the Oscar 7 mode-B beacon through Oscar 8.

table **1.** Notation for intersatellite distance problem. SSP = subsatellite point. Numerical values for Oscar **7** and Oscar 8 are shown in brackets.

parameter	satellite **1** [Oscar 71	satellite 2 [Oscar 81
radial distance (geocenter to satellite)	r_1[7,831.km]	r_2[7,281.km]
position vector (geocenter to satellite)	$\vec{r}_1$	$\vec{r}_2$
period (note: $P_1 \geq P_2$) (minutes)	P_1[114.945 minutes]	P_2[103.231 minutes]
orbital inclination	i_1[101.7 degreesl	i_2[99.0 degreesl
elapsed time since ascending node of satellite (minutes)	t	$t-\tau$
latitude of SSP at time indicated	$\phi_1(t)$	$\phi_2(t)$
longitude of SSP at time indicated	$\lambda_1(t)$	$\lambda_2(t)$
longitude of SSP at ascending node	λ_{10}(occurs at $t = 0$)	λ_{20}(occurs at $t = \tau$)

appendix

The mathematical derivation of the close-encounter curves is outlined in this appendix. Note that it is not necessary to read this section to use the close-encounter curves. To understand the following material, you need some background in spherical coordinates and three-dimensional vectors. Our analysis focuses on two satellites in circular orbits, as shown in fig. 3. The notation used is summarized in table **1.** Note that the subscript 1 is used to refer to the satellite with the longer period (higher altitude! The intersatellite distance, $s(t)$, is given by the magnitude of the vector $(\vec{r}_1 \quad \vec{r}_2)$

$$s(t) = \sqrt{(\vec{r}_1 - \vec{r}_2) \bullet (\vec{r}_1 - \vec{r}_2)}$$

$$= \sqrt{r_1^2 + r_2^2 - 2(\vec{r}_1 \bullet \vec{r}_2)} \qquad (1)$$

To evaluate $\vec{r}_1 \bullet \vec{r}_2$, express the Cartesian components of each position vector in terms of the spherical coordinates of the given satellite (θ = colatitude, λ = longitude). Both coordinate systems have their origins at the geocenter and rotate with the earth. The orthogonal unit vectors i, **j**, and k are defined as follows: i is along the line joining the geocenter to the intersection of the equator and the prime meridian,] is along the line joining the geocenter to intersection of the equator and the 90 degree east meridian, k is along the line joining the geocenter and North Pole. The position of satellite 1 is given by

$$\vec{r}_1 = r_1 \sin\theta_1 \cos\lambda_1\, i + r_1 \sin\theta_1 \sin\lambda_1\, j + r_1 \cos\theta_1\, k \qquad (2A)$$

The position of satellite 2 is given by

$$\vec{r}_2 = r_2 \sin\theta_2 \cos\lambda_2\, i + r_2 \sin\theta_2 \sin\lambda_2\, j + r_2 \cos\theta_2\, k \qquad (2B)$$

Transforming from colatitude, 8, to latitude, 0, $(0 = 90° - \theta)$

$$\vec{r}_1 = r_1 \cos\phi_1 \cos\lambda_1\, i + r_1 \cos\phi_1 \sin\lambda_1\, j + r_1 \sin\phi_1\, k \qquad (3A)$$

$$\vec{r}_2 = r_2 \cos\phi_2 \cos\lambda_2\, i + r_2 \cos\phi_2 \sin\lambda_2\, j + r_2 \sin\phi_2\, k \qquad (3B)$$

The inner product, $\vec{r}_1 \bullet \vec{r}_2$, of eq. **1** can now be evaluated

$$s(t) = \sqrt{r_1^2 + r_2^2 - 2r_1r_2[\cos\phi_1 \cos\phi_2 \cos(\lambda_2 - \lambda_1) + \sin\phi_1 \sin\phi_2]} \qquad (4)$$

The coordinates 0 and λ, for each satellite, appearing in the brackets on the right hand side of eq. **4,** are the same as those of the respective subsatellite points (SSPs):

$$\phi_1(t) = \arcsin\left[\sin i_1 \sin\left(360° \frac{t}{P_1}\right)\right]; \qquad (5A)$$

$$\phi_2(t) = \arcsin\left[\sin i_2 \sin\left(360° \frac{t-\tau}{P_2}\right)\right]; \qquad (5B)$$

$$\lambda_1(t) = \lambda_{10} + \frac{t}{4} + (-1)^{n_2+n_3} \arccos\left[\frac{\cos\left(360° \frac{t}{P_1}\right)}{\cos\phi_1(t)}\right]; \qquad (6A)$$

$$\lambda_2(t) = \lambda_{20} + \frac{t-\tau}{4} + (-1)^{n_2+n_3} \arccos\left[\frac{\cos\left(360° \frac{t-\tau}{P_2}\right)}{\cos\phi_2(t)}\right] \qquad (6B)$$

where

$$n_2 = \begin{Bmatrix} 0 & 90° \leq i \leq 180° \\ 1 & 0° \leq i < 90° \end{Bmatrix}$$

$$n_3 = \begin{Bmatrix} 0 & \phi(t) \geq 0° \begin{pmatrix} \text{Northern} \\ \text{Hemisphere} \end{pmatrix} \\ 1 & \phi(t) < 0° \begin{pmatrix} \text{Southern} \\ \text{Hemisphere} \end{pmatrix} \end{Bmatrix}$$

Here are the sign conventions: North latitudes are positive, south latitudes are negative, all longitudes are in degrees west, and longitude displacements toward the west are positive.

The sign convention for longitudes adopted in **eqs. 6A** and **6B** is the same as that used in the Satellabe, Oscarlocator, W6PAJ Orbit Calendar, and most other U.S. and Canadian Amateur literature. It's a very convenient convention for stations located between 0 and 180 degrees west longitude. Most non-Amateur literature, recognizing the computationai advantages of a right-hand coordinate system, is based on a different sign convention – east longitudes and displacements towards the east are regarded as positive. This approach was used in the best treatment of orbits available in the Amateur literature – Peter D. Thompson, Jr., "A General Technique for Satellite Tracking," QST, November, 1975, page 29.

Although the situation may sound confusing at first, it's really just a rninor problem once you're aware of its existence. Since both conventions have merit and are well established, the best course of action for radio Amateurs working on basic computations appears to be to use a right-hand coordinate system for computations and then, as a final step, transform to the U.S./Canadian convention.

To compute the intersatellite distance, s, as a function of time for a specific set of two orbits, **eqs. 5A, 5B, 6A,** and **6B** are substituted in **eq. 4.** The result is indicated symbolically by:

$$s = s[(i_1, i_2, P_1, P_2, r_1, r_2), \Delta\lambda_o, \tau, t]. \qquad \textbf{(7)}$$

You can see that s depends on a set of constants $(i_1, i_2, P_1, P_2, r_1, r_2)$, the parameter $\Delta\lambda_o \equiv (\lambda_{20} - \lambda_{10})$, which is a measure of the difference in longitudes at the ascending node, the parameter τ, which expresses the difference in time of the two ascending nodes, and the time t measured from the ascending node of satellite 1. Once the two satellites are chosen, the six constants are known and **eq.** 7 expresses the fact that the intersatellite distance is a function of three variables.

If you have a TI-59, HP-67, or similar programmable hand calculator, you can use **eq.** 7 in conjunction with equatorial crossing data to first compute AX, and τ for a specific set of two orbits, and then calculate the distance, s, every 2 minutes over the course of the orbits. Although this method works, it is very time consuming if you're trying to evaluate a large number of orbits for their suitability for interlinking experiments. This leads to the problem of finding some simple way of expressing **eq.** 7 in graphical form.

The AX, term in **eq.** 7 can be written:

$$\Delta\lambda_o \equiv \lambda_{20} - \lambda_{10} = \frac{\tau}{4} + \Delta\lambda^* \qquad \textbf{(8)}$$

where AX* is a slowly varying function describing the relative orientation of the orbital planes of the two satellites.

The 1979 W6PAJ orbit calendar, using data by Dr. Tom Clark (W3IWI), predicts that AX* will be 18.6 degrees on January 1, 1979, 39.0 degrees on July 1, 1979, and 19.4 degrees on December 31, 1979. If AX* can be treated as a constant, then **eq.** 7 will depend on only two variables, and a function of two variables can often be illustrated on a single graph as a set of curves. To test this approach, three graphs (each graph consisting of a set of close-encounter curves) were drawn for AX* = 18.6 degrees, 19.0 degrees, and 19.4 degrees. (All computations were performed on an HP-97 programmable calculator.) From these graphs, it was evident that a 0.4-degree change in Ah* had a negligible effect on the close-encounter curves. A single graph (fig. **2**) can therefore be used for all of 1979.

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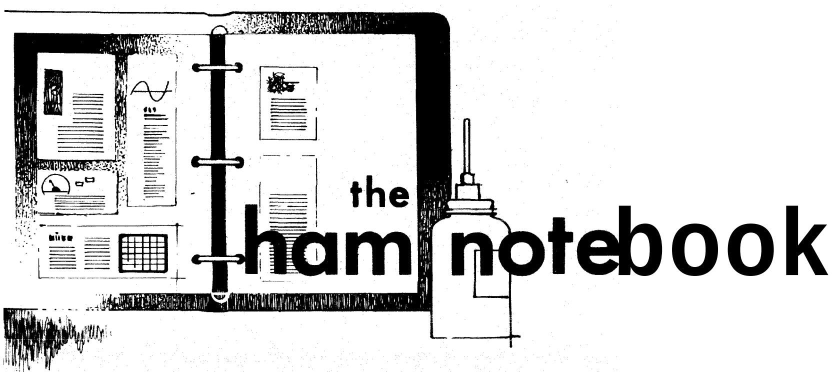
ham radio



a very simple synthesizer system

Some time ago my wife, Barbara, and I built a general-coverage receiver with a digital frequency readout. The local oscillator is a free-running MC1648, tuned by a variable capacitor with a small varactor for bandspread. Since the i-f is 10.7 MHz, the local oscillator tunes 13.7-40.7 MHz for the 3-30 MHz receiver range. The stability of this oscillator was adequate for normal operation, but when the receiver was left tuned to 30.0000 MHz for 24 hours the thermal variations in a normal room caused a slow frequency drift of about ± 1.5 kHz. This was most annoying. We dreamed of a synthesized local oscillator to alleviate this problem. Unfortunately all the synthesizers looked complicated and our small chassis was already pretty crowded, so for a long time we did without.

Recently, MacKeand[1] published a simple frequency-lock-loop circuit that reminded me of an earlier, more complex, circuit by Ryder[2]. Study of the short article by DeLaGrange[3] gave us a good insight into the theory of frequency-locked loops but no practical circuit for this application.

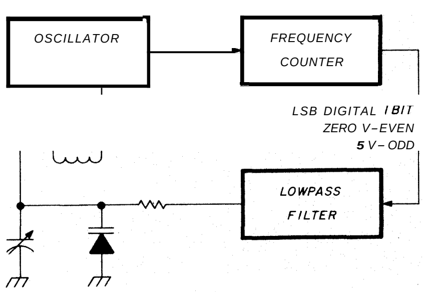


fig. 1. Basic frequency-locked-loop system.

Finally, I realized that a very simple approach could be taken, which I illustrate by the following example.

If I tune my receiver to 30.0000 MHz and it begins to drift up in frequency to 30.0001 MHz, all I have to do is sense this small change and apply it through the varactor to drive the frequency down. When it's driven down until the frequency counter reads 30.0000 MHz again, I can start forcing it back up again, hovering about the transition between 30.0000 and 30.0001 MHz. This could be accomplished simply by sensing the **1** bit of the latch on the least-significant bit of the frequency counter and applying it to the varactor through a lowpass filter. Then any odd LSB signal will drive the frequency slowly down, and an even count will drive it slowly back up.

What if the frequency were initially drifting down? When the counter reached 29.9999 MHz, the feedback scheme would drive it further down until the even count 29.9998 MHz came up. The **1** bit would change and the circuit would hover between 29.9998 and 29.9999 MHz. We could therefore lock onto any frequency in 200-Hz increments. If we invert the **1** bit before applying it to the lowpass filter, we can lock onto all the in-between 100-Hz points by taking the even-odd transition for lock instead of the odd-even. A block diagram of the scheme is shown in **fig. 1.** A slightly more elaborate diagram of the scheme used in our receiver is shown in **fig. 2,** and a complete schematic, except for the frequency counter, is shown in **fig.** 3.

Actually, the most important part is the use of the op amp U4B to perform the frequency lock. Although I show a free-run position, I never really use it anymore. The receiver just smoothly steadies itself on the nearest stable transition and otherwise behaves exactly as it did before we added the U4 circuit.

I'd like to emphasize that this scheme can be applied to any receiver with a digital frequency counter so long as it has a latched output, which I think all of them do. No special indicators of lock are needed, because proper operation is indicated by the receiver's slowly changing between 30.0001 and 30.0000 MHz – or whatever other frequency is selected. The actual frequency drift is only on the order of ± 10 Hz, and it is so slow that I can't detect it by ear. All the ICs

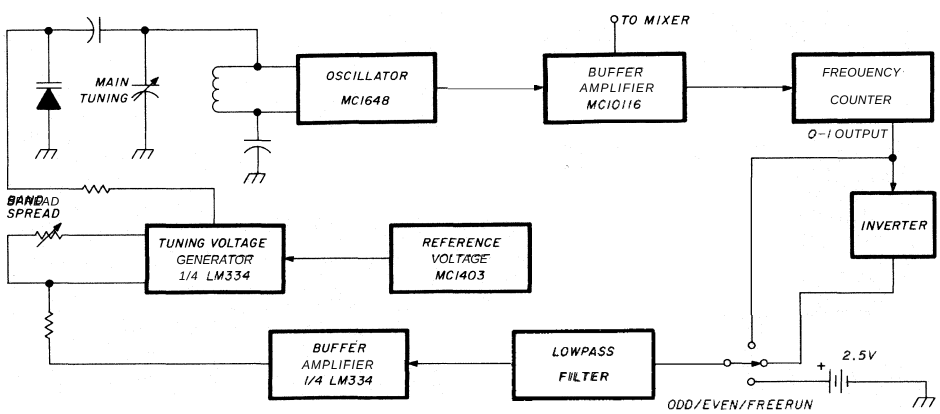


fig. 2. Frequency-locked local oscillator block diagram.

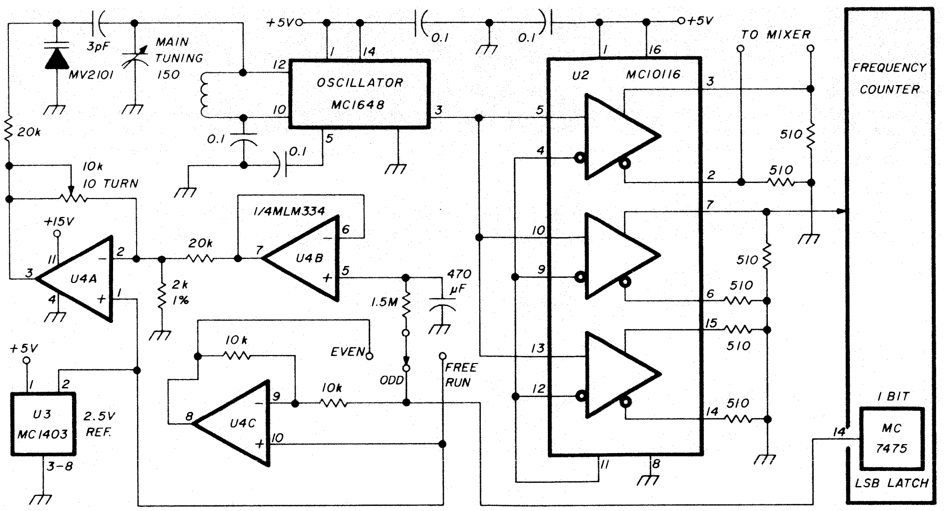


fig. 3. Frequency-locked-loop schematic.

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Pat O'Neil, AA7M, and Barbara O'Neil
Motorola Semiconductor, Inc.

Heath HW-101 sidetone control

I'm basically a night owl. If you're like me, you find many a late-night CW ragchew interrupted by friend wife complaining of "that loud beeping." I finally decided to correct the situation on my HW-101 by adding a sidetone volume control. Extra front-panel controls on radios tend to look somewhat tacky, so I decided that the addition must not be conspicuous.

After some pondering, I decided on a concentric af gain control as the solution. Heathkit provided the lever knob and knob insert. All that was needed was a switched dual control.

A call to the local parts distributor proved informative. They carried Clarostat controls in modular form; custom dual controls were no problem. The inside shaft would hold the original af gain-control knob; the rear pot, therefore, must be a 1-megohm replacement for the present control. (Don't forget the switch.)

I decided to control the sidetone by replacing R318 (100 kilohms) in the audio amplifier circuit with the front control in the dual assembly. Then, by connecting C311 to the wiper of the sidetone control, the amount of sidetone injected into the af circuit could be controlled. Hookup is straight-forward; I suggest, however, you use shielded cable for the interconnections.

The modification looked good, provided no surprises, and best of all – works great!

J. K. Davis, AD9M

XR-205 waveform generator as a capacitance meter

A 205 chip, a counter, and a calculator provide a means of measuring capacitance to within 1 or 2 per cent. The frequency is determined by a capacitance connected across pins 14 and 15 of the chip. *C* ***times*** *F* is a constant, *k*, with *C* in microfarads and *F* in Hz. This constant is truly constant over a very wide frequency range. The specification sheet gives this constant as 400, but I find it to be 260 with my generator, so this is apparently a nominal value.

I use a compression trimmer of about 200-1500 pF, permanently wired across pins 14 and 15, with parallel binding posts and short clip leads for connecting additional capacitance. The trimmer tunes across the i-f range.

I was fortunate in having several 1 per cent capacitors as standards, but only one is required. The first step is to determine the capacitance of the trimmer plus strays, C_{tr}, at a known frequency, after which the constant, *k*, may be determined from

$$k = FC_{tr} \quad (1)$$

The procedure for determining C_{tr}, the trimmer capacitance, is as follows:

Set the trimmer to about mid range and note the frequency, F_1. Connect a known capacitor and note the new frequency, F_2. Divide F_2 by F_1 and call this quotient *f*. Now calculate C_{tr} from the following formula:

$$C_{tr} = \frac{fC_s}{(1-f)} \quad (2)$$

where C_{tr} = trimmer plus stray capacitance (μF)

C_s = standard or known capacitance (μF)

$f = F_2 F_1$ (Hz)

Now calculate k from $k = F_1 C_{tr}$. Save F_1 for measuring unknown capacitors.

The measurement procedure, after k and C_{tr} have been determined, is to first set the frequency to F_1 with the trimmer capacitor. Connect the unknown capacitor and measure F_2. Divide k by F_2 and subtract $C_{tr} = C_{unknown}\ (C_u)$:

$$C_u = \frac{K}{F_2} - C_{tr} \quad (3)$$

As before, C is in μF, F in Hz. These calculations would be difficult without a pocket calculator but are easy with one. Assuming the trimmer, C_{tr}, is 300 pF, with 75 pF in parallel, k/F_2 would come out 0.000375 . . . on the calculator.

A half-dozen 1 per cent capacitors, from a few pF to 0.1 μF, were measured within 1 per cent of the nominal value. Indications are that the accuracy holds up to several hundred μF. EXAR-205 chips available from JAMECO.

W. S. Skeen, W6WR

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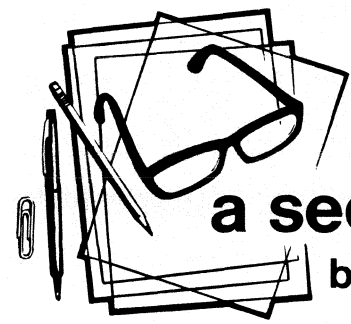


a second look

by Jim Fisk

Many of the problems that Amateur Radio has had to face over the past few years — and will continue to face in the future — can be traced directly to one cause: the demands of an increasing number of Radio Amateurs for finite space in the radio spectrum. More and more stations have been squeezed into the same bands, creating a congested mess for the Amateur operator, particularly during peak occupancy periods on weekends and holidays. In many cases the effects of crowded band conditions are aggravated by poorly designed ham equipment: transmitters which contribute broadband noise and splatter, and receivers which cannot cope with nearby strong signals. The technology to solve these problems is available, but very seldom is it put to work; and until Radio Amateurs demand improved equipment performance it probably won't be.

Herein lies the problem. How do you press for cleaner emissions from transmitters or push for receivers that will handle strong signals without overloading when there are no performance standards? And not only are there no performance standards, in general the published equipment specifications have not kept pace with technology. Gone are the days when receiver sensitivity and selectivity were the only things you looked for in a communications receiver; although today's receivers must contend with an abundance of closely-spaced strong signals, receiver "specmanship" has changed little since the days of vacuum tubes.

A few of the manufacturers have begun to provide data on dynamic range, intercept point, and blocking, but you cannot compare equipment from different manufacturers because they don't use the same standards or test procedures. And since there are no standards, receivers with identical operating specifications often have vastly different on-the-air performance. Intercept point is currently in vogue, and as one of my correspondents recently pointed out, "everyone suddenly has a '+20 dBm' receiver — whether it costs $250 or $5000!" This points up the need for standardized test procedures. Specifications for intercept point and dynamic range are meaningless unless the input signal separation is specified, and that crucial information is usually not available to the consumer. If it's not, ***caveat*** *emptor!*

The preponderance of "+20 dBm" receivers reminds me of the time a few years ago when every ssb transmitter on the market was advertised with third-order IMD down "more than 30 dB." It didn't matter whether the final was built with rf power tubes, TV sweep tubes, or transistors — the magic number for third-order IMD was always −30 dB. Then W6SAI and a few others pointed out that most of the TV sweep tubes couldn't do better than −22 dB in rf linear service, and some were as bad as −18 dB! That spelled the beginning of the end for TV sweep tubes in amateur transmitters; as new ssb transmitters were introduced, more and more were designed around tubes and transistors which were intended for linear rf service.

Receiver intercept point and transmitter IMD are only two of the problem areas. How about those built-in speech "processors" that often add so much distortion they make large sections of the band virtually useless? (And as an aside, what are we to do about the operators who refuse to turn them off even when they're told how bad they sound?) Established performance standards won't improve operating habits, but they would make it a lot easier to purchase equipment that meets your needs. And in the long run, a realistic set of standards will inevitably result in better Amateur equipment for all of us.

Jim Fisk, W1HR
editor-in-chief

comments

noise-figure measurements

Dear HR:

Bob Stein's article on noise-figure measurement in the August 1978, issue was very good but I disagree with his statement that, for equal signal and image channel gains, the error will be 3 dB. The correct expression is:

$$F_{double} = \tfrac{1}{2}\,(F_{single} + 1)$$

and can be found in *Radio Astronomy* by John Kraus. In decibels the proper difference is:

F_{double}	F_{single}
10 dB	12.8 dB
6 dB	8.4 dB
2 dB	3.4 dB

Stein's comment about putting a filter at the input of a preamp and not putting one after the preamp is very important. If one were to build a receiver with no filter after the preamp, the noise-figure meter would be correct for the entire receiver and best performance would not be achieved.

Charles H. Solie, WB5LHV
Houston, Texas

WB5LHV's letter referenced a book with which I was not familiar, and would have been welcome if for no other reason than bringing it to my attention. As it is, he certainly posed an argument which caused me a considerable amount of thought and research. To start with, the authorities for my discussion of image-response error in my article on automatic noise-figure measurements are as follows:

1. Albert Van der Ziel, *Noise: Sources, Characterization, Measurement,* pages 154-156.

2. George E. Valley, Jr. and Henry Wallman, *Vacuum Tube Amplifiers,* page 720.

3. Merrill I. Skolnik, *Introduction to Radar Systems,* page 365.

4. W.W. Mumford and Elmer H. Scheibe, *Noise Performance Factors in Communications Systems,* pages 45 and 61.

***Eq. 13** in my article appears in slightly different, but equivalent, form in reference 2, above. To clarify the situation brought up by WB5LHV, it is necessary to start with the Friis/IEEE expression of noise factor. By definition, it includes only that noise from the input termination which appears at the output via the principle-frequency transformation of a heterodyne system, and does not include spurious contributions from an image-frequency transformation. The single-sideband (channel) noise factor, F, is defined by the formula*

$$F = \frac{N_{to}}{GkT_oB} \qquad \textbf{(1)}$$

where N_{to} *is the total noise power output from the receiver*
G *is the gain of the receiver*
k *is Boltzmann's constant,* 1.38×10^{-23} *joule/°K*
T_o *is the standard reference temperature,* $290°K$
B *is the receiver bandwidth in Hz*

Since the noise output of the receiver is the sum of (a) the internally generated noise plus (b) the termination noise times the receiver gain,

$$F = \frac{kT_rB + GkT_oB}{GkT_oB} \qquad \textbf{(2)}$$

where T_r *is the receiver noise temperature (often shown as* GT_e*, where* T_e *is the effective noise input temperature).*

Consider now a single-channel receiving system in which the receiver has no image rejection (equal gains in the signal and image channels). If a narrow-band noise source were connected to the receiver input, the preceding expressions of noise factors would apply. However, if a broadband noise source is used, there will be noise applied to both channels, so that the double-sideband noise factor, F'*, will be given by the relation*

$$F' = \frac{N_{to}}{GkT_o(2B)} = F2 \qquad \textbf{(3)}$$

or

$$F = 2F' \qquad \textbf{(4)}$$

On the other hand, if a double-channel radio astronomy receiver is under consideration, there is signal information in both channels and the receiver noise contribution is divided equally between channels. Therefore, the noise factor of a double-channel receiver in this application is expressed as

$$F' = \frac{\tfrac{1}{2}(kT_rB) + GkT_oB}{GkT_oB} \qquad \textbf{(5)}$$

$$= \frac{kT_rB + 2GkT_oB}{2GkT_oB} \qquad \textbf{(6)}$$

$$= \frac{kT_rB + GkT_oB}{2GkT_oB} + \frac{GkT_oB}{2GkT_oB} \qquad \textbf{(7)}$$

*Then from **eq. 2**,*

$$F' = \tfrac{1}{2}F + \tfrac{1}{2} = \tfrac{1}{2}(F+1) \qquad \textbf{(8)}$$

*Thus, both relationships for F are true, but are defined differently, depending on the application. **Eq. 3** is applicable to a single-channel receiver which has no image rejection, because the receiver noise contribution will degrade only the signal channel, there being no signal information in the image channel. In the double-channel radio astronomy receiver, there is signal information in both channels, so that **eq. 8** defines the relationship between the single- and double-sideband noise factors.*

Since in radio communications we are concerned only with single-channel receivers, the discussion in my article is correct.

Robert S. Stein, W6NBI

antenna gain and directivity over ground

A discussion of Yagi gain and directivity, reference antennas, and the effects of planet Earth

The concept of antenna directivity and antenna gain is now several decades old; in fact, the terms are now quite familiar and are used by both technical and nontechnical people alike, apparently without much thought to what they mean — and also what they do not mean!

Directivity is commonly defined as the ratio of the maximum radiated energy flux density (at some "best" azimuth and elevation angle) to the average radiated energy flux density, averaged over the entire solid angle or sphere. Note that this property is determined solely by the complete three-dimensional spatial pattern of radiated energy. Directivity has a nice conceptual ring; it should, and presumably does, measure quantitatively the ability of the antenna to focus, concentrate, or direct its radiated output in a specific direction — compared with a reference antenna which has zero directivity (*i.e.*, equal output in all directions). This reference antenna is often referred to as an "isotropic radiator." However, this reference antenna is totally fictitious — nobody knows even in principle how to make such an antenna.* But it is a useful concept.

The gain of an antenna at any azimuth and elevation angle is normally defined as the ratio of actual maximum radiated energy flux density to that which would be produced by an isotropic radiator whose total radiated output power is the same as the antenna's input power. The gain of the antenna is then exactly the same as the antenna directivity multiplied by the radiation efficiency of the antenna. The radiation efficiency of an antenna is less than unity due only to conductor losses (and sometimes dielectric losses) and earth (ground) losses. In the remainder of this article I will assume these losses to be negligible and therefore will equate gain to directivity numerically, even though their definitions are different.†

It has been customary to object to the use of an isotropic radiator as the reference antenna because it is fictitious and physically unobtainable. Two suggestions have been made for alternative references:

*It is conceptually possible to approximate an isotropic source by the superposition of a very large number of independent and incoherent dipole sources whose orientations are uniformly distributed over the entire solid angle or sphere. It certainly would not be simple to make such a system, at least in the high-frequency region, nor to prove the accuracy of this approach to the isotropic assumption.

†Nearly all "normal" high frequency antennas have radiation efficiencies which closely approximate unity. This is not true for very short radiators, nor for antenna systems with significant ground losses.

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

(1) the infinitesimal dipole, whose pattern can be closely approximated by a very short linear radiator; and (2) the half-wave dipole. The infinitesimal dipole has a theoretical directivity or gain (over isotropic) of 1.5 (1.76 dB), and the half-wave dipole an isotropic gain of 1.64 (2.15 dB). Note, however, that these directivities or gains have been stated *only* for free-space conditions. Moreover, these free-space dipole reference alternatives are still fictitious and physically unobtainable in the real world, *where real antennas interact with the real earth.*

Antennas — a favorite topic of many, providing unlimited, and often heated discussions. Whether on the air or at any reasonable size gathering of amateurs, antenna facts and fallacies abound — quads vs Yagi, short verticals, EME arrays, the interests are endless.

Undoubtedly, one of the most popular subjects is the Yagi antenna. Everyone, the contester, the DXer, and even the person with just a casual interest, wants the maximum performance from their antenna, and as often is the case, maximum performance for the money spent. In many cases, however, the two do not equate; for its performance, that new five-element, 20-meter monobander on the 30-foot boom might as well be a dipole strung between a couple of trees. Is the manufacturer to blame? Not really, he's just building a marketable product. The fault lies with us, the user. Fact — for the most part, gain of a Yagi antenna is dependent upon boom length, with the number of directors being a significantly less critical factor. However, amateur fallacy would have everyone believe that the greater the number of directors, the better the antenna, regardless of boom length.

In a similar vein, consider the performance of a receiver in the Amateur Radio service. A few years ago performance, or actually lack of, was an often overlooked and misunderstood factor. Today, as a result of numerous articles in the Amateur Radio magazines, most amateurs are acutely aware of performance. This has resulted in a new generation of receivers from the manufacturers; receivers which now exceed previously unheard of performance standards.

Unfortunately, the Yagi antenna is far from being an easy subject. The antenna may appear to be very simple, but its performance depends upon a large number of interrelated variables, each factor affecting the others. As it turns out, the advent of the modern-day, high-speed digital computer has reduced the months of tedious calculations and emperical designs to a few minutes in front of a CRT terminal. This is not as simple as it may seem, for as we all should know, the computer is just a tool. Solving the problem still requires research and most importantly understanding.

As anyone who has heard the antenna talks by W2PV knows, Jim Lawson understands Yagi antennas. A look at Jim's station and its performance should be ample proof. On the personal side, Jim Lawson, W2PV, received his PhD in Physics from Michigan in 1939 and has worked for General Electric since 1945, the last fifteen years with Corporate Research and Development in Schenectady, New York. Jim's amateur career started in 1934 when he was first licensed as W9SSO. His other calls were W8QIU and the more well known WA2SFP from the 1960s.

As a preview of what's to come, over the next seven to eight months, *ham radio* will be printing a series of articles written by W2PV dealing with the design of Yagi antennas. Topics will include: computational methodology, simple Yagis, Yagi variations, ground effects, perferred designs, scaling and element tapering, and stacking. However, to start the whole series off in the correct perspective, this article will deal with the antenna in the real world and our perceptions of antenna gain.

Having known Jim for several years and having benefited many times from this material and his knowledge, I can say, for anyone with an interest in Yagi antennas, read the articles; they will be well worth waiting for. **K1XX**

Conceptually, one can still use the free-space (fictitious) reference antennas, even though the real antenna is used over earth or ground. This has the difficulty, however, that one cannot experimentally make or use any of the reference antennas. Furthermore, the gain of an antenna referred to a free-space standard, even though quite correct, gives a value which is unnaturally high for most technical users. Nevertheless, it is a value which can in principle be derived, for example, from a complete measurement of the space energy flux density pattern of an antenna. The (isotropic) reference radiation is simply the average flux density over the complete 4π solid angle (complete sphere), which is clearly just one-half of the average flux density over the 2π solid angle of the irradiated hemisphere. The half space below the conducting ground plane is, of course, not irradiated.

table 1. Yagi gain(s) in free space (dB).

reference	lower Yagi	upper Yagi	stacked Yagis
isotropic	10.28	10.28	13.37
infinitesimal dipole	8.52	8.52	11.61
half-wave dipole	8.13	8.13	11.22

Another possibility is to use a reference antenna, preferably one that can be easily constructed, and place it at the same height as the antenna, or perhaps substitute it for the antenna in its actual position. This concept is appealing and is popular* because it suggests a relatively simple measurement technique. Unfortunately, the measurement technique is *not* simple, either in practice or in concept; moreover, gain ratios of different antennas measured in this way are generally *not* the same as if referred to the free-space reference, and also *not* the same as the actual ratios of peak energy flux densities. There is also the nagging question of what the "position" of an antenna system is (where one should substitute a reference dipole). For example, is the "position" of a stacked array over ground its mathematical center? The basic reason for this confusing state of affairs is that a dipole reference antenna itself exhibits a gain (referenced to free space) which depends upon

*See, for example: The ARRL ***Antenna** *Book,* 73th edition, American Radio Relay League, Inc., Newington, Connecticut, 1974, page 43.

height over ground. Most importantly, the gain occurs at an elevation angle different from that of the test antenna, even though the antenna is located in the same position!

An example will illustrate the nature of this confusing conceptual problem. This example will approximate the situation for a large stacked Yagi system now in fairly wide use on the 14-MHz band. The Yagis are each constructed with six evenly-spaced elements on a boom 0.66 wavelength long. The lower Yagi is at a height over ground of 0.6 wavelength and the upper Yagi is at a height over ground of 1.6 wavelengths. I have calculated the pattern(s) and maximum gain(s) of these Yagis (individually and stacked) by methods which I shall not attempt to justify in this article,* but which I believe are essentially correct.

table 2. Reference half-wave dipole gain over earth (dB).

reference (free space)	height 0.6λ	height 1.6λ	height 1.1λ
isotropic	9.18 (25°)	8.52 (9°)	8.69(13°)
infinitesimal dipole	7.42 (25°)	6.76 (9°)	6.93 (13°)
half-wave dipole	7.03 (25°)	6.37 (9°)	6.54 (13°)

These gain figures, together with the elevation angles at which maximum gain occurs, are shown in **tables 1,** 2, and 3. **Table 1** shows the (hypothetical) free-space gain of the Yagi system (in decibels) referenced to three different free-space standards. **Table** 2 shows the (hypothetical) gain(s) of reference half-wave dipole(s) over ground at the same height(s) as the Yagi system (the "height" of the stacked array is taken as its mathematical center). I've also included the elevation angle at which maximum energy flux density occurs. **Table** 3 shows the gain(s) of the Yagi system referred to various standards, including half-wave dipoles at the same "height" over ground.

Constructing **table** 3 presents a fundamental conceptual difficulty which needs resolution. In comparing the ratio of energy flux densities of the Yagi with the (substituted) reference dipole, what should be the elevation angle or angles at which this ratio is taken? The energy flux densities of the Yagi system and the reference dipole *both* vary with vertical angle, *but vary differently.* One can now define gain in any of several ways:

A. The maximum ratio of Yagi flux density, F_Y, to reference dipole flux density, F_D (occurring at angle a_1);

B. The ratio of maximum F_Y (occurring at a_2) to F_D at that same angle;

C. The ratio of F_Y to the maximum F_D (occurring at a_3);

D. The ratio of maximum F_Y (at a_2) to the maximum F_D (at a_3).

It is interesting that perhaps the most logical definition is the first: unfortunately, a_1 will generally be a vanishingly small angle where the performance of both Yagi and dipole becomes vanishingly low, and where minute ground resistance and height effects make an actual experimental comparison highly inaccurate. Moreover, we are not really interested in using the Yagi at this angle. The most relevant measurement is probably 2, where the Yagi is used at its "best" elevation angle, a_2.

To show how confusing these quantities become, **table** 3 shows "gain" at a very low angle — say 1 degree — approximating case A. Also included are the angle for maximum F_Y (case B), the angle for maximum F_D (case C), the "gain" as the ratio of maximum F_Y to maximum F_D (case D), and finally, in case E, the stacked-Yagi system at an angle which obviously produces a remarkable "gain" figure. This pathological behavior is due to the choice of an elevation angle at a fairly deep null in the reference dipole's pattern.

The tables show a perplexing array of gain figures. Which of them is correct? Actually, they are all correct; *all* of the differences are caused by the behavior of the reference antennas. The strange behavior of a reference dipole over ground, as shown in **table 2,** is due to its change in radiation resistance caused by mutual coupling with the ground image. This is a well-known effect." The elevation angle at which the reference antenna's gain is maximum is generally larger than that for a Yagi (higher-gain) antenna at the same height due to the way in which the "natural" gain of the Yagi falls off at higher angles.

These tables also illustrate that the gain of a real antenna over real earth *can not* be stated unless its height is specified. Moreover, the concept of "stacking gain," *i.e.,* increase in gain due to stacking, is determinable only in free space. Over earth, stacking does produce an increase in gain — but increase over what? It is probably best to avoid the temptation to use any "stacking-gain" figure, but simply to refer directly to the gain of the antenna system, which should now be understood to depend not only on the antenna components (Yagis), but also on their physical locations above ground! The tables also show

"Computation methodology will be described in a forthcoming article

"See, for example: John D. Kraus (W8JK), Antennas, McGraw-Hill Book Co., Inc., New York, 1950, page 305; *The* ARRL Antenna Book, 13th edition, American Radio Relay League, Inc., Newington, Connecticut, 1974, page 54.

that big Yagis generally perform well over earth, but not quite by the same ratio as free-space gain.

What then should one use as a reference standard? It seems at the outset that the preferred definition should allow natural comparison of two different antennas (say the lower vs the upper Yagi in the previous example). That is, we require that the ratio of the maximum gains of the upper to lower Yagis is simply the ratio of maximum energy flux densities of upper to lower Yagis. This is true in principle, if we use one of the free-space references, but not true for the half-wave, "substituted" reference dipole! In other words, if one used what has become a popular idea – gain "measurements" through a substituted half-wave dipole – the results are guaranteed to be confusing and guaranteed not to represent, even conceptually, a true measure of peak energy flux.

table 3. Yagi gain(s) over earth (dB) and elevation angle of maximum energy flux.

reference	Yagi at 0.6λ	Yagi at 1.6λ	stacked Yagis
isotropic (free space)	15.00 (21°)	16.12 (9°)	17.06 (10°)
infinitesimal dipole (free space)	13.04 (21°)	14.36 (9°)	15.30 (10°)
half-wave dipole (free space)	12.65 (21°)	13.97 (9°)	14.91 (10°)
half-wave dipole at Yagi height			
A angle = 1° (see text)	7.05	7.80	10.60
B see text	5.82 (21°)	7.59 (9°)	8.97 (10°)
C see text	5.27 (25°)	7.59 (9°)	7.75 (13°)
D see text	5.61	7.59	8.36
E angle = 27° (see text)			21.65

If the substituted half-wave dipole is not to be used in gain measurements, how then can one go about experimentally measuring the gain of an antenna that is interacting with earth or ground? It is easy to see that this is a formidable problem. The maximum intensity angle must be determined by a test signal generator or test detector; this must be done at the correct elevation angle and at a range well beyond the near field of the antenna and well beyond the Fresnel zones. The test detector must be calibrated in absolute terms of energy flux per unit area, or, if absolute calibration is not possible, one must accurately measure the entire radiated pattern over the hemisphere, with sufficient accuracy to determine gain with the required precision. That this is a difficult undertaking is obvious; there appears, however, to be no other way of experimentally determining gain."

In view of these considerations, I would like to suggest that, conceptually, antenna gain be defined simply as the ratio of maximum radiated energy flux intensity at the best azimuth and elevation to the average radiated energy flux intensity over *all* angles, *i.e.*, the full 4π solid angle or full sphere. I suggest that this definition, in fact, is quite common; † it is consistent with using a free-space, isotropic reference standard, and it gives the right ratios of energy fluxes for different antennas and antenna combinations whether in free space or over earth. It can, and I believe should, be used uniformly in all situations (including over the conducting earth). It can, in principle, be measured by integrating the complete spatial energy flux pattern.

Please note that, in contrast, "gain" measurements through dipole substitution, in principle, will give wrong ratios of energy fluxes for different antenna combinations.

This isotropic gain definition (referred to free-space isotropy) will give much higher figures for gain than those to which we have become accustomed. But we can get used to that; after all, we are already used to the outrageous claims for gain made by commercial antenna manufacturers!

To summerize, note the following points:

1. Antenna gain or directivity should be referenced to a free-space standard in order to be useful in making meaningful comparisons.

2. Experimental "gain" measurements over the earth by (dipole or other) reference substitutions will give confusing results and incorrect comparison ratios.

3. Experimental measurement of gain over earth is exceedingly difficult and basically impractical.

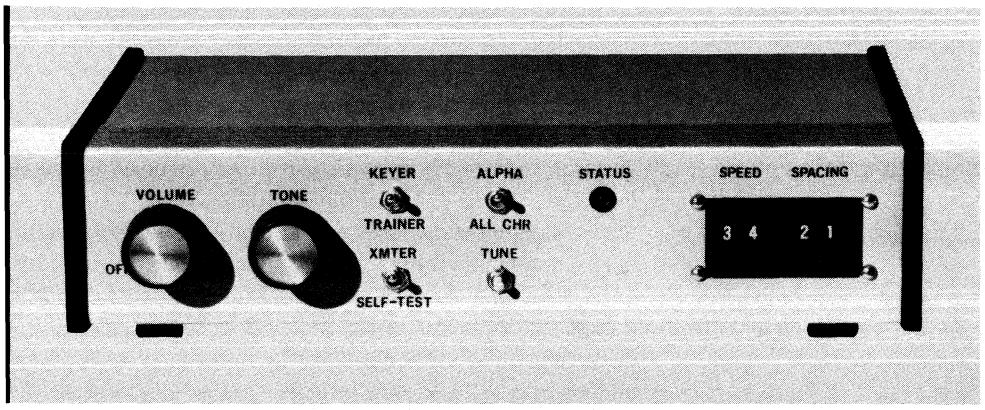
4. Directivity or gain can be calculated! The accuracy depends upon the proper mathematical characterization of the physical antenna (Yagi) and the use of sufficiently accurate computational methods. It is likely that the overall accuracy of gain calculation using modern methods significantly exceeds the practical accuracy of experimental measurement.

5. Large (Yagi) antennas perform well over earth, but not by the same ratio as in free space.

ham radio

"One can probably approximate the gain of an actual antenna by making measurements on a good model. While measurements are usually much easier to make on a (small) model, it is sometimes quite difficult to prove that the model is a faithful representation of the real thing and that all scaling laws are understood and properly applied.

†See, for example: John D. Kraus (W8JK), Antennas, McGraw-Hill Book Co., Inc., New York, 1950, page 17; "Reference Data for Radio Engineers," 4th edition, International Telephone and Telegraph Corp., page 703.



CW Trainer/Keyer

using a single-chip microcomputer

Theory and construction details for a keyer that serves as a training device and an iambic keyer with dot and dash memories

Have you ever wanted some code practice to help increase your speed to pass that elusive 13- or 20-wpm barrier, but found W1AW being clobbered by interference and you have all the code tapes memorized? This combination CW trainer/keyer could be just what you need.

As a trainer, the trainer/keyer sends random five-character code groups with selectable speed, spacing, and character set. There's no guessing at what speed the trainer/keyer is sending because the speed and spacing are digitally selected by front-panel thumbwheel switches. Any speed or spacing from 1 to 99 wpm can be selected in 1-wpm steps. With separate speed and spacing switches, the character speed can be, say, 22 wpm while the character spacing is only 9 wpm; any combination is possible as long as the speed is greater than, or equal to, the spacing. Thus, for practice to get past that 13-wpm barrier, set the character speed at 15 wpm and gradually increase the spacing until 15-wpm spacing is reached. This technique, pioneered by Russ Farnsworth (W9SUV) in his "Easy Method" records and tapes, is known from numerous code-learning studies to be the best for rapidly building up code speed. In addition to the selectable speed and spacing, the trainer's character repertory is selectable. The five-character code groups can be constructed from either the alphabet or from the alphabet plus the numbers and punctuation.

In the keyer mode, the trainerlkeyer performs as an iambic keyer with both dot and dash memories. As in the trainer mode, sending speed is digitally selectable in the same 1-wpm steps from 1 to 99 wpm. No more guessing at what speed you're sending!

microcomputer control

The trainer/keyer uses a new type of integrated circuit which will revolutionize Amateur Radio: a single-chip *microcomputer.* Notice that it's called a microcomputer rather than a microprocessor, which you've probably read about in equipment reviews or ads. There are several distinct differences.

By John Beaston, N6TY, 4415 Tilbury Drive, San Jose, California 95130

First, a microprocessor requires additional ICs to make a computer: RAM (random-access memory), EPROM/ROM (erasable-programmable or nonerasable read-only memory), and I/O (input/output). Ads for microprocessor systems (computers) usually show circuit boards crammed with these extra components. Microcomputers, on the other hand, have all the electronics from these additional components packed into a single piece of silicon – hence the name "single-chip" microcomputers.

Aside from the obvious difference of size, microcomputers are generally easier to use than microprocessors. A large part of microprocessor system design involves simply getting all the various components playing together. In a microcomputer, this task has already been accomplished by the microcomputer manufacturer. All you must do is tell the microcomputer what's to be done and supply the interface to the high frequency rig, RTTY gear, or whatever.

Now, microcomputers aren't going to make microprocessors obsolete. They are intended for slightly different applications. While microprocessors are perfect for applications such as small-business computers and high-speed communications, microcomputers are designed to bring the power of a computer to control-type applications. Microcomputer applications around the home might include a microwave-oven controller, an energy-management center, a repertory phone dialer, or a burglar/fire alarm system. In Amateur Radio, microcomputers are already appearing in applications such as scanning and remote-controlled high frequency and vhf rigs. Other applications that come to mind are self-tuning transmatches; automated tracking Oscar antennas; digital-station consoles; sophisticated accessories for RTTY, SSTV, and CW; and, of course, the CW trainer/keyer. Only imagination limits the applications.

Before going into the details of the trainer/keyer let's discuss microcomputers in general, since they are so new.

microcomputer basics

All microcomputers are similar internally. **Fig. 1** shows a structure that applies almost universally. For the sake of illustration, let's assume that this microcomputer is being used as a traffic-light controller.

Central processing unit. Looking at the function of the first block in the block diagram, the CPU (central processing unit) can be thought of as the master control sequencer – the brains of the microcomputer. It performs the actual counting of cars and changing of lights at the appropriate times. These actions result from following a list of instructions (the program) stored in the program memory. A typical program might be this: Leave the light green for street A until no cars have passed for 15 seconds,

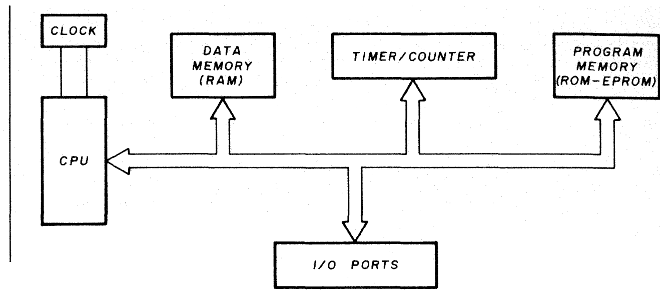


fig. 1. Typical microcomputer structure showing major functional blocks.

then give street B the green until either five cars are waiting on street A or until 30 seconds have elapsed, whichever occurs first.

I/O section. If our microcomputer is to execute this particular program it needs some way of detecting the presence of cars traveling on street A. Sensors imbedded in the roadway sense the presence of a vehicle. To feed the sensor information into the microcomputer, simply connect the sensor to one of the input ports in the I/O section (assuming voltage compatibility, of course). The CPU then reads information on an input port whenever told to do so by an instruction from the program memory.

Memory sections. Since the CPU must control each light in a traffic situation, it must remember which light is on in addition to other variables. Other such items might be the length of time a light has been lit, how long it needs to remain lit, and how many cars are waiting at street A while street B has the green. Each of these little pieces of information is stored in the microcomputer's data memory.

Let's compare the program memory and data memory, since they sound similar but have distinctly different functions, The program memory is simply a list of instructions in a format the CPU understands (machine language). The CPU reads and executes the instructions in sequence, one by one. For any given traffic intersection the list never changes. The CPU executes the same list of instructions over and over.

The data memory, on the other hand, holds data such as time or the number of cars waiting. Each location in the memory stores a particular piece of information. For example, assume location 3 in the data memory stores the number of cars waiting at street A, while street B is green. As soon as the light at B turns green, the CPU makes the contents of location 3 zero. Periodically the program memory instructs the CPU to check the car sensor, through the input port, to determine if another car has tripped

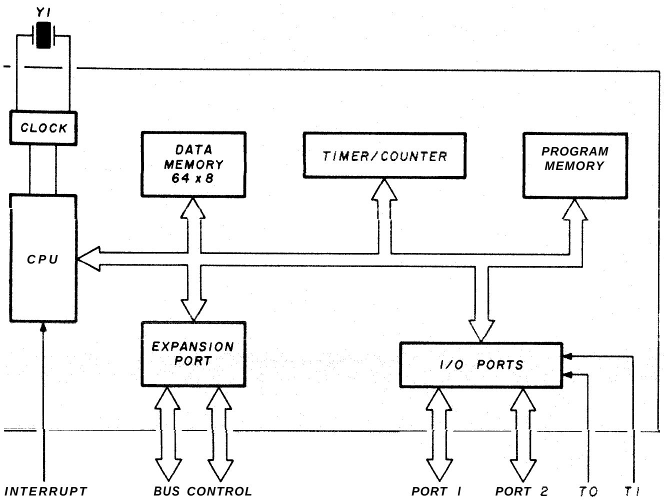


fig. 2. Block diagram of the Intel 8748 showing key elements

the sensor. If such is the case the program instructs the CPU to increment the number in memory location 3, then to test it to determine if the number is equal to 5. If so, the CPU should change the light at B to red and the light at A to green. If the number is not yet 5, there's no need to change any light, so the program just continues executing without changing anything. The data memory, unlike the program memory, never tells the CPU what operations to perform; the data memory just stores information.

When the CPU determines it's time to change a light, it does so through an output port in the I/O section. Normally, each light is connected to a separate line on the output port. At the appropriate time, the CPU turns a light on or off by switching the port pin associated with a particular light. Since an output port's output level is TTL compatible (either 0 or 5 volts), an external switch between the output port and the light is needed for the actual switching.

Clock. The last block in the diagram is the clock generator. This block determines at what times the CPU reads and executes new instructions from the program memory. All operations within the microcomputer are synchronized to this master clock. The frequency of this master clock is set by an external crystal. Typical frequencies are in the range of 1-12 MHz.

Program memories. There are two different types of program memories, either ROM or EPROM. As the name implies, the ROM-based program memory can-

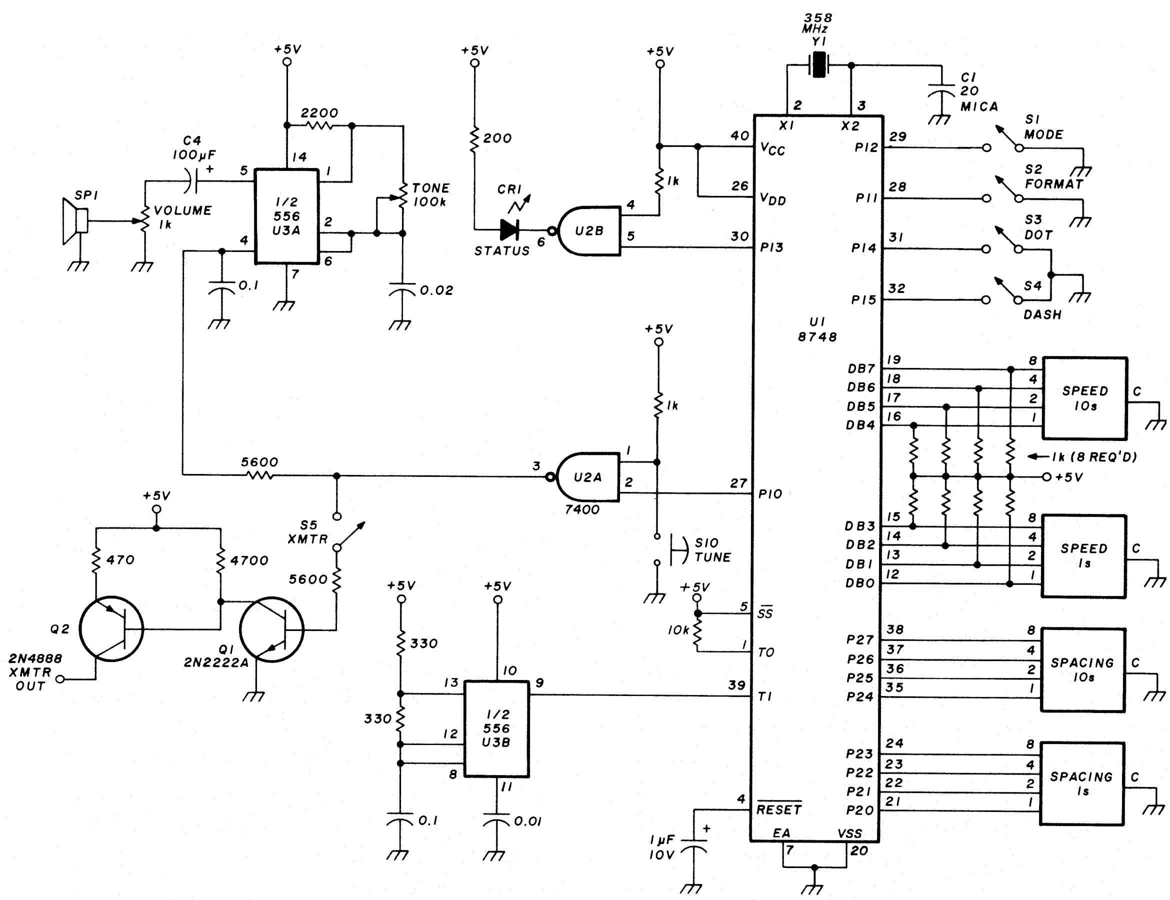


fig. 3. Schematic diagram of the CW trainer/keyer. All resistors are ¼ watt, 10 per cent tolerance. S3 and S4 and the contacts of an iambic paddle.

not be changed. The actual user program is placed into the memory by the microcomputer manufacturer at the time the device is fabricated. Since this involves special tooling, the user is generally required to pay a fee — plus submit an order for a large number of devices. If only one microcomputer is needed for a special project, a ROM-based microcomputer isn't the way to go.

Some microcomputers use EPROM for their program memory. EPROM technology allows the user to erase and reprogram the microcomputer at any time. These devices were originaily developed to help the user debug his program before committing it to ROM. However, EPROM-based microcomputers are perfect for one-of-a-kind projects such as we hams usually undertake. They can be changed at any time with no worry about big orders or tooling costs. So if there's a need for a particular gizmo, just use the EPROM version. It's like having your own custom IC!

Intel 8748

The EPROM microcomputer used for the trainer1 keyer is the Intel 8748. Its block diagram is shown in **fig. 2.** Notice the similarities to **fig. 1**. The 8748 has 1024 bytes of program memory and 64 bytes of data memory. (Each byte holds one instruction, or piece of data.) This may sound awfully small if you're familiar with microprocessors; however, remember that microcomputers are used mostly for control applications. Very few of these applications require more than 1000 bytes of program. But if your particular application requires more, the 8748 is easily expanded to 4000 bytes of program memory and 320 bytes of data memory using external components.

The 8748 also contains a total of twenty-six I/O lines. There are two 8-bit I/O ports (PORT1 and PORT2) that can be mixed as any combination of input or output lines. Another 8-bit port (BUS) either expands memory or is a simple input or output port. The remaining two lines are the test inputs, TO and T1. The CPU can test these inputs under program control. They also have special functions depending on the mode of the internal timer/counter.

The programmable timer on the 8748 is an 8-bit up

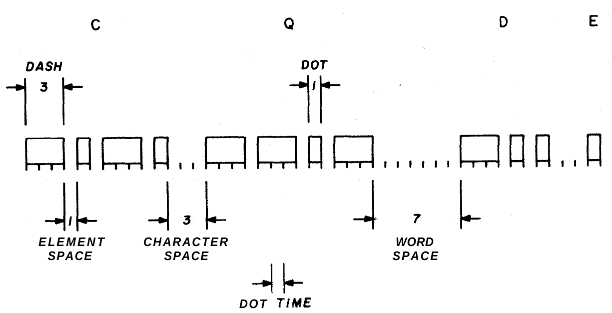


fig. 4. Morse code timing definition example for the Morse characters CQ DE.

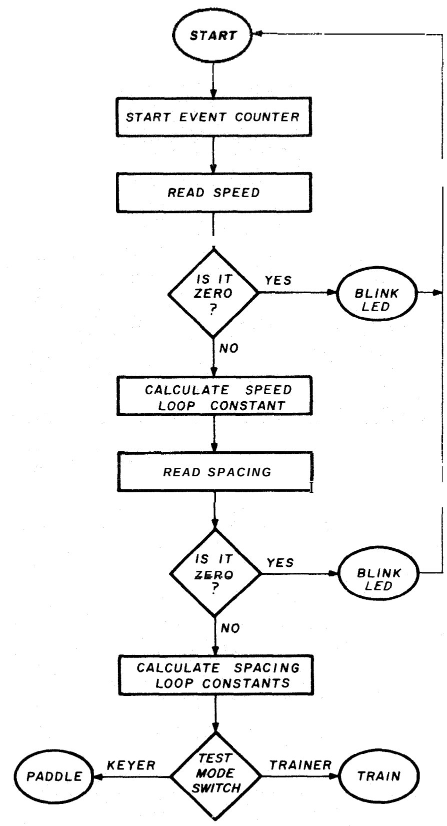


fig. 5. Flow chart for the START routine in the CW trainer/keyer software.

counter. This timer either measures time intervals or counts external events. The specific mode is selected by the program. In addition, the CPU can read, load, start, or stop the timer.

Rather than go into greater detail on the 8748, I'd suggest that any interested reader obtain The ***MCS-48 User's Manual*** from Intel's literature department. This manual provides all the hardware and software details for the 8748 as well as for several other single-chip microcomputers.

trainer/keyer circuitry

Fig. 3 shows the schematic of the 8748-based trainerlkeyer. The circuitry is straightforward. It requires only three ICs, a crystal, an LED, and a handful of switches, resistors, and capacitors. The thumbwheel switches determining the character speed (SPEED) and spacing (SPACING) connect to the BUS (DB7 = DB0) and PORT2 (P27 = P20) lines respectively. The trainer/keyer assumes BCD (binary-coded-decimal) coding for the speed and spacing input. The use of BCD thumbwheel switches is the easiest way to get the inputs into this format. Simple spst toggle or DIP switches could also be used; however, the BCD coding must then be done manually. Notice that no pull-up resistors are required on PORT2, although they are needed on the BUS inputs. PORT2 has the pull ups internally.

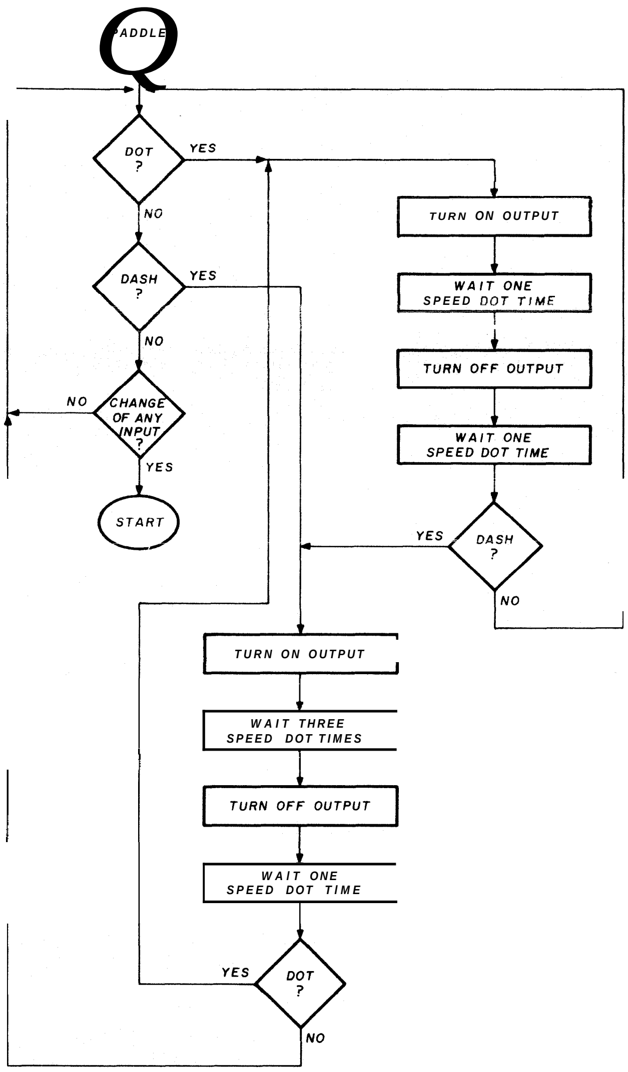


fig. 6. The PADDLE routine implements the iambic keyer section of the trainer/keyer.

PORT1 (P17 = P10) is the remaining 8-bit I/O port. Unlike the BUS and PORT2 ports which are only inputs in this application, both inputs and outputs are mixed on the various lines of PORTI. The inputs on PORT1 are the MODE, FORMAT, and DOT/DASH switches. MODE selects between the trainer and keyer modes. When MODE selects the trainer, the FORMAT switch determines the trainer's character set. With the FORMAT switch open, the trainer sends code groups containing only the alphabet. When FORMAT is closed, the alphabet along with numbers and punctuation form the code groups. If the keyer mode is selected through the MODE switch, a paddle supplies the inputs through the DOT and DASH lines. These inputs are independent, so either iambic or noniambic paddles may be used. The FORMAT and SPACING inputs have no effect when the keyer mode is used.

There are two outputs on PORTI. P10 is the actual CW output and P13 is a status indication output. The two-input NAND gate, U2A, buffers the CW output. The other input is a TUNE switch, which clamps the NAND output high when depressed. This NAND output keys the sidetone generator, U3A, and/or the transmitter. Switch S5 disconnects the transmitter while in the trainer mode if desired. The sidetone generator uses one-half of the 556 as a bistable multivibrator triggered by the CW output. The transmitter keying circuit shown is for grid-block-keyed transmitters. For other keying techniques, a 5-volt relay could replace this circuit.

The status indicator output, P13, notifies the user if an invalid combination of speed or spacing is selected. In normal operation the STATUS LED is lit continuously as a power-on indicator. If 0 wpm is selected, or if the spacing is greater than the speed, the LED flashes on and off until the condition is corrected.

The trainer uses the internal timer/counter in the event-counter mode to generate random numbers. These random numbers are eventually used to select the CW characters within the code groups. U3B supplies the events to be counted. As with U3A, this half of the 556 is wired as a bistable multivibrator. This oscillator output is tied to the 8748 T1 input. T1 is a dedicated input to the event counter when using the event-counter mode. Since U3B is free running, the event counter simply increments through its entire 8-bit range. As we'll see shortly, the program periodically reads the contents of the counter. This number then selects the next CW character to be transmitted. Random character generation is guaranteed because there's no synchronization between the program and the free-running oscillator.

The last piece of hardware is the crystal. The crystal supplies the basic time interval in which the CPU steps through the program in the program memory. Keeping cost in mind, a standard TV color-burst crystal, 3.58 MHz, was chosen.

the code

Before describing the software, let's review just how a Morse code character is developed. Every character is made up of elements: dots and dashes. One basic time unit relates all the elements and the spacing between them. Dots are one dot-time long, while dashes are three dot times in duration. Within a character, elements are separated by an inter-element space equivalent to one dot time. The space between characters within a word is defined as three dot times. Spaces between words are seven dot times. The basic time interval is a dot time. **Fig. 4** illustrates this timing for "CQ DE".

For any given speed, the trainerlkeyer uses the time for one dot element as the basic time unit. To convert from words per minute to this basic time unit, we use the equation found in the ARRL *Handbook:*

$$wpm = 2.4\ (dots/second)$$

Since in this formula a dot is made up of both the dot itself and the space between it and the next dot, the equation for the basic time unit becomes:

$$dot\ time\ (second/dot\ element) = 1.2/wpm$$

We'll call this basic time unit a "dot time," with the understanding that it is the time equivalent to a dot element itself, not the dot element and the following space. For example, at 20 wpm, the dot time is 1.2/20, or 60 ms.

The trainerlkeyer has a short delay program that takes 0.1 ms to execute. This program does nothing except wait for 0.1 ms. To get the time interval needed, say, for 20 wpm, we simply make this program execute 60 ms/0.1 ms, or 600 times. At one wpm, we need 1.2s/0.1 ms, or 12,000 times; while at 100 wpm the count is only 120. The trainerlkeyer uses this software technique to generate the time for each code element.

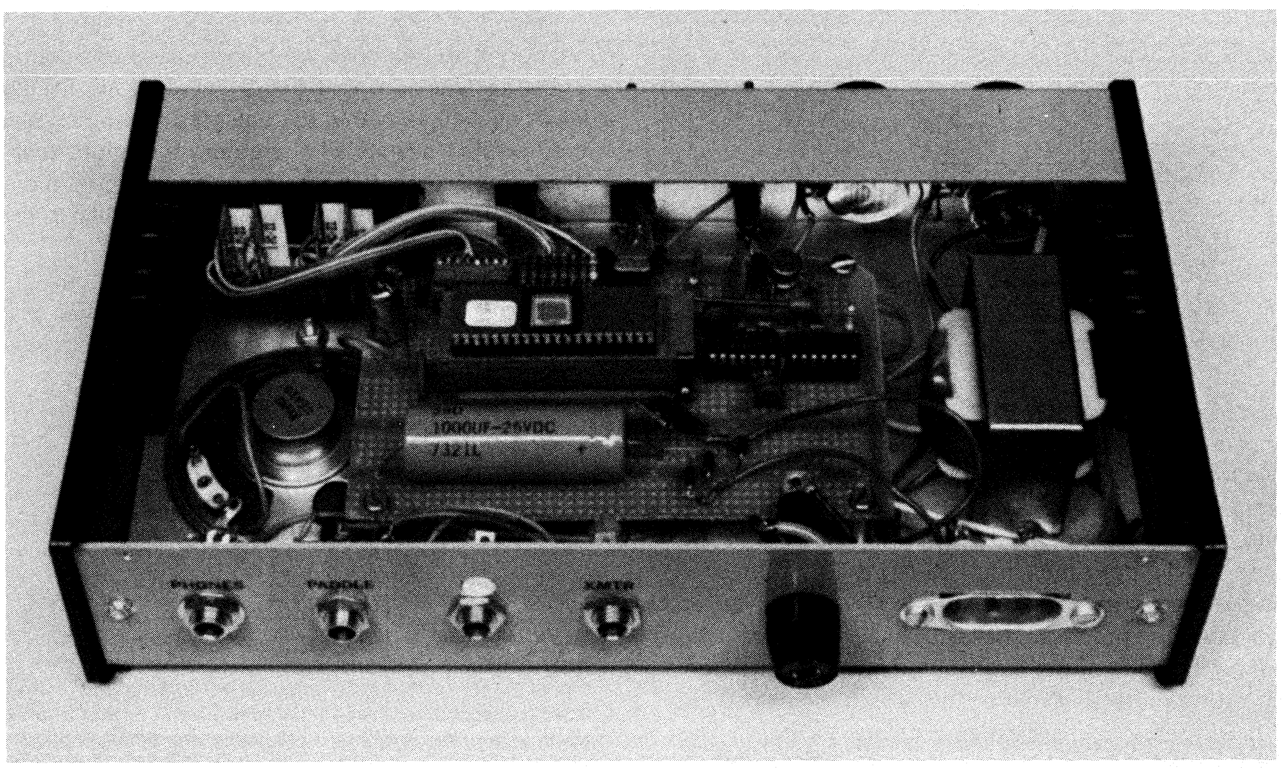
The program is the heart of the trainerlkeyer. It has three basic sections: start, keyer, and trainer. The flow charts for each of these sections are shown in **figs. 5** through **7.**

Program start. When power is turned on, the 8748 begins executing the program corresponding to the START flow chart, **fig. 5.** This routine first starts the event counter, then reads the speed selected through the SPEED thumbwheel switches. If the selected speed is 0 wpm, the program branches to a routine that flashes the STATUS LED once and returns to START. If the selected speed is something other than 0 wpm, the routine converts the BCD number into binary and uses this binary number to calculate how many 0.1-ms steps are needed to give the desired speed. This number of steps is called the SPEED loop constant. Once the SPEED loop constant is found, the same procedure is used for reading the SPACING thumbwheei switches and computing the SPACING loop constant.

Once both loop constants are known, the MODE switch is tested. If the trainer is selected, the program branches to the routine called TRAIN, **fig. 7.** If the keyer is selected, the PADDLE routine is executed, **fig. 6.**

Keyer routine. Looking at the latter first, PADDLE simply tests the DOT and DASH inputs. If the dot side of the paddle is pressed (the P14 input is 0), the routine sends a dot element by turning on the CW output, waiting in the delay loop for one SPEED dot time, turning off the CW output, and waiting out the

Internal view of the CW trainerlkeyer. As can be seen, this version was constructed on perf board rather than using a printed circuit board. Interconnections between the digit switches and the circuit board are made by using component carriers, which also hold the pull-up resistors *(photocourtesy WB6SFC).*



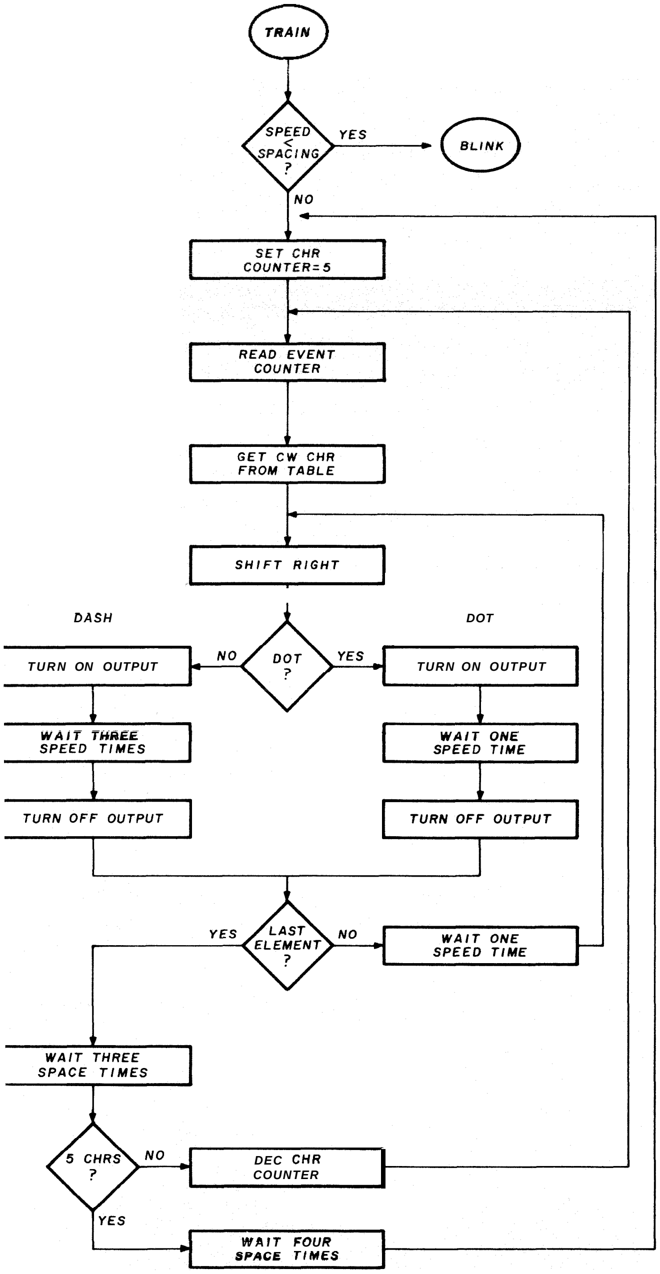


fig. 7. The trainer section of the trainer/keyer uses the TRAIN routine.

inter-element space for one more dot time. Then, to give the iambic nature to the keyer, the DASH input is examined to determine if the dash paddle is pressed. If so, a dash is sent by turning on the CW output for three SPEED dot times and waiting the one-dot-time inter-element space. Then the DOT input is tested again and the process repeats.

If neither side of the paddle happens to be pressed, the routine checks if either the SPEED, SPACING, or MODE inputs have changed. If one or more has changed the routine branches back to START to recalculate the loop constants, change modes, or both. If not, the DOT and DASH are tested again.

The one area not shown in the flow chart is how the dot and dash memories are incorporated. During the delay time of sending an element or an inter-element space, the input for the other side of the paddle is tested occasionally. If it's pressed, a software flag (special bit) is set to indicate that memory is needed. When the current element is complete, this flag is tested. If it's set, the opposite element is sent before continuing. No action is taken if it's not set.

Trainer routine. The trainer routine is similar to that of the keyer except that the inputs come from internally selected characters. These characters are located in a special section of the program memory called the character table. The total number of characters in the table is 128. The frequency of occurrence of each character in the table roughly reflects its occurence in everyday text. In other words, the table contains seven A characters, six Es, six Ns, *etc.,* while containing only two Js, two Zs, and so on. Each number and punctuation character has two entries.

When the trainer needs a new character, it reads the internal event counter, which is driven from the free-running oscillator. This number is used as an offset into the character table to select the next character. If the event counter happened to be 27 when it was read, the twenty-seventh character in the table (in this case it is a G) is the next character transmitted.

Since the character set is variable, some testing is needed of the number read from the event counter. These tests ensure that the offset is within the table limits as well as within the selected character set. If the FORMAT selects all characters the entire table is used. If only alphabet characters are selected the offset range is restricted to only that particular portion of the table.

Characters in the table are stored in a binary form equivalent to Morse code. This form specifies that a dot is represented as a binary 1 while a dash is a 0.

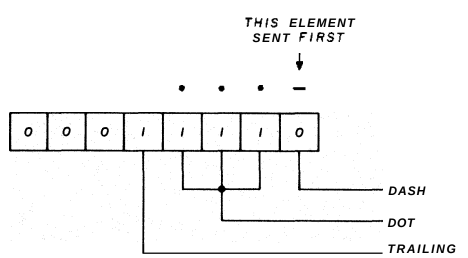


fig. 8. Binary format in the character table for the character B. All characters are read right to left and contain one extra trailing one. All remaining positions out of eight total are filled with zero.

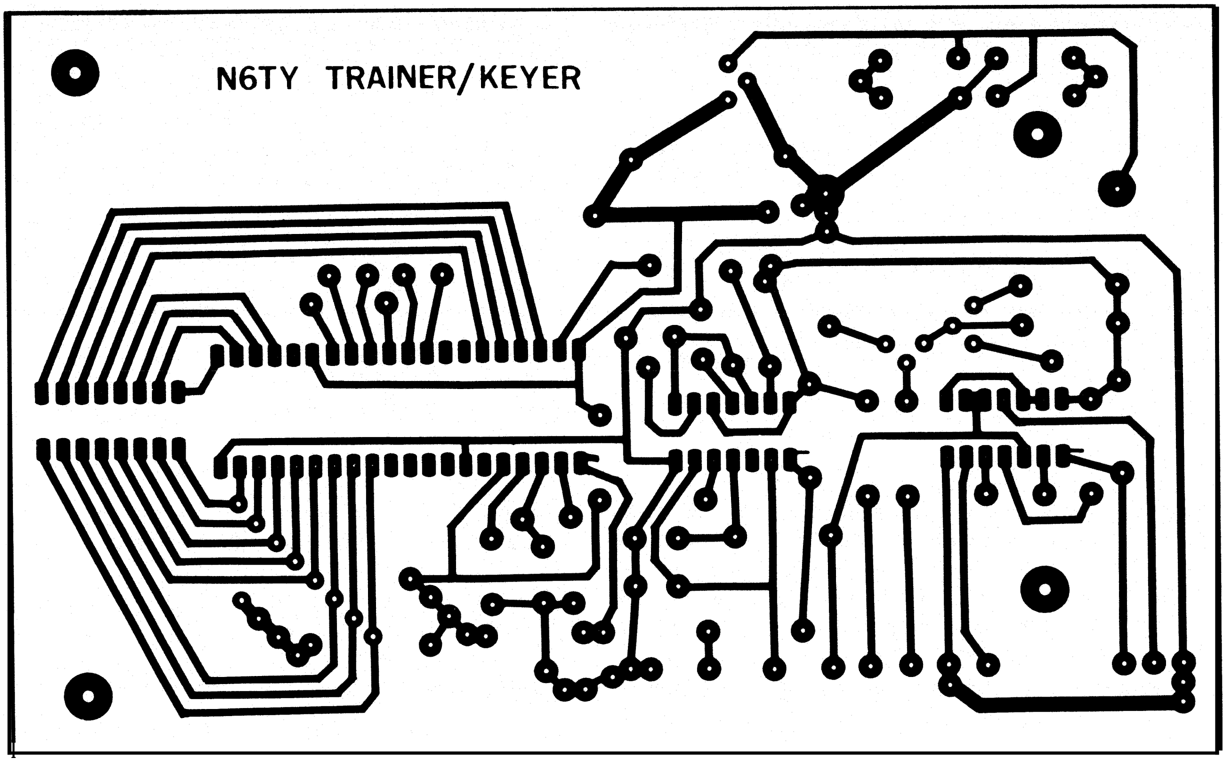
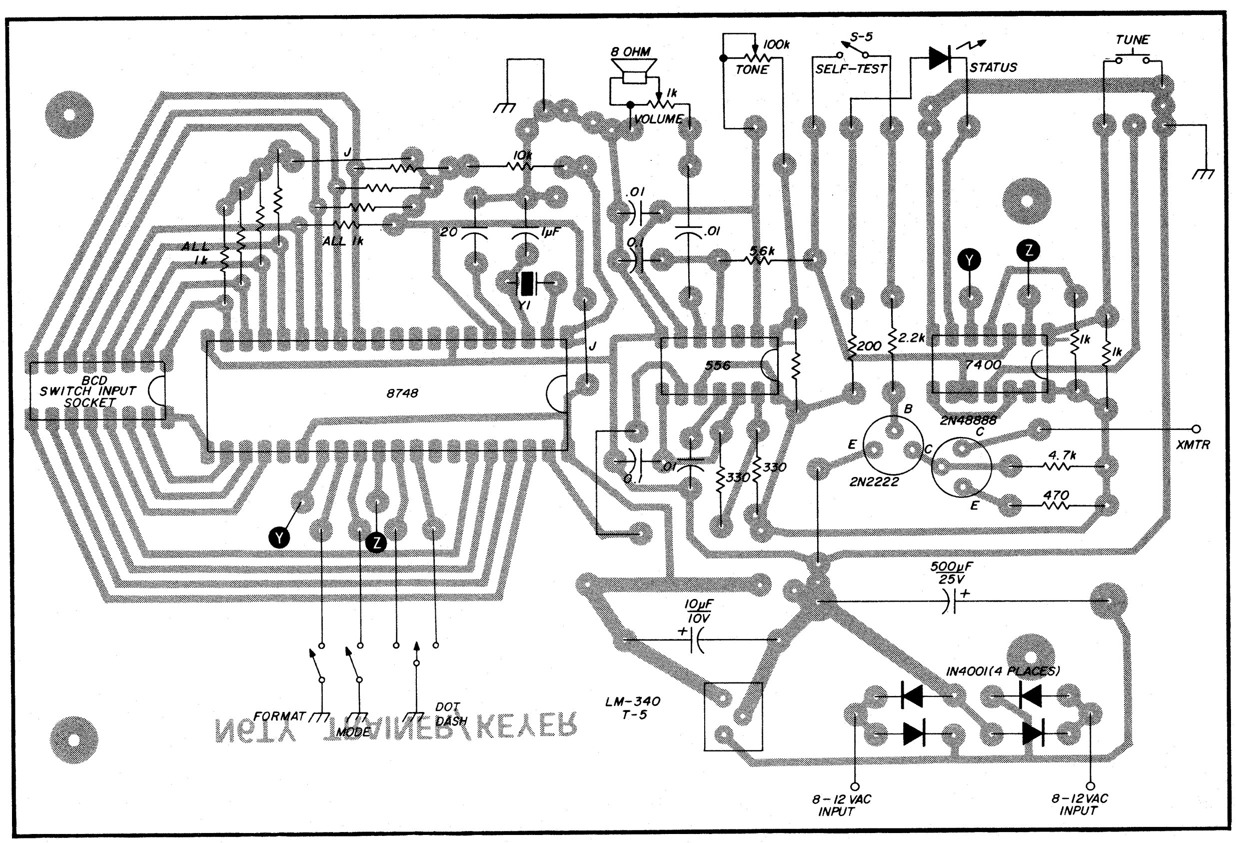


fig. 9. PC-board layout for the CW trainer/keyer above and component layout diagram below.



Each of the 128 entries requires one byte (eight bits). The characters are right-justified and contain a trailing 1. Fig. **8** illustrates the format for the letter B. To reproduce the Morse code equivalent, the trainer simply shifts the whole character one position to the right and tests the bit that falls off the right-hand end. If the bit is a 1, it sends a dot; if it's a 0 a dash is sent. This process repeats, bit-bv-bit, until all the positions contain zeroes except the right-most. When this occurs, the character is complete, and a new character is fetched from the table.

Now that we understand the basic operation of the trainer, let's look at its flow chart again (fig. *7).* Since the trainer uses both speed and spacing, the first operation checks for the validity of the selected combination. Speed greater than the spacing is invalid, and the STATUS LED blinks accordingly. If the combination is okay, the character counter is set to 5. This software counter keeps track of the number of characters remaining to be sent in a five-character code group. Next, the event counter is read, and a character is selected from the character table.

Character shifting is begun at this point. Let's assume the first element is a dash. The CW output is turned on and the delay routine waits three dot times as specified by the SPEED loop constant. The CW output is then turned off. Since the length of time until the next element depends on whether this element is the last in a character or is simply an intermediate, intercharacter element, the routine checks for the character-done condition of all bits zero except the rightmost. If not the last element, an inter-element space is needed, and the delay routine waits one SPEED dot time before the character is shifted for the next element. If the dash happened to be the last element, the delay routine then waits the intercharacter space of three dot units, using the dot time specified by the SPACING loop constant. Since a character has just been completed, the character counter is tested to see if the character was the fifth character in a character group. If not, the character counter is decremented by one, and the routine reads the event counter and gets the next character.

If the just-completed character was the last of a group, an additional four SPACING dot times are awaited, to give a total of seven SPACING dot times between groups, which corresponds to a word space. The routine then returns to reset the character counter to 5 and begin the next group.

construction

Construction of the trainerlkeyer is straightforward. PC and perforated board techniques work equally well. The printed circuit layout is shown in fig. **9** for those who would rather "roll their own." The only constraint in the construction is that the crystal and its companion 20-pF capacitors be located as close as possible to the 8748 IC. Be sure to bypass all inputs and outputs from the enclosure to minimize rf entering the enclosure. As for the power supply, the trainerlkeyer requires only a 5-volt supply. Any 5-volt supply capable of supplying 150-200 mA is sufficient. The PC board layout contains provisions for a diode bridge rectifier and three-terminal voltage regulator.

parts

Many of the larger mail-order IC houses stock 8748s. Otherwise, they can be purchased at any Intel distributor listed in the *MCS-48* User's *Manual.* Demand for the 8748 is quite high, so availability may be limited; be sure to call around. Several versions of the 8748 are available. These different versions are denoted by another digit after the 8748; *e.g.,* 8748-4. Any version will work in the trainerlkeyer. The 8748-8 is the lowest-cost version, so ask for it if there's a choice. The only difference between different dash-numbered parts is the maximum crystal frequency. The maximum for the -8 is 3.6 MHz, while a "no dash" device works up to 6 MHz.

One caution. These devices are unprogrammed and must be programmed with the trainerlkeyer software for use in this application. Intel distributors usually have a programming service available for a small charge, or an 8748 programmer can be constructed based on the timing shown in the user's manual. Programming requires a knowledge of the machine language program. Listings of this program are available from the author.

afterthoughts

The 8748 is an extremely flexible device. The number of applications for it within Amateur Radio are almost limitless. Other applications that have been built are a single-chip Morse code encoderldecoder and a WWVB digital clock that never needs to be set. As an example, the encoderldecoder allows you to receive and transmit CW directly on any Baudot or ASCII teleprinter, plus any terminal unit having RS-232 switching levels (ST-5 or ST-6). For transmit, the encoder accepts either Baudot or ASCII serial characters from the keyboard. Up to thirty-two characters are buffered and transmitted in CW at any speed from 1 to 99 wpm. On receive, the decoder adapts to any CW input speed all the way from 1 to over 100 wpm. Received characters are formatted in either Baudot or ASCII and sent serially to the printer. Why not pick up an *MCS-48* ***User's*** *Manual* and get into computer-controlled ham radio the easy way?

ham radio

dip meter converter

for very low frequencies

Using an ordinary dip meter at frequencies below 100 kHz can be a problem — this converter allows your meter to work accurately to 1 kHz

You're just starting to build some equipment for vlf. There's a tuned circuit that you *think* you've cut to about 100 kHz, but something seems to be wrong. You reach for your trusty dip meter and you find . . . .

If your dip meter resembles my ancient Millen, you find that it cuts off at 1.7 MHz. It doesn't come anywhere near vlf.

When faced with this disagreeable situation, I first devoted some thought to winding a few appropriate coils for the Millen. Then I toyed with the idea of starting from scratch with a VCO chip. Finally I tried a simple and rather obvious approach that worked beautifully.

By using a ripple counter **(fig. 1)** you can divide what you need by what you don't have. Starting with a dip oscillator that's well calibrated, you can go to a frequency as low as you like. No improvising of coil forms, no hunting around for calibration points, and no worries about the mechanical stability of a new tuning mechanism.

The 4040 counter chip divides by 2^1 through 2^{12}. This fits the approximately 2:1 tuning range of the typical dip oscillator and leaves no gaps in the spectrum. Some conditioning of the input is required, but the 4001 is inexpensive and readily available, and it can be used for this purpose.

All that's necessary to measure a resonant frequency is to feed an rf probe from the divider output. Connect the probe directly to the tuned circuit. The rf voltmeter measures the voltage drop across the circuit. When off resonance, the probe is essentially short circuited. At resonance, the test circuit looks like a large resistance, and the voltage rises accordingly. Since the probe is directly connected to the test circuit, it's possible to measure not only the resonant frequency but the resonant impedance and Q as well. Seems like a nice bonus to get all this with such a simple apparatus.

the circuit

The first section of the 4001 is biased into its active region for use as an amplifier. This unloads the dip oscillator and helps keep its original calibration intact. There are different ways of handling the bias problem, but the one shown in **fig. 1** proved to be thoroughly stable, and both the A- and B-series chips worked well. R1 provides a certain measure of input protection.

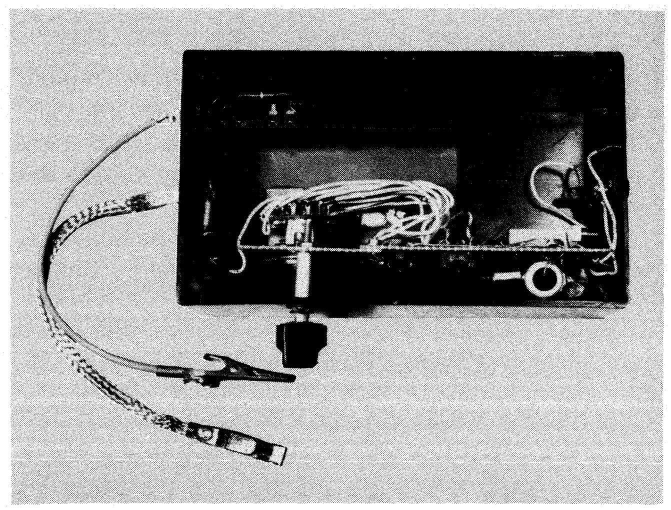
A characteristic of CMOS is that steady-state offsets propagate poorly through a series of gates. Tying together both inputs to the second gate assists the process under certain circumstances. In this case, I found that the third gate could be switched by manipulating the bias on the first gate. The steady-state output, however, could not be set anywhere except at a supply rail. The fourth gate wasn't needed, so it was tied off. Supply voltages can vary from about 7 to 15 volts. For a variety of reasons, 12 volts seemed to be a good compromise.

From the divider chip, a 12-position rotary switch selects the desired subharmonic and routes it to the four output terminals. Terminal A is for using the gadget as a utility squarewave generator. Terminal B is normally connected to terminal C either directly or through R4. This resistor is located outside the case, because it must be changed under certain circum-

By E. G. Von Wald, W4YOT, 932 North Federal Highway, Lake Worth, Florida 33460

stances, as discussed later. Terminals C and D form the probe. C2 and CRI, with filter R5C3, form the business end of the rf voltmeter. A sensitive, high-resistance indicator – such as an electronic voltmeter – should be used to read the voltage.

As mentioned, the 4040 divides by a factor as high as 2^{12}. When you apply this divisor to the lowest frequency on the Millen, you get an output of a little under one-half kHz. This seemed pretty far below that required for my purposes, so the 12th output on the chip was left unconnected. The open position on the switch allows the instrument to be used as an rf probe with external excitation. It has a resistance (at these frequencies) in excess of a megohm, shunted by several picofarads.



Inside the vlf dip meter converter. The case is plastic with an aluminum top panel lnot shown). Plastic is inexpensive and easy to work but mars easily. Static electricity is no problem.

construction

Construction is simple and, allowing for the difficulties of working with quasi microcircuitry, reasonably fast. Since CMOS lends itself to point-to-point wiring, the additional problems of working up an etched circuit board are avoided. Fiberglass perfboard with 2.6-mm (0.1-inch) hole spacing was used with no. 24 AWG (0.5-mm) solid wire. Stripped of its insulation, this wire laces nicely through two or three adjacent holes. When pulled up tight, it holds its position well, forming a stable anchor to the perfboard. Trim the wire about 5-mm (0.2 inch) above the adjacent socket terminal. Then use a soldering tool to bend the trimmed stub over against the terminal. Twist the wire so that it stays against the terminal by itself. Soldering becomes very easy this way.

The photograph of the instrument interior shows the mechanical layout. Metal shielding should be used to keep stray fields from the 4040 out of the rectifier circuit. Shielding was made from scrap flashing copper. The shield also serves as a mounting bracket. A spacer and bolt were inserted close to the switch to provide extra rigidity when switching.

A word about using resistors at radio frequencies. The resistance values tend to fall off as frequency is raised. Henney[1] gives some data for %-watt film- and slug-type resistors. **Fig. 2** summarizes this information for the frequency-resistance range likely to be encountered here.

adjustment

When the circuit has been wired and checked for accuracy, insert the chips and power up. Use a VOM set to about 10 mA in series with one of the supply leads. Set the selector switch about mid range.

Screw R2 all the way to one limit, then slowly bring it back past its midpoint. Supply current should

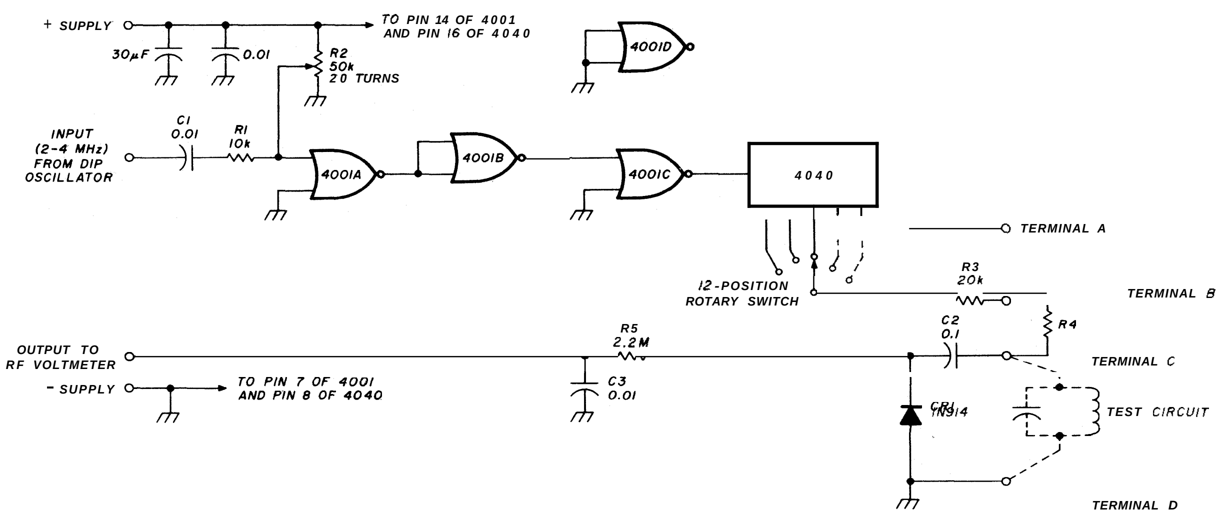


fig. 1. Vlf converter schematic. R4 will vary, as described in text.

be negligible until you get near the active region, when it will jump rather suddenly to several milliamperes. Even with a 20-turn pot, this adjustment range is narrow — % turn should carry you all the way through.

Leaving the pot set at the middle of the active region, short circuit terminal B to terminal C. Couple

than at resonance) at regular odd submultiples of the resonant frequency. This is why it's easiest to start hunting for resonance from the high end. It also provides a graphic illustration of just what the corners of square waves are made of.

You can use inductive coupling for resonance checking (but not for Q or impedance). Simply con-

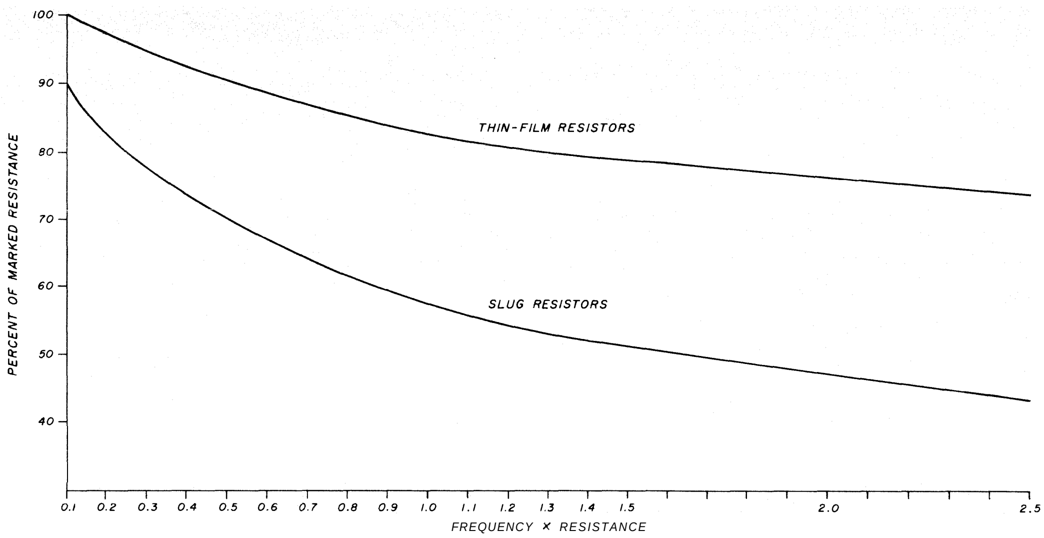


fig. 2. The product of frequency times resistance is shown on the horizontal axis. Frequency should be in MHz; resistance in megohms. The thin film resistors are clearly superior at radio frequencies.

the dip oscillator to the input through a 3-turn link at the end of a short length of RG-58/U cable.

Set the dip oscillator to 4 MHz. When the oscillator is switched on, there should be another jump in current to perhaps 7 or 8 mA with a 12-volt supply. No further adjustment is necessary unless the supply voltage is changed by an appreciable amount. Should the output voltage vary appreciably across any single output range, either R2 is not set right or else more coupling to the oscillator is needed. Keep in mind, though, that there are about 4 or 5 pF of shunt capacitance at the rectifier and it's not always negligible (compared, in this case, with the resistance of R3). At 2-MHz output, for instance, the open-circuit voltage will be quite low. However, when a tuned circuit is connected, this shunt capacitance is simply absorbed into the tuning capacitance. At these frequencies, the error should not be serious.

Once the device is functioning properly, connect the probe across a marked i-f transformer. Output voltage should fall toward zero. To find the resonant frequency, start from above the expected frequency and tune down. When you get close to the target, the output should peak up nicely. After locating the proper resonance point, keep tuning down the spectrum. You'll find peaks (considerably lower in voltage

nect a multiturn link directly across the probe terminals. Couple this link to the circuit in question. Output will still be a peak at resonance. Expect to use a lot of turns at these low frequencies.

operation

You can use the 20 kilohms of R3 by itself for routine frequency checking, with a short in the R4 position. Measuring circuit Q and resonant impedance is slightly more complicated. You have to make one or two voltage readings and do a little arithmetic.

The divider chip itself can be viewed as a generator with a low internal resistance. Shunting such a low resistance across a parallel-tuned circuit would greatly lower the apparent Q. The simplest way of avoiding this is to use a couple of megohms for R4 and make your measurements with a millivoltmeter.

If yours is a standard analog electronic voltmeter with a lowest range of 1½ volts, you can't do this. The rectifier response would be somewhere between linear and square law, but you wouldn't know where. The voltage readings would be completely unreliable. It's better in this case to keep the voltage to the rectifier high, assume that it's more or less linear, and make a correction for the shunting resistance. Here's how to do it.

Short the square-wave output terminal directly to the probe terminal. With no loads connected, switch the output to its lowest frequency range. Observe the voltage reading (use the dc scale). This should be around 4% volts with a 12-volt supply. Call this voltage E.

Remove the short. Connect a suitable resistor in the R4 position. 50 kilohms would be a good value to start with. Now connect the probe to the tuned circuit and measure the voltage reading at resonance. Call this V_0. For best accuracy, V_0 should be somewhere around half to three-fourths of E. You may have to hunt a bit to get a resistor that brings V_0 within these limits. The amount of resistance required varies according to the resonant impedance of the circuit. If it's very low, you may have to use the minimum (R3 alone).

Once you get V_0 within the proper limits, measure the Q in the usual manner by detuning the oscillator until you get $0.707\ V_0$ either side of resonance.

$$Q_a = \frac{F_{resonant}}{F_{high} - F_{low}} \quad (1)$$

Q_a is the apparent Q not the true Q. To find true Q:

$$Q = Q_a\left[\frac{1}{1 - \dfrac{V_0}{1.27E}}\right] \quad (2)$$

The factor 1.27 comes about because the E you measure is a square wave and includes harmonics. Some of them tend to cancel the peak of the fundamental waveform. When the tuned circuit is connected, it bypasses these harmonics and they no longer affect things. (To be precise, 1.27 happens to be the coefficient of the first ac term in the Fourier series expansion of the squarewave after being normalized to the dc term.)

To find the tuned-circuit resonant impedance, you can use the formula

$$Z_0 = (R3 + R4)\left[\frac{1}{\dfrac{1.27E}{V_0} - 1}\right] \quad (3)$$

These results should be accurate to within about 10 per cent or so, assuming you're using 5 per cent resistors properly derated. Frequency readings will be a few per cent on the low side because of the parasitic shunt capacitances. Decouplng the test circuit by feeding it through a large resistor will greatly reduce this error, but the peaks are harder to locate.

By the way, if you're accustomed to using a dip meter, you'll notice something rather odd when measuring with this device. The precise peak at resonance seems somehow more elusive than the dip used to be. This is not just imagination. With the dip meter, the oscillator frequency tends to lock in with the test-circuit resonant frequency. This produces a sort of "slot" that holds the dip down over a certain tuning range and makes it more easy to observe. The effect is quite pronounced with very tight coupling to a very high-Q circuit. Such grabbing of control by what you're trying to measure is entirely absent with this gadget. The result is a somewhat different "feel."

conclusion

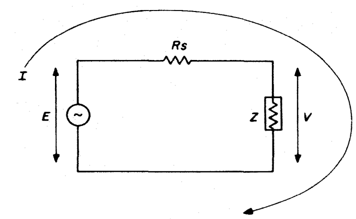
Despite its simplicity, this little indicator should be a big help in getting started at vlf. Now you ***can*** reach for your trusty dip meter — with the attachment — when working at 100 kHz. Or even 1 kHz!

reference

1. Keith Henney, *Radio Engineering Handbook,* McGraw-Hill, New *York,* 1959.

appendix

An approximate equivalent circuit, neglecting the internal resistance of the chip and the parasitic capacitances, looks something like this



$$E = I(R_s + Z)$$

$$V = IZ$$

At resonance $E_0/V_0 = (R_s + Z_0)/Z_0$

Hence $Z_0 = R_s/(E_0/V_0 - 1)$

As measured, E and V are real. Z is, in general, complex, which leads to complex equations. However, for Qs greater than about 10, the following reasoning is far simpler and leads to the same practical results.

First, define the true Q.

$$Q = Z_0/X_{L0}$$

When R_s **is in parallel with** Z_0 **(through the generator), the effective parallel resistance is**

$$(R_sZ_0)/(R_s + Z_0)$$

$$Q_a = [R_sZ_0/(R_s + Z_0)][1/X_{L0}]$$

$$= [R_s/(R_s + Z_0)]\,Q$$

Hence $Q = Q_a[(R_s + Z_0)/R_s]$

After substituting for Z_0 **and some fiddling, you get the formula given. This notation is for sine-wave voltages. If** E **is a square wave some adjustment must be made, as indicated in the text.**

ham radio

ground systems

for vertical antennas

Data to help you select the most efficient ground system for your vertical antenna

Over the past few years I've read and enjoyed the many articles about vertical antennas that have appeared in the Amateur magazines. Several of these articles refer to a very old research paper considered to be a classic in the field of ground system design.[1] I finally located this paper, which was written in 1937 by three engineers who worked for RCA in New Jersey. Several of their main points are summarized in this article, and the important results presented given in graphical form.

Using several different antenna heights, the RCA engineers measured the transmitted field strength at ground level, while varying the number and length of the radial ground wires. The radials were buried at about 15 cm (6 inches), with a test frequency of 3.0 MHz. One test, using radials laid on the surface of the earth, gave essentially the same field strength readings as when the radials were buried.

interpreting the graphs

First let's look at **fig. 1.** The actual measured field strength is plotted as a percentage of the maximum theoretical field strength for antenna heights from 10 to 90 degrees. Remember that one wavelength consists of 360 electrical degrees, so that 90 degrees equals 1/4 wavelength, while 45 degrees is 1/8 wavelength. Antennas taller than 1/4 wavelength were not used because of excessive height requirements. In **fig. 1A** each radial is 41 meters (135 feet) long, which amounts to 0.412 wavelength at 3 MHz. In **fig. 1B** each radial is 27 meters (90 feet) long (0.274 wavelength), while **fig. 1C** was plotted for 13.7-meter (45-foot or 0.137-wavelength) radials. **Figs. 2A** through **2D** show field strength as a function of antenna height for three radial lengths with the number of radials held constant.

What do these graphs mean? First, for any number of radials of any length, making the antenna higher will make it radiate more efficiently, although the graphs tend to flatten out in most cases once the antenna height reaches 1/8 wavelength or so. **Fig. 1C** shows that, even if many radials are used, the field strength doesn't exceed 80 per cent of the theoretical maximum value because the radials (just over 118 wavelength) are simply too short. Using only 15 radials in this case gives results almost as good as using 113 radials. This holds true even if the antenna itseif is very short.

If you use radials of about 1/4 wavelength (as in **fig. 1B**), the measured field strength increases to about 92 per cent of the theoretical maximum if 113 radials are installed. Increasing the length of the radials still further, to about 0.4 wavelength, as in **fig. 1A**, brings the measured field strength to 98 per cent of the theoretical maximum, if you again use 113 radials.

Looking at **figs. 1A**, **1B**, and **1C**, you can see that if you use only two radials for your ground system, the field strength will be *virtually identical* no matter how long the radials. Even if you use a full-size, 114-wavelength vertical, the field strength will be less than 65 per cent of the theoretical value. As the number of radials increases, the improvement in field strength is progressively greater (the curves become further apart) as the length of the radials is increased. As you've already discovered from **fig. 1C**, putting down more radials may not yield much improvement if the radials are *too short.* Similarly, using very long radials is a waste of time if they are too few in number, as shown in **fig. 2A. Fig. 2D** illustrates the benefits that can be gained by using long radials if you can install a lot of them.

examples

1. Suppose you have a 15-meter (50-foot) tower section that's base loaded with a low-loss inductor, and the system is resonant in the 160-meter "DX window" at 1825 kHz. The antenna height works out to be slightly less than 1/10 wavelength, or about 35 electrical degrees. Thinking that you'll probably need lots of radials to get good efficiency with such a short antenna, take a look at **fig. 2D,** which is for a ground system of 113 radials. Using 0.137-wavelength radials (about 22.5 meters or 74 feet) the efficiency is about 68 per cent; it is 87 per cent for 0.274-

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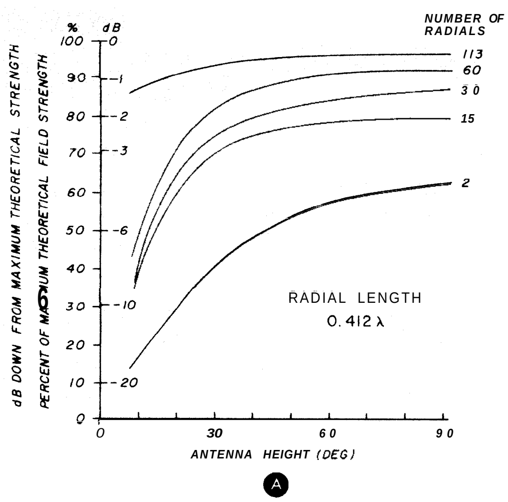
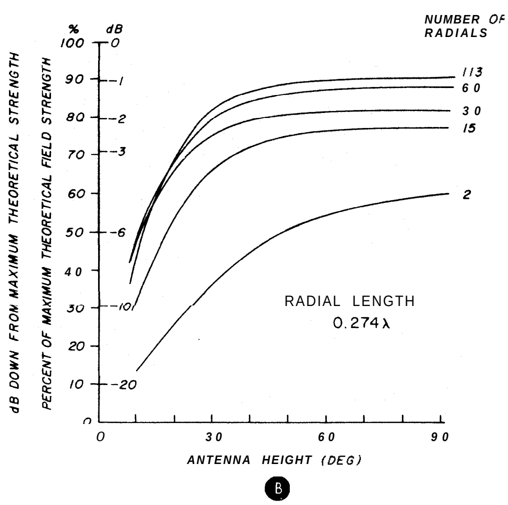
wavelength radials (45 meters, or 148 feet), and 96 per cent for 0.412-wavelength radials (68 meters, or 222 feet). By interpolation, 90 per cent efficiency would require 113 radials with lengths of about 0.32 wavelength (53 meters, or 173 feet).

Another approach is to use **fig. 1A**. About 96 per cent efficiency can be achieved by using 113 radials 0.412 wavelength long, as mentioned above. Using 60 of these radials yields 86 per cent efficiency; and, bv interpolation, 90 per cent efficiency is attained by





using 81 of these long radials. The first scheme requires 5948 meters (19,500 feet) of wire, while the second approach uses almost 5490 meters (18,000 feet).

2. Suppose you have a phased array of 114-wavelength verticals on 40 meters, and want 90 per cent efficiency. From **fig. 1B**, it seems that 60 radials, each 0.274 wavelength (12 meters, or 38 feet) will give the desired result. Or you can use **fig. 1A** to determine that 38 radials, each 0.412 wavelength (17 meters, or 57 feet), will also do the trick. Total wire length is 695 meters versus 662 meters (2280 versus 2170 feet)

3. Suppose you have a four-band trap vertical (40 to 10 meters) and your yard is such that the radials can't be any longer than 5.8 meters (19 feet). This amounts to about 0.137 wavelength at 7150 kHz, so **fig. 1C** applies.

Now it looks like it would be a waste of time to put in more than fifteen radials, because very little field strength is gained by adding extra ones. But this is a multi-band antenna, and those 5.8-meter (19-foot) radials are 0.274 wavelength at 14.2 MHz and 0.412

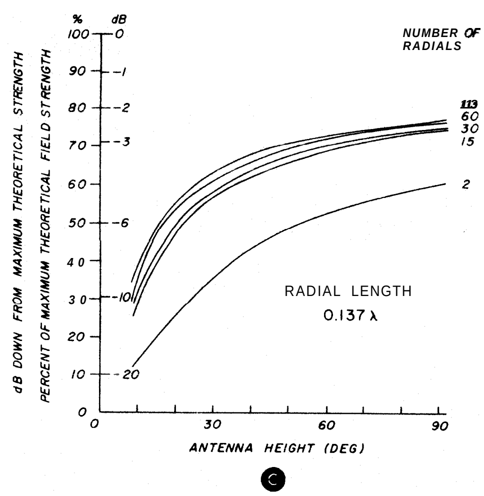


fig. 1. Design data for ground radial systems. Graphs A, B, and C show actual measured field strength as a percentage of maximum theoretical field strength for various numbers of radials 0.412. 0.274, and 0.137 wavelength long.

wavelength at 21.34 MHz, so **fig. 1B** applies for 20 meters, and **fig. 1A** applies for 15 meters. If you use sixty radials, each 5.8 meters (19 feet) long, you won't get much improvement on 40 meters, but the increase from 15 to 60 radials yields a gain in efficiency from 79 per cent to 91 per cent on 20 meters and from 77 per cent to 94 per cent on 15 meters. The results on 10 meters are better yet. Radials that are "short" on one band may be quite "long" on another higher band.

A tall antenna will have a broader bandwidth than a shorter antenna of the same diameter, so if a short antenna ***must*** be used, make it "fat." Instead of aluminum tubing or pipe, make the vertical radiator from tower sections. Alternatively, a wooden telephone pole can be used to support a wire cage built around it. A short, fat antenna may have a bandwidth as great as, or greater than, that of a tall, thin one. A top hat can be used to capacitively load a vertical and make it seem taller. Experiments by W2FMI show that a top hat of diameter ***D*** will increase the effective height of the antenna by $2D$.[2] A taller antenna will also have a higher radiation resistance.

There's nothing sacred about a 114-wavelength vertical, but an antenna of this height will have zero reactance, and therefore may be easier to match to the transmission line because its base impedance will be purely resistive. As the antenna is made taller, up to a maximum of about 5/8 wavelength, the elevation angle of the major lobe of radiation becomes lower, which should improve the DX capabilities of the antenna. If a very tall antenna is erected, vertical stacking through the use of a coaxial sleeve or other means may be used to achieve extremely low radiation angles.3

conclusion

Remember that these graphs were drawn from data taken at a specific location in New Jersey, and results may vary somewhat depending on local soil conductivity. However, the general results are still useful and may be used as a guide. It's important to note that for any given length of radials there comes a point at which it's useless to install any more of them.

If the radials are short compared with the operating wavelength, radiation efficiency will be low and relatively few radials will be needed to reach this point. If the radials are quite long, then many can be installed before reaching the point of diminishing returns. The "classic" ground system used by a-m broadcast stations consists of 120 radials, each 112-wavelength long, which gives a radiation efficiency on the order of 95-98 per cent.4

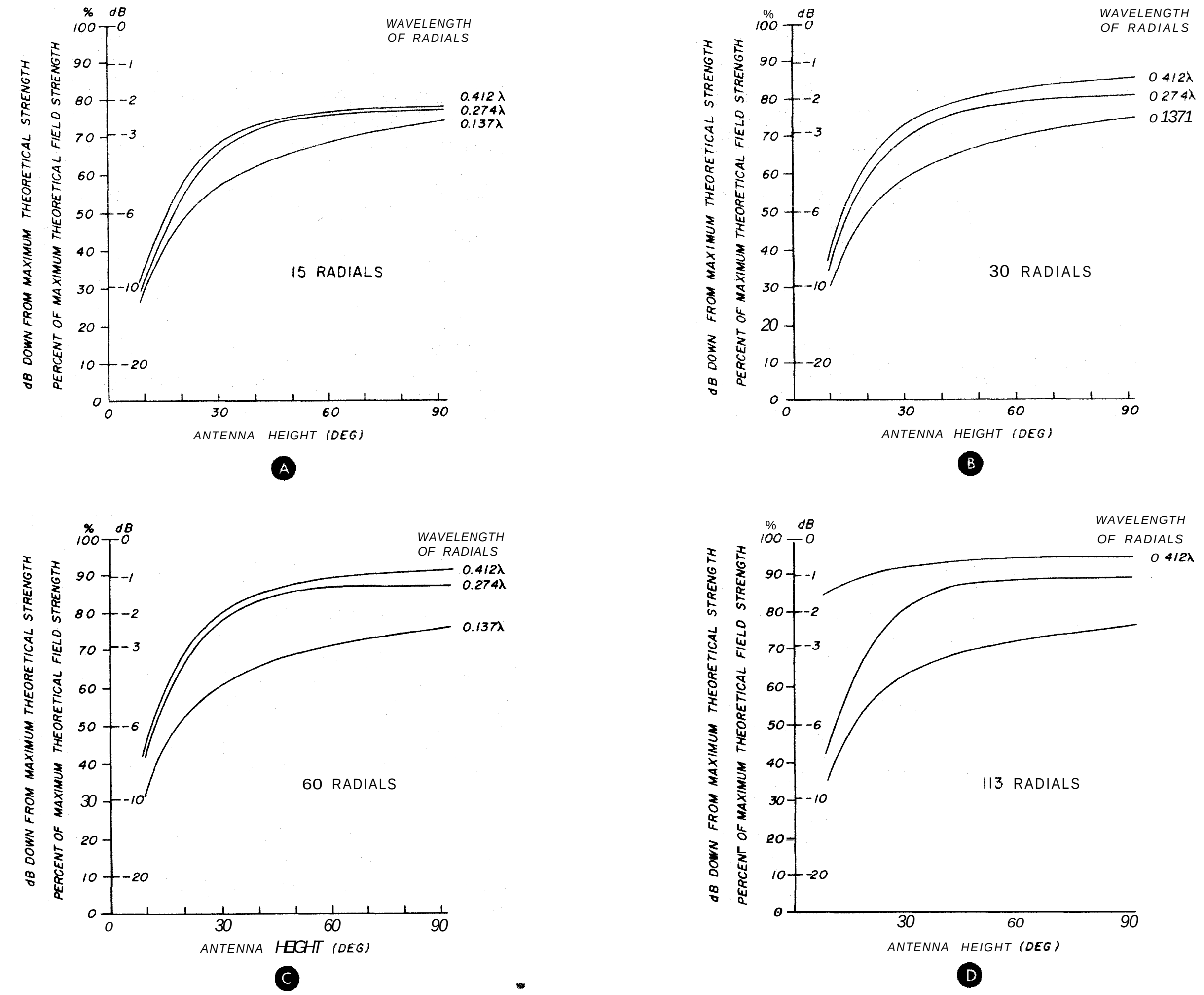


fig. 2. Performance comparison of fixed numbers of ground radials of various lengths. These graphs show measured field strength as a percentage of maximum theoretical field strength for 15, 30, 60, and 113 radials.

references

1. *Brown, Lewis, and Epstein, "Ground Systems as a Factor in Antenna Efficiency," Proceedings of the Institute of Radio Engineers, June,* 1937, *page* 753.
2. *Jerry Sevick, "Short Ground-Radial Systems for Short Verticals," QST, April,* 1978, *page* 30.
3. *Paul Lee, "Theoretical Aspects of Vertical Stacking," The Amateur Radio Vertical Antenna Handbook,* Cowan, 1974, *page* 43.
4. *Keith Henney, Radio Engineering Handbook,* McGraw-Hill, *New York, 5th edition, page* 20.

ham radio

all about traps and trap antennas

These days, almost everyone who enters the Amateur Radio hobby purchases or builds a bandswitching transmitter or transceiver covering 80 to 10 meters, allowing a degree of operating flexibility unknown in the days of single-band finals and plug-in coils. A logical extension of the bandswitching transmitter is an equally flexible antenna, capable of "instant" operation over the same bands the transmitter covers. Hence the trend to efficient, automatic, multiband antennas. One of the most popular, and technically sound, multiband antenna systems is the trapped antenna, both in its vertical and horizontal forms. Unfortunately, if ever there was confusion surrounding antennas, there's confusion about trap antennas; a great deal of mumbo-jumbo has been written about them.

In this article, I will attempt to set things straight with a discussion of some of the basic antenna concepts necessary to put traps in perspective, then I'll proceed to talk about both the trap dipole and the trap vertical. I'll also make some suggestions on how to feed, match, and tune the trap antenna, and discuss some possible problems you may encounter with harmonic radiation and television interference (TVI).

basic concepts

Many beginners do not realize the importance of using a good antenna, and, as a result, waste much of their transmitter power and spend much of their time unsuccessfully trying to make contacts. While virtually any piece of wire can be "loaded up" using a wide-range antenna coupler on any band, the performance of such a makeshift antenna will not likely set any DX records. Far better for single-band operation is the dipole, or "doublet," antenna, which is usually the simplest and most trouble-free kind you can use. The dipole is normally cut for the center of the desired band and fed with 50- or 75-ohm coaxial cable. A coax-fed dipole that is high and in the clear will usually work well over a range of ± 2 per cent of the center frequency before the mismatch even goes above 2:1.

A simple half-wave dipole is cut to frequency using the formula $F\ (in\ feet) = 468/frequency\ (in\ MHz)$.

Whereas the impedance at the ends of the antenna is quite high, approaching 3000 ohms or more, the center impedance of a high-frequency dipole at moderate heights runs about 50-75 ohms. This presents a good match for easy-to-handle coaxial cable. The other end of the transmission line (which may be of any reasonable length) is connected to the output connector of the transmitter or transceiver. Thus, a good match is also effected between the transmission line and the transmitter, which normally has a pi network output circuit designed to handle line impedances under about 100 ohms.

Problems arise when you try to use a dipole far from its design frequency: For example, using one cut for the center of the relatively wide 80-meter band (3750 kHz) at the high end, 4000 kHz. The impedance will then be reactive, since the antenna is no longer perfectly resonant, and it may result in a moderately high SWR and loading problems at the transmitter. Also, if you try to use a dipole that's cut for one band on another band, you may develop a severe transmission line mismatch (SWR) of up to 20 to 1 or higher. If you were to load up an 80-meter dipole with 40-meter rf from your transmitter, for example, you would find that the antenna no longer acts like an ordinary dipole, but rather like two half-wave antennas fed at their endpoints. The impedance might be around 3000 ohms, resulting in a mismatch to 75-ohm coax of about 40 to 1!

Thus, the dipole is essentially a single-band antenna, although there is an exception. At odd harmonics of its fundamental frequency the antenna's center impedance is low, so that it can be fed with low-impedance coax and work on certain higher-frequency bands. A 40-meter dipole, for example, can be used on 15 meters (21 MHz being an odd harmonic of 7 MHz); this fact is also made use of by trap manu-

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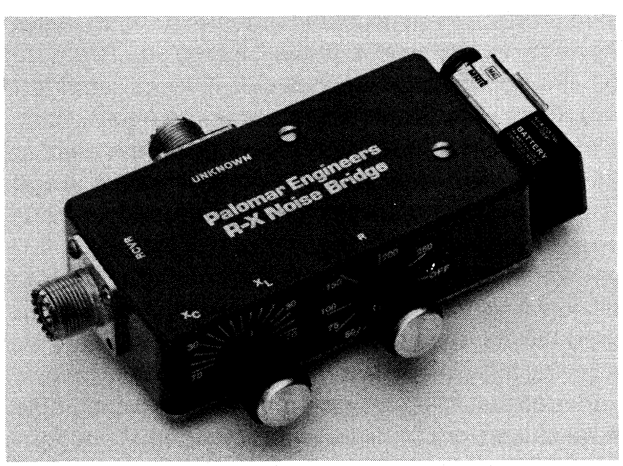
facturers to make possible operation on several bands using a minimum number of separate traps. In fact, the horizontal trap dipole is really an ingenious and versatile adaptation of the basic dipole or doublet.

What about verticals? I'll cover trap verticals later, but first I want to discuss basic vertical antenna concepts. The vertical is popular on all the high-frequency bands, and especially on the three highest bands (20, 15, and 10) for DX work. The simplest vertical is a quarter-wavelength long, fed "against ground" or connected to "artificial" groundplane radials tied to the base of the antenna. The groundplane, or earth, acts as a sort of mirror image for the antenna, allowing it to be a quarter-wavelength long rather than a full half-wavelength. The vertical's feedpoint impedance is usually between 25 and 40 ohms, so it offers a fair match to 50-ohm coax.

The vertical's popularity stems from two factors. The first is that it is a space saver; all one needs is vertical space, rather than a long horizontal antenna "run." If buried ground radials rather than aboveground radials are used, no horizontal space at all is required. (Of course, to be effective, the vertical should be installed as far as possible from other objects, such as house and utility wiring, and the ground system should be as extensive as possible.)

The second factor promoting the popularity of the vertical antenna for high-frequency work is that it produces a low angle of radiation, which places maximum signal near the horizon for good effect when working DX. When compared with a horizontal antenna mounted 10 to 15 meters (30 to 50 feet) high, the vertical will usually perform better over longer hauls, roughly 950 to 1300 km (600 to 800 miles) and beyond. On the other hand, the horizontal will usually give a better account of itself over shorter distances. The choice of vertical versus horizontal antennas for the high frequencies depends a great deal on what you want to use them for -- and also on individual preference. Overall, I prefer the horizontal on 80 and 40 meters (for short- and medium-haul work), and the vertical on 20, 15, and 10 for longer-path DX.

Traps and completed trap antennas may be fine tuned to resonance using an rf noise bridge such as the one shown here. The traps may also be adjusted using a grid-dip oscillator, and overall antenna performance may be checked with the aid of an SWR (standing wave ratio) bridge *(photocourtesy Palomar Engineers).*



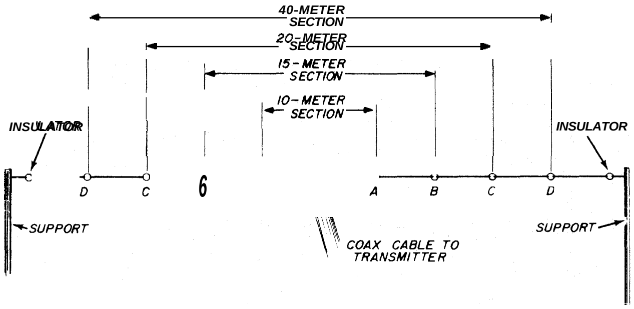
variations on a dipole

Lack of space has prevented many an Amateur from putting up any satisfactory antenna for the high-frequency bands, let alone install multiple antennas for each band on which he might want to operate. In many cases, an antenna length of about 30 meters (100 feet) is all that can be managed on a small city or suburban lot, not quite enough for operation on 75 and 80 meters. The trap antenna offers promise in this case, since its traps act as loading coils to "shorten" the required length to between 30 and 35 meters (100 and 110 feet) for 80-meter operation, and the electrical characteristics allow the antenna to be fed easily on the higher bands by an untuned (coax) feedline.

Trap operation is actually quite simple. Each is a parallel-tuned LC coil and capacitor circuit which is installed in the antenna to "divorce" the remainder of the antenna from the section on the "inside" of the traps. The principle involved is that an inductor and capacitor in parallel, when tuned to a given frequency and installed in a line, present a near-infinite impedance to rf current at that frequency. In effect, the parallel circuit acts to "trap" that particular frequency, acting as though it were the end of the antenna for that frequency. At all other frequencies (both above and below the trap's resonant frequency), the trap is a short circuit so that rf passes through it as though is were not there.

Several trap dipole configurations are popular. In a simple trap scheme, only one pair of traps is required for operation from 80 through 10 meters. In this case, a basic flat-top length of about 32.4 meters (108 feet) is used for 80-meter operation, the shortening being due to the electrical loading caused by the traps. On 40 meters, the traps insulate, or "divorce," the outside wire sections so that the antenna looks electrically like a 40-meter dipole. On the higher bands — 20, 15, and 10 meters — the trap dipole works out to electrical lengths that are roughly odd multiples of half-wavelengths (such as three half-waves on 20, five half-waves on 15, and seven half-waves on 10). Thus, use is made of the same principle that allows a simple 40-meter dipole to work on 15 meters.

In a more complex trap arrangement, a pair of traps (one on each side of antenna center) is required



selected trap and trap antenna suppliers

Antenna Supermarket, Post Office Box 1682, Largo, Florida 335450

Butternut Electronics Co., Route 1, Lake Crystal, Minnesota 56005

Cushcraft Corp., Box 4680, Manchester, New Hampshire03108

Electrospace Systems, Inc., Post Office Box 1359, Richardson, Texas 75080

Hy-Gain Electronics Corp., 8601 Northeast Hwy. 6, Lincoln, Nebraska 68505

Mosley Electronics, Inc., 4610 Lindbergh Blvd., Bridgeton, Missouri 63044

New-Tronics Corp., 15800 Commerce Park Drive, Brookpark, Ohio 44142

Pace-Traps, Box 234, Middlebury, Connecticut 06762

Unadilla-Reyco Div., 6743 Kinne Street, East Syracuse, New York 13057

Western Radio, Box 400, Kearney, Nebraska 68847

Wilson Electronics, 4288 S. Polaris Avenue, Las Vegas, Nevada 89119

fig. 1. Typical **80-10** meter multiple trap antenna using four pairs of traps. Section A-A forms the 10-meter antenna; the traps at the end of this section consist of resonant **inductor/capacitor** circuits which isolate the rest of the antenna when operating on 10. Sections B-B, C-C, and D-D work similarly on 15.20. and 40 meters. The full antenna resonates on the 80-meter band, but is slightly shorter than a full-size dipole due to the loading effect of the coils. Antennas can also be constructed using fewer traps, as explained in the text. The antenna should be installed at least 10 meters (30 feet) above ground; plastic clothesline or rope can be used as halyards. The trap antenna can be fed with either coaxial cable or with 72-ohm twinlead, which should be run from the antenna at right angles as far as possible; slight tension on the **feedline** will minimize swing.

for each band (except the lowest). For example, to cover 80, 40, and 20 meters, you would need two pairs of traps — one for 40 and another for 20. A pair is not required for 80, since the antenna itself resonates on that band. For five-band coverage from 80 through 10 meters, you would use four pairs of traps. In this scheme, the antenna is resonated by the single-band traps as a "true" half-wavelength dipole on each band, rather than as some multiple of a half-wavelength as in the simpler arrangement. Any combination of traps can be selected to make a "custom" multiple-band trap antenna, such as one covering 160 and 20 meters, or 80, 20, and 15 meters, to cite but two of many possible examples. Fig. **1** shows trap arrangement in a multiple-trap antenna.

Either trap scheme is capable of delivering good results. In the single-trap version, you can adjust the antenna length for the two lowest bands, but exact resonance on the higher bands (such as 20, 15, and 10) may not fall where you want it to fall and a high SWR may result. On the other hand, single trap antennas are lightweight; if matching becomes awkward, the feedline length can be adjusted to allow the antenna to take power, or an antenna coupler to be used. The multiple trap kind has the advantage that it can be adjusted for a low SWR on each band, but interaction between the traps can make adjustment tricky. Also, the many traps involved can introduce some system loss, and the antenna is heavy. In either case, trap antennas are not broad-band antennas, and they will show an increasing SWR as they are operated farther away from the center frequencies. Usable bandwidth is typically several hundred kHz, depending on the band, the design of the traps, and the physical antenna length involved. Although the ARRL *Radio Amateur's Handbook* and *Antenna Book,* as well as numerous Amateur magazine articles, describe trap construction, in some respects it's a good idea to purchase commercial traps, because the required low-loss construction and weatherproofing can be tricky.

Antenna installation is more critical with trap antennas than single-banders, especially since trap resonances are set assuming a high and clear antenna and they can easily be upset if these conditions aren't met. It's worthwhile to mount the trap dipole as high as possible, at least 10 meters (30 feet), If possible, the antenna should be a wavelength or more away from buildings and other obstructions, especially metallic towers and utility lines. It should also be well clear of tree limbs. Plastic clothesline (the kind without a metallic core) or rope can be used to support the antenna ends; the transmission line should be run away from the antenna at right angles for as far as practical and can be stapled directly to the house siding or run through TV-type standoff insulators in the shack. Do not try to use a trap dipole in a very limited space by bending down the ends at right angles; this will usually upset trap operation and detune the antenna, as well as unpredictably distorting the radiation pattern.

If space is a problem, however, you can run the flat-top in a horizontal-V or inverted-V shape with good results; many Amateurs actually prefer the V, believing that there is some apparent gain on the higher bands, directivity is enhanced, and the angle of radiation (for DX) is lowered. Inverted-V configurations have the advantage that the antenna requires only one high center mast, as the ends are sloped downward and can be suspended from convenient lower supports. There is no hard and fast rule about the angle of the V, but 90-120 degrees is normally used; it should not be less than about 75 degrees. Such arrangements are perhaps the best way to get on the air on several bands from a small city lot, yet allow for some directivity and gain which can be helpful in competing on the higher bands. **Fig.** 2 shows some popular V configurations.

What about antenna placement and directivity? Generally speaking, since the trap dipole is, essentially, a dipole, the signal pattern will be bi-directional and roughly doughnut shaped, with maximum signal perpendicular to the wire direction (90 degrees). Maximum radiation angle will be about 30 degrees from the horizontal. These figures assume an antenna height of about a half-wavelength, although fairly similar patterns will result if the antenna is at least an eighth to a quarter wavelength high. As a practical matter, directivity will not be pronounced on 80 meters and will be only slightly noticeable on 40. On the higher bands (20, 15, and 10), however, the antenna becomes much more directional, and optimum radiation departs from the doughnut-shaped pattern to become more of a cloverleaf, with maximum radiation lying about 30 degrees off the ends. This is especially true of single trap dipoles where the antenna operates at some multiple of a half-wavelength.

Thus, if you were to run your antenna east and west, maximum signal would be radiated southwest, northwest, northeast, and southeast. If possible, orient the antenna so that the four main lobes of the cloverleaf lie in the most favorable directions from your particular location for working DX on these bands. If you use the V or inverted-V configuration, the antenna becomes more sharply directional, with some gain apparent on the higher frequencies. Some typical radiation patterns are shown in **fig.** 3.

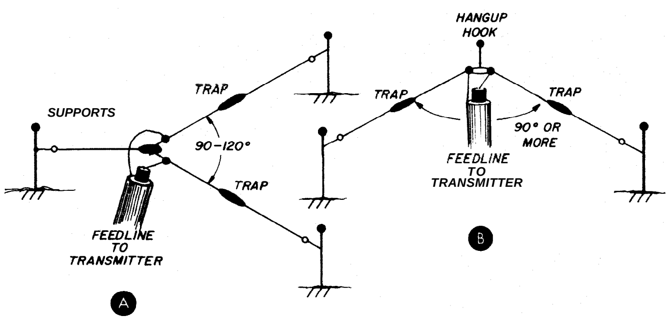


fig. 2. Diagram at (A) shows a V trap dipole, which can allow relatively long antennas to fit on small city and suburban lots. Three supports are required; the antenna is more directional than a straight dipole and may exhibit a few dB gain, particularly on the higher frequency bands. Diagram (B) shows an inverted-V trap dipole. which can be hung from a high center mast and the ends run to lower supports to form an approximate 90-degree angle at the apex. This arrangement is a favorite with **DXers** for improved low-angle radiation characteristic. The two configurations may be combined to form a so-called sloping-inverted V but the radiation pattern may become rather unpredictable.

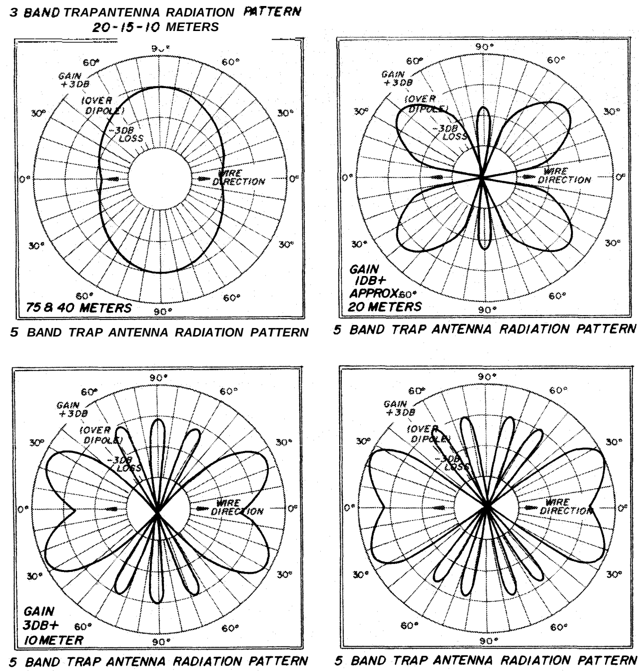


fig. 3. Typical trap dipole radiation patterns. The familiar doughnut-shaped patterns tend to become cloverleaf-shaped on the higher bands, where directivity is much more pronounced. The radiation pattern will vary from the patterns shown for V and inverted-V configurations. Vertical radiation patterns depend largely on the antenna's height above the ground or the structure on which the antenna is mounted. For best results, the antenna should be installed at a height of about 10 meters (30 feet) or more.

A logical extension of the trap antenna is the multi-band rotatable beam, usually in the form of a tribander covering 20, 15, and 10 meters. The tribander uses the same trap principles in resonating the director and reflector elements of the beam to give the antenna its directionality, front-to-back ratio, and gain figure. Few commercial trap beam antennas are available for 80 and 40 meter operation due to their unwieldy size, but for the high-frequency DXer, the three-band trap beam is probably a "best bet" if separate, full-size beams can't be installed.

trap verticals

The trap vertical's operation is very similar to that of the trap dipole; it works on the same principle.

As I've indicated, the quarter-wave vertical is a single-band antenna. But, like the dipole, it can be put to use on odd harmonics of its resonant frequency, so you can use the same antenna on, say 40 and 15

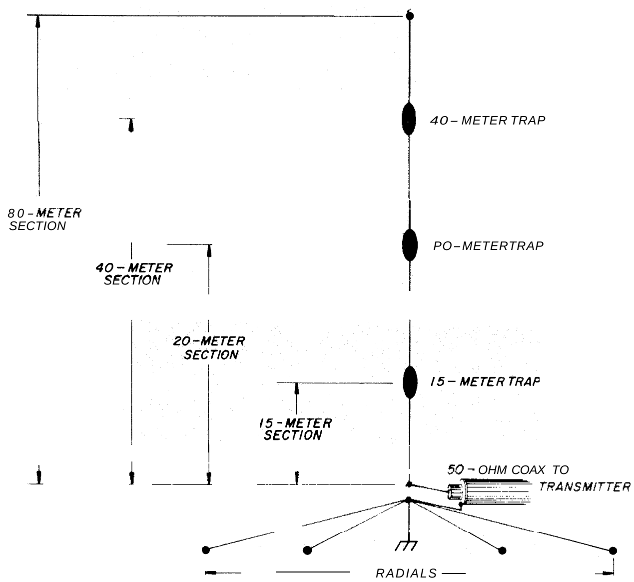


fig. 4. Typical trap vertical antenna configuration. The example shown operates on four bands, 80 through 15 meters, and uses three traps; many other variations are possible. Each trap isolates the higher sections as required to simulate a resonant quarter-wavelength at the operating frequency. A good radial ground system or groundplane is essential for good efficiency in the vertical antenna.

meters because of this special relationship. Besides the fact that the vertical is essentially a one-band antenna, its length for the lower bands, such as 160 and 80 meters, is much too long for most tastes. Many hams feel comfortable working with a length up to about 15 meters (50 feet), but not the 19.5 meters (65 feet) of an 80-meter vertical or the 39 meters (130 feet) of a full-sized 160-meter "skypole." As in the case of the dipole, the addition of the traps has a shortening effect on the antenna, so that the typical 40-10 meter trap vertical is somewhat shorter than 10 meters (33 feet). Verticals that are to be used on 80 or 160 also are longer for good efficiency on these bands, but shorter sticks can be used by installing add-on base loading coils at the expense of radiation efficiency. One trap per band (for all bands except the lowest) is usually installed to cause each section of the antenna to act as a quarter-wave vertical (although fewer traps may be used in certain designs). Thus, an 80 through 15 meter antenna is about 15 meters (50 feet) high and uses three traps in order to get four-band performance, as shown in **fig. 4;** sometimes 160 meters is added by means of a base loading coil or "resonator." The ground radials effectively provide the "missing half" of the antenna.

Like the standard single-band vertical, the trap radiates most of its power omnidirectionally at low angles above the horizon, and thus is a good choice for DX work. Unfortunately, many beginners have had poor results using trap verticals, but this is usually the result of mounting the antenna too close to signal-robbing obstructions, or not using a good ground system, or trying to work with a "too-compact," highly shortened vertical (especially on the lower bands, such as 80 and 160).

The trap vertical is a bit more sensitive to mounting position and grounding than single-band types; trap operation is easily upset by proximity effects. The antenna should be installed well clear of buildings, rain gutters, trees, and utility wires for good operation and low SWR. A good ground is extremely important for efficient performance. This means using six to twelve radials buried at least 15 cm (6 inches) in the ground, one or two ground rods at the base of the antenna, and possibly a direct connection to the house's cold-water piping. Without a good ground, much of the transmitter's power will be dissipated as heat, and the ground will not provide the proper mirror effect necessary for good results. If it is impossible to get a good ground connection, or if there are too many power-absorbing objects near the antenna when mounted at ground level, you can construct an artificial ground system, known as a groundplane, using at least four wires, or radials, connected together at the base of the antenna and running away from it like the spokes of a wheel. Usually, four quarter-wavelength wires are used, but, in the case of the multiband trap vertical, these would only be a quarter-wavelength for one band. In this case, you could run several quarter-wavelength wires for the lowest band to be used, and add several shorter "random" lengths which would take care of the higher bands. Using an artificial groundplane with a trap vertical is necessary in some congested locations and when the antenna must be installed atop a high building. In any case, be sure to carefully follow the antenna manufacturer's suggestions for mounting and grounding whenever installing a trap vertical, since the ground system is an integral part of the antenna.

You should also be aware that there is another type of multiband vertical that is technically related to the trap but is a good deal less expensive due to its much simpler mechanical construction. This is the base-loaded vertical, which uses a section of aluminum tubing usually between 4.8 and 10 meters (16 and 33 feet) in length; it has a tapped loading coil connected to it at the ground end. By making adjustments to the coil, the antenna can be resonated and matched closely on each band. However, since there are no traps, the antenna does not automatically switch bands, but requires that the operator change tap settings outdoors at the antenna when switching bands. (Remote switching arrangements are possible using relays controlled from the radio shack; these can become quite complex, however, for five-band operation.)

feeding, matching, tuning, and harmonics

The trap dipole can be fed with 50- to 75-ohm coaxial cable (using either RG-58/U or RG-59/U for moderate power levels — RG-8/U or RG-11/U if you're running a full kilowatt); the antenna impedance at resonance will usually fall in this range. The larger coax will also have the lowest loss, and, if you use a cable having a polyfoam center insulating material, it will have about 30 per cent greater power-handling capacity and a similar reduction in signal loss. If you must use a long feedline, use the larger-size cable with foam insulation.

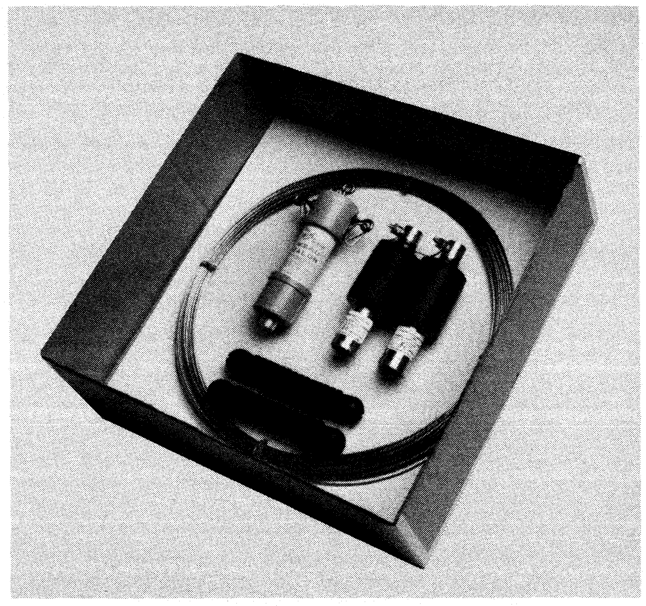
Since the dipole antenna is a balanced, or symmetrical, type, if you are using coax feed you may want to use a 1:1 balun transformer as the center insulator; this is not absolutely necessary, but using a balun can help equalize rf flow and prevent antenna current from flowing down the outside of the coaxial cable, causing distortion of the radiation pattern and possibly TVI. Commercial baluns are small devices resembling center insulators; most of them also have a convenient hang-up hook for V configurations and a coaxial connector.

You can also use 72-ohm twinline as the lead-in; it is less expensive than coax, and, since it is balanced, it does not require the use of a balun at the antenna — although you may still want to use one at the transmitter to mate with the coax output of most transmitters and transceivers. This type of line is frequently "lossier" than coax, however, and it should not be used on long runs, say over 45 meters (150 feet), unless special low-loss, transmitting-type twinline is used.

The vertical trap antenna should be fed with 50-ohm coax, and, if the cable is to be buried, the heavier RG-8/U type is preferred. A balun is not used with the vertical since it is an "unbalanced" type of antenna and works quite well when fed directly by coax.

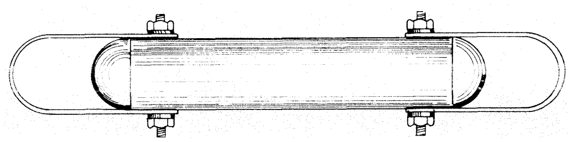
I should point out that any multiband antenna, trap types included, involves compromises to allow it to cover several bands. It is not possible to get a perfect match on each and every band, or from band edge to band edge. Some advertising literature leads one to believe that the trap "match" is perfect over all bands, but this is not so. In most cases, it is necessary to adjust the traps, shorten or lengthen sections of the antenna flat-top, or even change transmission line length to get uniform transmitter loading on all bands.

What is important is to get a reasonable match across all the bands that you want to use; if the feedline isn't too long, losses will not be excessive and the antenna will work well even on the higher bands with SWRs of 4:1 or 5:1, although you may want to use an antenna tuner to facilitate loading.



Representative trap multiband antenna coils. The coils electrically isolate portions of the antenna flat-top depending on their resonant frequency. Electrically similar but mechanically different coils can also be used in connection with vertical antennas for multiband operation *(photo courtesy Unadilla/Reyco)*.

When using traps, you may find that one or more bands show a higher SWR than desired. Since most commercial traps are pre-tuned and sealed, this means that you must adjust antenna section lengths to "tweak" the SWR into shape. Reyco, in its product literature, gives a simple procedure for adjusting the traps in a dipole: After constructing the antenna according to the recommended dimensions, each band is checked for SWR (starting with highest band) and the point of lowest SWR is noted. If the resonant frequency is too high, the center sections are increased slightly in length until the SWR is lowest at the desired operating frequency; if too low, the lengths are decreased. The same procedure is followed on each band, working from the high to low, adjusting the wire sections between traps. Using the one-pair type system, you can use this procedure for the two lowest bands, but adjusting the lengths for best operation on, say 80 and 40 meters may "throw out" the SWR on one of the higher bands. These effects can't be overcome, but they can be minimized by cutting the feedline to certain lengths. Doing this will not affect the true SWR, but can help the transmitter to "see" a good match. You should find that starting with coax lengths of about 13.5, 25.5, 33, and 40.5 meters (45, 85, 110, and 135 feet), or twinlead lengths of 22.5, 30, 33, and 39 meters (75, 100, 110, or 130 feet) should reduce loading problems on the higher bands; the feedline can be lengthened or shortened as necessary to get a good compromise match on all bands. It may not be possible to get a perfect match on all bands, however, and trying to get the match down from, say, 2:1 to 1:1 on each and every band can be extremely frustrating and is probably just not worth the effort in terms of improved ability to get out.



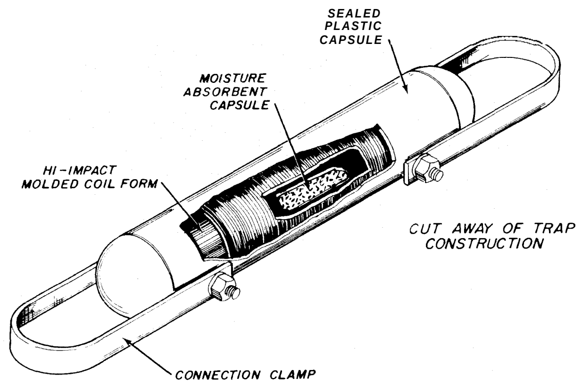


fig. 5. Representative exterior sealed trap construction is shown at (A). In this case, the trap itself is encapsulated in high-impact plastic which offers protection against the elements. The antenna wires are attached to the trap by means of two metal U-clamps. An interior view of the same trap is shown at (B). Traps designed for multiband beams and verticals are electrically similar, but are usually encased in or wound over fiberglass forms for added strength. For lowest loss, there should be no metal inside the rf field of the trap coil.

Trap verticals are tuned in similar fashion, but instead involve either sliding the aluminum element lengths, tuning the traps themselves, or adjusting a special top-hat section. In most cases, you can get good results using nothing more than your SWR bridge in adjusting commercial trap antennas, but if you build your own traps (whether for dipole or vertical use) you will have to first adjust them for resonance using a grid-dip oscillator or rf noise bridge, then seal them against weather before installation. **Fig.** 5 shows typical trap construction.

There is a potential problem in using multiband trap antennas that should be carefully considered, and that is the problem of harmonics. While a single-band antenna will reject even harmonics of the operating frequency, that rejection just isn't there in the trap antenna, and, in fact, any harmonics present are efficiently radiated. This is an especially critical problem when using a five-band antenna system, since harmonics of even 80-meter signals will be radiated nicely through at least ten meters; this problem has caused many unsuspecting Novices operating on 80 and 40 to receive FCC citations for radiatinq out-of-band signals. While most pi network output circuits have good harmonic suppression, if they are loaded too heavily (such as when trying to get a high-SWR antenna to take power), their harmonic rejection can be destroyed and second-, third-, or higher-order harmonics are passed on to the antenna and radiated.

You can make a rough check on your own harmonics by having a friend listen to your signal on harmonically related frequencies; if he is located a short distance from your home, your signal should be received very weakly, if at all, on harmonics. This is not an infallible test, however, and a more certain approach to harmonic suppression is to use an antenna coupler or tuner in the transmission line between your transmitter and the antenna. The tuner will add a great deal of selectivity to the antenna system, causing harmonics to be reduced to an acceptable level. In addition, the tuner also helps the transmitter to "see" a near-perfect 50- to 75-ohm match and thereby load more consistently from band to band. This is a real plus when operating with a moderately high SWR on the transmission line, frequently the case when using multiband antennas. In addition if the coupler is installed so as to be in the circuit on receive (automatically the case when using a transceiver), it will add a good deal of front-end receiving selectivity as well, helping to prevent i-f image signals and very strong local stations on other bands from coming through and cross-modulating the receiver's front end.

Not to be overlooked is the increased potential for TVI (television interference) when using multiband antennas. Certain conditions may cause the antenna to radiate vhf harmonics of your high-frequency signal, particularly on the lower TV channels (2 through 6). The short, well-matched transmission line between the transmitter and antenna coupler or tuner makes an ideal spot to install a lowpass filter, as shown in **fig. 6;** placing it there prevents any possible danger to the filter from high SWRs or upsetting its operation by SWR mismatches. The use of a good lowpass filter (which in itself can provide 60 dB or more harmonic attenuation) in connection with an antenna tuner should keep you out of trouble with both the FCC and your neighbors.

summary

Trap antennas, both horizontals and verticals, offer much to the ham who has space limitations and cannot erect separate antennas for each band. Traps are capable of excellent performance if installed and adjusted properly, but one should keep in mind that certain trade-offs are involved in their design; this fact must be recognized when interpreting their performance.

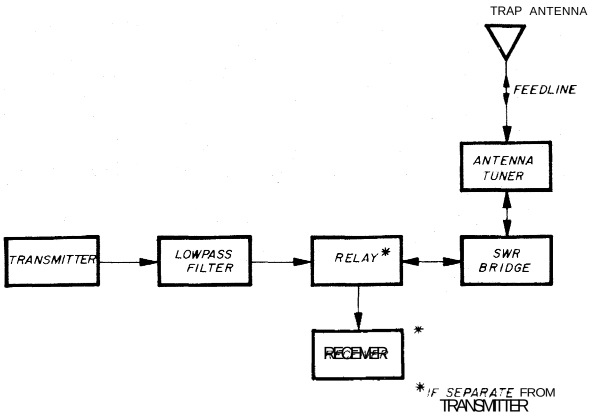


fig. 6. Suggested multiband trap antenna system feeding arrangement. Due to the inherent nature of trap and other multiband antennas, harmonics are easily radiated. For this reason, you should use a lowpass filter and antenna coupler to suppress these harmonics. The antenna relay shown is required only if a transmitter is used, that is, if you are not using a transceiver. If you use a separate receiver, by plating the antenna relay as shown the antenna coupler is used to good advantage on receiving as well. While either 72-ohm twinlead or coax may be used to feed the trap antenna, coax is usually the best choice. In adjusting the antenna length, move the SWR bridge to the antenna side of the coupler; once the adjustments are made, place the bridge as shown for routine tuneup.

Trap antennas can take the place of five or more individual antennas. Remember that your trap is doing a big job for you, so install and treat it right. If you do, you can expect excellent performance from it.

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ham radio

More Details? CHECK--OFF Page 110

curing frequency drift in the Swan 350 transceiver

A frequency counter, some deliberation, and a handful of capacitors will help tame frequency drift in this popular rig

I recently undertook the task of reducing the more than 1-kHz frequency drift in K6OWA's Swan 350C transceiver. Since then I've had several queries from others who own these reliable rigs. The availability of a frequency counter made this project easy to implement. This article summarizes the techniques I used so that other Swan 350 owners can reduce frequency drift to a tolerable level. The final drift (or lack of it) achieved by this electronic surgery depends on your determination and patience.

analysis

Most VFOs drift to a lower frequency. However, K6OWA's Swan 350 had a positive frequency drift. Most drift occurs in the VFO coil, caused by changes in coil dimensions with temperature. Drift in the Swan 350 is likely aggravated by the seven tubes clustered around the VFO box. Lack of ventilation traps the heat inside, thus extending the time before drift levels off. On-off cycling over a period of years will gradually stretch the coil wire to the point where it will not return to normal – something like the "set" of a fishing pole. Compensation originally found adequate will no longer keep the drift within reason.

It's interesting to compare specifications of maximum frequency drift in present-day rigs. Several well-known units list 1 kHz, some less than 300 Hz, and a very few 100 Hz. My TS-520, rated at 2-kHz, came out with 75 Hz. Both the Alda 103 and the Swan 100MX are rated at less than 100 Hz. The ARRL's *Amateur Radio Handbook* (1976), on pages 166 and 169, shows homebrew VFOs. One stabilizes in 1½ *minutes* at 15 Hz, while the other levels off in only 30 *seconds* with a 25-Hz drift. Truly amazing.

mechanical considerations

Since mechanical ruggedness is synonymous with low frequency drift, begin by tightening *all* nuts and bolts around and in the VFO compartment, including the five screws holding the band coils to the chassis. In some cases the VFO circuit board is underneath the chassis and not in the can. Then apply Lubriplate to the bandswitch bearings to reduce torsion on

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these parts. Thoroughly clean all eight wafers with contact cleaner and remove lint and dirt from the variable cap plates with a pipe cleaner.

compensation

The next step, compensation, will be facilitated by the use of a 7- or 8-digit frequency counter. Absolute accuracy isn't imperative because the counter records approximate drift only. Begin with the chassis lying on its final-amplifier side, and connect a short piece of wire to pin 1 of the VFO amplifier, V1, (see your schematic). This will permit attachment of a counter probe with less chance of short circuits to other leads.

Set the bandswitch to 20M and the dial to 14,200 kHz. When the set is turned on, the counter will read roughly 8700 kHz, as noted on page 14 of the Swan manual. During the first run, connect a voltmeter to read the – 10 volt bus from zener diode 1N2974. If this voltage is steady during a 45-minute test the zener need not be replaced. (A faulty diode will cause shift, however.) Leave the cover on the VFO box to simulate normal conditions. Have a sheet of paper ready to log data each five minutes from the moment the set is turned on. At least 30 minutes should elapse – certainly long enough to see a leveling off or maximum drift.

Turn off the Swan 350 and let it cool for the next run. Meanwhile make a graph of the first run as in **fig. 2,** curve A. K6OWA's Swan 350 was still increasing in frequency after 45 minutes (past 1200 Hz). If experience convinces you that drift occurs on only *one* band, refer to your schematic and **fig. 1,** noting capacitors 1711, 1713, 1718, 1720, and 1723. These negative temperature coefficient caps compensate each band, the idea being that each coil needs individual compensation, and C1709 in series with the variable cap takes care of things in general. Most likely this won't be the case – cap C1709 will be removed and replaced, thereby compensating all bands.

You're now ready to make a substitution for C1709. Remove the VFO cover plate, probe, and voltmeter leads. Unsolder the rf and dc leads from beneath the chassis. (Two nuts and washers under the chassis and two nuts, washers, and two spacers hold the PC board to the chassis.) With the Swan 350 lying on its side, remove the PC board. This will position the PC board horizontally and will make board removal and replacement easier because the hardware won't get lost in the equipment innards. The spacers especially have a nasty habit of getting lost if removed with the board in a vertical position. Unsolder C1709 at the variable cap (C1706) and carefully pull the PC board from the compartment.

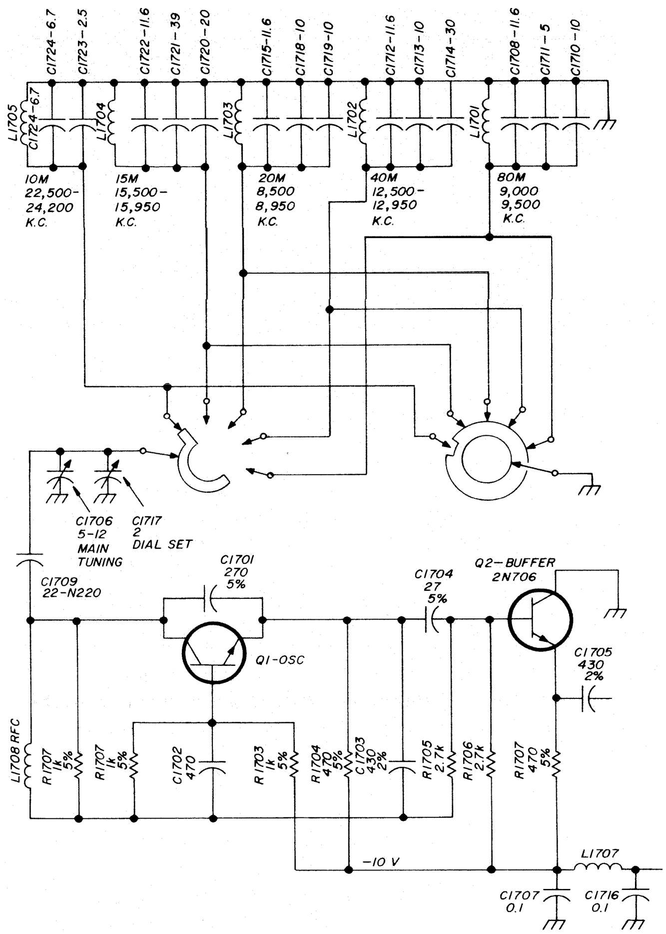

fig. 1. Partial schematic of the Swan 350 transceiver showing the VFO section. Capacitor C1709 is the subject of the modifications in this article.

compensating capacitors

Capacitor C1709 is really an unknown and is suspect. Discard it since it's incapable of compensating for the inherent frequency drift. The Swan manual specifies C1709 as a 22-pF capacitor with an N220 temperature coefficient, meaning 220 parts per million. (Capacitor C1709 in K6OWA's rig was an *N150,* obviously tailored to compensate *that* unit.)

A second trial run must now be made. I chose a parallel combination of 10-pF/N150 and 12-pF/NPO.

The N value of this new combination is

$$N_{eq} = \frac{10}{22} \times 150 = N68 \qquad (1)$$

Put everything back, tighten the board securely, reconnect the counter, and begin anew with another run.

Fig. 2, curve C, shows a *negative* drift (more than 300 Hz). The next, and you hope final, attempt must

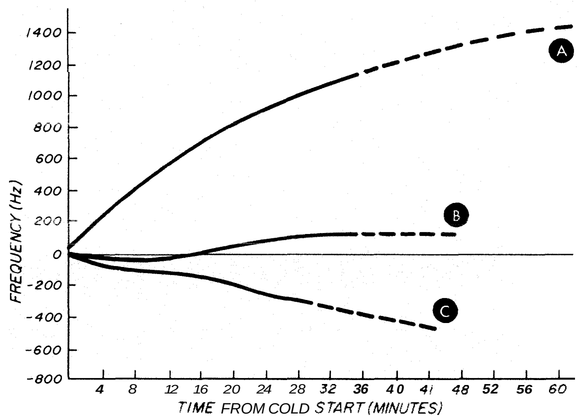

fig. 2. Test data showing frequency drift as a function of time for several test runs on the Swan 350 transceiver. Capacitor C1709 must be tailored for individual cases. Naturally, any long-term drift will approach zero. (A): Frequency drift of the as-built Swan 350 transceiver extrapolated to 1500 Hz after a warm-up period of one hour. C1709 was a 22-pF/N150 compensating capacitor. (B): The third test run, using an N120 capacitor for C1709. The measured frequency drift is 105 Hz after 35 minutes from a cold start. (C): A trial run using substitute N68 capacitor for C1709. The measured frequency drift is 335 Hz after 30 minutes from a cold start.

have C1709 with an *N* between 68 and 150. I tried a 12-pF/N220 and a 10-pF/NPO resulting in N120.

Out comes the oscillator board; again watch out you don't lose hardware. Replace the VFO top cover and connect the counter after soldering the N120 combination in place. The set will have cooled off sufficiently during this process for the next run to be from a cold start. Somewhere in this operation you'll take out time for lunch, so let the Swan 350 cool off — too much time and effort will have been expended to take chances on a set that hasn't reached room temperature.

The last run will perhaps be what you'll settle for in performance. Ralph, K6OWA, and I were quite elated when we finished the third run and plotted curve B (**fig. 2**), ending up with frequency drift close to 100 Hz. Any set that stabilizes in 30 minutes or so with that amount of drift is a pleasure to operate. This Swan 350 certainly takes no back seat to many new sets off the assembly line. Perhaps at some convenient time, another trial could be made using a 12-pF/NPO and 10-pF/N220 combination:

$$N_{eq} = \frac{10}{22} \times 220 = N100 \qquad (2)$$

Mathematically, any VFO can be compensated to achieve no drift. This particular Swan 350 might need an N105 or perhaps an N95; this must be determined experimentally. Generally you must weigh the advantages gained with close-to-zero drift and the effort expended.

When you've made mods to satisfy your standards, you must refer to the *Swan* manual, page 14. With the counter reading the VFO frequency, adjust the trimmers for each band to correspond with the dial-frequency/oscillator-frequency table. Follow manual instructions.

obtaining parts

If you live near an electronics emporium procurement of capacitors to conduct the tests will be simple; if you're on a Pacific atoll, you must provide yourself with enough NEG caps to experiment with. In our case, it seemed logical to use half of the 22-pF total value in an NPO (thus the 10- or 12-pF values), while the other half can be used with negative N values to suit. One of each of N220, N150, and N110 would be a starter and would probably bring the frequency drift within reason. This evaluation must be made on the second trial with a new cap value.

The direction of drift would dictate whether higher or lower N values will work. Two or three attempts should give a good idea of the necessary value. If a supplier has a poor selection, combinations other than 10 pF and 12 pF could be tried. Eight pF and 14 pF could be combined in a ratio that might do the job.

closing remarks

The technique outlined applies to any VFO, tube or transistor, in which an increase in temperature causes a frequency change. Getting rid of the heat in a Swan 350 is impossible; the set is already quite well ventilated, so an extra muffin-pan blower won't suffice! Thermal insulation of the outside VFO walls from the heat of adjacent tubes might help. However, I feel that the lack of circulation in the VFO itself is the cause of excessive frequency drift. A perforated chassis and sidewalls would allow heat to escape, whereas the Swan 350s, as designed, trap heat inside the box.

Further suggestions on drift problems are covered in many articles in the Amateur Radio literature: The ARRL handbooks and the new *Solid State Basics* from *ARRL* also elaborate on design of VFOs to minimize frequency drift.

acknowledgment

Neither K6OWA nor I have a frequency counter, so credit must be given to Reed Craven, WB6BFK, for the use of his counter, thus making this interesting project possible.

ham radio

close look at amateur fm

Should the FCC regulations that combine F3 and A3 emissions be changed? Here's one Amateur's viewpoint

Casual observers of Amateur Radio practice have found that operators seldom use more than three types of emission and frequency bands, even though Amateurs are allowed as many as thirteen emission types on seven bands, twelve types on six bands, eleven types on two bands, six types on four bands, five types on one band, and two types on one band (160 meters). Most Amateurs use two types of emission: CW (type A1) and a-m or SSB phone (type A3), on several of the bands listed and largely ignore frequency and phase modulation (type F3). Mobile operators, however, make good use of fm on the vhf and uhf Amateur bands. Fixed-station operators have little use for fm on the high-frequency (3-30 MHz) bands, even though A3 and F3 emissions are listed together on virtually all subbands except 160 meters.[1]

amateur fm

The disuse of fm on the 15-, 20-, 40-, and 80-meter bands is because of deterrents such as lack of suitable transmitters and receivers and the FCC's bandwidth limitation. Paragraph 97.65c, under the heading of *Emission Limitations,* reads: "On frequencies below 29.0 MHz and between 50.1 and 52.5 MHz, the bandwidth of an F3 emission (frequency or phase modulation) shall not exceed that of an A3 emission having the same audio characteristics; and the purity and stability of emissions shall comply with the requirements of 97.73[1]." This paragraph deserves special attention.

Amplitude modulation contains only one pair of sidebands. The audio-frequency limit for voice communications is nominally 3 kHz; therefore, the a-m bandwidth can't exceed 6 kHz. However, fm contains one or more pairs of sidebands, equally spaced from the carrier frequency or adjacent sidebands. Therefore the fm bandwidth can't be less than, and may greatly exceed, the a-m bandwidth.

We note that FCC specifies *fm bandwidth,* not frequency deviation, which is not proportional, although some Amateur publications[1] give a simple rule of thumb that fm bandwidth equals twice the maximum frequency deviation plus the maximum audio frequency. One book[1] states that this rule of thumb doesn't hold for narrowband fm, and defines sliver-band, narrowband, and wideband deviations as 2.5, 5, and 15 kHz, for which bandwidths are 6, 13, and 33 kHz approximately. The last two of these bandwidths are far below accurate values of 16, 22, and 48 kHz, respectively, which further analysis has proved.

bandwidth

Bandwidth is a complex function of deviation and audio frequency, not of deviation alone. It depends on a factor known as modulation index or deviation ratio – the ratio of maximum deviation to maximum audio frequency, which is a pure number. There's no simple relationship, such as a direct proportion between bandwidth and deviation, although bandwidth generally increases with deviation in a nonlinear manner for a fixed audio frequency. An increase in audio amplitude at any audio frequency increases the deviation, the number of sideband pairs, and the bandwidth. Also, an increase in audio frequency at any given deviation increases the spacing between the

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sidebands, which increases the bandwidth. The overall effect of changes in audio amplitude and frequency on bandwidth is often hard to evaluate, as, for example, the case of complex speech waves where the amplitude of the various frequency components varies somewhat inversely with the frequency.[2]

deviation

The fact that bandwidth is not a function of deviation alone is shown by the following example. Assume two cases with the same deviation of 10 kHz but different audio frequencies of 5 and 10 kHz respectively. In the first ease, four pairs of significant sidebands (at least 1 per cent of carrier amplitude) are 5 kHz apart, so the bandwidth is 4 x 2 x 5 kHz, or 40 kHz. In the second case, only three pairs of significant sidebands occur but are twice as far apart, so the bandwidth is greater (3 x 2 x 10 kHz). Since $60/40 = 1.5$, it appears that bandwidth is not in direct proportion to audio frequency for a fixed deviation.[3]

In fm broadcasting, the maximum permissible deviation is 75 kHz and the maximum audio frequency (for high-fidelity music) is 15 kHz, so the modulation index might approach 5 (75/15) if both values occur simultaneously. Similarly, in Amateur wideband fm, the deviation may be 15 kHz and the audio frequency may be limited to 3 kHz (for speech only); so the same modulation index would occur. A ratio of five signifies eight sideband pairs; hence bandwidths of 240 (8 x 2 x 15) and 48 kHz respectively are indicated. These values are extremes, which are not realized in practice because complex music and speech waves have smaller amplitudes at the higher frequencies. So less deviation and fewer sideband pairs occur than the indicated values.

narrow-bandwidth case

If we consider a narrow bandwidth of 6 kHz with a 3-kHz audio-frequency limit, as in Amateur a-m, there can be only one pair of significant fm sidebands. This corresponds to a modulation index of no more than 0.4 and a deviation of no more than 1.2 kHz ($1.2/3.0 = 0.4$). Under these conditions, the fm carrier retains 96 per cent of its unmodulated amplitude, and the sidebands have only 4 per cent of the amplitude they would have at a modulation index of 2.4 (or 5.5) – the ideal condition wherein all carrier power is converted to sideband power.

If the FCC had specified a 3-kHz deviation limit instead of an a-m bandwidth limit, a better condition would prevail. A modulation index of unity (3/3) would yield three pairs of sidebands, 18 kHz bandwidth (3 x 2 x 3 kHz), 76.5 per cent carrier amplitude, and 23.5 per cent sideband amplitude.[4]

Even the latter imaginary case wouldn't offer much of an advantage over the former real case, because power is proportional to the square of the amplitude (voltage or current), hence the sideband power is only 5.5 per cent (100 x 0.2352) of the ideal power.

From this discussion it follows that Amateurs can't make much use of narrowband fm wherein the bandwidth may not exceed 6 kHz. Accordingly, the FCC listing of type F3 along with type A3 emissions on bands below 29.0 MHz is misleading and serves no useful purpose, so it should be dropped. On the other hand, if a deviation of 5 kHz or 22 kHz bandwidth were permitted on certain Amateur bands, as or! certain commercial *frequencies,* fm operation would be quite feasible, because the carrier amplitude would be only 41 per cent, and the total sideband amplitude would be 59 per cent.

table 1. Amateur fm characteristics.

M	%F	P	ΔF	(BW)
0-0.4	100-96	1	0-1.2	0-6
0.5	94	2	1.5	12
0.83	83	–	2.5	16
1.00	77	3	3.0	18
1.67	41	–	5.0	22
2.00	22	4	6.0	24
2.40	0	5	7.2	30
3.00	26	6	9.0	36
3.33	34	–	10.0	38
4.00	40	7	12.0	42
5.00	18	8	15.0	48
5.50	0	–	16.5	51
6.00	15	9	18.0	54
6.67	25	10	20.0	60
7.00	30	11	21.0	66

Notes:

f = 3 kHz, fixed

M = modulation index = $\Delta F/f = \Delta F/3$, so AF = 3M

%F = carrier amplitude in per cent from reference 2

P = pairs of sidebands of 1% minimum amplitude (reference 2)

AF = deviation (±) = fM = 3M

BW = bandwidth – 2fP = 6P

(BW ranges from 1.4 to 2.0 times rule-of-thumb bandwidth, 2 AF + 3 and 1.5 to 4.0 times 2 ΔF.)

These points may be emphasized by reference to **table 1,** which may be converted to a series of curves if desired.

references

1. *The Radio Amateur's License Manual,* ARRL, Newington, Connecticut, 1977, pages 66-67 and 92.
2. August Hund, *Frequency Modulation,* McGraw-Hill, New York, 1942, pages 22 and 23.
3. Ed Noll, W3FQJ, *First-Class Radiotelephone License Handbook,* Sams and Bobbs-Merrill, New York, 1970, pages 139 and 140.
4. *FM and Repeaters for the Radio Amateur,* ARRL, Newington, Connecticut, 1972, pages 50 and 169.

ham radio

high-current regulated dc power supply

A high-current, regulated power supply incorporating rf protection, current limiting, and over-voltage protection

The present popularity of the two-meter fm band has attracted many Amateurs to that portion of the spectrum. Many hams use two-meter rigs in double-duty service, both in mobile and fixed-station applications. In either case, 12 Vdc is required for input power. For mobile use, the automotive electrical system provides input power; however, in fixed station use the rig requires a power supply which converts ac power to nominal 12 Vdc with current capacities ranging from about 3 to 10 amperes. Such a power supply should contain the following features:

1. Voltage Regulation — For steady-state regulation, the output voltage should remain within 1 per cent of the desired value over the full range of output current capability. Instantaneous load changes, or dynamic regulation, commonly encountered when keying or unkeying the transmitter, should not produce excessive voltage overshoot or undershoot. Additionally, the settling time required for return to steady-state regulation conditions should be minimal, on the order of milliseconds.

2. Current Limiting — The power supply should be protected against excessive output current. The current should be automatically limited to a preselected safe value.

3. High-voltage shutdown — In case of a power supply fault, which would apply unregulated dc to the rig, a protection circuit is required to disable the power supply output within milliseconds after fault condition detection. The radio being powered is thus protected against excessively high voltages which could permanently damage semiconductors or other components.

4. Rf protection — The voltage regulation and other control circuits should be rf bypassed to eliminate the possibility of voltage instability or other adverse effects caused by strong rf fields.

The power supply described in this article incorporates the outlined features. By using the specified components you can construct a power supply which is totally adequate for 30- or 40-watt transmitters.

circuit description

This power supply uses a standard series-pass circuit controlled by a 723 monolithic IC regulator. The high-voltage shutdown circuit is controlled by a μA741 operational amplifier. All regulation and control circuits are contained on one circuit card, compatible with a 15-contact edge connector.

The power circuit consists of a transformer, bridge rectifier, pi-section filter, and series-pass transistors. This is a standard circuit commonly used in many regulated power supplies.

Operator controls consist of an ON/OFF toggle switch, a pushbutton switch to reset the high-voltage shutdown circuit, and appropriate indicators.

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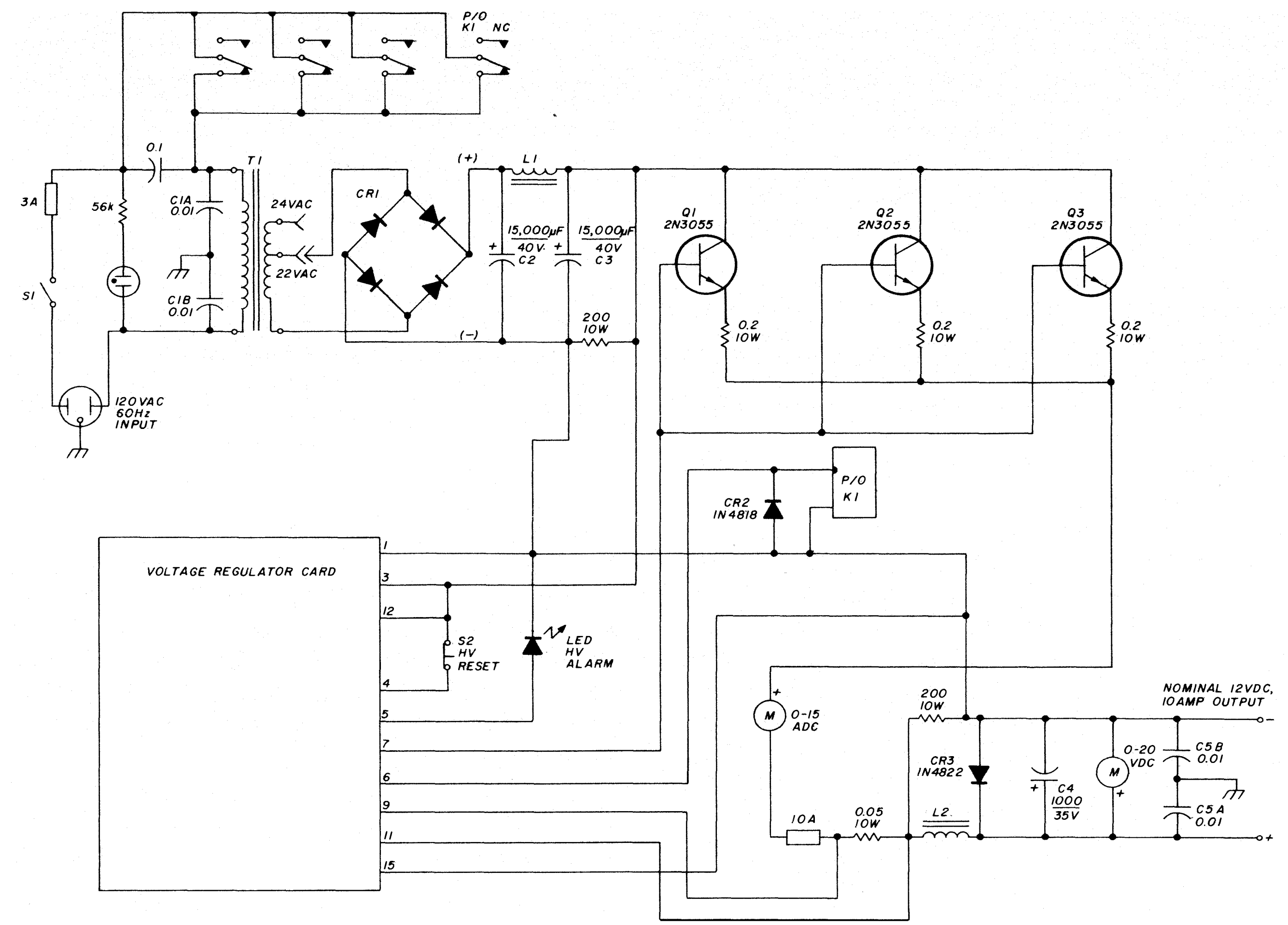


fig. 1. Schematic diagram of the chassis-mounted components. **CR1** is a **15-amp, 50-PRV** bridge rectifier; CR2 a 1.5-amp, **200**-PRV diode **(1N4818)**; and CR3 a 1.5-amp. **600-PRV** silicon diode **(1N4822)**. **K1**, a **4pdt**, 24-VDC relay, is available from Potter and Brumfield as type **KHP17D11** or Allied Control type **T163X-147**. The choke, **L1**, is a Stancor C-2688 with a **10-mH**, 12.5-amp rating. The rf choke. **L2**, is wound with fifteen turns of number 14 AWG **(1.6-mm)** stranded insulated wire on a Genelex **G49S/191** toroid. The power transformer. manufactured by Triad **(F-79U)**, supplies 24 **Vac** at 10 amps. The bridge rectifier, **CR1**, and the pass transistors, **Q1** through **Q3**, are mounted on finned heatsinks on the rear of the enclosure. For CRI, use Termalloy part 6169, and for the pass transistors part 6423.

The panel-mounted meters allow the operator to monitor output voltage and current. Voltage and shutdown level adjustment potentiometers on the circuit card are accessible through the front panel.

Power Circuit. The power circuit handles ac input, rectification, filtering, and dc-output functions (see **fig. 1**). The ac-input circuit contains ON/OFF switch S1, input fuse, and power transformer T1 as major functional components. The indicator assembly, connected across the primary of T1, illuminates when input power is applied. C1A and C1B provide noise and rf suppression at the primary of T1. The relay contacts of K1 are normally closed to allow current flow through the primary of T1. If the high voltage shutdown circuit activates at any time, the contacts of KI open to interrupt power to T1, thereby protecting the dc load from high output voltage.

CR1 rectifies the low-voltage ac supplied by the secondary winding of T1. The resultant pulsating dc is filtered by C2, C3, and L1. The filtered, unregulated dc supplies power to components on the circuit card and is applied to the collectors of series-pass transistors Q1-Q3.

The series-pass transistors perform the regulation function in the power circuit, under control of the regulator circuit. Under changing line and load conditions, the drive signal applied to the base-emitter circuit of transistors Q1-Q3 varies automatically. The changing drive signal serves to increase or decrease the collector-emitter voltage on Q1-Q3 as output voltage, output load, and ac input voltage levels dictate. The result of automatically varying the collector-emitter voltage on Q1-Q3 is to hold the emitter output voltage at a nearly constant level. This emitter output voltage is used as the regulated dc output.

The dc output circuits comprise metering, fusing, and suppression functions. A dc ammeter and voltmeter provide output current and voltage level monitoring capability. L2, wound on a toroid core, serves as an rf choke to decouple the power supply from any stray rf which may appear on the dc leads. C5A and C5B provide further rf suppression at the power supply output terminals. CR3 and C4 suppress voltage spikes generated by instantaneous load changes that could falsely trigger the high-voltage shutdown circuit.

Control Circuits. The control circuits, located on the circuit card (see **fig. 2**), comprise voltage regulation, current limiting, and high-voltage shutdown circuits. Various functions of the control circuits are described as follows.

Voltage regulation and current limiting functions are performed by U1, a monolithic IC voltage regulator. A voltage divider, R2, R3, and R4, is essentially connected across the power supply output. The setting of R3 primarily determines supply output voltage. R5 and R6 provide temperature compensation for U1, with C2 and C4 providing rf bypassing. Operating power for U1 is derived from the unregulated dc through isolation diode CR1, while C3 provides input power filtering. The output of U1 is applied to driver transistor Q1 through the normally closed contacts of K1. Q1 essentially increases the current-handling capability of U1. The output of Q1, as derived from the emitter circuit, drives the series-pass transistors in the power circuit.

The current limiting point is determined by the value of R6. When a voltage drop of 0.6 volt appears across R6, U1 begins current-limiting action.

The high-voltage shutdown circuit is composed of the comparator and relay-driver circuits. U2 is used as a comparator. The reference input of approximately 7.3 volts is derived through R12, CR2, and CR3. CR2 provides a stable, regulated voltage for the reference input applied to the inverting input of U2. CR3, in series with CR2, provides temperature compensation for the comparator circuit. The output voltage of the power supply is sampled in a voltage divider composed of R9, R10, and R11. A portion of the sampled output voltage is applied to the noninverting input of U2 through the wiper contact of R10. During normal operation, the reference voltage at the inverting input of U2 is higher than the sampled voltage at the noninverting input; therefore, the output of U2 is low and Q2 is biased into cutoff. In the event power suppiy output voltage rises to a sufficiently high level, the sampled voltage becomes higher than the reference voltage. The output of U2 changes to a high level, which biases Q2 into conduction. The resulting voltage developed across R17 triggers CR6 through steering diode CR7. When CR6 is triggered into conduction, K1 in the power circuit and K1 on the circuit card are energized and the following actions occur:

1. The output of U1 is disconnected from Q1, and the base of Q1 is grounded to drive it and the series-pass transistors into cutoff.

2. Power is applied to an LED, giving visual indication of a high-voltage shutdown condition.

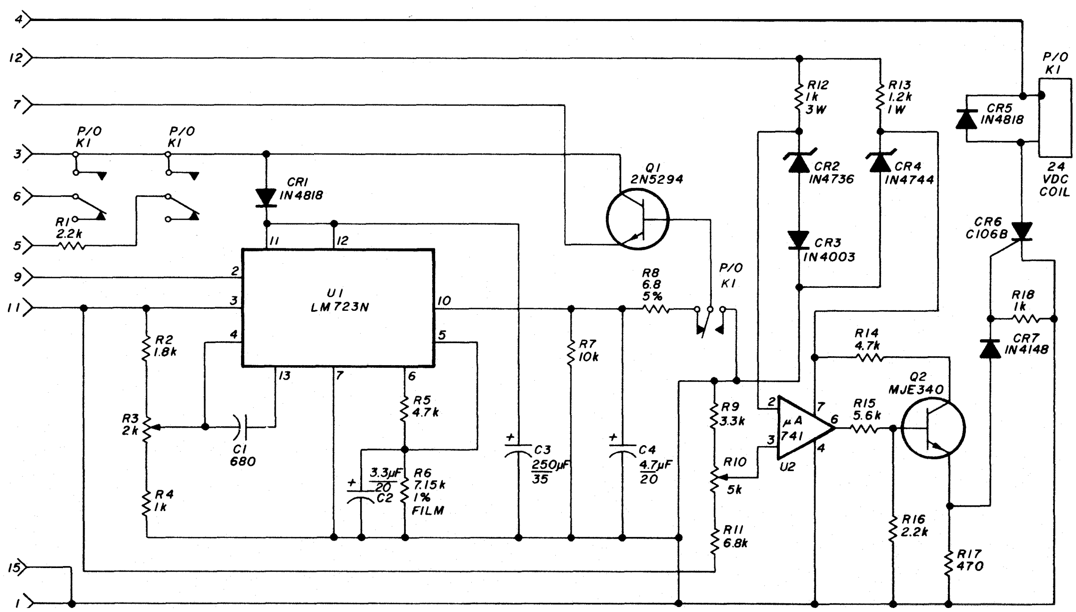
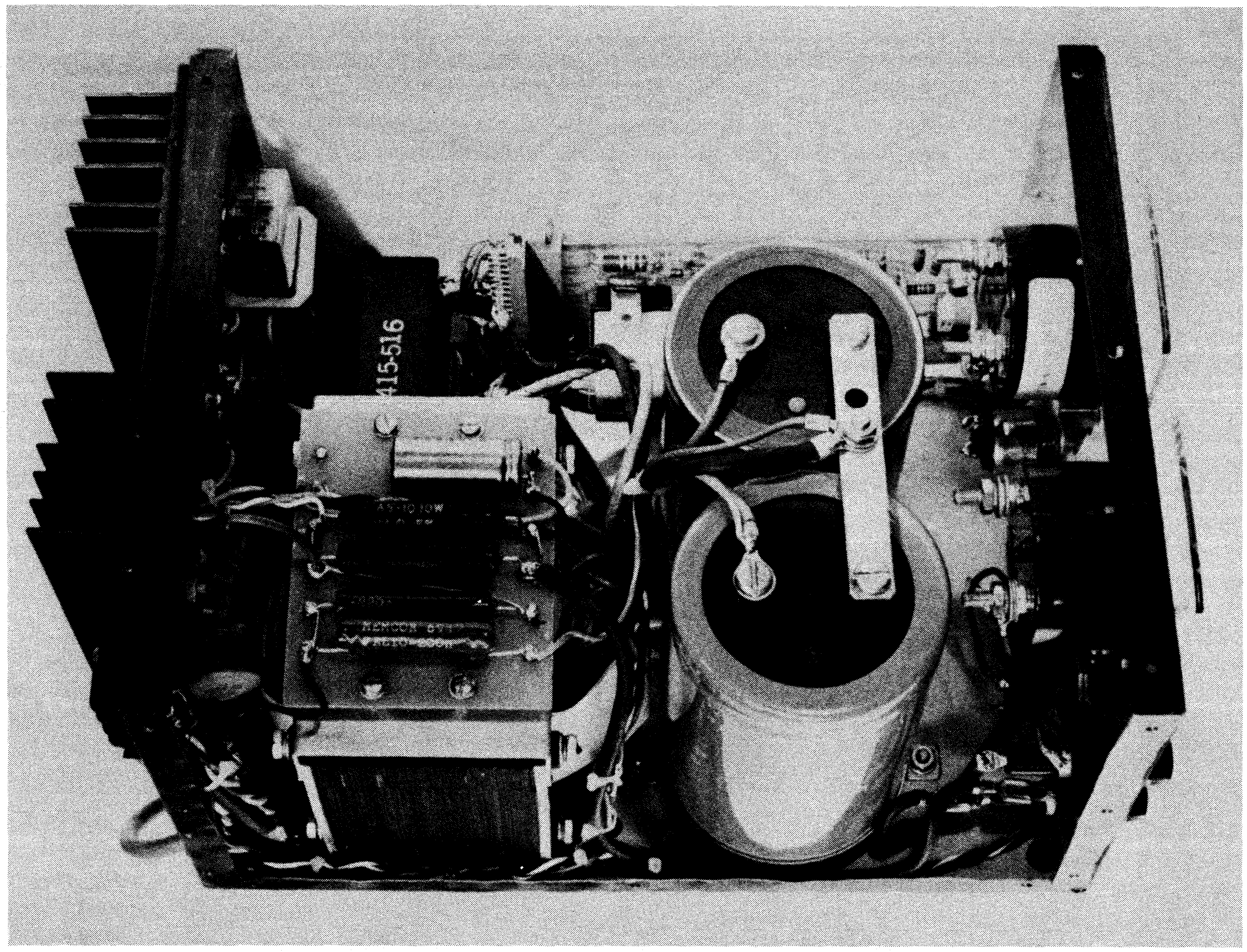


fig. 2. Schematic diagram of the voltage regulator and high-voltage shutdown circuits. CR2 is a 6.8-volt, 1-watt zener diode; CR3 is a 1N4003 (1 amp, 400 PRV); CR4 is another zener (15 volts, 1 watt); and CR6, the SCR, is rated at 4 amps, 200 PRV (C106B). The relay, K1, is the same type of relay used in the chassis-mounted components. The circuit card accepts a Potter and Brumfield relay socket. type 9KH2, for a KHP17D11 relay. Except as marked, all resistors are 10 per cent tolerance. ½ watt.



View of the author's power supply. The capped access holes, in the lower righthand corner of the front panel, allow access to the high voltage shutdown level and voltage adjustment controls mounted on the circuit board.

3. The primary circuit of the power transformer is opened to remove input power from the supply.

The approximate time from detection of a high voltage condition to supply shutdown is 3 milliseconds.

construction

Parts placement and general layout of the power supply are not critical; however, standard construction techniques and practices should be used when building the supply. Particular care should be taken in one area – lead lengths of rf bypass and decoupling components on the output terminals should be as short as possible. Additionally, meters should not be located near the power transformer or filter choke.

The particular construction technique chosen is at the discretion of the builder, but the power supply should be totally enclosed and shielded when assembly is complete. The supply shown in the accompanying photographs was built in a homemade cabinet to reduce construction cost.

Adequate ventilation must be provided for the power semiconductors, especially the series-pass transistors. The heatsinks specified in the parts list are adequate to handle device dissipation, but heatsinks of smaller size are not recommended. The series-pass transistor heatsink should be mounted outside of the power supply cabinet, on the rear panel, to ensure adequate ventilation. A thin coating of silicone grease, or other suitable thermal joint compound, must be applied to the mounting surfaces of all power semiconductors and insulating washers.

A circuit card pattern for the control circuits is shown in **fig.** 3. Pads are provided to make the circuit card compatible with a standard 15-contact edge connector; however, the circuit card can be permanently mounted and hardwired into the complete circuit.

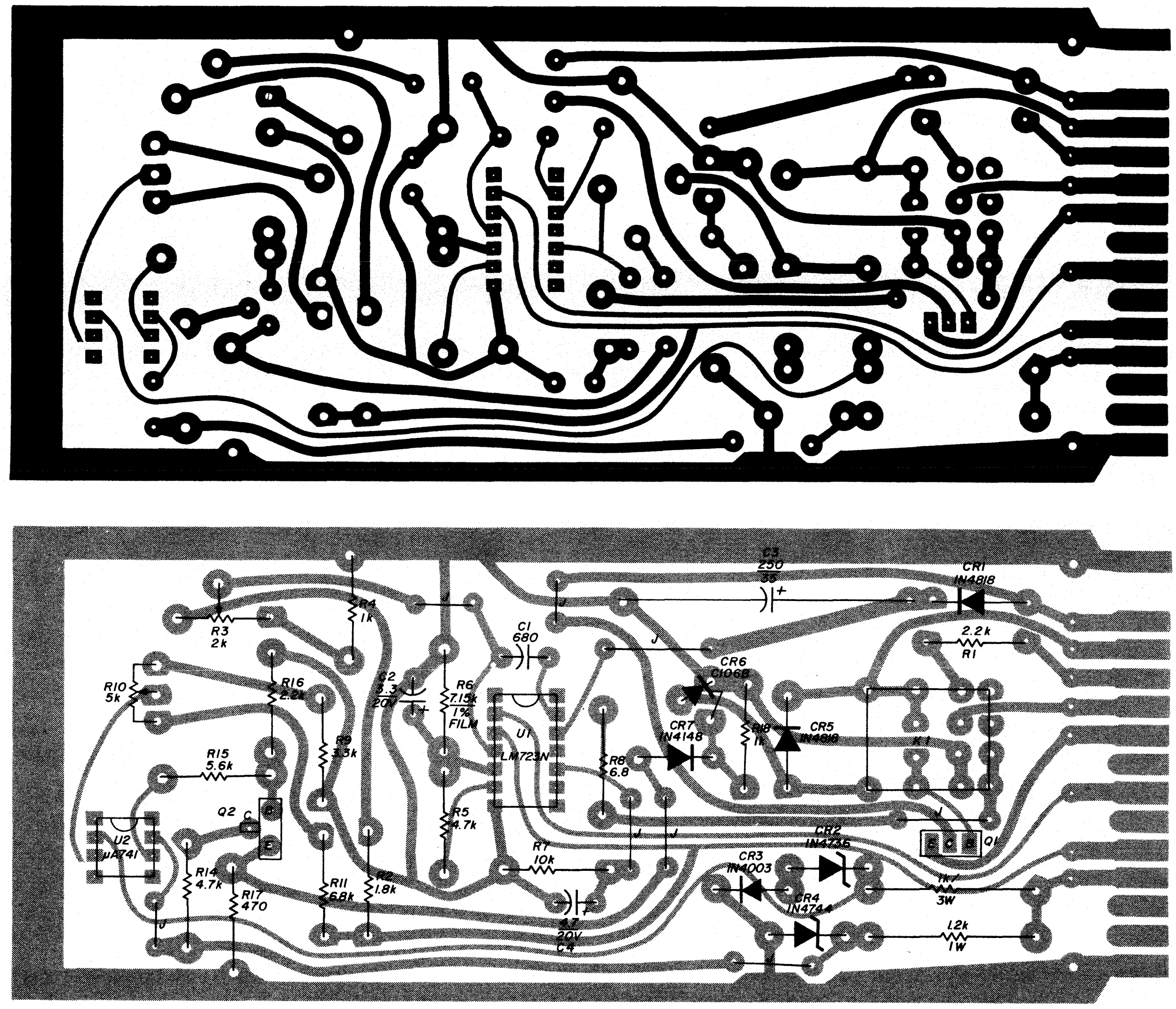


fig. 3. The circuit-board layout for the voltage regulator is shown above, with the parts placement diagram shown below. The edge pins will mate with a 15-contact circuit card connector, Amphenol 225-21531-101.

initial startup and adjustment

1. Before initial power supply startup, verify that all wiring is correct. If a tapped power transformer is used, ensure that the secondary voltage is no higher than 24 Vac.

2. On the regulator circuit card, adjust R3 and R10 fully clockwise.

3. Close S1 to start the power supply. Output voltage should be 10 Vdc or slightly less.

4. Adjust R3 on the regulator circuit card to 15 Vdc with no load at the output terminals. Turning the adjustment screw counterclockwise increases output voltage.

5. Slowly turn the adjustment screw of potentiometer R10 counterclockwise until the high voltage shutdown relays close. The LED should illuminate at this time, and supply output voltage should drop to zero.

6. Adjust R3 one turn clockwise, then press S2 to reset the shutdown circuits. Output voltage should now be present at the output terminals.

7. Slowly adjust R3 counterclockwise while noting the voltage at which the shutdown relays close. This voltage should be approximately 15 Vdc; if not, adjust R10 as necessary. The shutdown level can generally be adjusted to any desired level within the range of 13-16 Vdc. Individual operating requirements will determine the exact level.

8. Repeat Steps 6 and 7 if necessary to obtain the desired high voltage shutdown level.

9. Adjust R3 one or two turns clockwise, then reset the shutdown circuits by pressing S2.

10. Adjust R3 for the desired power supply output voltage. Adjustment is now complete.

ham radio

design considerations for linear amplifiers

Building a high-power, high-frequency linear amplifier and companion power supply is an interesting, challenging, and constructive project. And don't let anybody tell you that you can't do it! Many new hams approach equipment construction with great timidity. Be assured it isn't all that difficult. The toughest part of the work, in fact, is finding the necessary components. This is where flea markets, classified advertisements, surplus stores, and the junk box of neighboring Amateurs play an important part. You can find all the stuff you need, it just takes a little perseverance.

Before you begin bending metal, punching holes, and wiring, you should have the amplifier completely designed on paper, as outlined in the previous articles of this series. Once this task is done, you can make a parts list and start rounding up the components. You should also start thinking about the physical layout and assembly of the amplifier.

amplifier layout and assembly

Modern design indicates that the linear amplifier be enclosed in a metal cabinet, or box, that is shockproof, rf radiation-proof, compact, and easy to build. Many people build their linear amplifiers on a readily available aluminum chassis and then box up the chassis with aluminum sides and top to form a complete enclosure. This is not a bad idea. The cost is low and the chassis forms a platform and underchassis area that is hard to duplicate with simple tools. Once the enclosure is built, holes are drilled in it for leads and cables, control shafts, and for cooling air to enter and escape. Components within the box are positioned so that rf leads are short and direct and power wires are not coupled to the strong rf field within the box (see **fig. 1**). By paying attention to mechanical detail and armed with a knowledge of circuit design and a dose of common sense, the average Amateur can build a linear amplifier that looks good and works as well as the book says it should.

The object of using an all-metal amplifier enclosure is to keep the strong rf currents and subsequent harmonics within the box. Since these currents travel only on the surface of metal, the box can be made "electrically tight." Whenever a hole is made in the box, or a conductor brought into it, a leakage point is established through which rf energy can escape. It is important that these "rf holes" be reduced to a minimum in number and size, and that their effect upon circuit operation be controlled.

openings into the enclosure

Holes, leads, and shafts break the rf-tight amplifier box. Large holes for ventilation can be used without harm, provided they are screened so that air can enter and leave but rf energy cannot escape. Perforated metal sheet, having many closely spaced holes, is the best screening material to place over the openings. Copper wire window screen is not as effective because of wire corrosion which produces a film of insulating oxidation between the individual wires at the crossover points.

If a perforated sheet is to be used, it may be made by drilling lots of holes in the enclosure wall. Or the hole pattern can be drilled in an auxiliary plate placed over the ventilation hole. If such a plate is used, it should be bolted or riveted to the enclosure with a bolt-to-bolt spacing of about 2.5 cm (1 inch) so that rf energy cannot leak out through the crack between the surfaces. Mating surfaces between the metals should be clean and free of paint **(fig. 2)**. A screened ventilation opening should be about three times the size of an equivalent unscreened opening, since the screening material reduces the area of air passage.

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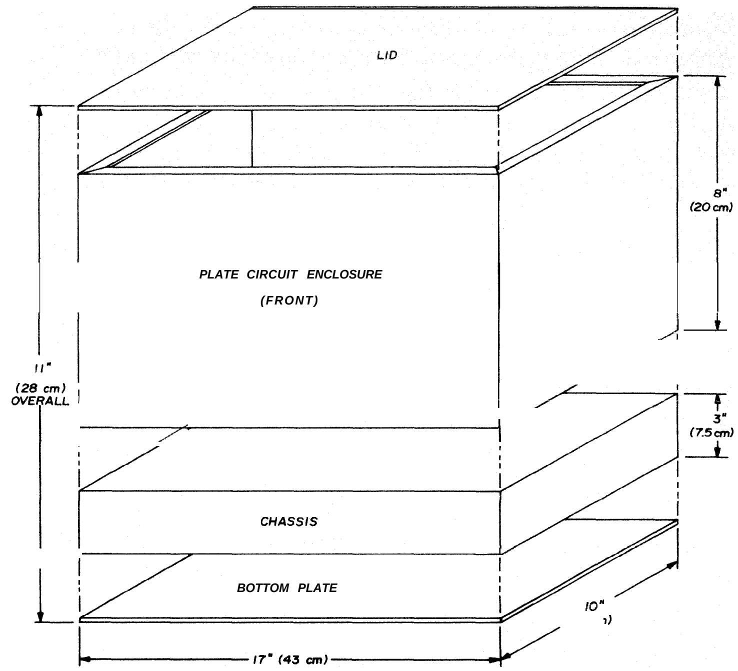


fig. 1. A representative amplifier enclosure. Basic unit is an aluminum chassis with bottom plate. The plate circuit enclosure is made of aluminum stock. For ease in assembly, aluminum channel angle stock is pop-riveted around outer edge of chassis to mate with the bottom of plate circuit enclosure. Angle stock is run up each corner and riveted to the four plates. Additional angle stock runs around the inside edge of the top of the enclosure and is tapped for 6-32 (M3.5) screws, which hold the lid on. Lid may be made of perforated metal for ventilation or may be modified from a solid aluminum sheet as discussed in the text. Additional angle stock may be required to hold bottom plate in place and to make the under-chassis area relatively air-tight. Blower to cool the tubes is mounted on the rear apron of the chassis. The completed enclosure is mounted behind a relay rack panel for appearance.

Control shafts passing into an rf-tight box should be made either of phenolic-insulating rod or of metal, grounded at the point of entry by means of a spring contact (**fig.** 3).

Long, narrow slots in the enclosure should be avoided, or else shunted with a ground strap every few inches; otherwise the opening tends to act as a "slot antenna" through which harmonic energy can readily pass — more easily than through a much larger circular hole, in fact.

Meters mounted in a wall of the shielded box pose a problem, as they are a source of prolific rf leakage. Unless the body of the meter is shielded and the leads well bypassed, it is more prudent and less time consuming to mount the meters outside the enclosure and to filter the meter leads running into the box.

pass-through leads

Careful attention must be paid to power and meter

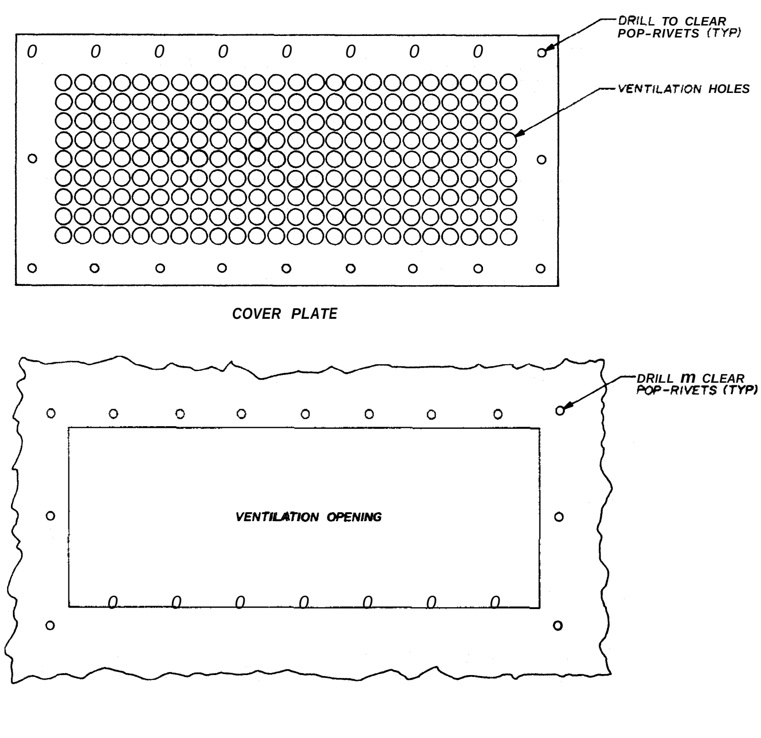


fig. 2. Ventilation holes are cut in sheet aluminum by means of a nibbling tool. Cover plate is cut slightly larger and drilled for ventilation holes. Plate and sheet aluminum are then drilled together for holes to place pop-rivets. If it is necessary to remove the cover plate for insertion of tubes, the plate may be held in position with sheet metal screws or by 6-32 (M3.5) nuts and bolts (provided assembly is such that you can get your hand inside the enclosure to hold the nut in place). Most side ventilation plates are fixed in position; top cover plates are removable.

leads entering and leaving the rf-tight box. Harmonic currents inside the box can easily flow out of the enclosure on these leads or even on the outer shield of a coaxial line if the shield is not properly grounded at the point of entry **(fig. 4)**. Unshielded leads entering the box must be carefully bypassed and filtered at the point of entry to prevent rf energy from escaping from the box and flowing down the leads. A combination of bypass capacitors and small filter inductors will close off this escape route. The inductor must have ample capacity to carry the current flowing in the lead. Feedthrough-style capacitors are often used in low-voltage power and metering leads.

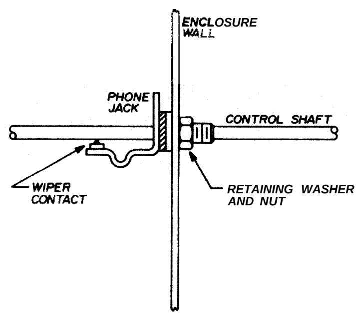


fig. 3. A single-circuit phone jack makes a good grounding device for a 6.5-mm (¼-inch) diameter shaft. The jack is mounted in the shaft hole, which is drilled out to accept the jack. The wiper contact of the jack rides on the shaft as it is rotated. The contact arm of the jack is grounded to the enclosure wall. Jack is positioned so that wiper arm is inside amplifier box.

amplifier wiring within the enclosure

Wiring within the rf-tight box can couple to rf energy because of the storing field within the box. Any lead in the box can pick up fundamental and harmonic energy and feed it outside the enclosure **(fig. 5)**. On the other hand, the lead can pick up rf energy from an outside source (your exciter, for example) and leak it into the box causing amplifier instability. The solution for this problem is to bypass or filter all internal power and control leads at each end, dress them close to the chassis, and keep them physically remote from areas of high rf energy.

All these precautions may seem more complicated and time-consuming than they really are. Unfortunately, most circuit diagrams leave off much of the important rf bypassing circuitry since it tends to clutter up the diagram; its existence may be only briefly mentioned in the text. And the filter circuitry is often left out of commercially produced units as a cost-cutting measure.

When you build your own amplifier, you can afford to take the time and do things the right way. Always remember that holes, shafts, and leads are sources of rf leakage from an rf-tight enclosure and, unless protected, are a direct invitation to TVI, harmonic radiation, and amplifier instability. Sadly enough, many modern amplifiers on the market look like they're in an rf-tight enclosure, but, in reality, they are only sitting in an attractive dust cover.

practical amplifier layout

A simple to understand and practical parts layout for a representative high-frequency linear amplifier using two 3-5002 tubes is shown in **fig. 6.** The layout can be adapted to other tubes. The assembly consists of an aluminum box made up from a standard chassis. A bottom plate pressurizes the underside of the chassis and a blower is mounted on the rear apron of the chassis. Air is introduced under the chassis and is expelled through the tube sockets and air chimneys. The heated air from the tubes escapes through the perforated top and side areas of the plate circuit compartment.

The meter and control circuits are placed outside the shielded enclosure. Wiring for these circuits is not critical, and is done with 600-volt insulation hookup wire. High voltage wiring is done with test-

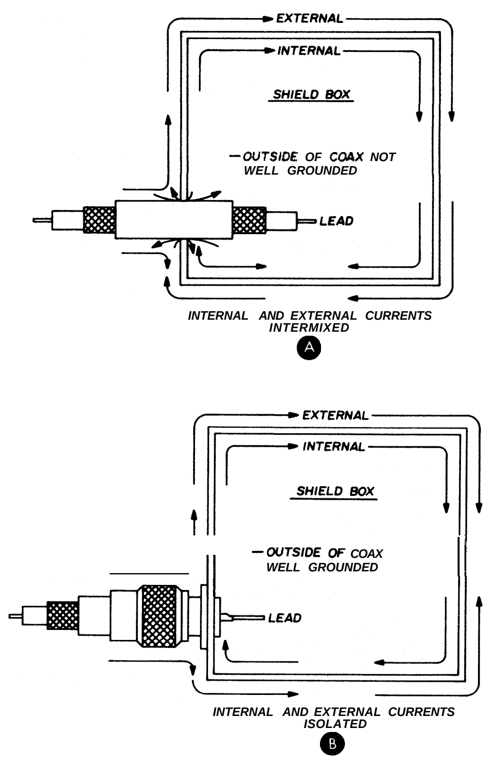


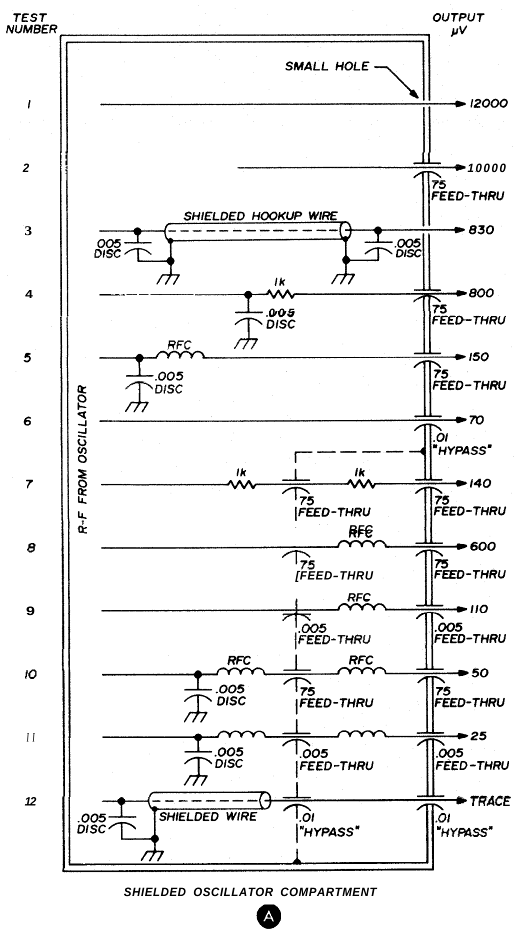
fig. 4. Improper termination of coaxial line can destroy effectiveness of the shield (A). Rf currents within the enclosure can escape via the outside shield of the line as it passes through the hole. Properly grounding the shield of the coaxial line to the box (B) ensures isolation of currents within the box. Rf currents outside the box are also prevented from entering the box.

prod wire of the type used for instrument test leads (10-kV insulating rating) or equivalent high-voltage cable. TV-type capacitors are used for lead filtering (fig. *7).*

Low voltage leads enter the amplifier enclosure via 1-kV feedthrough capacitors, which are also shunted at the point of entry with a larger value of capacitance to suppress low-frequency rf energy and transients which can pass through most feedthrough capacitors with little attenuation. A simple home-made rf choke and bypass capacitor are placed on each lead inside the enclosure. Note that all capacitors used on the 120-volt ac power line should be rated at 1.6 kV in accordance with the Electrical Underwriter's Code. Don't use run-of-the-mill disc bypass capacitors on the power line, as it is a source of random voltage transients which can easily puncture the standard 600-volt-working capacitor.

Power leads from the panel controls to the terminal strip on the rear of the amplifier must either pass through the box or go around it. It is easy to pass through the enclosure without breaking the rf seal with a short section of 1.3-cm (0.5-inch) diameter thin-wall electrical conduit with wall fittings on each end of the section (fig. **8).** An rf tight passageway through the amplifier for various wires is formed by the conduit. Coaxial fittings are used for the input and output rf connections and are mounted to the wall of the box.

The mouth of the blower is covered with a small piece of copper window screen. While not the best material, this screen doesn't reduce the air flow as much as a perforated metal sheet would do.



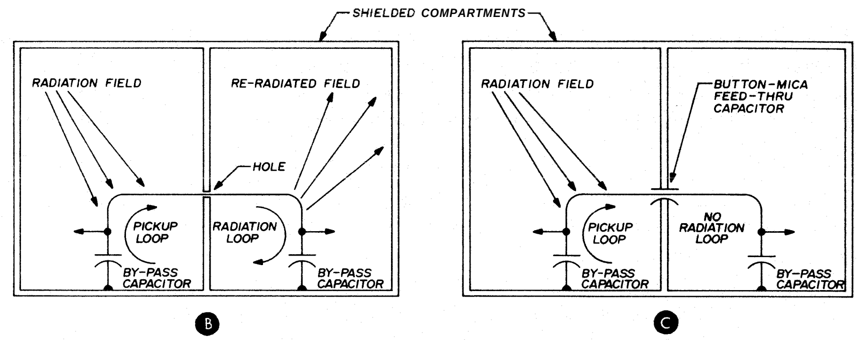


fig. 5. Tests of lead-filtering techniques (A). A signal generator having an output of 12,000 µV was placed in a metal box and rf leakage via various paths was measured. Very complex shielding was required to remove the last vestige of signal from the power lead. A combination of rf choke and capacitors, such as tests 8 or 9, does the job in adequate fashion. A very effective filter (not shown here), consists of a 0.001-µF feedthrough capacitor with a series rf choke. Both ends of the choke are additionally bypassed with a 0.01-µF disc capacitor. Energy can be conducted from one area to another as this test shows. Lead-through hole in partition (B) conducts energy from one compartment to the other. Proper bypassing (C) attenuates leakage. (Courtesy Radio Publications, Inc.)

While all this hum-drum filtering and bypassing might seem like overkill, it is the ***only*** way to achieve an amplifier that is rf radiation proof and one that keeps rf energy where it belongs, The rf energy leaves the box only via the output circuit where harmonics can be suppressed by means of a suitable lowpass filter before they reach the antenna. Without the filtering and bypassing, the harmonics suppressed in the antenna circuit would pass down the power leads or be radiated directly from the amplifier circuitry.

B-plus safety switch

A quick way to kill yourself is to remove the amplifier cover and fiddle around inside the box when the high voltage is turned on. Even the best of us might forget to turn things off and disconnect the amplifier from the supply before work is performed. A B-plus shorting switch will pay big dividends in operator longevity. It is simple to make (fig. **9).** The shorting ring is made of spring brass and is depressed when the amplifier lid is in place. When the lid is removed, pressure is taken off the shorting ring and it makes a direct contact between the high voltage circuit and the chassis. This short circuit results in a blown line fuse if the amplifier is inadvertently turned on when

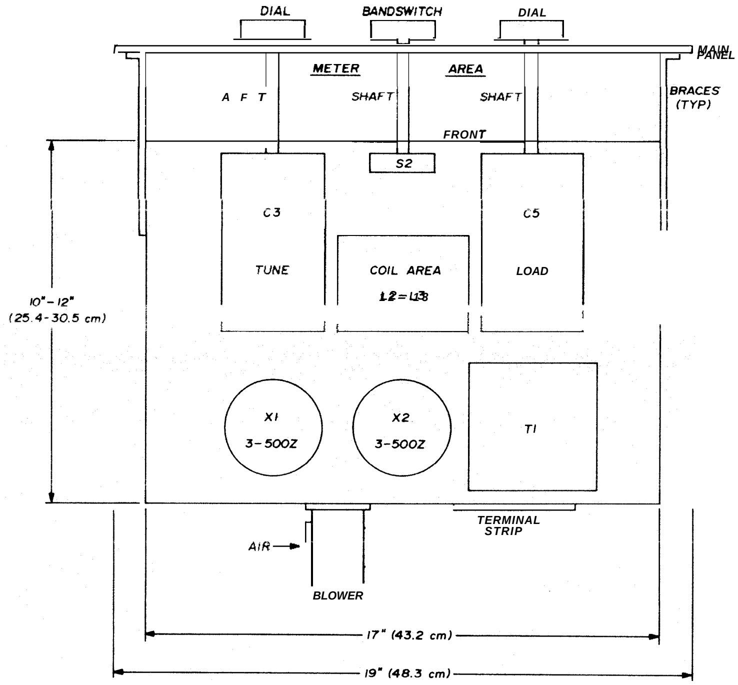


fig. 6. A practical layout for a linear amplifier using two 3-5002 tubes. This is representative of a layout using any popular tube or tubes available for the Amateur service. A 43-cm (17-inch) chassis is used, with depth chosen to allow proper placement of components. Tuning and loading capacitors are mounted symmetrically on the main panel with the plate bandswitch between them. Panel meters are placed across lower portion of the panel. An area for the plate coils lies between the two capacitors, immediately behind the bandswitch. The **3-500Z** tubes, air system sockets, and chimneys are near one rear corner of the chassis with the air blower placed on the rear apron between the sockets. To the side of the tubes is the filament transformer. To reduce transformer heating caused by infra-red radiation from the tubes, the transformer (which is normally black in color) is given a coat of white stove enamel. This reflects the heat from the tubes and reduces transformer operating temperature. The **fixed**-tuned cathode input circuit and bandswitch are located beneath the chassis, and the switch control shaft is brought out to the panel. Some Amateurs gang the input and output circuit band-change switches, but this is not necessary. The bottom of the amplifier is sealed with a metal plate, and the top area is made up of perforated aluminum sheets to permit ample tube ventilation.

the lid is removed. It also makes sure the filter capacitors are discharged before hands can be poked inside the amplifier.

metering circuits

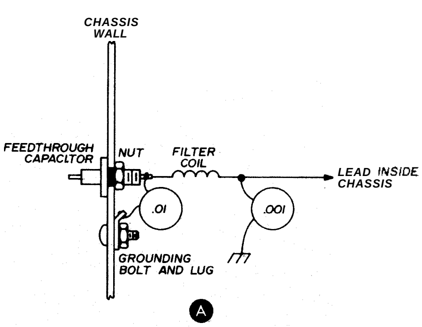
When you have power tubes in your linear amplifier that may cost upwards of $100, it is a smart and thrifty idea to take good care of them. As far as metering goes, it is wise to monitor both grid and plate current (and screen current if a tetrode tube is used) plus filament voltage. And a plate voltmeter is a handy thing and necessary if you run close to the maximum power level.

The meters are mounted outside the rf-tight box to remove them from the strong rf field of the amplifier. A single meter may be used as a matter of thrift to measure either grid or plate current if an appropriate switching circuit is employed, such as shown in **fig. 10,** where one meter does the work of two.

For economy and simplicity, a 0-1 mA dc meter is used. The scale will read 0 to 100 mA for grid current and 0 to 1000 mA for plate current. The scale need not be re-inked, since the user merely adds zeros to the reading to get the exact current value.

The meter is converted into a simple voltmeter circuit by a series-connected resistor. This voltmeter then reads the voltage drop across a shunt resistor placed in the circuit to be monitored. The whole circuit is inexpensive, accurate, and easy to make up. It does not require precision resistors – inexpensive one-percent metal film resistors will do (or carbon resistors in pinch).

As an example, suppose a 3.9-kilohm series resistor (a standard value) is used. The 0-1 mA meter is now turned into a voltmeter which reads 3.9 volts full



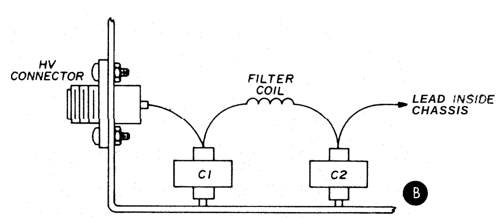


fig. 7. A simple filter circuit **(A)** for low voltage leads is made up of a 1000-volt feedthrough capacitor (**Nytronics CP09A3** style, **0.01-μF,** with case grounded and mounting stud). These, or similar capacitors can often be found in surplus stores. The disc capacitors are 1 **kV (Sprague 5GA-D10,** or equivalent). The filter coil is ten turns of no. 16 AWG (1.3 **mm),** 1.3-cm **(½-inch)** diameter spaced to 3.2 cm **(1¼** inches) long. The coil and capacitors are placed inside the chassis. High-voltage filter circuit (B). A high-voltage chassis connector (**Millen** or equivalent) is used. The capacitors are **500-pF, 10-kV** TV-type, with stud mounts. Some Amateurs use a high-voltage coaxial connector for the B-plus lead and run the high voltage in RG-8/U coaxial cable with the outer sheath grounded as a safety factor.

scale. All that is necessary now is to design a shunt which will produce a 3.9 volt drop across it at the desired full scale reading of the meter. Let's say we want 100 mA (0.1 amp) full-scale deflection for grid-current measurement. The shunt resistor (by Ohm's law) is:

$$\text{Shunt resistor (ohms)} = \frac{E}{I} = \frac{3.9}{0.1} = 39\ ohms$$

This, also, is a standard resistance value. If you want

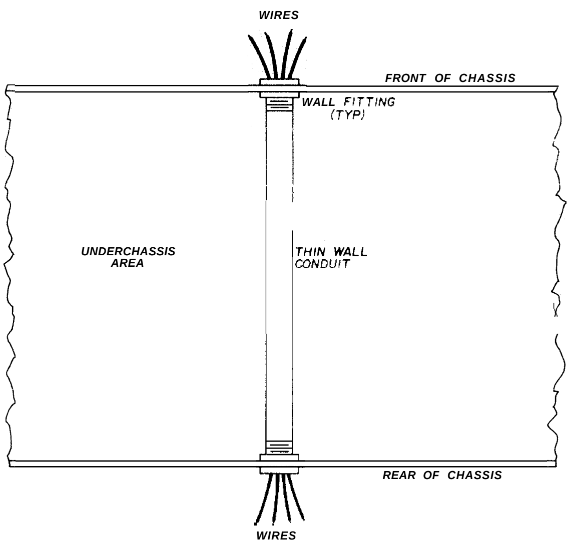


fig. 8. Wires can be passed through an underchassis area by using a short length of thin wall electrical conduit as a passageway. Conduit is attached to the chassis walls by means of metal wall fittings. Conduit and fittings can be purchased at electrical contractor or large home improvement store.

to read plate current at 1000 mA (1 amp) full-scale, the appropriate shunt resistor is:

$$\text{Shunt resistor (ohms)} = \frac{3.9}{I}$$

$= 3.9\ ohms$ (a standard resistance value)

Simple, isn't it? No expensive precision resistors are needed and everything is figured out by simple mathematics. Other full-scale meter readings can be worked out by changing the value of the shunt resistor.

inexpensive perforated metal sheet

A good way to make a ventilated rf-tight metal box is to use perforated aluminum sheet stock found in many hardware stores and home improvement centers. Ideally, the holes should be small and closely spaced so that it seems as if there is more open space than solid metal. As an alternative, you can make your own perforated sheet from solid aluminum

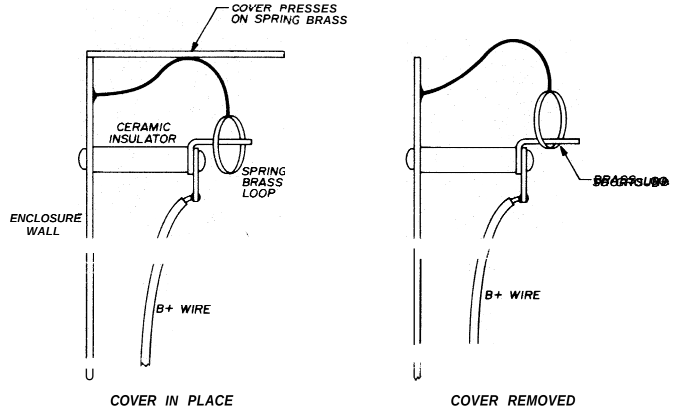


fig. 9. Inexpensive high-voltage shorting switch. The B-plus lead is connected to a ceramic standoff insulator. A short length of brass rod projects out from the insulator. A shorting ring made of spring brass loop encircles the rod as shown in the left illustration. When the lid is in place, the loop is centered around the rod. When the cover is removed the spring brass straightens out and the loop is offset, shorting the B-plus wire to ground.

sheet and an electric drill. The trick is to make up a drilling jig out of a small steel plate (**fig. 11**). This sounds like doing it the hard way, but once the jig is made, it can be used rapidly and can be reused time and time again. It is a worthwhile addition to the home workshop. The jig is held in position on the sheet with a pair of C-clamps and the holes easily and quickly drilled with an electric drill to the pattern you wish.

amplifier layout

If you look through the various Amateur magazines and handbooks (particularly those of the pre-1970 era) you'll see plenty of homebrew linear amplifiers. They bear a remarkable similarity as far as layout goes. Indeed, so do most of the linear amplifiers currently on the market. Time spent in seeing how others solved their problems is a big asset when

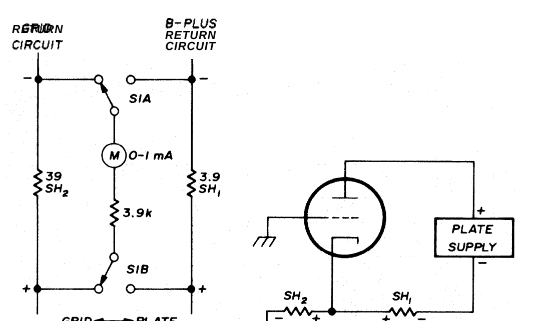


fig. 10. Meter-switching circuit (**A**), using an inexpensive 0-1 **mA** dc meter to measure either grid or plate current. In the grid position, the full-scale reading of the meter is 100 **mA**. In the plate position, the full-scale reading is 1000 **mA**. A simplified amplifier circuit (**B**) showing the dc current paths of the meter circuit. Note that the B-plus supply is "above ground" by virtue of the meter shunts.

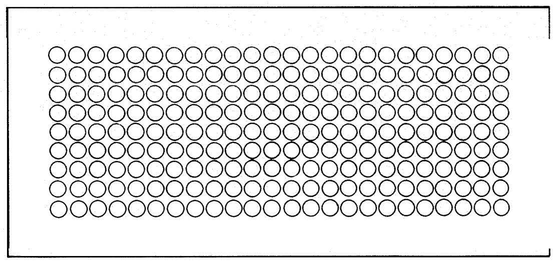


fig. 11. A drilling jig made of sheet metal 3.2-mm (118-inch1 thick is handy for making your own perforated stock for areas requiring ventilation. Jig is about 20 cm (8 inches1 long and 7.5 cm (3 inchesl wide. Holes should be about 6.5 mm (1/4 inch) in diameter.

it comes to laying out the components for your own linear amplifier.

You should lay the parts out on the chassis before you start drilling holes and bending metal. Some Amateurs make a cardboard mock-up of their amplifier and slide the components around in a three dimensional layout to make sure that one part does not mechanically interfere with another and that the dials fall on the panel in a symmetrical pattern.

Once general parts placement has been ascertained, the sides, back, bottom, and top of the enclosure can be laid out and cut from sheet aluminum. The finished parts can be held together by means of bent-over edges on the sheets or by means of aluminum angle stock cut to fit. Some people use nuts and bolts to hold everything together, while others use sheet metal screws or pop-rivets. The top of the enclosure is held in position with removable screws so that it can be taken off for tube installation.

The amplifier box is supported from the panel by spacer rods cut long enough to leave space for the meters between panel and amplifier. Shaft extensions can be used to couple the panel controls to the control shafts extending from the amplifier wall.

If your assembly is completely knock-down and the chassis plate is replaceable, the amplifier circuitry and tube complement can be changed at will while still retaining the panel, circuitry, and main body of the amplifier. But you'll never need this, since hams rarely rebuild their equipment!

recommended reading

The new 21st edition of the Radio Handbook is now available and has a greatly expanded section on design and construction of linear amplifiers, Photographs show many different designs using popular power tubes. The new Radio Handbook is available from Ham Radio's Bookstore. Also read "A Beginner's 50 Watt Rig" by Bill Wildenhein. W8YFB, in the July and August, 1978, issues of Ham Radio Horizons, This is a goldmine of design, construction, and layout information. You should also read "Custom Design and Construction Techniques for Linear Amplifiers Using the 8877," by Merle Parten, K6DC, in the September, 1971, issue of *QST*. A reprint of this article can be obtained at no cost from the Amateur Service Dept., EIMAC, Varian Division, 301 Industrial Way, San Carlos, California 94070. Ask for bulletin AS-45.

ham radio

ITAL TECHNIQUES

counters and weights

Counting in the digital sense is really a sequence of flip-flop state changes. The sequence repeats and the number of state changes required for one sequence is termed the division ratio. Thinking in terms of state changes is a clue in understanding counter operation. To make sense of certain flip-flop counters, the bit weight system is used.

states and weights

Each flip-flop in an array represents a bit of data. A counter always has the same number of bits, but each bit may be 1 or 0 depending on the sequence. A chain, or cascade, of flip-flops will have an orderly sequence of bit state changes, but it is still difficult to interpret the changes into decimal notation.

Assume four flip-flops in cascade. The maximum number of states is sixteen. Weights are assigned to each bit. The least-significant bit, or LSB, will have a weight of one; it represents the input flip-flop, which changes the fastest. The next bit has a weight of two; the next four. The most-significant bit, or MSB, has a weight of eight. Mathematically, the bit weight is 2^n, where n is the bit significance.

Converting binary states to decimal notation involves adding up the weights of any bit that is a 1. Forget any 0 bits. **Table 1** has four-bit binary states with equivalent decimal weights. Note that the maximum decimal number is fifteen. What happened to sixteen? Simple. An all-zero binary state is decimal zero, so all sixteen are accounted for. A decimal sixteen would require five binary bits with the MSB at 1, the remainder at 0.

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Table 1 also gives hexadecimal notation. This is common in microprocessor state designation, and some four-stage counter packages are called hexadecimal counters.

simple counter chain

Fig. 1 has three negative-edge, clocked, JK flip-flops in cascade. J and K inputs of all are tied high, so each stage divides by two. With three in cascade the total division, or count, is eight.

This simple cascade is a ripple-through counter since each successive stage state change is dependent on the previous stage propagation delay. A ripple-through counter should not be used at high speed for selecting a particular state.

Suppose you wanted to select a decimal 2 state. The flip-flop states would be $\bar{A}B\bar{C}$ (A and C low, B high). For a short period of time this same state would occur after the fourth negative clock edge after A had toggled low but B was still high. It is a very short time, but the select gate might pass this "glitch."

A solution is to make the counter synchronous using anticipated carry. Carry in counters is the state change output, that can cause the next stage to toggle. You can see that carry anticipation is possible by examining all previous flip-flop states and the clock before a toggle occurs; they are all high.

table 1. Four-bit binary states with decimal weights.

	binary states				hexadecimal
decimal	MSB			LSB	notation
0	0	0	0	0	0
1	0	0	0	1	1
2	0	0	1	0	2
3	0	0	1	1	3
4	0	1	0	0	4
5	0	1	0	1	5
6	0	1	1	0	6
7	0	1	1	1	7
8	1	0	0	0	8
9	1	0	0	1	9
10	1	0	1	0	A
11	1	0	1	1	B
12	1	1	0	0	C
13	1	1	0	1	D
14	1	1	1	0	E
15	1	1	1	1	F

Synchronous counter modification is shown in **fig.** 2. Additional AND gates set up the next flip-flop stage so that all change state at the same time. Carry out will have the same width as the high clock state but occurs only every eight clock periods.

Difference in state change time is caused only by differential flip-flop propagation delay (quite small) or the AND gate delay. The latter may be compensated for by adding inverters to the first stage outputs for any state select gating.

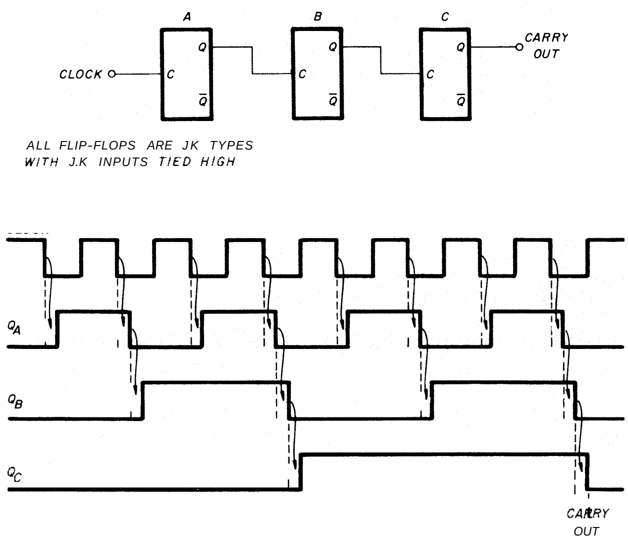


fig. 1. Cascade of three JK flip-flops with waveforms.

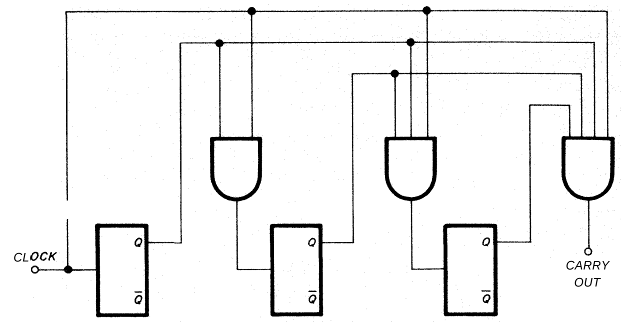


fig. 2. Synchronous three-stage binary counter using anticipatory carry.

decade counting

A minimum of four stages are necessary for division by ten. JK flip-flops can be used, and the control inputs will enable a binary state sequence from decimal 0 through decimal 9. This type of counter is called a binary-coded-decimal, or BCD, and is shown in **fig.** 3.

This is a cascade of divide-by-two (stage A) and a divide-by-five (last three stages). Ten is divisible by two, so state feedback isn't required for the first stage. Note that $\bar{Q}_D$ is made to both J and K inputs of stage B; as long as it is high (Q_D low), B will toggle on every negative edge of Q_A. Stage D will not toggle since G1 holds its J input low until a decimal 7 is reached.

At the decimal 7 state, both J and K of stage D are

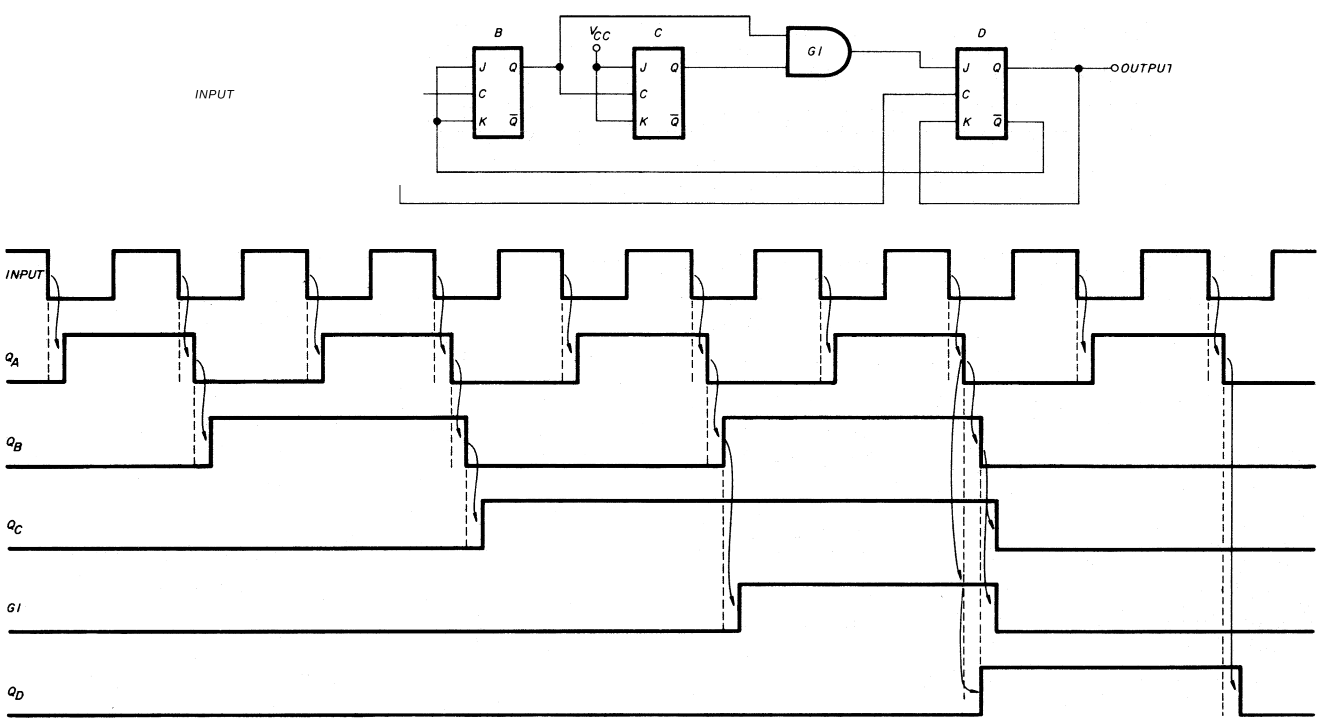
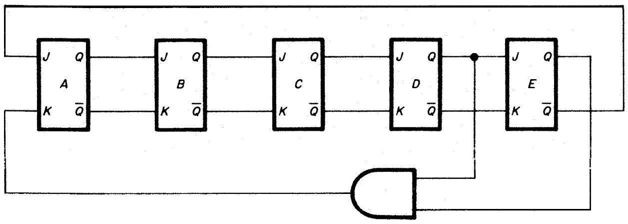


fig. 3. Binary-coded-decimal decade counter



INPUT IS MADE TO ALL C̄ FLIP-FLOP CLOCK INPUTS

TRUTH TABLE

INPUT	A	B	C	D	E	GATE
	Q OUTPUTS					
0	0	0	0	0	0	0
1	1	0	0	0	0	0
2	1	1	0	0	0	0
3	1	1	1	0	0	0
4	1	1	1	1	0	0
5	1	1	1	1	1	1
6	0	1	1	1	1	1
7	0	0	1	1	1	1
8	0	0	0	1	1	1
9	0	0	0	0	1	0
10	0	0	0	0	0	0

fig. 4. Five-stage, self-correcting decade ring counter.

high; D is set up to toggle. It did not toggle when G1 went high, since the control input must be present before a clock arrival. The eighth clock will make A, B, and C all low. D goes high from Q_A, since G1 has set up the toggle condition. Stage B is now inhibited from its J and K inputs made low from $\overline{Q}_D$.

The tenth clock will toggle stage D again, making Q_D low. This action removes B's inhibit, but B does not toggle, since the inhibit was still there when Q_A went low. If this is confusing it won't hurt to review the JK rules given in reference 1.

ring counters

These are shift registers modified by output-to-input state feedback. A shift register is a cascade of JK or D flip-flops with the Q and $\overline{Q}$ of one stage feeding the J and K inputs of the next (Q to D only for D flip-flops). The clock is common to all stages, and any input to the first stage will shift through all stages at each clock edge.

A divide-by-ten ring counter is shown in **fig. 4** with the state truth table. It's common in CMOS and is sometimes called a "Johnson," or "switched-tail," counter. The latter name is from inverted output state feedback. This one is called self-correcting from the AND gate connection.

At power-on you cannot be sure that all flip-flops reach one of the desired ten states. With five stages, the maximum number of states is 32 (2^5), so the counter must be able to shift out of an undesired state. One such pattern is 01101, and you can try it out on scratchpaper with and without and AND gate.*

Ring counters are inherently synchronous and are a bit easier to decode into specific states. **Fig. 4** will decode all ten states into decimal using only two-input gates for each state. A disadvantage is the number of flip-flops, which must be one-half of the count.

odd-modulo ring counters

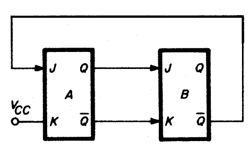
Modulus is another name for division ratio. Odd-modulo ring counters may have advantages over cascades with different state feedback. Divisions of three, five, seven, and nine don't require extra gating for correction. **Fig. 5** shows two counters with state tables.

A modulo-7 counter is created by adding a flip-flop before stage A. A modulo-9 adds two flip-flops. In both cases the Q output is connected to the next J, $\overline{Q}$ to K. The last three stages in divisions of five, seven, and nine will go through the modulo-5 pattern, skipping the 001 state between 011 and 000. All are self-correcting.

ssb quadrature counter

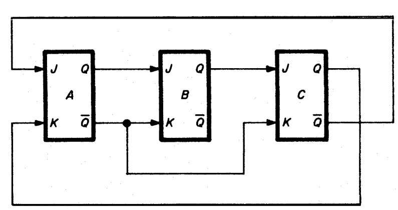
Fig. 6 is a variation of a ring counter and has been used in several phasing-type, single-sideband designs. It's made from one high-speed, dual-D package and uses all four binary states. Quadrature (90-degree)phase is maintained over a wide frequency range.

A disadvantage is that the VFO (clock input) must be four times the output frequency, and differential



DIVIDE BY THREE

INPUT	A	B
0	0	0
1	1	0
2	0	1
3	0	0



DIVIDE BY FIVE

INPUT	A	B	C
0	0	0	0
1	1	0	0
2	1	1	0
3	1	1	1
4	0	1	1
5	0	0	0

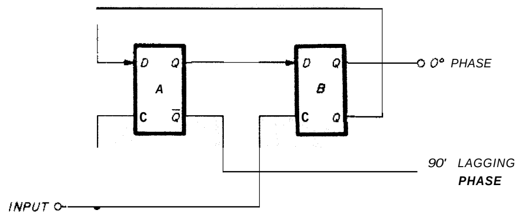
INPUT IS MADE TO ALL C̄ FLIP–FLOP CLOCK INPUTS

TRUTH TABLE

INPUT	A	B	C	D	E	GATE
	Q OUTPUTS					
0	0	0	0	0	0	0
1	1	0	0	0	0	0
2	1	1	0	0	0	0
3	1	1	1	0	0	0
4	1	1	1	1	0	0
5	1	1	1	1	1	1
6	0	1	1	1	1	1
7	0	0	1	1	1	1
8	0	0	0	1	1	1
9	0	0	0	0	1	0
10	0	0	0	0	0	0

fig. 5. Self-correcting odd-modulo ring counters.

*See appendix



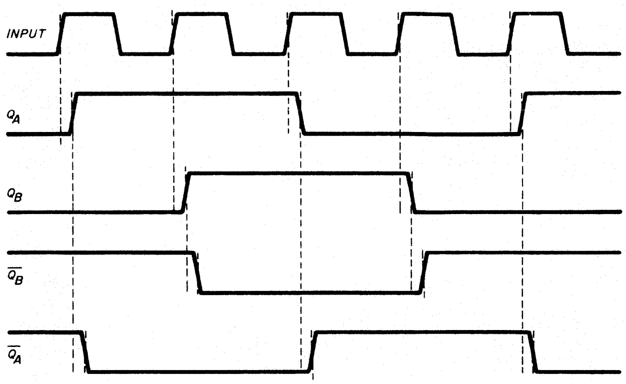


fig. 6. Two-stage SSB quadrature counter.

delay may limit the 90-degree difference. Differential delay in one package is never specified. It must be low since one degree of phase error in only 0.397 nanoseconds at 7 MHz. Output loading should be equal to minimize differential delay, and mixer output filtering must be used to eliminate harmonics.

direct set and clear

Both CMOS and TTL packages have a variety of different flip-flop and counter arrangements. Direct set (or preset) and clear (or reset) may be separate or common or in a combination. Specification sheet data should be studied carefully for each package to make certain all functions and pinouts are understood.

appendix

Starting the decade ring counter without the gate (Q_E directly to K_A) at a state of 01101 will produce this sequence: 00110, 10011, 01001, 00100, 10010, 11001, 01100, 10110, 11011, and back to 01101. It never arrives at a desired state. Adding the gate breaks the sequence at 11001. Both J and K inputs of the first stage are now 0 and it holds at a 1 state; the remaining stages shift through for the next state of 11100, a desired state. Remaining states are in the desired sequence. The worst glitch state is 01100 with the gate. It will go through 10110, 11011, then 01101 for the pattern given above.

reference

1. Leonard Anderson, "Digital Circuits — Propagation Delay and Flip-Flops," *ham radio,* March, *1979,* page 82.

ham radio

the weekender

broadband power-tracking VSWR bridge

One of the problems that most hams have experienced with commercial power and VSWR meters has been the maker's inability to provide adequate isolation between the forward and reflected power sampling ports over a wide bandwidth. There is also the problem of having to recalibrate the VSWR meter when the output power is varied. The power/VSWR meter described in this article has been designed to be truly independent of these problems. My goal was to design a dual-directional coupler with a flat response to at least 55 MHz, and a directivity greater than 30 dB. Three couplers were designed to meet these requirements, with coupling factors of 30, 24, and 20 dB. The 30-dB coupler can be used with transmitting systems having outputs up to 1000 watts. The 24-dB coupler can be used with systems below 200 watts, and the 20-dB coupler can be used with a 100-watt limitation. The 24-dB coupler turns out to be the most practical for average ham use.

circuit description

rf section. The rf sampling and detection circuit shown in fig. **1** was perfected with the use of a network analyzer having the capability of resolving amplitude variations of less than 0.1 dB over a 1 to 500 MHz frequency range.

In order to obtain the isolation between the forward and reflected power sampling detectors, two properly phased transformers are required. The toroid for the transformers is of "H" type magnetic material with a diameter of 9.5 mm (0.375 inch) and a thickness of 3 mm (0.125 inch), large enough to safely pass 200 watts without saturating the core. The primary of each coupler consists of a 2.5-cm (I-inch) piece of 0.141-inch OD semi-rigid coax with the solid copper outer jacket used as an electrostatic shield. It should be noted that the jacket is soldered to the groundplane of the printed circuit board on only one side of the toroid. Soldering on both sides would result in a shorted turn and actually degrade the performance of the coupler. The secondary of each transformer consists of fifteen turns of no. 31 AWG (0.2-mm) enamelled wire evenly spaced around the core. This provides 24 dB of coupling.

It is very important in winding the secondary of the transformers that the wire be spaced as evenly as possible. I found that having the turns spaced too closely caused the high-end performance to roll off at a much lower frequency than desired. The final design has a coupling response of 23.920.1 dB from 1 to 200 MHz, and a roll-off of 0.15 dB at 50 MHz. The insertion loss is negligible. The directivity (isolation between the forward and reflected power ports) is greater than 35 dB from 1 to 30 MHz rolling off to 25 dB at 200 MHz.

The rf detector uses a pair of germanium diodes as a positive halfwave rectifier. The output is then filtered and passed to the analog tracking circuit. In order to avoid large offsets in the rectified voltage, the diodes are matched as closely as possible in the center of the high-frequency band. The setup shown in fig. **2** allows matching to within 10 mV.

table 1. Values of return loss for different VSWR values.

VSWR	return loss (dB)	P_{out} = 1 kW	P = 100 watts
		P_{sec} (mW)	
1.05:1	32.2	0.6	0.06
1.1:1	26.4	2.3	0.23
1.2:1	20.8	8.3	0.83
1.5:1	14.0	39.8	3.98
2.0:1	9.5	112.0	11.20
3.0:1	6.0	251.0	25.10
4.0:1	4.4	363.0	36.30
infinite	0.0	1000.0	100.00

$$\textit{Return loss (dB)} = -10\log_{10}\left[\frac{VSWR-1}{VSWR+1}\right]^2$$

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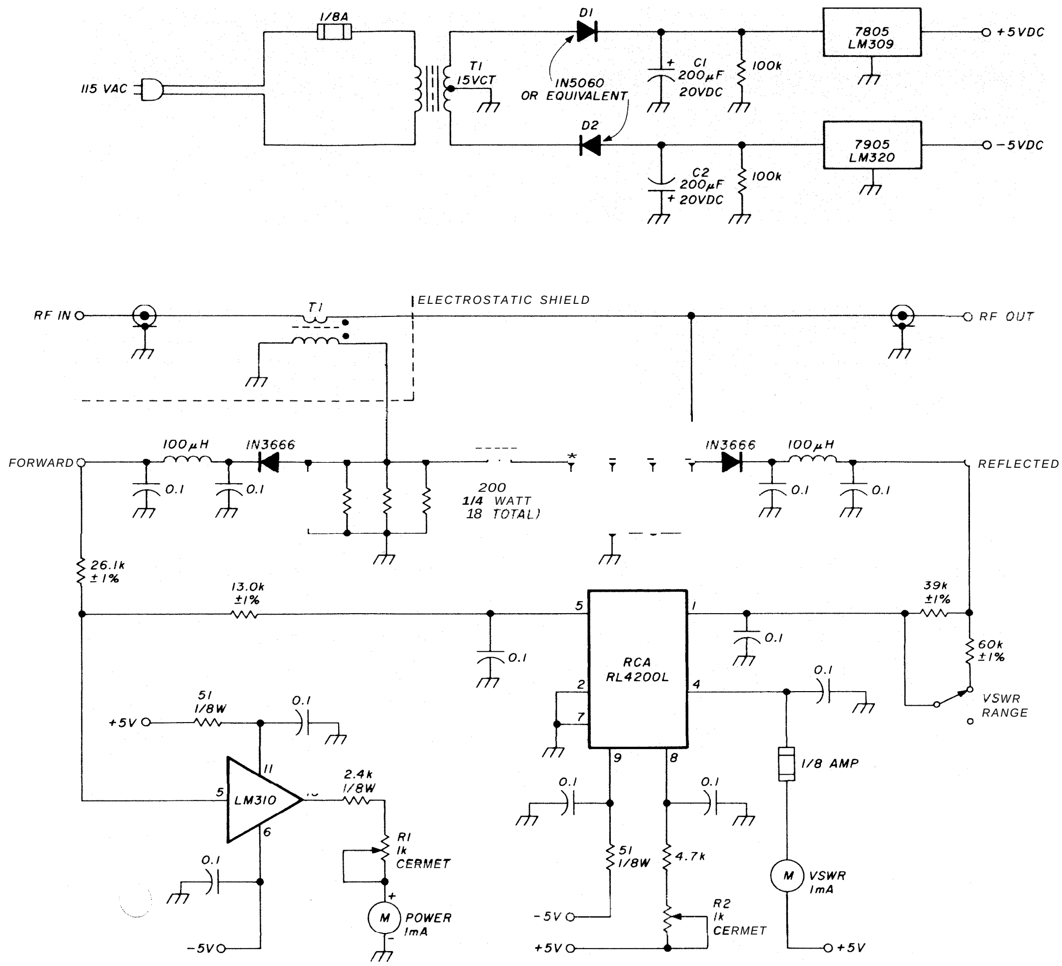


fig. 1. Schematic diagram of the broadband VSWR bridge and power supply. As specified in the text, T1 and T2 are wound to provide a coupling factor of either 24 or 30 dB.

30-dB coupler

This coupler is useful for high-power operation, that is, up to 1000 watts. With this looser coupling, adequate VSWR tracking for very low values of reflected power will be limited to approximately 100 watts of transmitter output. In other words, the VSWR will indicate properly as the power is varied from 100 to 1000 watts. Tracking very low values of VSWR becomes a problem as the output is reduced. The value of power in the reflected wave (see **table 1**) is determined by:

$$P_{sec} = P_{out}\,(dBm) - \text{return loss}\,(dB) - \text{coupling}\,(dB)$$

The only differences in the construction of this coupler are the meter faceplates and the coupling transformers. The transformers consist of the same one-turn primary as the 24-dB coupler, but the secondaries have thirty-one turns of no. 30 AWG (0.25-mm) enamelled wire even spaced around a type-H, magnetic-material toroid 12.5-mm (0.5-inch) diameter by 5-mm (0.188-inch) thick.

It will be necessary to construct the bridge in two sections separated by a shielded compartment. The printed circuit board can be cut in half, separating the couplerfrom the tracker. The board was designed with this purpose in mind, to obtain greater isolation, if needed, and to operate the coupler at some remote location in the coax.

A word of caution in winding the transformers – be very careful to avoid nicking the enamelled wire. This could result in a short to the outer shell of the semi-rigid coax used as the primary. I would suggest the use of clear epoxy on the entire transformer after assembly.

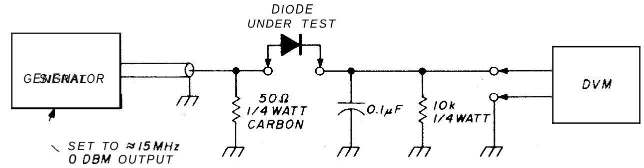
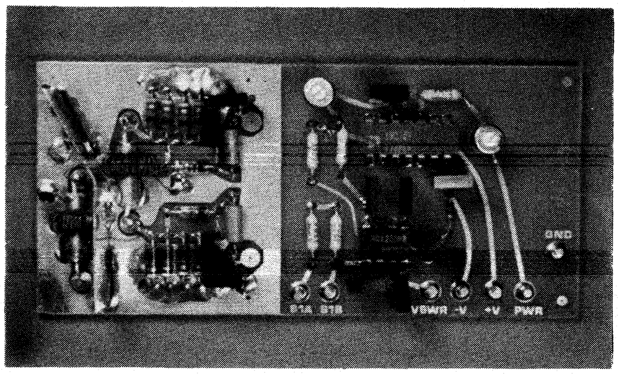


fig. 2. Setup for determining diode characteristics to select closely matched diodes as the detectors.



View of the circuit board used in the broadband VSWR **bridge. The board can be cut between the rf and current tracking portions for remote operation.**

Current tracker. This portion of the bridge makes the overall performance of the unit truly automatic. It is based upon the RC4200, a four-quadrant multiplier that is used as a current-ratio comparator. A reference current, established at pin 8, is used in the internal bridge portion of the IC. Forward and reflected currents from the diodes are fed to pins 1 and 5 and used internally to track the current ratio by the formula $I_4 = (I_8 I_2)/I_5$. It becomes evident that using this configuration will allow a constant VSWR to be displayed over a fairly wide dynamic power range, the limitation actually being the diode detectors and not the 4200. Thus, VSWR readings remain constant when transmitter power is varied from 200 watts down to 10 watts over the frequency range of 1.8 to 148 MHz.

The forward power is monitored by the use of a voltage follower fed from a 3-to-1 resistive divider. This divider network is necessary to keep the rectified voltage below the +5 volt supply.

I decided to have an expanded range for VSWR with 4:1 full scale being a reasonable choice. This is accomplished by shunting the 39-kilohm resistor with an additional 60 kilohms. The circuit of **fig. 1** can be modified to include a peak rms capability for monitoring the output power. **Fig.** 3 shows the modification.

alignment and testing

The VSWR is aligned using the setup in fig. **4.** The transmitter is first tuned up into a single dummy load. Then, with four loads in parallel, the antenna tuner is adjusted to reestablish a 50-ohm load to the transmitter. With the range switch in the 4:1 position, R2 is adjusted for a full-scale meter reading. Power is then noted on the series power meter, and R1 adjusted for the same reading on the power meter of the bridge.

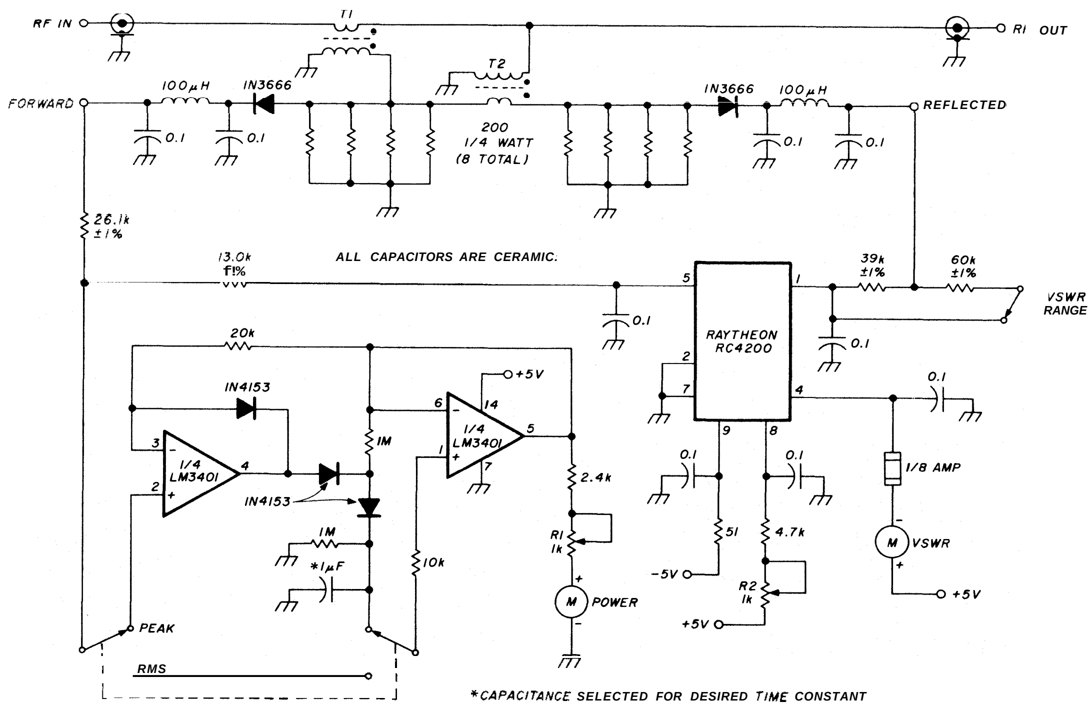


fig. 3. Schematic diagram of a similar VSWR **bridge which incorporates a peak-rms power readout.**

Next, one of the four loads is removed and VSWR readings are taken using both positions of the switch. The readings should exhibit a ratio of 3:1. This completes the testing and the bridge is now ready to be installed in the line permanently.

construction

This VSWR bridge is constructed on a single 10 x 5 cm (4 x 2 inch) printed circuit board, which is mounted on stand-offs inside an aluminum box. The rf is brought into and out of the box with short pieces of RG-58, stripped on one end for soldering to the terminal and groundplane of the printed circuit

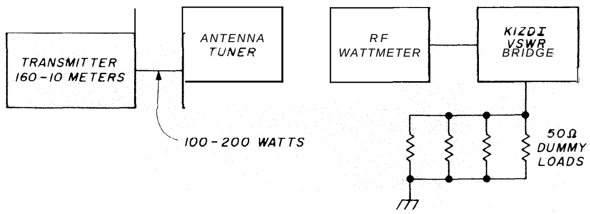


fig. 4. Test set up to calibrate the VSWR **bridge.**

board. The other end uses a male BNC connector connected to a bulkhead feedthrough connector. Two feedthroughs are mounted on the back of the main cabinet, one being used for rf in, and the other for rf out.

The meters are 1-mA full-scale movements with custom-designed scales. The face plates were removed and soaked in a solvent to remove the paint. They were then repainted with white spray. The next step was to tape the base plate to a pad of paper, and with the aid of a little geometry and a compass, a new arc was drawn on the white paint. The scale divisions were computed by changing power into a current ratio and finding the antilog.

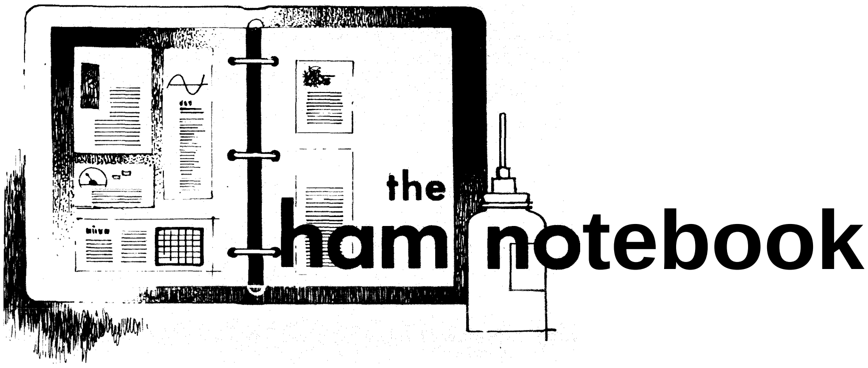
broadband transformer design

It is very important in broadband torodial transformer design to select a material that has a high permeability. This results in a transformer that, when measured on a vector impedance meter, will display a high real part of the impedance and a small reactive part, as indicated by a small phase angle reading. The low frequency roll-off is determined by the permeability of the material and the number of turns used, with a minimum number being determined by the impedance levels that the transformer will be working into. The high frequency roll-off is determined, for the most part, by the interwinding capacitance of the wires.

acknowledgments

I would like to thank Mark Stevens, WA1WSV, for his original ideas on the toroidal transformer approach, and also Eric Blomberg, N1BF, who helped design the current tracking portion of the bridge.

ham radio



Heathkit Micoder Touch-Tone pad adapted to low-impedance input

A quick and easy method of adapting the Micoder to transmitters such as the ICOM 22A, 22S, and 230 is to simply replace the audio-input pot (500 ohms) with a 10-kilohm pot. In the ICOM 22A this pot is R62, which is located behind the microphone connector. In the ICOM 230 it's R34, which is located on module AF.

The Micoder is designed for a 10-kilohm output load. For the ICOM 22A/22S, a Radio Shack 10-kilohm pot (part number 271-218) will work if bent slightly to allow the case to close. I used the 10-kilohm pot (Heath part number 10-1039, R122), which was left over when I converted my HD-1982 to an HD-1984. This conversion, which is available from Heath stores for about $9.00, provides a crystal-controlled oscillator that is quite stable and requires no adjustment or alignment.

Incidentally, there are two levels of Heath conversion kits. One has a zener for regulation and one does not. If yours has the zener, load resistor R105 should be changed from 820 to 470 ohms for the Micoder and to 1200 ohms if used with the Heath HT.

I set the Micoder level control, R103, at 50 per cent range. The new 10-kilohm pot should be set by trial and error with another station while on the air.

I also taped the removed pot (580 ohms) to an inside cover so that it could be easily retrieved for restoration to the original low-impedance microphone input.

E. L. Linde, WB2GXF

off-the-air S-line monitoring

While modifying the S-line for full break in (QSK) may not be desired by many, there are those who wish to monitor the off-the-air CW signal of the 32S-( ) rather than listen to a sidetone. This modification may be made by placing the 75S-( ) FUNCTION switch to OPR. However, this results in a very strong signal. You're then required to constantly adjust the RF GAIN to compensate for this strong signal, as well as the weaker received signal from the desired station. With the addition of two readily available components, off-the-air monitoring can be more pleasant.

Refer to **fig. 1**. With S1 in the NORMAL position receiver operation is normal. In the MONITOR position, R1 (Radio Shack 271-215) is added to the receiver's rf gain circuit, During receive, R1 is shorted by the contacts on the 32S-( ) VOX relay, K1. When the transmitter is keyed, normally closed K1 contacts open, inserting R1. RI's resistance is adjusted only once to provide a comfortable monitoring level in the speaker or headphones; CW SIDETONE is disabled by Slb. If you have difficulty obtaining a pot with a dpdt switch or if only an spst switch is available, Slb may be eliminated and the CW SIDETONE cable can be disconnected.

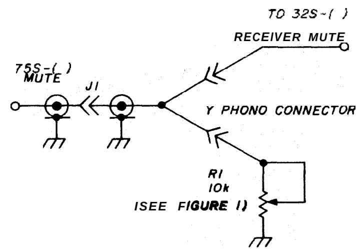


fig. 2. A "no-holes chicken-connection" may be made at the 75S-( ) MUTE jack for those not wanting to make any mods to their S-line gear.

The small size of the Radio Shack

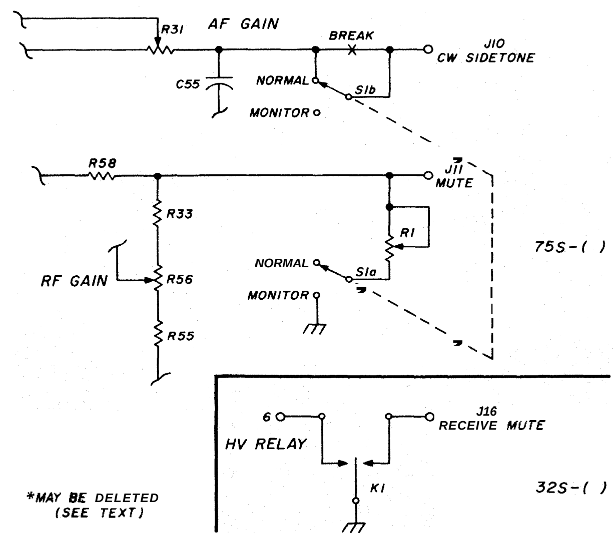


fig. 1. Radio Shack 271-215 pot for off-the-air monitoring level (R1) is added to the 75S-( ) rf gain-control circuit.

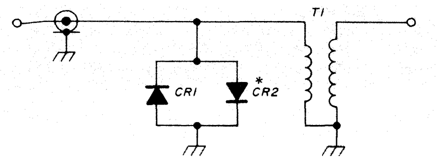


fig. 3. Rf input protection diodes added to the 75S-( ) antenna jack.

pot permits mounting it on the front panel between the MODE switch and the RF/AF GAIN control. The mounting hole should be centered laterally and located 79 mm (3-3/32 inches) from the top of the front panel. If a larger pot is used, it may not fit in that location but may be mounted on a small aluminum bracket and secured behind the PTO control by one of the PTO mounting screws. With front-panel mounting the adjacent chassis hole may be used for the leads. Shielded wire should be used for the CW SIDETONE lead to prevent hum pickup.

For those who may be hesitant to apply any modification to their S-line (or who may want to check the effectiveness of the addition), the added pot may be "chicken connected" by using a Y-type phone connector at the 75S-( ) MUTE jack, J-11 (fig. **2**).

I recommend the addition of two germanium diodes across the receiver antenna jack, J5, as rf protection devices (**fig. 3**).

Paul K. **Pagel, N1FB**

improved memory for the Yaesu FT-227R Memorizer

I like the FT-227R. It has all the features I enjoy in a 2-meter rig: It's fully synthesized over 4 MHz; it has 5-kHz offset tuning; it operates either up or down 600 kHz (or on so-called oddball splits); and, of course, it has simplex capability. Its power is also right: 10 watts for repeater operation or you can run 1 watt for local contacts. Included are tone-burst circuitry and provisions for CTSS and full-tone access.

After six months of operation I found that the rig will also work in Teletype service. We have a net that runs AFSK at 170 Hz in the New York City area. (Nothing gives as good an indication of how well a rig works as 20 minutes of key-down operation.)

Everything seemed to be what I wanted. But — when I closed down for the night I found that the Memorizer *forgets!* Whenever I took the rig into the house, or just out of the car, I found this to be true. I looked into the instruction manual one evening, and out came the answer.

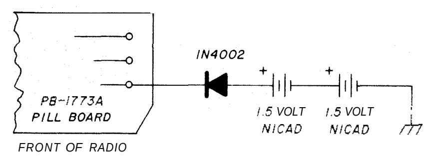


fig. 4. Mods to the Yaesu FT-227R Memorizer 2-meter rig to store a frequency after the radio is disconnected from its power source.

Remembering that my LED watch could remember time with just a low-power battery, I thought perhaps the same idea would work in the FT-227R. A check of the schematic shows that when the rig is connected to a battery, as in an automobile, and the memory switch is pressed, 5 Vdc is constantly applied to the phase-locked-loop (PLL) unit. This is the PB-1773A unit in the book. The 5 Vdc was derived from the 13.5-volt line from the auto battery. A further check showed that the ICs in the PLL were of the CMOS type and would probably work on very little voltage if pushed. The solution to the problem was simply adding 3 Vdc and a 1N4002 diode (fig. **4**). These mods allow a small voltage on the PLL circuit through a diode to prevent higher voltage from ruining the two AA cells. It's that simple.

construction

The first step in making the addition is to remove the five screws and put aside the bottom cover (this is the one with the speaker). Looking into the unit you'll see the PLL circuit in back of the front panel. In my unit the PLL circuit was covered by a fiber insulating cover.

Remove the three screws that hold the PLL circuit cover. Remove the cover. With the rig facing to the right, look at the upper left-hand corner of the board. You should see three traces running toward the "bottom," or down the long length of the board. Follow the third trace in from the end. It should go to pin 16 of all the ICs. This is the power trace and the one that we want

Solder a 100-150 mm (4-6 inch) wire to it. Then find a ground point on the board. Anywhere will do. Solder another wire of the same length to it. Replace the cover and the three screws on the PLL board. Observe good construction practice and use a very small (25-watt or so) soldering iron.

This should leave you with two wires sticking out of the PLL board with the cover on. Next solder two AA batteries in series. Solder the anode of the diode (unbanded end) to the positive side of the "top" battery, and the cathode (banded) side to the wire coming from the PLL board connected to ground. This will put a positive voltage on the PLL at all times.

Now solder the ground wire from the PLL board to the negative end of the batteries. That's all there is to the eiecrricai mods.

I wrapped the batteries and diode with PVC electrical tape and located them in an empty area of the board.

Total drain on the system is only 3 mA, so the batteries should last for quite a while. I've had the system in operation for some months and haven't yet replaced the batteries.

With this system you can have memory for the low, low price of under $0.65. This may be one of the best modifications available for the money. Now, if you have one frequency in memory and another on the dial, the radio will remember both frequencies, even if the power is off and the rig is removed from the automobile.

Stephen Mendelsohn, **WA2DHF**

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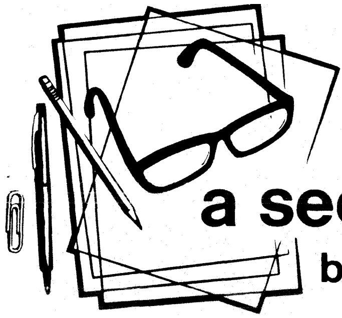


a second look

by Jim Fisk

If you're active on the high-frequency amateur bands, you have probably formed your own idea of what it would be like to operate from a foreign country. You don't need many DX entries in your logbook before you begin to see some trends: power input, types of equipment that are preferred in various places, and the antennas that are the most popular. Have you ever wondered how those same DX operators visualize American radio amateurs?

Writing in a recent issue of *Break-In,* the official journal of the New Zealand Association of Radio Transmitters, Harry Bourne, ZL1OI, provided some of the answers. While making contacts with more than 2500 amateurs in all callsign districts of the United States and Canada on 15 and 20 meters, Harry collected a good deal of interesting data on transmitter input power and antennas. He found, for example, that 13 per cent of the stations used less than 100 watts, 59 per cent used between 100 and 500 watts, and 28 per cent of the operators used more than 500 watts; he also found that the average power input on the 14-MHz band is higher than on 21 MHz.

In the antenna department, ZL1OI's survey showed that 48 per cent of the American amateurs use Yagi beams at heights of 30 to 80 feet (10-25 meters), 21 per cent use verticals (either ground mounted or as elevated ground planes), 13 per cent run quads, often at rather low heights above ground, and 13 per cent depend on half-wave dipoles. The remaining 5 per cent use a variety of antenna types including Zepps, delta loops, vee beams, rhombics and indoor antennas.

ZL1OI's logbook reveals further interesting results; signal reports, for example, confirm that antennas have a far greater effect on signal strength than transmitter input power — and it is much more effective to improve the antenna than it is to increase power. This will come as no surprise to serious DXers, but it's reassuring to have it confirmed by a DX station. And the excellent propagation conditions we've been experiencing for the past few months have made it possible for amateurs to achieve good DX results with low input powers, especially if they have a good antenna system. One afternoon not too long ago I hooked up with a G3 who was running 150 milliwatts input on CW; he reduced power to 35 mW and we easily exchanged signal reports on ssb. That's roughly 100,000 miles per watt! And just recently I worked 7X2BK on 28 MHz using 200 mW and a 3-element beam.

When propagation conditions are good and the high-frequency bands are as hot as they have been so far this year, directional antennas are not so important for increasing signal strength as they are for reducing interference from directions other than that of the desired station. With a power input of 200 watts, excellent DX results can be obtained with simple vertical or dipole antennas, or single quad or delta loops. If you're unable or unwilling to install a larger or more sophisticated antenna system, you may not be able to crack that big DX pileup on your first call, but with good operating techniques and patience you'll be able to work any station in the world on CW. On phone it's more difficult, but only because the competition is tougher and the interference is horrendous!

If you want to improve your station performance, the message is clear: spend your budget on your antenna ***system,*** not a linear amplifier, and remember that includes not only the antenna, but the ground system and the transmission line. If you're using inexpensive coaxial line, or cable that's several years old, you may be surprised to find that you can greatly increase your effective radiated power by simply installing RG-213/U or other high-quality coax.

If your budget won't allow a new antenna, try to increase the height of the one you already have; you may be able to double your signal strength by raising your antenna above nearby objects. And if your antenna is ground mounted, increase the number of radials; aluminum electric-fence wire is ideal and costs about a penny a foot. Unless you're already using a Yagi on a 100-foot (30-meter) tower, dollars invested in your antenna system will give you more bang for the buck than dollars spent in any other part of your ham station. Keep that in mind as you get your station ready for the coming DX season. Nearly all the propagation forecasters agree that band conditions this fall and winter will be better than they have been in twenty years — and conditions may not be as good for another twenty!

Jim Fisk, **W1HR**
editor-in-chief

comments

propagation predictions

Dear HR:

I have been advised that orders for the government publications on *Ionospheric Predictions* cited on page 30 of my article in the April issue of *ham radio* are no longer available from the Superintendent of Documents. I scouted around and, courtesy W9OWZ, discovered that photocopies can be obtained from National Technical Information Service, Post Office Box 1553, Springfield, Virginia 22151. Here are ordering information and prices:

Volume I COM-73-50654 "General Instructions"	$ 3.00
Volume II COM-73-50655 "Sunspot Number = 12"	$11.75
Volume III COM-74-50041 "Sunspot Number = 110"	$11.75
Volume IV COM-74-50042 "Sunspot Number = 160"	$11.75

Henry G. Elwell, Jr., N4UH
Cleveland, North Carolina

voltage-regulator noise

Dear HR:

I very much enjoyed W1HR's article on Gunnplexers in the January Issue.

However, I would like to bring something to your readers' attention in reference to the suggested 723 voltage regulator. This regulator employs internal zener regulation, and zeners being inherently noisy, can contribute to system signal-to-noise ratio (SNR) degradation. I have been able to increase the signal-to-noise of a studio-transmitter link (STL) receiver by just short of 3 dB and unmask a VCXO's actual distortion of less than 0.2 per cent by simply by-passing pin 5 of the 723 with a 10 μF capacitor and placing a 47-kilohm resistor between pins 5 and 6. Motorola indicates this addition in one of their application notes; however, its importance is not stressed, nor followed on in other application notes.

I have experienced no such problems with W1HR's suggested Fairchild device, the 78MG. I understand the 78MG regulators are not internally zener regulated. This should be considered by those who are looking toward the ultimate in noise figures, distortion, and SNR.

Dave Clingerman, W6OAL
RF Project Engineer
Moseley Associates, Inc.

Dear HR:

Over the past few years, I have monitored the 160-190 kHz band listening for the large number of stations that are supposed to be running beacons and scheduled transmissions. Since I have never positively identified any of these signals, it appears that either the wrong frequency was being monitored or the signal was too far down into the noise.

A discussion with N6GN concerning this problem resulted in the idea of using the sixth subharmonic of a one megahertz crystal as a standard operating frequency. The resulting 166.666...kHz signal would be very exact since the 5th, 10th, or 15th harmonic of the 1 MHz crystal could be set to be "zero beat" with one of the WWV signals.

I would he very happy to schedule anyone in the San Francisco area on 166.666 kHz.

Dick Bingham, N6HZ
4880 Burnside Rd.
Sebastopol, California 95472

anodize dyes

Dear HR:

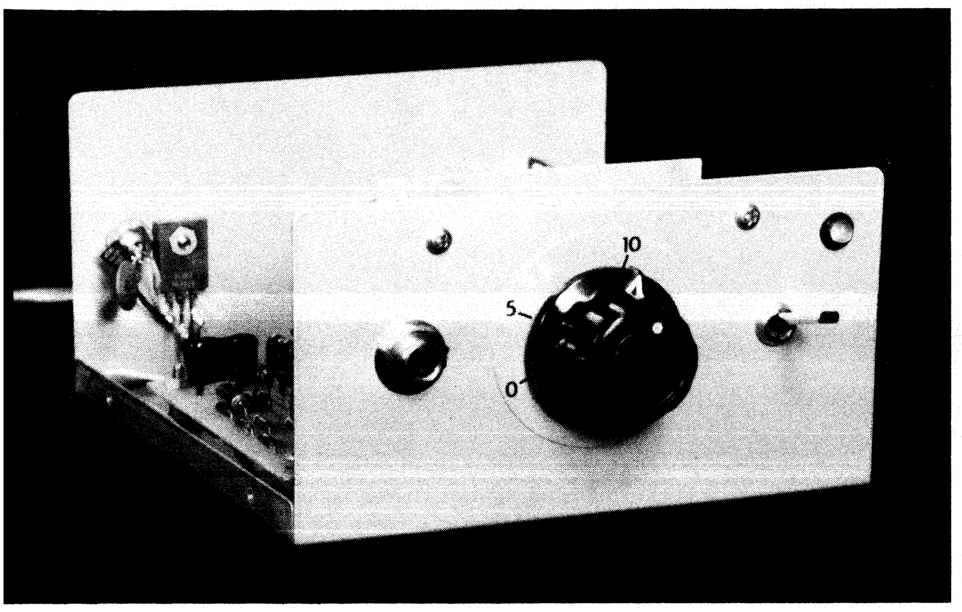
The article on anodizing in the January, 1979, issue of *ham radio* mentions several sources of dyes which can be used, including "drugstore" fabric dyes. The trade name dyes, *RIT* are typical of this group, are low cost, and have the reputation of being repeatable. These dyes are used warm, 50-60°C; the dyed surface is then sealed by boiling.

Bob Haviland, W4MB
Daytona Beach, Florida

note of acknowledgment

The April, 1979, issue contained an article entitled "The Jammer Problem: Some Interesting Solutions," page 56. This article was adapted from "A Contribution to the Mathematical Theory of Big Game Hunting," by H. Petard, Princeton, New Jersey, which originally appeared in *American Mathematical Monthly* (1938). The article has since been reprinted in *A Random Walk in Science* (1973), compiled by R. L. Weber and edited by E. Mendoza.

We thought that Petard's article on big game hunting would make an interesting basis for an adaptation geared to the very real problem of intentional interference in the Amateur bands. Thanks to Jim Kirkpatrick, WB7BUP, for the background information on the original piece. **Editor**



split-band speech processor

Design and construction details of a split-band audio speech processor that features up to 15 dB of clipping and low distortion

Speech processing, especially for SSB, can be a relatively inexpensive means of improving the effective "talk power" of a voice modulated transmitter. Much has been written about various devices and methods that can be used to gain this increase in effective talk power. The devices used have ranged from simple audio compressors to rf envelope clipper-filters. All of these devices attempt to reduce the peak-to-average ratio of the speech or rf waveforms, thereby overcoming the peak power limitations of the transmitter. Generally, the degree of improvement is proportional to the complexity of the processing method; the simpler circuits offer minimal improvement while the more complex effect substantial improvement.

This article will not attempt to present all the theory involved in speech processing; however, the interested reader is referred to excellent articles by Fisk,[1,2] Kirkwood,[3] Moxon,[4] and Schreuer[5] for more detailed overviews of the subject.

Until recently, rf envelope clipping has generally been accepted as the most effective SSB processing method. Distortion products are small, generally consisting only of intermodulation products. The primary disadvantage of rf processing is the circuit complexity involved, and the necessity of modifying the associated transmitter. When modifying the transmitter is out of the question, a processor using the audio-SSB-audio (Comdel) approach can be used. In this method, an SSB signal is generated, peak limited (clipped), filtered, and then demodulated back to an audio signal which then modulates the transmitter.

My initial efforts were directed toward designing and building a unit of this type. A breadboard model was constructed and evaluated under laboratory conditions. Performance was very good, and distortion was held to under 10 per cent at 20 dB of clipping. The circuit was, however, excessively complex. It required an audio preamplifier, two balanced modulators, an oscillator, a clipper, an rf amplifier, and an expensive mechanical or crystal filter.

By Wes Stewart, **N7WS**, 1801 East Canada Street, Tucson, Arizona 85706

At this point, Jim Metzger, W7TKR, suggested that I try the split-band approach. He had done some work with the process with considerable success and Fisk[2] had written in glowing terms about a similar unit available commercially from Maximilian Associates. This was inducement enough to build a breadboard model for evaluation.

basic circuit

Fig. 1 is a simplified block diagram of the split band clipper. The input signal is applied to an agc-controlled preamplifier which then drives the first set of bandpass filters (BPFs). The filters split the audio spectrum into four narrow bands which are tnen clipped and directed into the second set of BPFs, where the harmonics generated by the clipping process are filtered off. These filtered signals then go to the combiner stage where they are reassembled into the desired output.

Input amplifier. The design of the input amplifier is not particularly critical. The gain required will depend on the output amplitude of the source, the gain (if any) of the BPFs, and the limiting threshold of the clipper stages. If a very low output microphone is used, low noise may be of some importance. If, as in my case, active bandpass filters are used, the amplifier will also have to exhibit low output impedance. Automatic gain control is also desirable, as it helps maintain a high average clipping level, which in turn insures maximum talk power improvement.

Bandpass filters. As pointed out by Fisk, the optimum design for BPFs is a compromise between several conflicting requirements. Overshoot or ringing due to the near squarewave input from the clipper must be minimized, skirt selectivity should be good, and phase shift through the passband must be smooth and predictable. The latter point becomes important when the design of the combiner is considered, as will be seen later. Other very important factors to be considered are circuit complexity and reproducibility.

After pondering all of the above points, I decided on a two-pole Butterworth active filter. The Butter-

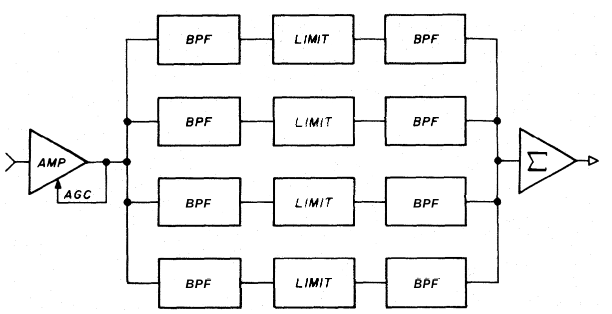


fig. 1. Block diagram of a split-band audio speech processor.

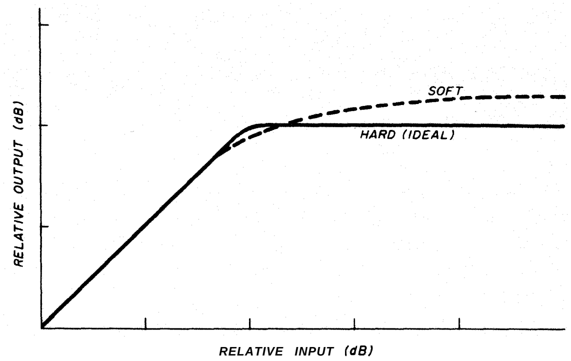


fig. 2. Comparison of "hard" vs "soft" limiting. Soft limiting is undesirable because of the uncertainty of the threshold point, making it hard to maintain constant output from the processor.

worth is not optimum when considering only impulse response and phase shift; however, when used in a low-Q configuration, it is a good compromise between filters with these attributes and those possessing superior skirt selectivity.

The final circuit is configured as a multiple-feedback type.[6] These filters are relatively insensitive to component variations, allowing the use of 5 per cent tolerance components and inexpensive operational amplifiers. Detailed design data for the selection of center frequency, gain, and Q will be given later.

Peak clipper. The clipper may seem to be one of the least critical parts of the circuit, but, in fact, its requirements are quite stringent. One of the most important factors in the performance of the clipper is that of clipping symmetry. Perfect symmetry insures that only odd harmonics are generated; second-order products would be too much for the two-pole filters to handle. An important point is that the only place clipping should occur is in the clipper. Clipping or limiting elsewhere in the circuit cannot be easily controlled and must be avoided. This may seem easy to do, but if the clipping threshold is too high, limiting may occur in a preceding stage when large amounts of clipping are in use. For example, if a clipping threshold of one volt is used and 20 dB of peak clipping is desired, the preceding stage must be able to have an output voltage swing of 20 volts peak-to-peak. If this stage is running off a single 12-volt power supply, this will of course be impossible.

Another important aspect is that of how "hard" the limiting is. Many of the circuits initially examined, which included limiting differential amplifiers, shunt-diode clippers, and operational amplifiers with shunt diode feedback, had rather "soft" limiting characteristics. That is, the threshold was ill-defined and the slope of the transfer function continued to change over a wide range of input levels. Fig. 2 graphically shows the difference between hard and soft limiting.

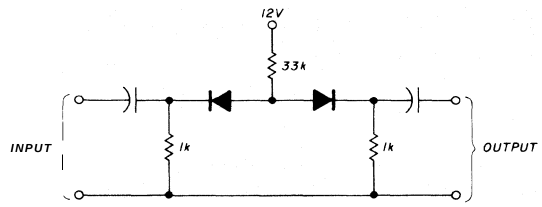


fig. 3. Circuit diagram of the amplitude iimiter (clipper) used in the final design. With the resistor values shown, the output will be limited to approximately 300 mV p-p.

Soft limiting is undesirable because it makes it difficult to maintain a constant peak output level.

The circuit finally selected for this application, as best satisfying the above requirements as well as using a minimum of parts, is shown in fig. 3. This will be recognized as a variation of the old series automatic noise limiter used in receivers. By suitable selection of resistor values and bias voltage, the clipping threshold may be adjusted over a wide range.

The performance of this circuit is demonstrated in fig. **4.** This is a multiple-exposure oscilloscope photograph taken of the output of the clipper. The inner, near sinusoidal, trace was obtained by increasing the input signal until a 3-dB increase caused only a 2-dB change in output. This point was defined as the clipping threshold. The middle trace represents a further input increase of 4 dB, and the outermost trace was obtained with a total input overdrive of 15 dB. The photograph shows the nearly flat peak output and the exceptional symmetry. A further test of symmetry was made by examining the frequency spectrum of the clipper output with a Hewlett-Packard 302A wave analyzer. With 15 dB of clipping, the second harmonic remained more than 40 dB below the fundamental output.

Combiner. The combiner has the job of taking the four BPF outputs and putting them back together again while maintaining their original phase relationships. Improper phasing will result in excessive passband ripple being generated. As described by Fisk, the Maximilian unit incorporates phase shift networks before the combiner to compensate for the phase shifts through the BPFs. As will be shown later, these networks can be eliminated by the judicious selection of filter characteristics and the use of a simple summing and differencing amplifier.

circuit description

Fig. ***5*** is the complete schematic of the system. The input is applied to Q1, an FET source follower, used to match high impedance microphones. The follower output drives U1, a Plessey SL1626 gain-controlled amplifier. This IC maintains a nearly constant output of slightly less than 100 mV RMS over an input range of 1 to 100 mV.

The SL1626 is used as recommended by the data sheet, except for the addition of R6 and C10, which are necessary to suppress a high-frequency oscillation. R4 lowers the sensitivity about 20 dB and may be unnecessary in some applications. Front panel adjustment of the clipping level is possible via R7.

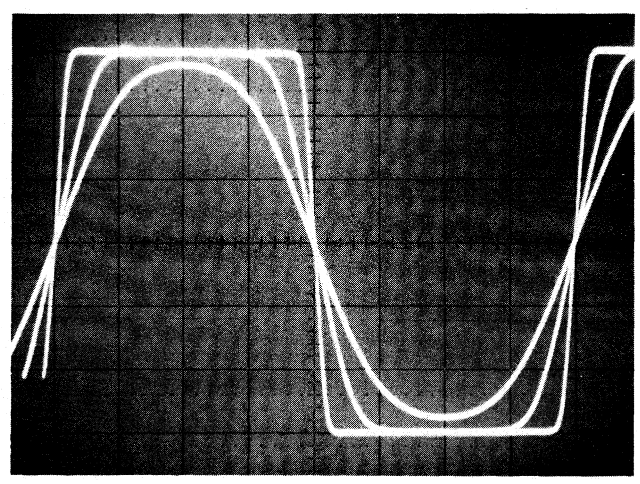
Amplifier U2A, one section of an LM324, develops a small amount of additional gain and serves as a low-impedance source to the following BPFs. The resistors used on the outputs of all the LM324s are necessary to eliminate cross-over distortion.:

All of the bandpass filters are operated at the same gain and Q; only the center frequency (f_o) differs from channel to channel. For simplicity, all capacitors are of the same value; the center frequency is adjusted by choice of resistor values. Using the given values, the overall frequency response wil! be approximately 350 to 3000 Hz at – 6 dB, with no greater than 3 dB of passband ripple. If other cutoff frequencies are desired, **appendix 1** gives the equations necessary to calculate new values of f_o and Q. **Appendix 2** gives the equations for calculating the parts values for the individual filters.

The clipping stages, as described earlier, use a pair of forward-biased diodes. With the bias resistor values shown, the clipped output will limit at about 300 mV p-p. The shunt-bias resistor values are kept low enough to insure that the input impedances of the second BPFs remain fairly constant even when the clipping diodes turn off.

The second set of BPFs are identical to the first. Their outputs are combined in another section of an LM324, which delivers the system output through a resistive divider. By adjusting the resistor values, the output amplitude can be set approximately the same as that of the microphone, allowing the clipper to be

fig. 4. Performance of the clipper stage shown in fig. 3. The sinusoidal trace was made at the threshold point (1-dB clipping). The middle trace shows 5 dB of clipping, and the outer trace was made with 15 dB of clipping. The vertical sensitivity is 50 mV/div.



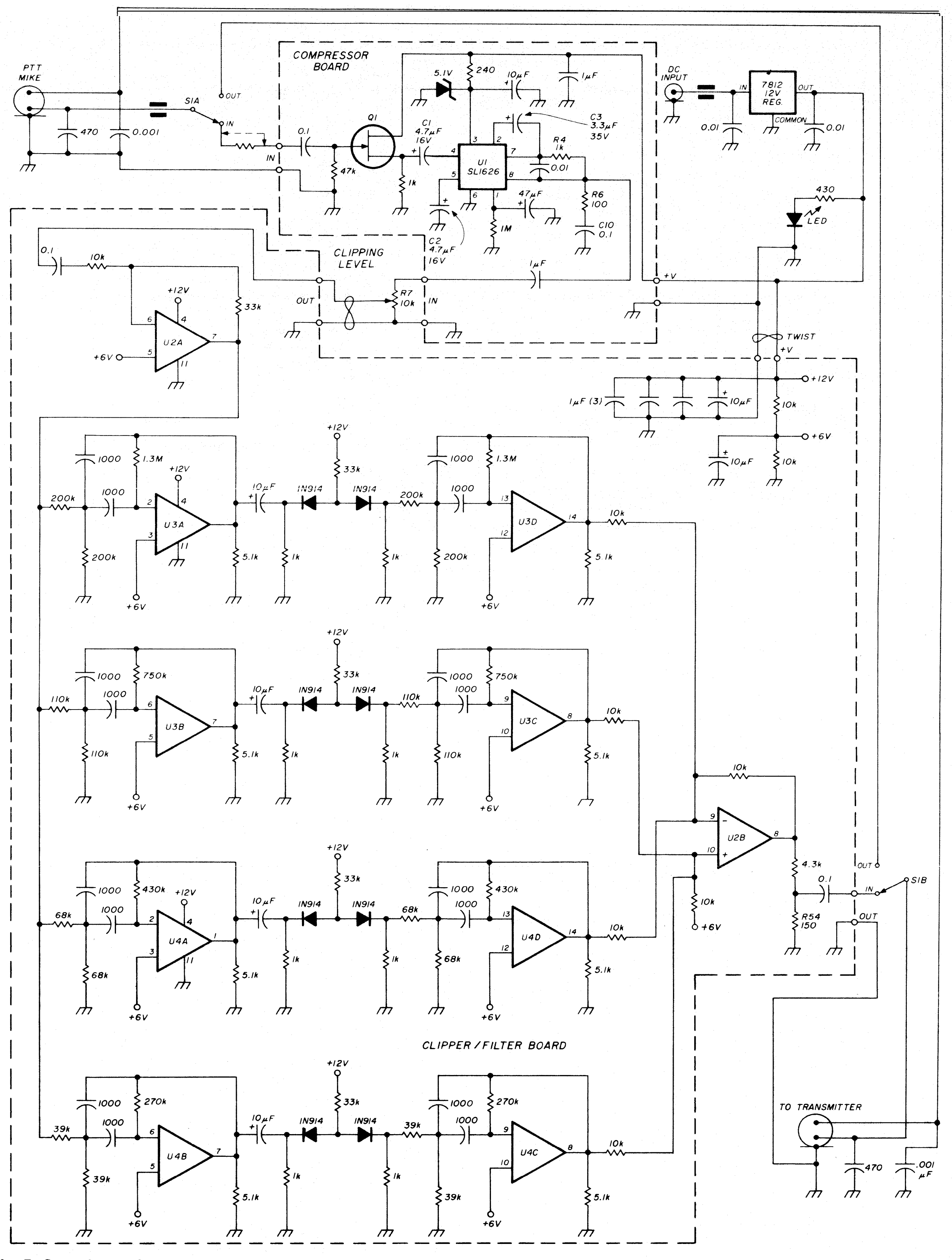


fig. 5. Complete schematic diagram of the split-band audio processor. Q1 is a 2N4392 or equivalent. U2, U3, and U4 are LM324s. C1 and C2 are dipped tantalum capacitors (RS 272-1409). C3 is also a dipped tantalum (RS 272-1408). All other polarized capacitors are tubular tantalums or electrolytics. The remaining capacitors are ceramics, with the exception of the 1000-pF capacitors, which are 5 per cent dipped micas. All resistors are 1/4-watt, 5 per cent, carbon composition.

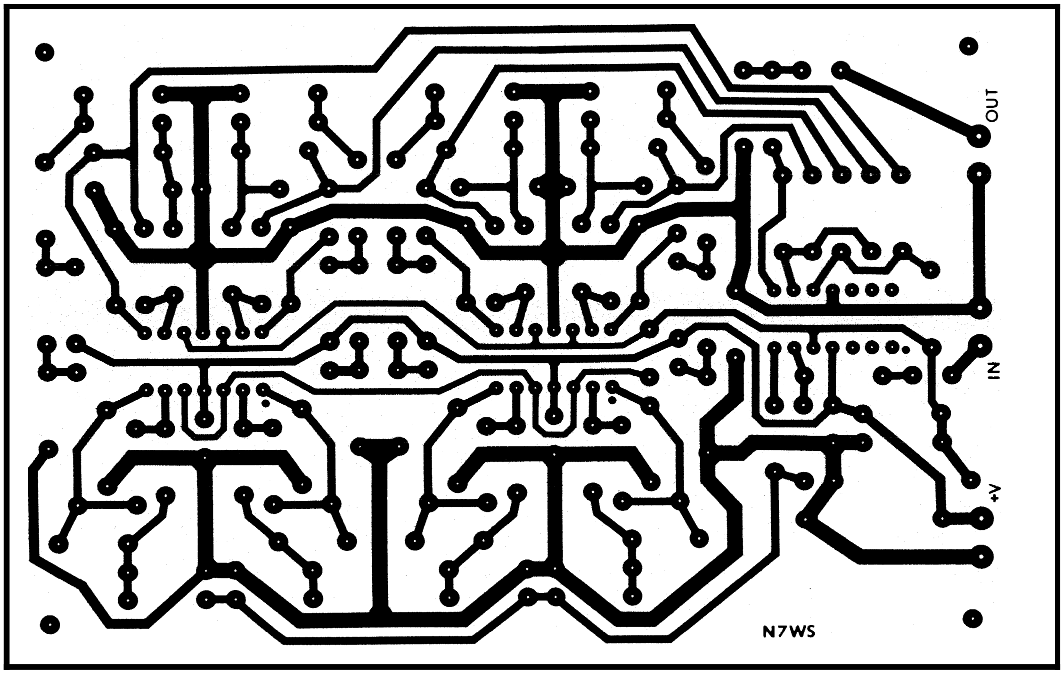


fig. 6. Full-size printed-circuit layout for the audio clipper board.

switched in and out without a gain change in the transmitter. Note that the BPF outputs are alternately connected to the plus and minus inputs of the combiner. The next section will demonstrate why this is done.

Assuming the equations shown in the appendix were used to determine the f_o and Q of the individual filters, adjacent filters will share a common -3 dB frequency. **Eq. 1** demonstrates how an input signal at this frequency is shifted $+45$ degrees in one channel and -45 degrees in the other:

$$\theta = 90 - \arctan\left(\frac{2Qf}{f_o} + \sqrt{4Q^2 - 1}\right) - \arctan\left(\frac{2Qf}{f_o} - \sqrt{4Q^2 - 1}\right) \quad \textbf{(1)}$$

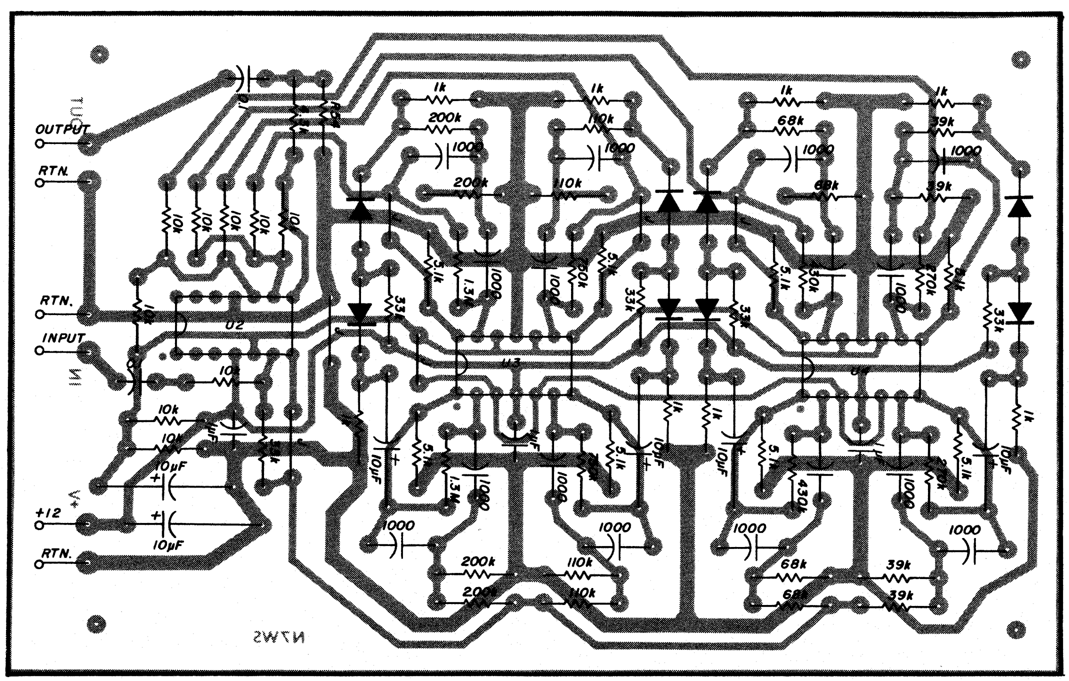


fig. 7. Component location on the audio clipper board.

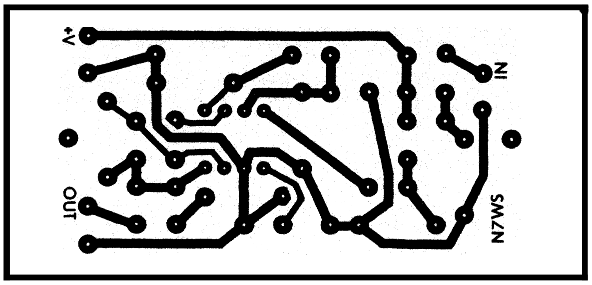


fig. 8. Full-size circuit board layout for the compressor board of the split-band processor.

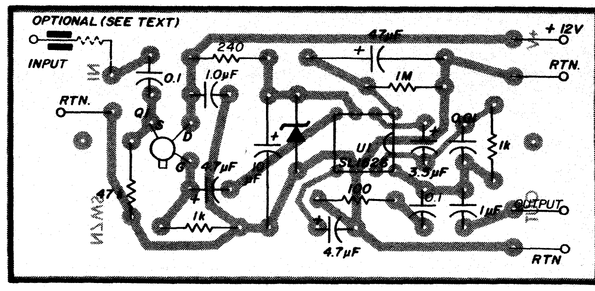


fig. 9. Parts placement diagram for the compressor circuit board.

where θ is the phase shift in degrees
f_o is the filter center frequency
f is the frequency of interest

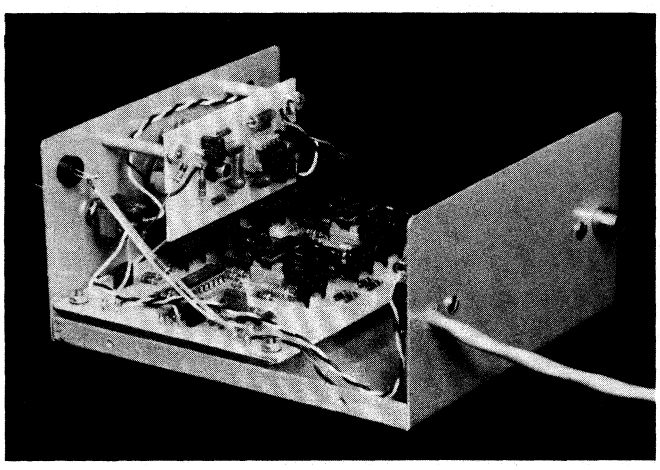
After cascading the two filters in each channel, this shift will be doubled to ± 90 degrees. Clearly, if these two signals are vectorially added, their sum will be zero because they are of equal amplitude but 180 degrees out of phase. A simple solution to this problem is to invert the phase of one signal. This is effectively what is done by the combiner.

Solving **eq. 1** for other frequencies will yield a phase error that increases with distance from the -3 dB point. This error is less important, however, because the amplitude difference also increases, so the larger signal dominates when the summation is made.

construction

For added versatility, the circuit is constructed on two etched circuit boards; the input compressor on one, the clipper-filter on another. This allows either one to be used alone in other applications. **Figs. 6** and **8** are full-sire layouts of the foil sides of the two boards, **while figs. 7 and 9** show the component placement. These boards have been laid out with considerable attention to preventing ground loops. A hand-wired board should be built with the same attention.

fig. 10. View of the prototype split-band speech processor. The circuit boards are mounted using metal spacers and machine screws. Room is available for mounting an ac power supply; an external supply was used for this model.



The prototype shown in **fig. 10** was constructed in a Radio Shack enclosure (270-253). Sufficient space remains for the inclusion of an ac-operated power supply. **Fig. 11** is a schematic diagram of a suitable supply. Liberal use of ferrite beads and bypass capacitors on all leads entering the enclosure eliminates any chance of problems with rf interference.

performance

As **fig. 12** shows, the frequency response is very close to what was calculated, despite the use of 5 per cent components. By adjusting R7, the clipping level can be varied from 0 to 15 dB. Greater amounts of clipping can be had by increasing the gain of either U2 or the BPFs, or reducing the clipping stage bias to lower the clipping threshold.

Caution should be exercised before deciding on greater amounts of clipping, however. This could turn out to be too much of a good thing. Increased clipping does continue to reduce the peak-to-average ratio, but at the same time distortion increases rapidly. This is shown graphically in **fig. 13.** As pointed out by Moxon,[4] most of the improvement is obtained by the first 6 dB, with little to be gained by increased amounts. My on-the-air tests seem to indicate that 10 to 12 dB is about optimum with this system. All of this is rather subjective, but the whole topic of speech intelligibility and recognition is pretty subjective, so take it for whatever it's worth.

Total harmonic distortion was measured with an HP 331A distortion analyzer at various frequencies and clipping levels. The results of these measurements are shown graphically in **fig. 13.** As the figure indicates, distortion begins to rise rapidly as the clipping level approaches 15 dB.

These measurements were of necessity made with single frequency inputs which represent worst-case conditions. Because clipping is occurring on every

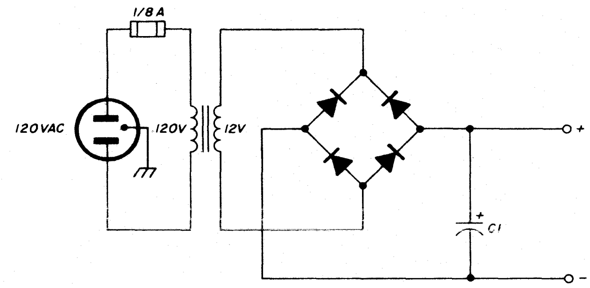


fig. 11. Schematic diagram of an ac power supply suitable for use with the processor. The transformer can be a Radio Shack 273-1385, the diode bridge a 276-1151, and C1 either 272-1019 or 272-1032.

half cycle, harmonic generation is maximum. With speech, clipping occurs much more randomly, with proportionally less total distortion.

On-the-air tests have been extremely gratifying. Reports have indicated substantial increases in apparent signal strength without noticeable distortion or loss of naturalness as long as the clipping level was held around the 10- to 12-dB point. Some loss of naturalness seems to occur above this point, but up to 15 dB, the sound is still not too objectionable. No tests have been run at levels in excess of 15 dB.

operation

Operation is very simple. The agc amplifier holds the clipping level constant, relaxing the operator requirements considerably. Some adjustment of the input sensitivity may be necessary if the microphone used has either a very high or very low output. While the dynamic range of the compressor will handle a higher input, the rise in background noise between speech pauses will be annoying to the listener. In this case, a series resistor may be added to the input which, in combination with R1, forms an attenuator. In the case of a very low-output mircophone, increasing the value of R4 will increase the sensitivity. Highest gain occurs with R4 omitted entirely.

On the output side, changing the value of R54 will control the maximum output level. This interacts with the audio gain control on the transmitter, so

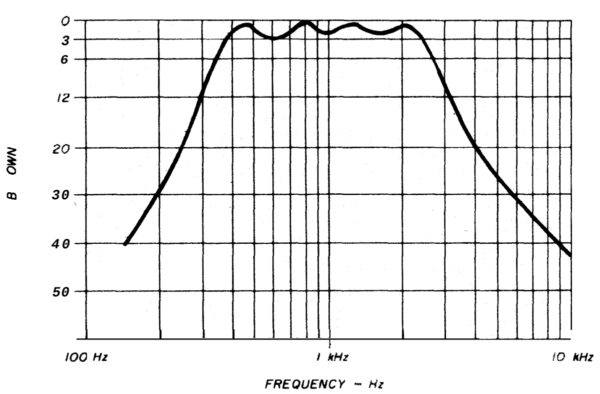


fig. 12. Measured audio response of the speech processor.

corrections can be made either place. I tried to pick a value that allowed the clipper to be switched in and out without having to readjust the microphone gain each time.

Finding the best setting for the microphone gain is best done with the aid of an oscilloscope on the transmitter output. With the clipping level set to maximum, adjust the transmitter gain so the peak output just approaches the level achieved with full carrier or excitation. If no oscilloscope is available, I find that just whistling into the microphone and setting the gain to the point that just activates the transmitter ALC works out very well. If you are not going to use the maximum amount of clipping available, then do the adjusting at the clipping level you intend to use. Even the best of clippers will nor maintain a completely flat output vs input characteristic. Therefore, if you adjust your gain at 15 dB of clipping, then reduce it to 10 dB, your peak output will drop a little.

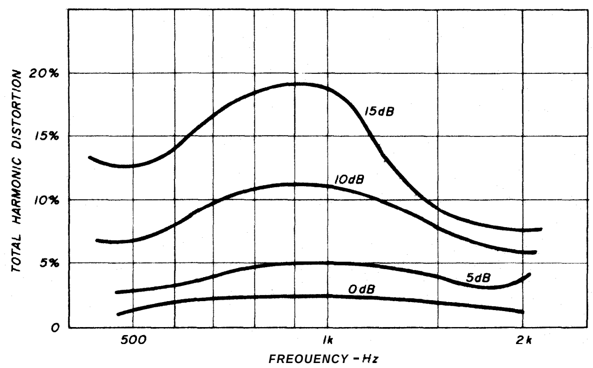


fig. 13. Total harmonic distortion vs clipping level. These curves were made with single tone inputs. Average distortion with speech input should be lower.

This effect can be explained as follows: As shown in fig. 4, sine waves subjected to 15 dB or so of clipping take on the appearance of pretty good square waves. As mathematical analysis can show, a square wave is composed of a fundamental frequency and all of its odd harmonics. We try to filter out these harmonics and retain only the fundamental. Unfortunately, the peak amplitude of this fundamental component is larger than the peak amplitude of the square wave by a factor of $\frac{4}{\pi}$, or 2.1 dB.[8] It is this factor that causes a continuing increase in output despite the use of a "perfect" limiter.

I want to express my thanks to Jim Metzger, W7TKR, for his technical advice, to Frank Baker for his circuit-board layout genius, and to Don Scheick and Norm Keopfer for their assistance in the preparation of the circuit boards. Additional thanks go to the many others who offered advice and encouragement, to Norma Putney for the typing of the manuscript, and to my wife, Terry, for the many hours spent away from family affairs during this project.

appendix 1

For new passband limits, the values for Q and f_o can be found as in the following example:

1. Define the low frequency -6 dB point, f_L (350 Hz)
2. Define the upper frequency -6 dB point, f_L (3000 Hz)
3. Find the multiplying coefficient, L

$$L^4 = \frac{f_H}{f_L} = \frac{3000}{350} = 8.571$$

$$L = \sqrt[4]{8.571} = 1.711$$

4. Find the individual filter cutoff frequencies

$$f_L = 350\ Hz$$
$$Lf_L = 599\ Hz$$
$$L^2 f_L = 1025\ Hz$$
$$L^3 f_L - 1753\ Hz$$
$$L^4 f_L = 3000\ Hz$$

5. Find the individual center frequencies

$$f_{o1} = \sqrt{(350)(599)} = 458\ Hz$$
$$f_{o2} = \sqrt{(599)(1025)} = 784\ Hz$$
$$f_{o3} = \sqrt{(1025)(1753)} = 1340\ Hz$$
$$f_{o4} = \sqrt{(1753)(3000)} = 2293\ Hz$$

6. Determine required Q

$$Q = \frac{f_o}{BW}$$

$$Q_1 = \frac{458}{249} = 1.839$$

$$Q_2 = \frac{784}{426} = 1.840$$

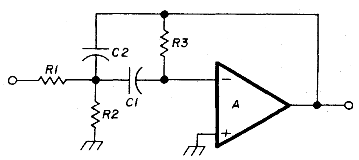
$$Q_3 = \frac{1340}{728} = 1.841$$

$$Q_4 = \frac{2293}{1247} = 1.839$$

Use $Q = 1.84$

appendix 2

The multiple feedback bandpass filter shown below may be designed by the following method (example in brackets):



Choose: $C = C_1 = C_2$

$$[C = 1000\,pF = 10^{-9}F]$$

Let: $H = \dfrac{|A_o|}{Q}$

$$[H = \frac{3.39}{1.84} = 1.84]$$

where A, = desired gain

$Q = Q$ from **appendix 1**

Calculate: $K = 2\pi f_o C$

$$[2\pi \cdot 458 \cdot 10^{-9} = 2\ 878 \cdot 10^{-6}]$$

$$R_1 = \frac{1}{HK}$$

$$\left[\frac{1}{1.84 \cdot 2.878 \cdot 10^{-6}} = 188.8k\right]$$

$$R_2 = \frac{1}{K(2Q - H)}$$

$$[2.878 \cdot 10^{-6}(1.84) = 188.8k]$$

$$R_3 = \frac{2Q}{K}$$

$$\left[\frac{3.68}{2.878 \cdot 10^{-6}} = 1.28M\right]$$

This completes the calculations; the final step is to select the nearest 5 per cent standard resistor values. If, as in the above example, A, equals Q^2, R_1 will equal R_2, which minimizes errors due to tolerance variations.

The following program, written for an HP 25 calculator, will speed the design of the BPF:

HP-25 Program Form

Title Multiple Feedback Bandpass Filter

Switch to PRGM mode, press ☐ PRGM , then key In the program

DISPLAY LINE	CODE	KEY ENTRY	X	Y	Z	T	COMMENTS
00							
01	31	↑					enter f_o
02	02	2					
03	61	X					
04	15 73	gπ					
05	61	X					
06	24 00	RCL 0					
07	61	X					defines K
08	23 06	STO 6					
09	24 05	RCL 5					
10	15 03	g ABS					
11	24 04	RCL 4					
12	71	÷					
13	23 07	STO 7					
14	24 06	RCL 6					
15	61	X					
16	15 22	g 1/x					defines R1
17	23 01	STO 1					
18	24 04	RCL 4					
19	02	2					
20	61	X					
21	24 07	RCL 7					
22	41	–					
23	24 06	RCL 6					
24	61	X					
25	15 22	g 1/x					defines R2
26	23 02	STO 2					
27	24 04	RCL 4					
28	02	2					
29	61	X					
30	24 06	RCL 6					
31	71	÷					defines R3
32	23 03	STO 3					
33	13 00	GTO 00					
34							
35							
36							
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49							

REGISTERS	
R_0	C
R_1	R1
R_2	R2
R_3	R3
R_4	Q
R_5	A
R_6	K
R_7	H

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ham radio

antenna design for omnidirectional repeater coverage

The problem of good coverage with vhf antennas on towers with large cross-sectional areas is resolved in this article

This **is** the story of how one club obtained uniform coverage in all directions with a repeater antenna mounted on the side of a very wide tower. Perhaps the solution will help others with the typical problems of side-mounted antennas.

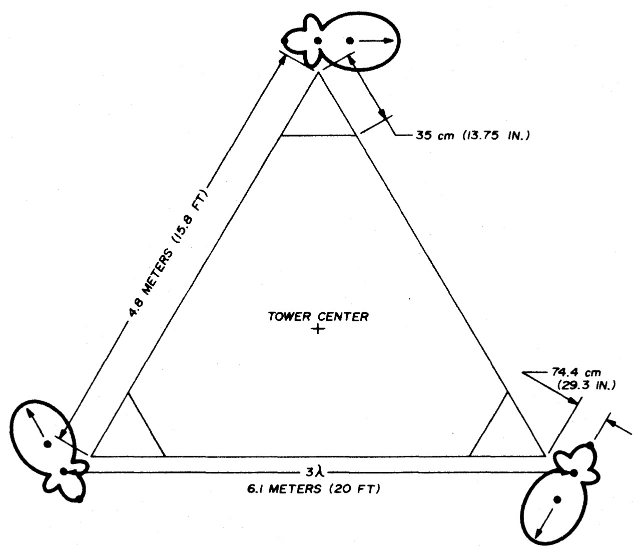
the problem

The difficulty that the Western Illinois Amateur Radio Club (WR9AEA) faced was not an unusual one on side-mounted repeater antennas. Coverage was not uniform in all directions; there were many peaks and many nulls. In some directions range was disappointingly short. Unless the repeater antenna is mounted on top of a structure this situation is typical, because a side-mounted antenna pattern usually has peaks and nulls resulting from the interference, reflections, and absorption of the structure. The local TV station, unfortunately, wouldn't let us put our array on the top of their tower above the TV antenna.

An interesting aspect of the WR9AEA problem was the large cross-sectional area of the TV broadcast tower we're using. The triangular shape is 4.8 meters (15 feet, 10 inches) on each side. Although this tower is very wide for a 244-meter (800-foot) structure, the problem and solution are relevant to both smaller and larger structures.

solution

The solution needed was some type of antenna array all the way around the supporting structure. Minimum coupling to the tower and uniform illumination of the horizon with good input vswr were required. A search of Amateur Radio reference materials yielded no answers. At this point the club



NOTES
1. 63.5 mm (2-1/2 IN.) 00 TUBING AT ALL TRIANGLE CORNERS.
2. ANTENNA HEIGHT' 223 METERS (731 FTJ ABOVE GROUND.
3. OPERATING PARAMETERS:

FREQUENCY (MHz)		WAVELENGTH. CM (IN.)
TRANSMIT'	14703	204 (80.3)
RECEIVE:	14763	203 (79.9)

4. RMS GAIN:
2 LAYER ≈ 0.8 dB
3 LAYER ≈ 3.8 dB

fig. 1. Tangential-fire antenna array using Yagis attached to a tower of large cross section. Note that the main lobe of each radiator is perpendicular to the tower and that free space exists in front of, and to the rear of, each pattern. The resultant radiation pattern of each antenna is summed so that the overall pattern is essentially omnidirectional.

By James R. Ruxlow, N9SN, 8 Elmwood Drive, Quincy, Illinois 62301

president, Tom, W9NJV, approached a local professional antenna engineer, Ron, W9NOO. Ron is very well respected for his many years of designing vhf and uhf broadcast antennas.

As usual, Ron knew what to do. He suggested a "tangential fire arrangement" for mounting antennas on the very large triangular tower. Of course, we didn't know what he was talking about; but as if often the case with someone who really knows his subject, Ron was able to make it simple for us.

description

By "tangential" Ron meant that the radiators would have their maximum radiation on a tangent, or at right angles, to the tower. This seems a little unusual at first, because we normally think in terms of an antenna radiating straight out from a tower. But here, if you're standing at the center of the tower, the maximum energy is pointed off to one side rather than straight out. To obtain constant signal amplitude in all directions, one radiator is placed on each leg of the tower. Notice from **fig. 1** that the main lobe of each radiator is prependicular to the tower and there is free space in front of, and to the rear of, each radiation pattern. The tower structure is off to one side of the radiator, so there's a minimum of coupling and distortion.

pattern sum

To obtain omnidirectional coverage it's necessary for the pattern from one radiator to add to the next, so that the resulting sum is as close as possible to a circle. **Fig. 2** illustrates this concept of the addition of the patterns. (In this figure the patterns are drawn to a very large scale, and the tower triangle to a very small scale, to represent the addition that takes place in the far field.)

The ideal individual radiation patterns would have a *6* dB beamwidth of 120 degrees. The half voltage (−*6* dB) intensity of one radiator would then coincide with the half voltage radiation of the next. If the components from adjacent radiators are in phase, they will then sum to equal the full intensity. **Figs.** 3 and **4** illustrate the development of this concept. Since the cosine function has a value of one-half ±60 degrees, the desired pattern shape is referred to as a cosine pattern." The repeater antenna is vertically polarized, so our concern is the pattern

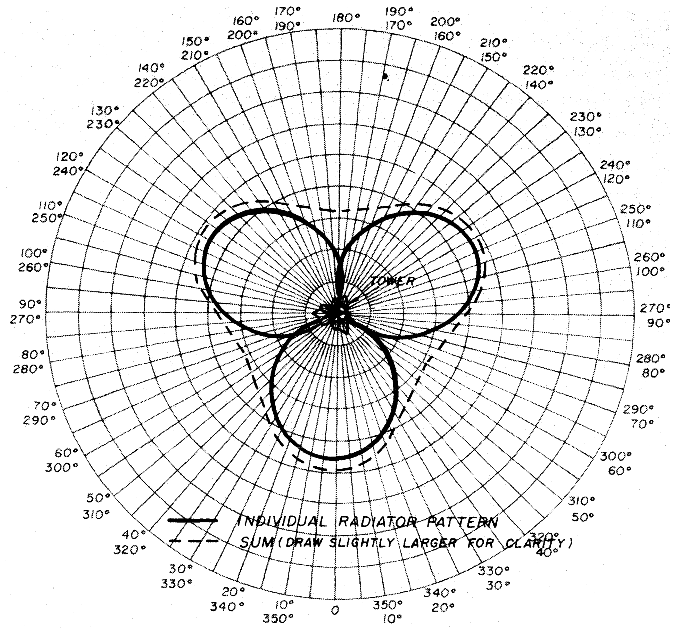


fig. 2. Development of the cosine radiation pattern resulting from three antennas fed with the proper phasing system. The sum of the patterns approaches a circle.

in the plane perpendicular to the radiating elements (**H** plane). Other patterns lend themselves to four or more radiators around a tower.[1, 2]

radiators

Ron told us that the desired cosine-shaped pattern is approximated by the typical short Yagi antenna. We decided to use on each leg of the tower a five-element Yagi manufactured locally. This beam

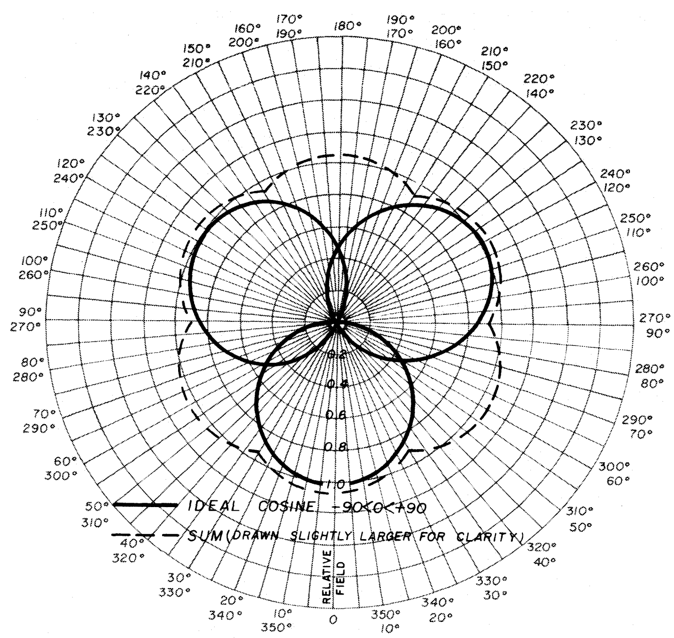


fig. 3. Ideal cosine pattern of three antennas fed in phase. The cosine of the angle, Ø, lies between −90 and +90 degrees.

"Another variation of this concept is the $\cos^2$ pattern, which was developed for vhf antennas on ballistic missiles. The same problem existed: the requirement for omnidirectional coverage with minimum attenuation from antennas mounted on the side of a huge mass of metal (the missile). Much time and effort went into the development of the $\cos^2$ antenna, which is now standard for range safety and telemetry electronics on large rocket launch vehicles. Some of the early work on these antennas was done by the engineering department of the Convair division of General Dynamics for the Atlas missile in the late 1950s. **Editor.**

has standard dimensions with about 9 dB gain. It is very well constructed to take the rigors of being mounted 163 meters (535 feet) in the air. This was an important consideration, because nobody was interested in climbing up there – or paying a professional to go up there – in *windy,* cold weather to tighten a bunch of flapping aluminum.

spacing between radiators

For the amplitudes of the patterns of the radiators to add, it's necessary for the phasing and spacing to be correct. In our case, each Yagi was fed in phase

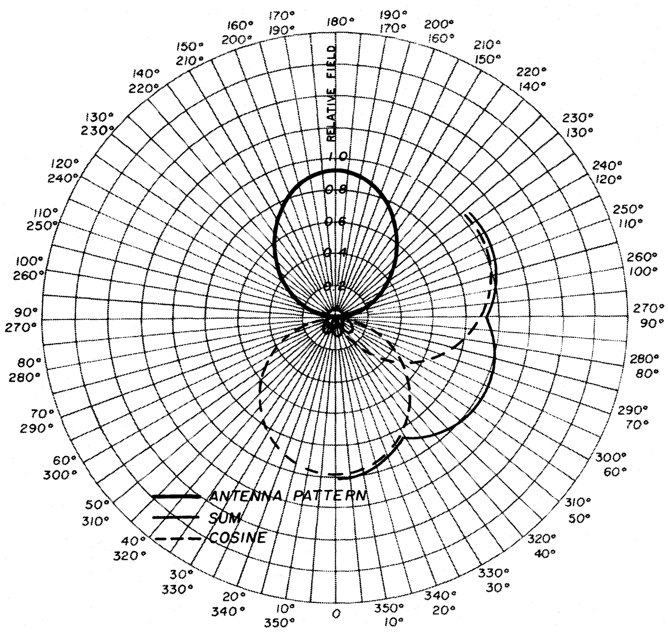


fig. 4. The development of the cosine pattern of n antennas fed in phase. This data is from a pattern recorder used during tests of the WRSAEA two-meter system.

through an equal length of feedline. The center of radiation (driven element) of each beam must be an integral number of free-space wavelengths apart. This requirement assures that the energy of each element will add correctly with energy from the next element. This is represented by the dimension $n \cdot \lambda$ (n times lambda), **fig.** 5. To suspend the Yagis at least one-half wavelength from the tower legs, the spacing worked out in our case to three wavelengths (see **fig. 1**). Ron pointed out that there are techniques for spacing the radiators at any multiple of one-third wavelength.3

gain of the array

At this point some of us got enthusiastic about the gain of this concept. After all, with three 9 dB Yagis the gain should be high, right? Wrong. When the patterns add up to a circle, the average gain drops to that of a half-wavelength dipole. It was hard for

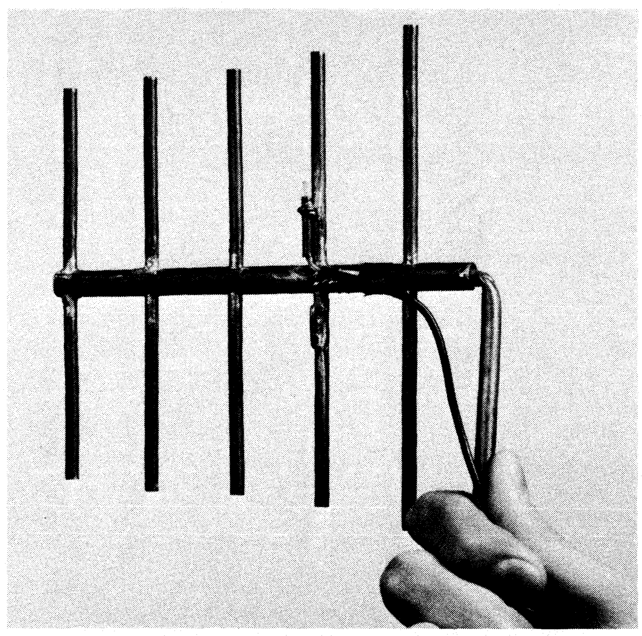
Antenna-mounting hardware consists of stainless steel and heavy-gauge aluminum.

some of us to get this through our thick heads, but the single stack or "bay" of three radiators around the tower yields to gain equal to that of a reference dipole.

Ron pointed out that the addition of a second level, or bay, of three more Yagis, stacked one wavelength above, would double the gain and give 3 dB over a reference dipole. So we decided to build a two-bay system with three Yagis per bay.

scale-model tests

To make sure the thing would work, Ron and his collaborator, Joe Donovan, tested a scale model of



Scale model of one of the antennas used for tests.

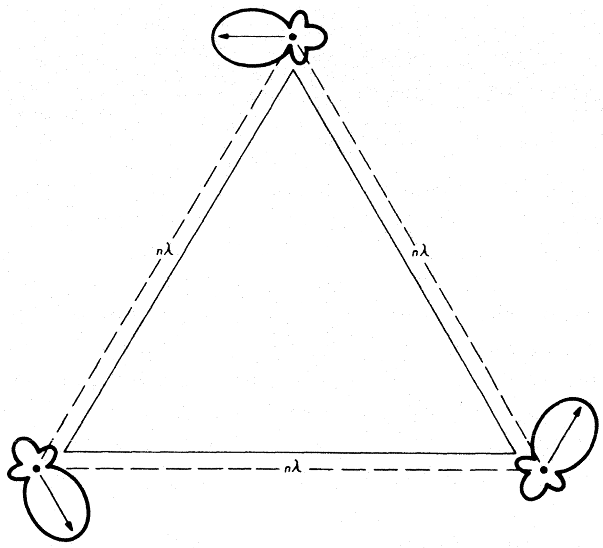


fig. 5. The center of radiation from each beam must be an integral number of free-space wavelengths apart so that the energy of each element (antenna) will add with energy from the next. This is represented by the dimension $n\lambda$, where λ is the spacing in wavelengths.

the tower cross section and Yagi elements. A convenient test frequency for their scale-model antenna was 955 MHz. At this frequency the models are small enough to be easily rotated by a powered turntable. A continuous plotter automatically recorded the pattern shape. **Fig. 6** shows the pattern with three Yagis pointed straight out, or a radial-fire arrangement. Note the nulls down to 20 dB below maximum and broad areas of poor gain. This type of poor circularity is typical of many side-mounted vhf antennas. **Fig. 7** shows the pattern of the tangential-fire configuration used for our new array. The circularity is ±3 dB or better. In other words, the gain in any direction is no more than 3 dB from the average.

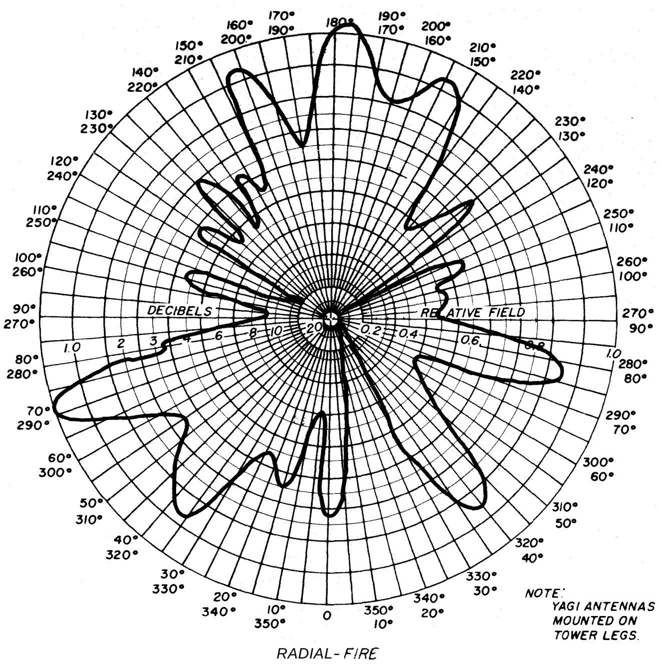


fig. 6. Radial-fire pattern in which three Yagis are pointed straight out from the tower. Note nulls and broad areas of low gain. The poor circularity is typical of many side-mounted vhf antennas.

power divider

A power divider to feed the six Yagis in phase from a single feedline was the next design task. A quarter-wavelength transmission-line transformer is perhaps the most simple technique. If all six Yagis are matched to 50 ohms and fed through convenient, equal lengths of feedline, the feedlines can be paralleled at a single point. Six 50-ohm loads in parallel result in an impedance of 8.3 ohms. In other words, we need an impedance transformation of six to one.

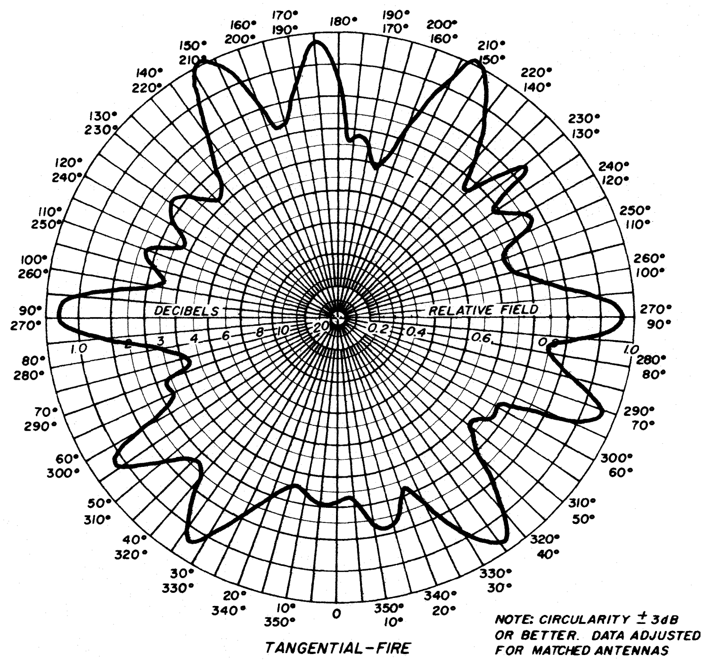
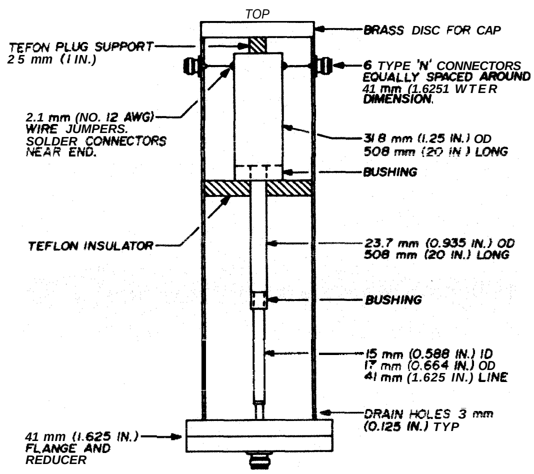


fig. 7. Radiation pattern of the tangential-firs arrangement used at WR9AEA. Circularity is ±3 dB or better, which means that the gain in any direction is no more than 3 dB from the average.

The design curves in Chapter 22 of reference 4 shows about a ±5 per cent bandwidth at a vswr of 1.2 for a six-to-one transformation with a single 1/4-wavelength transformer. The usual equation, $Z = \sqrt{(Z1)\ (Z2)}$ or $Z = \sqrt{(50\ (8.3)}$, tells us that 1/4 wavelength of transmission line, with a characteristic impedance of 20.4 ohms, would match 8.3 to 50 ohms. However, the design curves also show that, by making the transformation in two steps, the bandwidth at a vswr of 1.2 can be increased to ±20 per



NOTES: I
1. 41 mm (1.625 IN.) RIGID LINE WTER DIMENSION. 38.8 mm (1.527 IN.) ID.
2. OVERALL LENGTH 1.2 meters (4 FT)

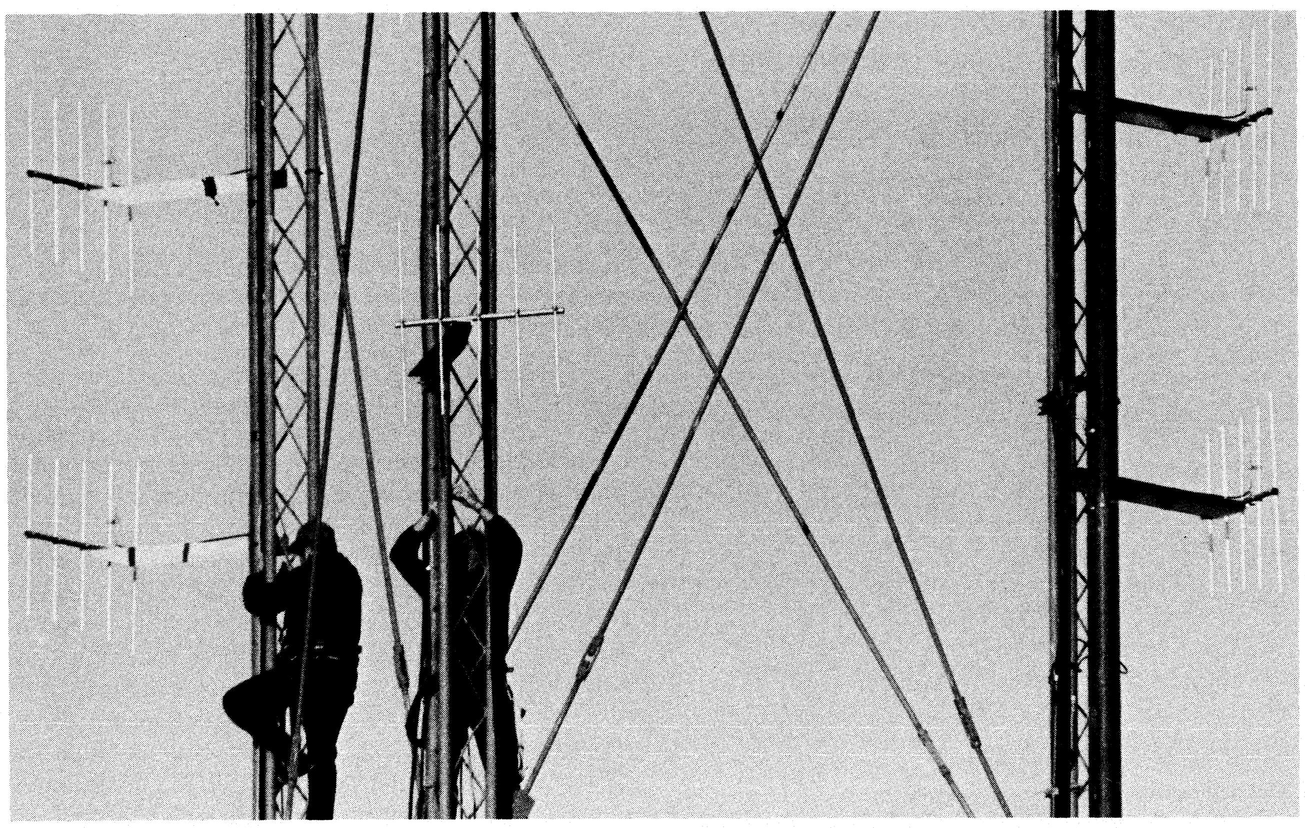
fig. 8. Construction details of the power divider used with the WRSAEA antenna.

cent. This configuration is less sensitive to inaccuracies and changes in the load impedance. Two 1/4-wavelength sections in series match the 8.3-ohm load to an intermediate value to 20.4 ohms, which is in turn matched to 50 ohms. One 1/4-wavelength section has an impedance of $\sqrt{(50)\ (20.4)} = 32$ *ohms* and the other $\sqrt{(20.4)\ (8.3)} = 13$ *ohms*.

divider construction

Fig. 8 shows the construction of the power divider. The 114-wavelength sections are coaxial. Therefore the usual formula $Z = 138\ log\ (db)$ was used to calculate the ratio of the diameter of the outer to inner conductors. it was convenient to construct the outer shell from a piece of 41-mm (1.625-inch) rigid coax line. The 50-ohm type N input was constructed from a 41-mm (1.625-inch) flange, a 41-mm (1.625-inch) reducer and a short section of 41-mm (1.625-inch) inner conductor. The six outputs are type N connectors spaced equally around the circumference at the opposite end. The center conductors of the six type N outputs are connected in parallel with short lengths of 2.1-mm (no. 12 AWG) solid copper wire to the end of the last 1/4-wavelength inner conductor. Some routine lathe work was necessary to construct the inner conductors, brushings, Teflon supports, and end cap.

Rex, K9ZJV, put his workshop facilities to the task of constructing the divider. Initial testing showed a very flat vswr of about 1.22 over the whole 2-meter band. To bring the device up to professional stan-



The WRSAEA array on an fm broadcast antenna tower. Array is at the 20-meter (65-foot) level for testing. The heroes doing their thing for the cause are NSSN, left, and W9NWN.

dards, a stub was added to the input transmission line to reduce the vswr to less than 1.1 from about 142 to 151 MHz. See **fig. 9.**

full-scale tests

The Yagis, mounting hardware, feedlines, and power divider were then mounted on the TV tower at the 20-meter (65-foot) level. Jim, N9SN, and Dave, W9NWN, performed these tasks of installing and adjusting the antennas. This work provided a very important check of all parts of the system before the critical full-height installation. A check of the pattern was made by comparing the signal received from the array with that from a reference Yagi, hand-held out from the tower in the direction of the field-strength meter. Although this method of checking a pattern isn't accurate, seventeen measurements in all directions showed no major peaks or holes. Once all the minor mechanical bugs were corrected, a professional climber was hired to install the array several wavelengths below an fm broadcast transmitting antenna, approximately 163 meters (535 feet) high.

predicted coverage

Ed, W4HTP, calculated the predicted coverage using broadcast techniques. FCC 50/50 curves calculate coverage exceeding 50 per cent of the time in 50 per cent of the potential receiving locations. The calculations consist of two steps: prediction of field strength from the repeater transmitter and determination of field strength required by various configurations of fixed and mobile stations. The results are shown in **table 1.** (Reference 5 and 6.)

table 1. Predicted coverage of the WR9AEA two-meter repeater antenna array for various receiving antennas. Predictions are based on 50 per cent of the potential receiving locations for 50 per cent of the time. Distances are for receiving the repeater.

Example	receiving equipment	required field strength at 9 meters (30 ft)	Distance km	(miles)
1	1/4-wavelength rooftop mobile	+21 dBμV/m	77	(48)
2	5/8-wavelength rooftop mobile	+18 dBμV/m	85	(53)
3	Ringo at 9 meters (30 ft)	+1 dBμV/m	144	(90)
4	11-element beam at 12 meters (40 ft)	−12 dBμV/m	216	(135)
5	2 stacked 11-element beams at 24 meters (80 ft)	−21 dBμV/m	280	(175)

results

Results have been excellent. Coverage in all directions seems to bear out the predictions. Mobile coverage is 72-88 km (45-55 miles); fixed stations at 160 km (100 miles) check in regularly. There appear to be no holes in the pattern. All bad spots seem to be explained by local terrain. We hope our experience and the references will help other groups to obtain omnidirectional repeater service.

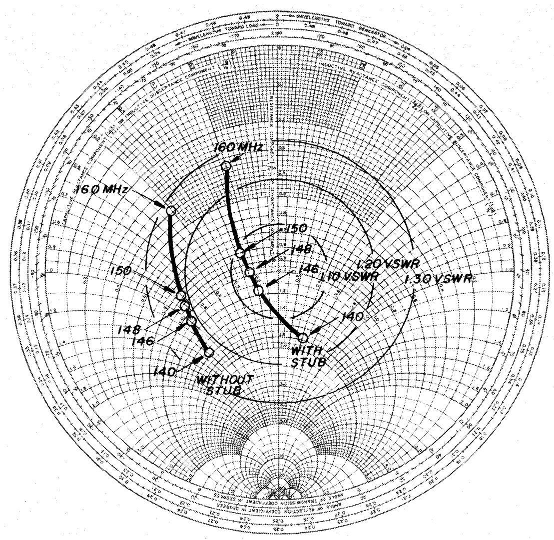


fig. 9. Response of the power divider with and without a stub on the transmission line.

acknowledgments

The Western Illinois Amateur Radio Club wishes to thank Ron Fisk, W9NOO, for his professional guidance. We also thank the club president, Tom, W9NJV, for his help during this project. Tom deserves public recognition for his constant, active leadership in getting everybody to work together. Photo credits to to Roger Humke, WA9KRG.

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ham radio

exposure to radio-frequency generating equipment: is it safe?

Review of the literature has produced some interesting observations on the hazards of EMR — here's a report of the latest information on this controversial subject

Those of us involved with electromagnetic radiation (EMR) for business or pleasure have, in the last few years, become disturbed by certain questions, which grow louder and more insistent as time passes. These questions address the safety, or lack of safety, associated with exposure to EMR. The principal target of these inquiries has been in the area of microwave radiation, but there has been increased interest in the matter of safety within the more usual communication frequencies: uhf, vhf, and high frequency. These questions have been amplified, perhaps more than they deserve, by the press; and it is becoming harder and harder to ignore them.[1] You find yourself wondering whether the push-to-talk button and antenna on a handheld vhf transceiver might better be replaced by a trigger and gun barrel respectively.

While under most conditions we can't "feel" EMR, we're becoming more aware of its presence. This is an unwelcome occurrence, since most of us have always thought of shortwave radiation as being nothing if not safe. We all live most of our lives within a constant "fog" of EMR of many frequencies. In fact, we're constantly affected by the information transmitted by radio waves. The suspected damage that may be caused by rf energy varies from the "blahs" to cataracts, heart disease, cancer, impotence, and birth defects. What's worse, while most of this is unproven, it hits the lay press with the impact of fact.

By Steve Kraman, MD, **WA2UMY**, 2901-B Candlelight Way, Lexington, Kentucky 40502

Even the more responsible reviews of the subject offer no assurance that we may continue our use of radio with impunity.[2,3]

Distrust of the establishment also has had a hand in promoting doubt among the rf-irradiated public. We mistrust our government when we hear that the Russians have a standard for safe rf fields that is one thousand times smaller than ours, and that they believe that many adverse mental health and physical effects can be produced in man by low-dose rf exposure. We read that the Russians have been beaming microwave radiation (for questionable reasons) at the U.S. embassy in Moscow for years; then we hear that more cases of cancer may occur among the embassy workers than can be explained by chance. When the United States government chalks this up to chance, we wonder.[4]

The purpose of this article is not to reassure you that there's nothing to worry about; not quite enough information is available for that yet. Nor is the article intended to alarm, since I don't believe that the known facts justify that either. The following text reviews what is known to date about EMR and its effects on biological systems. It separates fact from fiction and suggests a reasonable response by those who must swim in the sea of EMR that surrounds us.

history

Interest in the biological effects of EMR has waxed and waned considerably since the first true electromagnetic field was generated by Hertz in 1888. Some research was done in 1891 by D'Arsonval and Tesla, but this was essentially the only work done in this area until the 1930s. Before this, in 1891, when electric lights were installed in the White House for the first time, they were not placed in the rooms that the President used frequently, because they were considered potentially dangerous. During the 1930s interest in rf radiation began to grow, spurred by the development of high-power transmitting techniques.

But World War II caused this research to grind to a halt in the favor of work of a more certain and conventionally destructive nature. At the end of World War II we had a new toy to learn about and with which to experiment — nuclear energy. However, this knowledge didn't stop progress in the field of radar and communications. The existence of high-power rf-generating equipment began raising safety questions, principally within the military community.

The Tri-Service study. In 1956 the Tri-Service program was established, coordinated by the Air Force. Its purpose was to conduct research to determine the biological effects of nonionizing radiation. This research effort lasted four years and four annual conferences were held. The outcome of the Tri-Service program was to suggest that there was no evidence implicating levels of electromagnetic radiation beiow 100 mW/cm^2 in damage to living tissues.

The implication that EMR could cause damage only in its capacity to heat was clear. It's of importance, however, that very little, if any, of this research was done at levels below 100 mW/cm^2 and, indeed, most of it was between the power levels of 300-400 mW/cm^2. The Tri-Service study, then, addressed only the problem of thermal effects and assumed this to be the only danger. The government accepted this opinion, and, partly as a result, little further research was done in this country from 1960 to 1970.

Federal EMR legislation. During the present decade, interest in the biological effects of EMR has escalated steadily, primarily because of technological advancements that have resulted in increased exposure to EMR by the general population. The skyrocketing popularity of CB radio and microwave cooking account for much of this increased exposure. The use of high-power communications and radar equipment has also heightened concern for personnel in the military. Additionally, several pieces of federal legislation have stimulated research in the field by calling for protection of the public from all sources of radiation, including ionizing, nonionizing, sonic, and ultrasonic devices. These federal acts are the Radiation Control for Health and Safety Act of 1968, the National Environmental Policy Act of 1969, and the Occupational Safety and Health Act of 1970. They require that users of EMR-generating equipment demonstrate the safety and effects on the environment of their equipment. Yet another factor that helped spur American researchers was the presence of a large body of Soviet-bloc literature that points to conclusions much different from those of the Tri-Service program and imply that very low levels of EMR (by our standards) could be dangerous. These studies, while for the most part poorly controlled, carried out, and reported, could not be totally ignored, since many of our own studies suffer from the same shortcomings.[5] The Russian studies are covered in more detail later.

Much has been published on the hazards of exposure to electromagnetic radiation, from low-power, low-frequency equipment to high-power microwave devices. The author of this article is a medical doctor and a Radio Amateur interested in this controversial subject. He has researched the available literature (domestic and foreign) on the subject. This article sums up the results of that research. The conclusions imply that Amateur Radio transmitting equipment probably does not impose health hazards on humans provided certain safety considerations are observed. **Editor.**

physical characteristics of EMR

EMR must be distinguished from ionizing radiation (x-ray, nuclear), since its effect on molecular structure is much different. Nuclear radiation causes no significant heating of the irradiated object. Instead, its photon energy is sufficient to disrupt the atomic bonds, thereby causing ionization and damage to the molecular structure. If this molecule is part of a living cell, it may become damaged, die, or its genetic material may be changed. EMR doesn't cause these effects, because the photon energy of even microwaves is so small that it causes no ionization. The energy absorbed, however, can increase the speed of molecular vibration, thereby causing an increase in temperature. The more energy absorbed, the more heat produced. To date, all known damage by EMR *seems to be* the result of this heating. This effect is clearly seen through the window of a microwave oven.

table 1. Properties of electromagnetic waves in biological media*

			muscle, skin, and tissues with high water content						
frequency (MHz)	wavelength in air (cm)	dielectric constant ϵ_H	conductivity σ_H (mho/m)	wavelength λ_H (cm)	depth of penetration (cm)	reflection coefficient air-muscle interface τ	ϕ	muscle-fat interface τ	ϕ
1	30000	2000	0.400	436	91.3	0.982	+179		
10	3000	160	0.625	118	21.6	0.956	+178		
27.12	1106	113	0.612	68.1	14.3	0.925	+177	0.651	−11.13
40.68	738	97.3	0.693	51.3	11.2	0.913	+176	0.552	−10.21
100	300	71.7	0.889	27	6.66	0.881	+175	0.650	−7.96
200	150	56.5	1.28	16.6	4.79	0.844	+175	0.612	−8.06
300	100	54	1.37	11.9	3.89	0.825	+175	0.592	−8.14
433	69.3	53	1.43	8.76	3.57	0.803	+175	0.562	−7.06
750	40	52	1.54	5.34	3.18	0.779	+176	0.532	−5.69
915	32.8	51	1.60	4.46	3.04	0.772	+177	0.519	−4.32
1500	20	49	1.77	2.81	2.42	0.761	+177	0.506	−3.66
2450	12.2	47	2.21	1.76	1.70	0.754	+177	0.500	−3.88
3000	10	46	2.26	1.45	1.61	0.751	+178	0.495	−3.20
5000	6	44	3.92	0.89	0.788	0.749	+177	0.502	−4.95
5800	5.17	43.3	4.73	0.775	0.720	0.746	+177	0.502	−4.29
8000	3.75	40	7.65	0.578	0.413	0.744	+176	0.513	−6.65
10000	3	39.9	10.3	0.464	0.343	0.743	+176	0.518	−5.95

Importance of EMR absorption. Depending on the EMR frequency, the size and character of the target, and the presence of other objects, a certain amount of energy will be absorbed and the rest will be reflected or refracted. Only the energy absorbed by the object affects it, and this has been one of the problems in microwave research. Most studies conducted to assess the effects of microwave radiation on biological subjects measure the field strength of the electromagnetic field, even though the actual amount of energy absorbed is unknown. Not only is the absorbed dose important, but so is the *size* of the subject and the EMR *frequency,* since the depth of penetration decreases as the frequency is increased.

An animal may handle a heat load more easily if the heat load is applied to its skin, where air cooling occurs, than if the heat load is developed internally within vital organs that are cooled only by blood circulation. In this way EMR research is years behind nuclear research, which has long recognized and used the *rad,* a unit of absorbed energy, when referring to exposure to radioactive materials. The *roentgen,* a unit of emitted energy, is fine for describing the generating equipment but says little about its effect on the person receiving the radiation. Recent studies in the field of EMR have used methods to calculate or measure the actual absorbed dose of radiation.

Near- and far-field considerations. Another factor affecting the field density is whether the object is in the near or far field. The far field is that distance (generally more than one wavelength) beyond which the electric and magnetic fields are coherent and in phase. The field impedance is constant in the far field, so that measurement of either the electric or magnetic component will be proportional to and will determine the power density. In the near field, however, the electric and magnetic fields are out of phase, and it becomes more difficult to measure power density. In this situation it's more convenient

"Tables reprinted from "Non-Ionizing Electromagnetic Wave Effects In Biological Materials And Systems" by C. C. Johnson and A. W. Guy, *Proceedings of the IEEE,* Vol. 60, No. 6, June, 1972.

to measure volts per meter (V/m).[6] This is of little consequence with microwaves, because the far-field situation exists at distances of only a few meters or so from the antenna. However, at most communications frequencies, the object in question is often in the near field. At a wavelength of 80 meters for instance, the entire dwelling and many of the neighbors may be in the near field, and this makes research into near-field phenomena quite important.

The so-called nonthermal effects of EMR are the more controversial aspects of the subject. The present research push is to discover if these effects exist and, if so, whether they offer significant health hazards. This research is necessary because most, if not all, past studies reporting to show nonthermal effects ignored regional temperature changes caused by concentration within objects. The studies were poorly or not controlled, entirely anecdotal in nature (and therefore impossible to evaluate), or were so incompletely described that they can't be duplicated.[7]

The scientific method demands that, as much as possible, all factors other than that being evaluated be accounted for and set aside to attribute a possible effect to a certain cause. This is impossible to do with certainty, so the importance of statistical analysis (to determine the possibility that chance alone caused a certain effect), and repetition by other experimenters, is of extreme importance if you don't want to be led astray. This is the essence of the scientific method, and while well understood for decades, it's often overlooked by scientists who may be in a hurry and feel secure that they can be objective.

The Soviets recognized the effects of EMR on human nervous tissue as far back as 1937, when Turlygin found that excitability of the central nervous system was increased when a spark oscillator was switched on near the subject's head (not a totally unexpected effect). Since then, with the exception of the war years, Soviet-bloc literature in this area has poured out in increasing quantities.

While many effects have been reported, those most frequently noted involve the central nervous system. These reports include frequencies from 30 to 30,000 MHz and power ranges of microvolts to tens of milliwatts/cm^2. Unfortunately, and as previously mentioned, most of these reports lack data without which intelligent evaluation is impossible; *i.e.*, frequency, power, waveform, orientation of the body with respect to the beam, and type of experimental animal used. Many of the reports involving people exposed to EMR quote a wide range of subjective complaints such as headache, weakness, depression, trembling, chest pains, inhibition of sex drive, inability to make decisions, general tension, and sense of anxiety. Other more objective findings reported are asthma, fast or slow pulse rate, high or low blood pressure, and **EKG** changes.

The belief that these ailments are being caused by EMR exposure is so strong in the Soviet Union that exposed workers can get the day off with pay if they complain of them. There is, however, a question regarding the willingness of Soviet plant managers to admit they've been exposing workers to higher-than-permitted levels of electromagnetic radiation for fear of losing their jobs. In the realm of parapsychology,

table 2. Properties of electromagnetic waves in biological media*

		fat, bone, and tissues with low water content							
						reflection coefficient			
						air-muscle interface		muscle-fat interface	
frequency (MHz)	wavelength in air (cm)	dielectric constant ϵ_L	conductivity σ_L (mho/m)	wavelength λ_L (cm)	depth of penetration (cm)	τ	ϕ	τ	ϕ
1	30000								
10	3000								
27.12	1106	20	10.9-43.2	241	159	0.660	+174	0.651	+169
40.68	738	14.6	12.6-52.8	187	118	0.617	+173	0.652	+170
100	300	7.45	19.1-75.9	106	60.4	0.511	+168	0.650	+172
200	150	5.95	25.8-94.2	59.7	39.2	0.458	+168	0.612	+172
300	100	5.7	31.6-107	41.0	32.1	0.438	+169	0.592	+172
433	69.3	5.6	37.9-118	28.8	26.2	0.427	+170	0.562	+173
750	40	5.6	49.8-138	16.8	23	0.415	+173	0.532	+174
915	32.8	5.6	55.6-147	13.7	17.7	0.417	+173	0.519	+176
1500	20	5.6	70.8-171	8.41	13.9	0.412	+174	0.506	+176
2450	12.2	5.5	96.4-213	5.21	11.2	0.406	+176	0.500	+176
3000	10	5.5	110-234	4.25	9.74	0.406	+176	0.495	+177
5000	6	5.5	162-309	2.63	6.67	0.393	+176	0.502	+175
5800	5.17	5.05	186-338	2.29	5.24	0.388	+176	0.502	+176
8000	3.75	4.7	255-431	1.73	4.61	0.371	+176	0.513	+173
10000	3	4.5	324-549	1.41	3.39	0.363	+175	0.518	+174

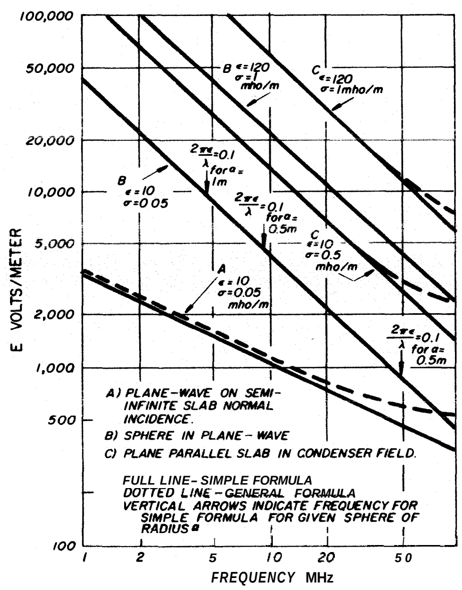

fig. 1. Variation with frequency of electric-field strength to produce rate of rise of temperature at 1°C/hr, in: (A) The skin depth of semi-infinite layer of lossy dielectric in plane-wave field, (B) Sphere of lossy dielectric in plane-wave field. (C) Slab of lossy dielectric in capacitor field.

many Russian experimenters believe not only in esp, but also that it is effected through microwave transmission and reception from brain to brain!

The Eastern European literature describes microwave effects on many constituents of the blood, on gland functions, eyes (cataracts), and reproduction (sterility, altered development of the fetus, altered sex ratio of births [with girls predominating]).[7] These results over many years have led the Soviet-bloc countries to adopt a maximum permissible dose (MPD) of EMR of 0.01 mW/cm^2 (one one-thousandth of the U.S. MPD). While this is looked upon with considerable disdain by Western scientists, the remarkable consistency of the many reports can't be discarded out of hand. This has led many U.S. investigators to run better controlled studies to try to prove or disprove these reports.

Recent studies, mostly from the West but also notably from Poland, have shown higher degrees of control and sophistication and are probably more reliable than older reports. More accurate generating and measuring devices are being developed and used, and we are becoming more aware of the physics of EMR and how the conditions of the experiments affect the rf fields produced and power absorbed. Grants from the National Institute of Health and Public Health Service, among other agencies, have spurred research in this field. Most of the work has been with microwaves, and I will outline some of it here.

cataracts

It's generally accepted that microwave radiation can cause cataract formation (opacities in the lens of the eve), Reports of this affliction occurring in relation to radar work or exposure to microwave ovens have been fairly well documented. Also, many animal studies have been made to determine the basis of this effect. While all is not known with respect to this matter, it seems quite certain that the formation of cataracts results from the heating of the lens in a strong microwave field (that is, a field not associated with correct use of properly operating equipment).

A fairly well documented case was that of a woman whose microwave oven leaked considerable radiation while the door was being opened. The leve! of radiation was 40-60 mW/cm^2 while opening and 1-2 mW/cm^2 while closed. She used it for years that way and developed cataracts in both eyes described as "typical microwave cataracts."

It must be mentioned that not all ophthalmologists agree that such cataracts are typical, and many people develop cataracts with age. Other reports of documented cataract formation are those of radar workers when abnormally over exposed (looking into the waveguide).

The optic lens is susceptible to selective heating because circulation is practically nonexistent. Therefore, it has limited capacity to dispose of heat loads. There is no evidence that nonheating levels of microwave radiation can cause damage, and the levels generated by a properly operating microwave oven are far below this point (1 mW/cm^2 when new; 5 mW/cm^2 when used — measured at 5 cm (2 inches) from the door).[8]

effect on the nervous system

Studies of the effects of microwave radiation on the brain have been spurred by the large number of reports in the Soviet literature of emotional and performance changes and the frequent reports of people who can "hear" microwaves — specifically radar. These studies conclude that many persons indeed can hear pulsed microwave energy. But they hear it at the frequency of modulation and cannot detect CW radiation. The actual cause of this is still not clear, but it may be due to selective heating and cooling of certain nerve cells in the ear or brain causing vibration that is detected as sound.[9] Actual brain damage has been demonstrated only with levels of radiation far in excess of the present U.S. safety limit.

Studies on the activity of mice subjected to low-power microwave radiation have been contradictory (some show no effect, some show decreased activity). These have been carried out at different frequencies and power densities so that further research is certainly needed to clear the air.

blood-forming system

Several studies have been done to assess the effects of microwaves on the blood and especially on the white blood cells, which are responsible for protection against infection. Most of these studies show little or no effect and the importance of this is questionable so far. More research is also needed in this area.[10]

reproduction

Reproduction research is of obvious importance because of the known sensitivity of the fetus to subtle changes in its environment. Caution in this matter is extreme. Witness the fact that there is not even one drug that is known to be safe for use during pregnancy. A potential risk is always assumed. Radio-frequency energy in the microwave range, however, is probably not a danger to pregnant women, because the energy is absorbed by more superficial tissues (penetration of EMR decreases as frequency increases). The effect on male fertility is real, however, since the testes are heat sensitive and must exist in an environment cooler even than body temperature to produce sperm. This is the reason for their location. Heating of the testes by microwaves or anything else will cause sterility, but this is temporary unless extreme heating occurs or exposure lasts over many months to years. So, while all the information is not in, there seems to be little or no danger to reproduction from current microwave exposure levels.

uhf, vhf, hf studies

Much less work has been done at these frequencies than at the microwave level. However, it's important to explore this area, since we're exposed more to EMR in this part of the spectrum and such energy can penetrate deeper into the body than microwaves do. As previously mentioned, the near field is more significant at lower frequencies because it occupies more space.

behaviorial effects

Little has been done in this area. One study of interest exposed rats to low intensity (0.5 mW/cm^2) 300-920 MHz radiation for 40 days. While the rats were probably in the near field (and this was really not accounted for), certain effects were noted:

1. Lower levels of activity

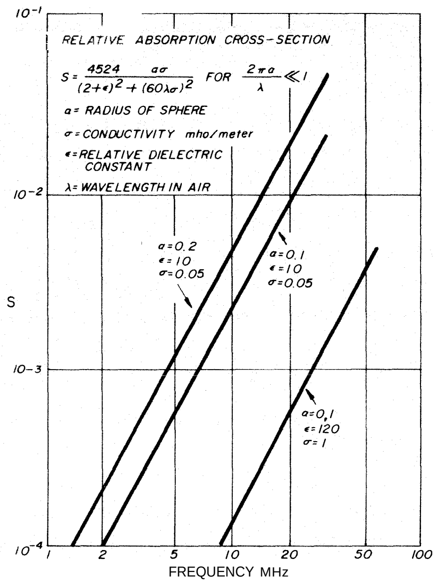

fig. 2. Variation with frequency of the radiation absorption cross section of a sphere of lossy dielectric.*

2. Greater emotionality

3. Longer period of time needed to recover from an electrically induced convulsion

4. Longer time to learn to swim in a water maze

5. Difference in weight of the adrenal gland[11]

This study raises more questions than it answers.

Another study measured egg production of hens exposed to various frequencies of emf from 260 MHz - 2.435 GHz at levels of power not thought to produce heating effects. All hens except those exposed to 915-MHz radiation layed fewer eggs during the experiment.[12] What this means in practical terms is far from clear.

The next study is perhaps more helpful to us. It was carried out by S.T. Rogers in England in an attempt to estimate the danger (if any) to shipboard personnel exposed for long periods of time to emissions of the shipboard radio. Since exposure under these conditions is mainly under the near field, it had to be determined how this exposure would compare to far-field exposure where the power density may be easily measured.

The present radiation power limit of 10 mW/cm^2 is based on the increase of body temperature of a person by l-degree centigrade, while one-half the body surface area is exposed to the source of electromag-

"Graphs reprrnted from "Radio Frequency Radiation Hazards To Personnel At Frequencies Below 30 MHz" by S J Rogers, *Biological Effects And Health Implications Of Microwave Radiation* (Symp Proc Med College of VA, Richmond, VA, Rep RRH/DBE 70-2) pages 222 232. September, 1969 netic radiation. This is considered to be the highest safe exposure. When exposed to near-field radiation from a whip antenna, power density measurements become very complex and are difficult to relate to far-field density.

Rogers contends that "the electrical properties of human tissues show that they resemble lossy dielectrics and that any heating due to rf radiation would be a function of the electric component of the field." His theoretical and experimentai approach to this subject is complex and elegant, and I refer those with more curiosity and a mathematical inclination to the original article.

Rogers concludes that, to cause a temperature increase of 1 degree centigrade per hour in a test liquid in a near field, a field strength of about 2840 V/m would be necessary. He further states that, to allow for a margin of error, a field strength of 1000 V/m would be a convenient and reasonable limit. To convert this to practical terms, the electric field was measured at various distances from a whip antenna radiating at an output power of 1000 watts at frequencies from 2.1225-21.480 MHz. At all frequencies above 4.455 MHz, the field strength at 1.5 meters (5 feet) from the antenna was less than 100 V/m, but this rose sharply as distance decreased. At 4.455 MHz the field strength at 1.5 meters (5 feet) was exactly 100 V/m, and at 2.1225 MHz it was about 500 V/m.[13]

On the basis of Rogers' findings, there would be relative safety in most situations an Amateur Radio operator may find himself in — the most caution to be observed at the lowest frequencies. It cannot be concluded, however, that distortion of the field by other objects would not focus rf energy to higher intensities than expected.

conclusion

Obviously, much work remains to be done in the field of EMR, its effects on biological systems, and on the safety of those exposed to it. I believe that it will be many years before anyone can say with adequate experimental support that our use of EMR is safe to us and future generations.

I think it safe to say that the lack of clear, nonthermal effects of EMR, despite many studies searching for it, supports the conclusion of the Tri Service program, which in 1960 said that "no data" was obtained to invalidate the safety level of 10 mW/cm^2."

We should remember, however, that distorted rf fields may focus power within objects, and that certain organs, and the fetus, are more susceptible to thermal damage.

I feel fairly secure in the use of Amateur Radio equipment in the way it's commonly employed, *i.e.*, high-power equipment radiating through antennas outside of the shack and some distance in the air, and low-power vhf and uhf transceivers used close to the body.

I urge avoidance of the following situations, due to knowledge of danger or insufficient studies:

1. Avoid high-frequency, **high-power** equipment with antennas in the shack within 3 meters (10 feet) of living areas.

2. Avoid direct radiation to the eye by a transmitter in the microwave region ("looking down the horn").

3. Avoid prolonged close contact with any antenna radiating more than minimal amounts of energy.

4. Women in the early months of pregnancy, or those who may become pregnant, should avoid contact with strong hf, vhf, and uhf fields.

I believe these are reasonable precautions that should cause no one much hardship, while allowing continued enjoyment of Amateur Radio equipment.

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ham radio

uhf and microwave frequency counters

A discussion of frequency counters and counting techniques for use above 500 MHz

Today's frequency counters have upper frequency limits ranging from 1 MHz to 24 GHz. Much has been written in the various Amateur Radio journals describing counters that perform up to 500 or 600 MHz, but relatively little has appeared heretofore explaining the techniques by which higher frequencies are measured. This article is intended to supplement one previously published[1] and to explain the methods used today, and in the past, that permit uhf and microwave frequency measurements to be accomplished.

Although most lower-priced counters that can measure frequencies in the 500-MHz region use prescaling, state-of-the-art digital components in use today permit *direct* counting to well over 500 MHz. Frequency counters with ranges greater than this arbitrary, if not completely accurate, 500-MHz limit employ one of the following frequency-extension techniques:

1. Prescaling, which can extend the frequency range to about 1.5 GHz (although indications are that frequencies over 2 GHz will be practical within a year)

2. Transfer-oscillator down-conversion, which can extend the frequency range to over 40 GHz

3. Heterodyne down-conversion, which can extend the frequency range to about 18 GHz.

Prescaling is the simplest and most familiar technique used to extend the range of a direct counter. It entails scaling, or dividing, the input frequency down to one which is within the frequency range of the direct-counting logic in the counter. The dividing factor may be any integral number. If the prescaler is external to the counter, it will usually divide by ten or one hundred, so that the frequency can be read directly from the counter after you have mentally multiplied the counter reading by ten or one hundred, as applicable. If the prescaler is built into the counter, it may scale by any integral factor.

The advantage of using an external prescaler is obvious — it permits extending the frequency range of an existing counter at relatively low cost. Its disadvantages become equally obvious after it has been used. First, there is the necessity of mentally moving the decimal point, since the counter is actually displaying the divided input frequency. Second, one digit of resolution is lost for every decade of scaling. For example, a 900,000.208-kHz signal measured with a scale-by-ten prescaler will read 90,000.021 on a counter having a I-second gate time (1-Hz resolution). Multiplying by ten yields a frequency of 900,000.21 kHz; the I-Hz resolution is lost by scaling. It can be re-established only by increasing the gate time by a factor of ten, provided the counter has that capability.

If the prescaler is an integral part of the counter, mentally scaling the frequency and moving the decimal point is eliminated, since this will be accomplished in the counter when the mode is changed from direct count to prescaled count. Nevertheless, the loss of resolution remains. It can be reduced however, by scaling by a factor of less than ten and simultaneously increasing the gate time by the same factor, as shown in fig. 1.

Suppose that the internal prescaler divides the

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input frequency by four. If the time-base frequency is also divided by four, the gate time is increased by the same factor and there will be no change in the number of signal pulses gated through to the decade counters. Thus, prescaling is accomplished with only a fourfold increase in gate time and no loss in resolution.

Switching from direct to scaled operation may be carried out in one of three ways. If a single input connector is used, the counter mode is generally manually switched. If two separate input connectors are employed, one for low-frequency signals and the other for high-frequency inputs, the counter mode may be switched either manually or automatically when the input signal is present at the high-frequency port.

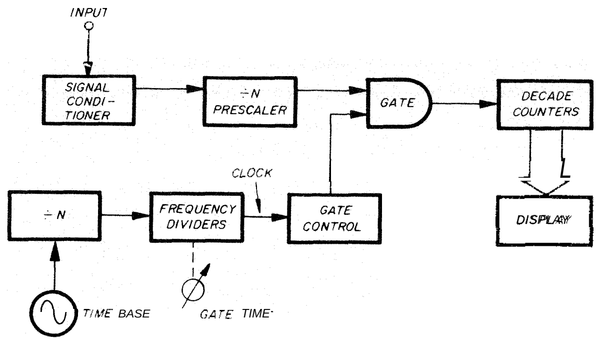


fig. 1. Block diagram of a frequency counter with an internal prescaler.

Fig. 2 shows the block diagram of a counter which employs automatic switching between separate direct and prescaled input connectors. The switches at the time-base output are actually logic circuits, but are shown as conventional switches to simplify the diagram. If there is no signal applied to the high-frequency input, or if the signal amplitude is below a pre-established level, the switching logic will connect the time-base oscillator directly to the frequency dividers. In that state, the counter will function in its direct-count mode.

When a signal of sufficient amplitude is applied to the high-frequency input, the threshold detector actuates the switching logic to connect the time-base output through the divide-by-N circuit before it reaches the frequency dividers. The prescaler output is fed to an appropriate point in the low-frequency signal conditioner. Thus, the counter is switched to its scaled mode automatically whenever a usable signal is connected to the high-frequency input.

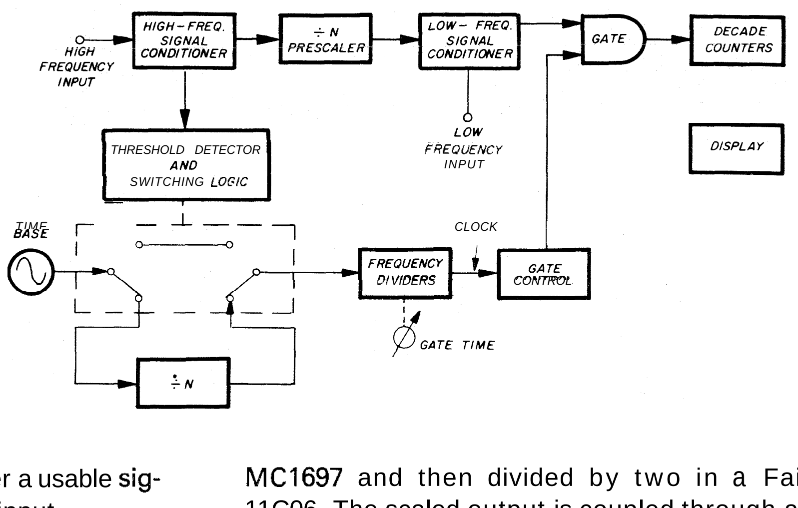


fig. 2. Block diagram of a frequency counter which employs automatic switching from direct to scaled count. The switches shown schematically are actually logic circuits rather than mechanical switches.

A circuit of this type, which I have incorporated in my homebuilt nine-digit counter, is shown in **fig.** 3. Since the direct-count limitation of the counter is about 200 MHz, it was necessary to scale by eight in order to achieve the design objective of 1300 MHz using available devices. The count limit is actually over 1450 MHz.

The prescaler input impedance is nominally 50 ohms. The use of a 3-dB pad between the input and the Amperex ATF417 amplifier keeps the input VSWR at less than 2.1:1 over its entire frequency range. The sensitivity of the prescaler is between 10 and 25 millivolts rms (depending on frequency) between 100 and 1000 MHz, rising to 100 millivolts at 1300 MHz. The decrease in sensitivity is attributable to two factors. First, the ATF417 is designed to cover the 40- to 860-MHz range, so that its gain drops off from 25 dB at 860 MHz to approximately 15 dB at 1300 MHz. Second, the Motorola MC1697 is guaranteed only to 1000 MHz, although it typically clocks to over 1500 MHz. However, as might be expected, the threshold level increases above 1000 MHz.

As is apparent from the schematic, the amplified signal from the ATF417 is divided by four in the MC1697 and then divided by two in a Fairchild 11C06. The scaled output is coupled through a small capacitor to a suitable point in the low-frequency signal conditioner. Because I did not want to add a negative supply for the ECL integrated circuits in the prescaler, I chose to power the devices from the +5 volt and +24 volt supplies already in the counter.

This necessitated capacitive coupling between the MC1697 and the 11C06, since the former requires a supply of 6 to 7 volts. Bias at the clock input of the 11C06 is optimized by means of the 2.5-kilohm pot.

Automatic gate-time switching is accomplished by dividing the clock frequency by eight when an input signal of sufficient amplitude is applied to the prescaler input. An LM311 comparator is configured so that when there is no prescaler input, the positive dc at the comparator's noninverting input exceeds that at the inverting input and keeps the output high, This inhibits both the 11C06 (via pin 9) and the 7493 (via pin 3), and also enables a path from the clock input terminal to the clock output terminal through two gate sections of a 7402, which has no effect on the clock frequency.

When an input signal is present at the prescaler input, a portion of the amplified signal from the output of the ATF417 is sampled and applied to a Hewlett-Packard 5082-2835 hot-carrier diode for rectification. The resultant negative dc, applied to the noninverting input of the comparator, causes the comparator output to go low. This enables the 11C06 and the 7493, and inhibits the direct path between the clock input and clock output terminals. The clock frequency is scaled in the divide-by-eight section of the 7493 and applied to the counter logic from the clock output terminal.

The MC1697 is prone to false counting below 100 MHz and when the input signal amplitude is too low. To prevent false readings, the comparator voltage reference is set by a 1-kilohm pot at the inverting input to establish a threshold level below which the prescaler is inhibited and above which erroneous readings will not occur.

transfer-oscillator down-conversion

One of the earliest methods of measuring frequencies in the uhf and microwave regions was by means of the manual transfer oscillator. The transfer oscillator was completely separate from the counter. It consisted of a stable VFO (typically 100 to 200 MHz), a harmonic mixer, and a zero-beat indicator, usually a cathode-ray tube.

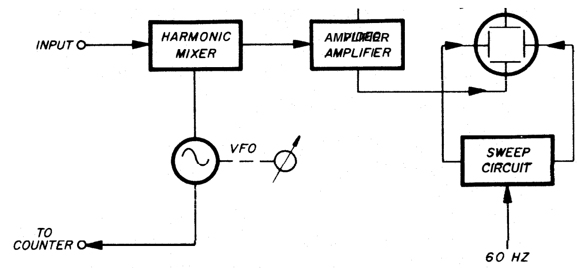


fig. 4. Simplified block diagram of a manual transfer oscillator. The frequency range of the VFO must be within the frequency-measuring range of the counter used in conjunction with the transfer oscillator.

A simplified block diagram of a transfer oscillator is shown in **fig. 4.** The input signal is connected to one input of a harmonic mixer, and one output of the VFO is routed to the other input of the mixer. A second VFO output is connected to the counter, which obviously must be capable of measuring the VFO frequency. The harmonic mixer serves both as a mixer and a harmonic generator, mixing the input signal with the fundamental VFO frequency and with harmonics of the VFO generated within the mixer. The VFO is tuned to the lowest frequency to produce an output from the mixer that is within the passband of the video amplifier. This produces a display on the cathode-ray tube, whose horizontal sweep is usually derived from the ac line frequency. The VFO is then carefully tuned for a zero-beat indication on the CRT, and the fundamental VFO frequency is read on the counter.

If the approximate frequency of the input signal is

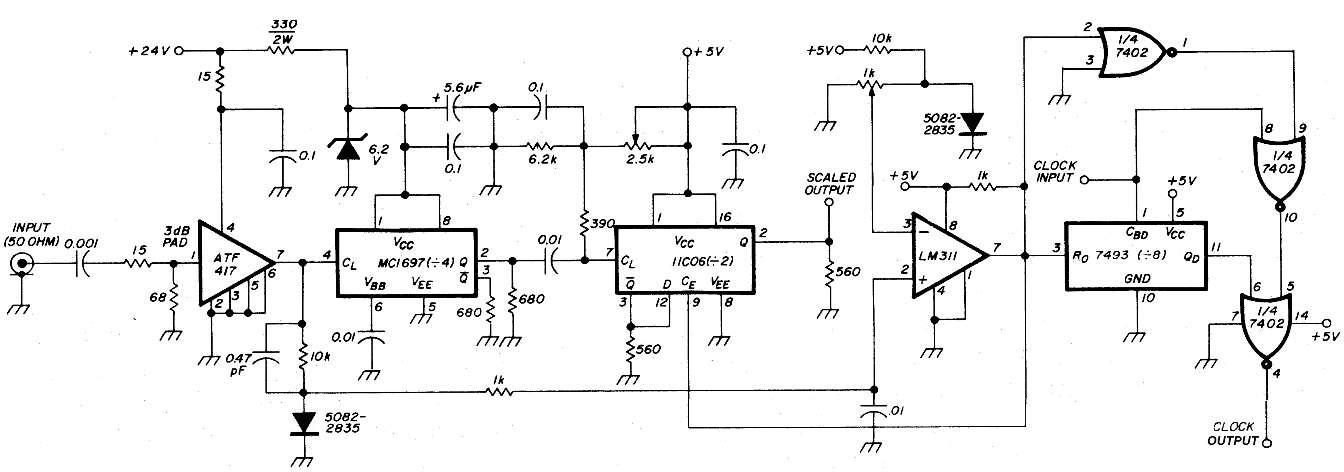


fig. 3. Schematic diagram of a 1300-MHz prescaler which incorporates circuits for automatically switching the gate time when an input signal is applied.

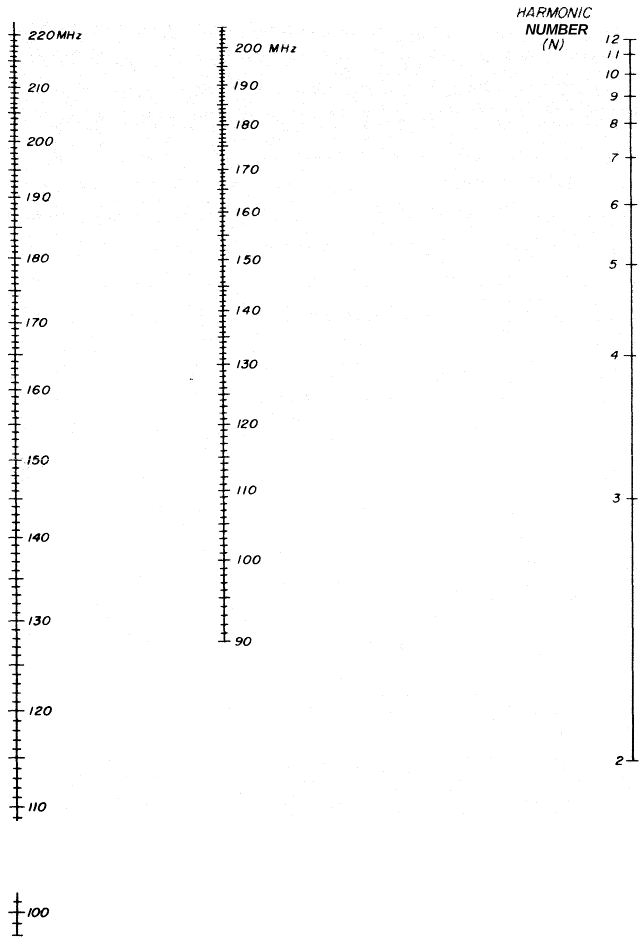


fig. 5. Nomograph for determining the harmonic number of an unknown frequency between 400 MHz and 2 GHz when measured using a manual transfer oscillator whose VFO tunes from 90 to 220 MHz (courtesy Hewlett-Packard *Company*).

known, and it is a relatively low multiple of the VFO frequency so that there is no ambiguity in determining the harmonic number, the input frequency is calculated by multiplying the counter frequency reading by the harmonic number. However, if the unknown frequency is much higher than the VFO frequency, it becomes necessary to determine the VFO harmonic with which the input signal has been mixed. This entails an even more time-consuming procedure of measuring two adjacent fundamental VFO frequencies whose harmonics produce a zero beat, and then determining the input frequency or harmonic number, as follows.

If f_X is the input frequency, f_L is the lower of the two adjacent VFO frequencies, and f_H is the higher VFO frequency, then

$$f_X = \frac{f_H \times f_L}{f_H - f_L}$$

The harmonic number may be determined from the following equations, where N_L is the harmonic number of f_L, and N_H is the harmonic number of f_H

$$N_L = \frac{f_H}{f_H - f_L}$$

$$N_H = \frac{f_L}{f_H - f_L}$$

The harmonic number may also be determined from the nomographs of **figs. 5** and **6** (extracted from reference 2) for the two preceding equations.

The modern transfer-oscillator frequency counter performs essentially the same procedures, but does so automatically. **Fig. 7** is a much simplified block diagram of such a counter. The automatic transfer oscillator consists of two channels, a lock channel and an N-computing channel. The input signal is split in a power divider and applied to one input of the lock harmonic mixer and to one input of the N harmonic mixer. A low-frequency, voltage-controlled oscillator (VCO 1) is swept from its minimum to maximum frequency, typically 100 to 200 MHz, until an output is obtained from the lock harmonic mixer which will pass through the lock video amplifier. The

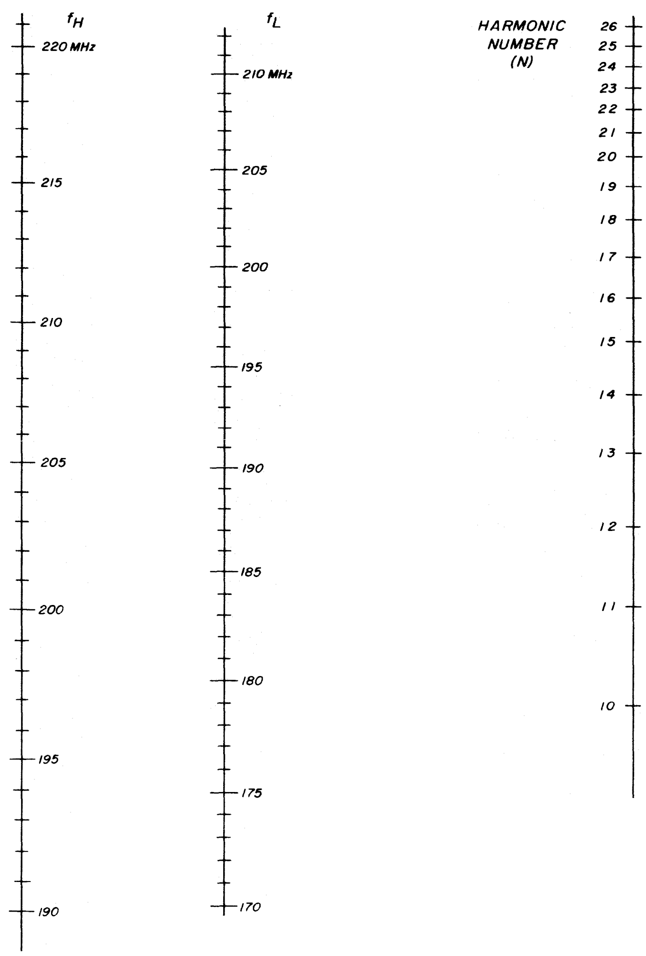


fig. 6. Nomograph for determining the harmonic number of an unknown frequency between 2 and 5 GHz when measured using a manual transfer oscillator whose VFO tunes from 170 to 220 MHz (courtesy Hewlett-Packard Company).

signal from the video amplifier is applied to one input of a phase detector, and a reference signal derived from the time base is fed to the other input of the phase detector. Since the output of the phase detector controls the VCO sweep generator, VCO 1 will be phase-locked to the input signal. The output of VCO 1, when so locked, will be 1/N times the input frequency.

A second voltage-controlled oscillator (VCO 2) also provides a signal, via the N harmonic mixer and the N video amplifier, which feeds one input of the N mixer. This signal, when mixed with the reference signal derived from the time-base oscillator, results in an output from the N mixer which is proportional to the harmonic number, N. The N counter then increases the gate time by a factor equal to N. (Note the similarity to the prescaling counter in this respect.) Thus, the counter will provide a direct readout of the input frequency in terms of N times the frequency of VCO 1, whose output is fed to the direct-counting circuits after being converted to the appropriate logic level by the Schmitt trigger.

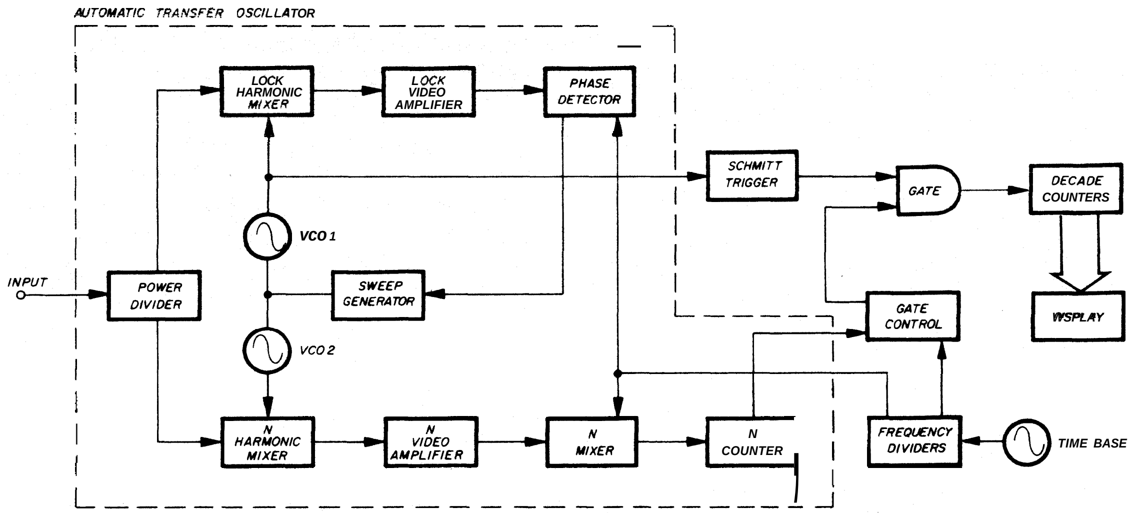


fig. 7. Simplified block diagram of an automatic transfer-oscillator frequency counter.

heterodyne down-conversion

The concept of heterodyning a high input frequency down to one within the range of a low-frequency counter is one that should be completely familiar to anyone with a basic knowledge of electronics. Implementing this concept, however, requires that the heterodyne oscillator frequency be known to the same degree of accuracy as the counter time base if accurate frequency measurements are to result. This is accomplished both in manual heterodyne down-converters and in automatic heterodyne counters by generating the heterodyne frequency from the counter time base.

A simplified block diagram of a typical manual heterodyne converter, which may be either a separate instrument or a plug-in unit, is shown in **fig. 8.** It will accept any frequency between 1.1 and 10.1 GHz and down-convert it to one within the 100-MHz range of the counter connected to its output. Down-conversion is realized by applying the unknown frequency to one input of a microwave **mixer,** with a known frequency fed to the other input of the mixer.

The known frequency is derived from the time-base oscillator in the counter through frequency multiplier and harmonic generator circuits. The output of the harmonic generator is a comb of frequencies which are multiples of 200 MHz and are fed to the harmonic selector. This circuit is a tunable cavity whose *Q* is high enough to select only a single frequency from the comb input and whose dial is calibrated in terms of the 200-MHz harmonics between 1 and 10 GHz. Obviously, it is possible for the input frequency to heterodyne with either of two adjacent 200-MHz harmonics to produce a beat frequency of less than 100 MHz. However, the lower of the two adjacent harmonics will produce a heterodyne frequency equal to the input frequency minus the harmonic frequency, while the higher harmonic will result in a heterodyne frequency equal to the harmonic frequency minus the input frequency. Since the former will result in a counter reading, which, when added to the selected harmonic frequency is the input frequency, it is the desirable one to use. This is accomplished by always tuning the cavity from the low-frequency end until the indicator shows the first output. The indicator responds to any output from the amplifier which is in the counter's frequency range, so that the lowest harmonic can be selected and harmonic ambiguities eliminated.

Because the tunable frequency is a harmonic of the counter time base, determining the unknown frequency is dependent on only the tuning dial calibration for the selected harmonic; this calibration need only be sufficiently accurate to discriminate between adjacent harmonics. Therefore, as long as you are certain that the lowest harmonic has been selected, operation and frequency determination using a manual heterodyne down-converter is somewhat simpler than the same process involving a manual transfer oscillator.

To iterate a point made previously, the measurement accuracy of the heterodyne conversion process is essentially the same as that of the basic counter because the harmonic frequencies are derived from or phase-locked to the time-base oscillator. Because of the problem of sweeping and selecting the appropriate harmonic, an automatic heterodyne converter became realizable only with the advent of the electrically tuned YIG (*Y*ttrium-*I*ron-*G*arnet) filter.

The YIG filter consists of a single-crystal sphere of yttrium-iron-garnet in a controllable magnetic environment. The ferromagnetic resonance of such a sphere in an rf field can be varied by changing the magnetic field, and therefore can be controlled electrically. An rf signal can pass through the filter when the signal frequency is the same as the ferromagnetic resonant frequency; all other frequencies will be greatly attenuated. Thus the YIG filter is actually the heart of an automatic heterodyne counter, a block diagram of which appears in **fig. 9.**

The counter time base is multiplied and drives a harmonic generator, much the same as in the manual heterodyne converter. The comb output of the harmonic generator feeds the input of the YIG filter, with the filter output applied to one input of a microwave mixer. The unknown input frequency is fed to the second input of the mixer The filter control circuit drives the control magnet coils in the YIG filter so that

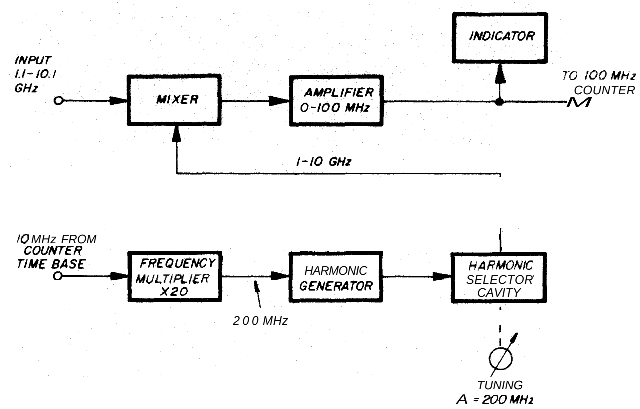


fig. 8. Simplified block diagram of a manual heterodyne down-converter used with a 100-MHz counter to measure frequencies between 1.1 and 10.1 GHz.

the resonant frequency of the filter is swept from its lowest to its highest frequency. The mixer output, generated from the lowest harmonic frequency to pass through the filter, is amplified, converted to an appropriate logic level by the Schmitt trigger, and passed through the gate to the decade counters.

The output of the video amplifier is also applied to an in-band signal detector, whose output inhibits the filter control sweep and keeps the YIG filter at the acquisition frequency. The signal detector also controls the signal acquisition logic, which further controls the filter tuning as required for successive measurements. The signal acquisition logic also controls the gate and other logic circuits (not shown on the block diagram) in the counter so that a direct reading of the input frequency is displayed on the counter readout.

a look at today's technology

Although there are a considerable number of uhf

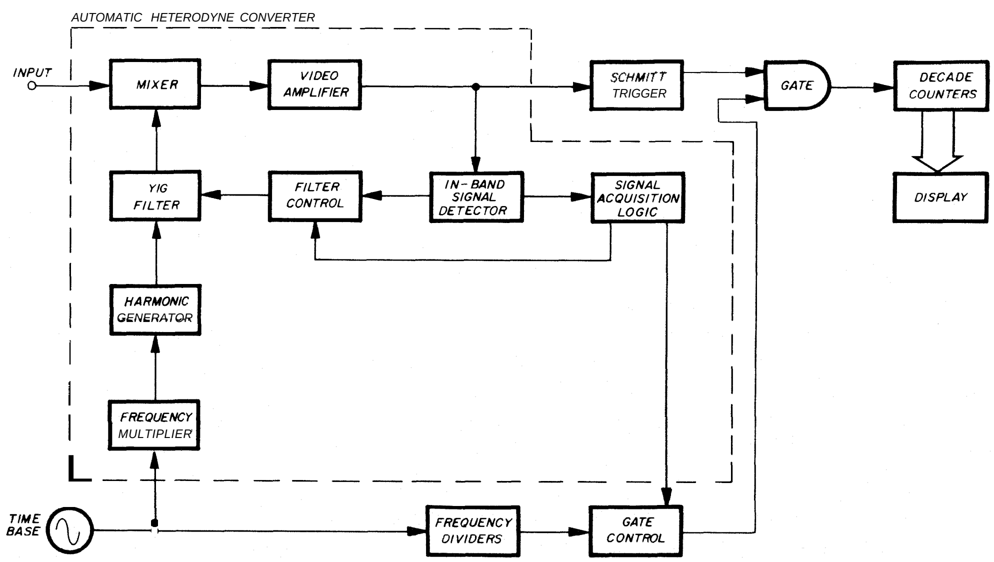
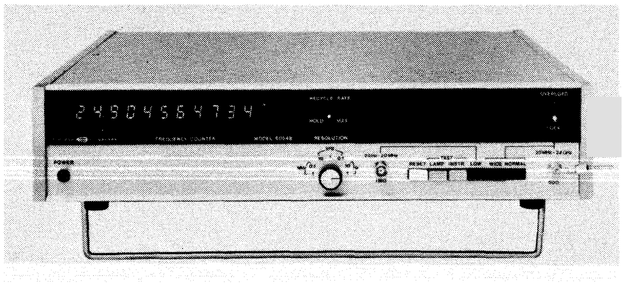


fig. 9. Simplified block diagram of an automatic heterodyne-converter frequency counter. The heart of the system is the electrically-tunable YIG filter.



The Systron-Donner model 6054B employs a FLACTO™ (Frequency Locked Automatic Computing Transfer Oscillator) to permit measurements, with 1-Hz resolution, of frequencies as high as 24 GHz (courtesy Systron-Donner Corporation).

counters available in today's market, the over-10-GHz microwave counter field is dominated by three manufacturers: EIP, Hewlett-Packard, and Systron-Donner. Since microwave counters employ the latest technology, a brief look at some typical instruments should be of interest to readers who are not employed in the microwave electronics industry.

The Systron-Donner model 6054B covers a frequency range of 20 Hz to 24 GHz. It is an automatic transfer oscillator type of counter which employs a circuit designated by Systron-Donner as an ACTO™ (Automatic Computing Transfer Oscillator). A simplified block diagram of the ACTO circuit is shown in **fig. 10.** The similarity to **fig. 7** is apparent, with the input signal split and applied to the lock and N channels. However, only a single VCO is used for both frequency and harmonic (N) determination.

The N-computing channel is used to determine the value of N in the following manner. As can be seen from the block diagram, the VCO output is fed to the fr shifter, as is a ·kł signal derived from the counter time base. The frequency shifter is a single-sideband generator that produces one sideband which is 1 kHz higher than the phase-locked VCO frequency. This frequency-shifted signal is routed to the N harmonic mixer and heterodyned with the unknown input frequency. The resultant mixer output, which will pass through the N video amplifier, has a frequency of N times 1 kHz; this signal is applied to the N computer. The N computer digitally compares the video amplifier output frequency with the 1-kHz reference and generates a signal which corresponds to the harmonic number, N. The signal is further processed and applied to the gate-control circuit to increase the gate time by a factor equal to N. Thus, the counter will provide a direct readout of the input frequency, in terms of N times the phase-locked VCO frequency, which is fed to the direct-counting circuits after being converted to digital levels by the Schmitt trigger.

The counter also employs a Frequency Locked Automatic Transfer Oscillator (FLACTO™), which is a modification of the ACTO technique. The frequency-lock feature permits the counter to tolerate very high levels of frequency modulation and makes the measurement virtually immune to the rate of modulation. Additional details may be found in reference 3.

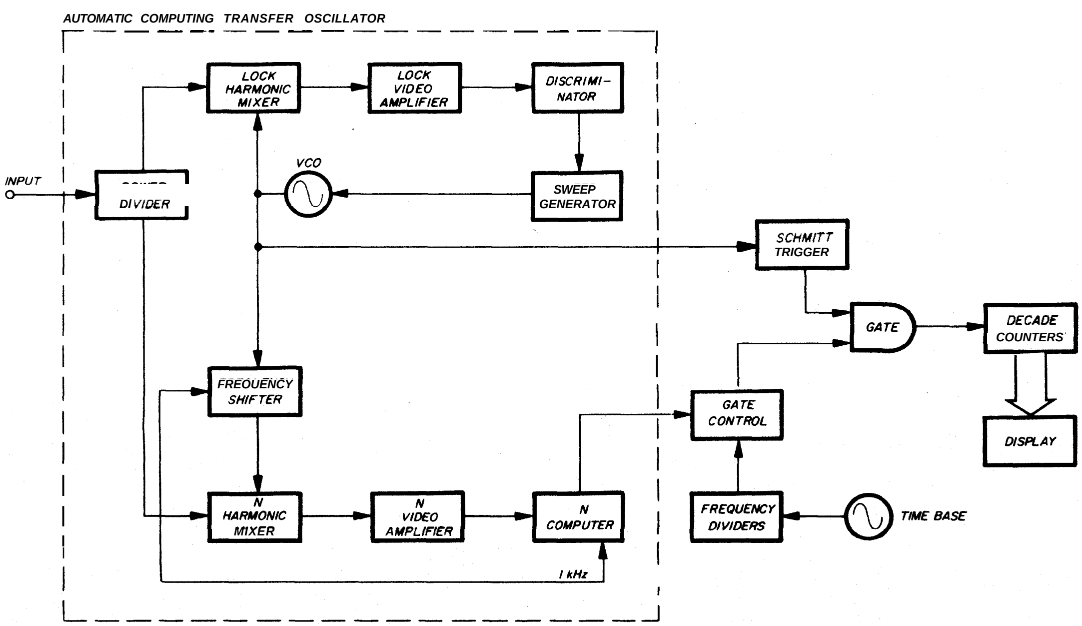


fig. 10. Simplified block diagram of the Systron-Donner ACTO™ (Automatic Computing Transfer Oscillator) down-converter. This is an automatic transfer oscillator which requires only one VCO.

The model 6054B has two signal inputs. One is a high-impedance, direct-counting input for 20 Hz to

20 MHz; the other is a 50-ohm input for signals between 20 MHz and 24 GHz. When the latter is used, one of two operational modes may be selected. In the normal mode, the local oscillator is locked to an internal reference, which results in a high resolution reading in the shortest period of time (1-Hz resolution for I-second sampling). In the wide mode, the local oscillator will harmonically track the input frequency, which enables it to track swept or frequency-modulated signals.

A new technique, known as harmonic heterodyne conversion, is used in the Hewlett-Packard model 5342A microwave frequency counter. This conversion scheme is a hybrid of the heterodyne and transfer-oscillator down-conversion circuits in that the counter acquires the input frequency in the manner of a transfer oscillator, but measures the frequency as does a heterodyne converter.

A block diagram of the harmonic heterodyne down-converter appears in **fig. 11.** In this arrangement, the conversion oscillator is a programmable frequency synthesizer locked to the counter time base. The synthesizer output is applied to a sampler, as is the input signal. The microprocessor increments the synthesizer until one of the inputs from the sampler is in the counting range of the direct counter. At that time, the signal detector generates a signal that causes the microprocessor to cease incrementing the synthesizer, and the amplified sampler output frequency is counted; this frequency is the input frequency divided by a harmonic number, N.

To determine N, the microprocessor increments the synthesizer to cause a small frequency change.

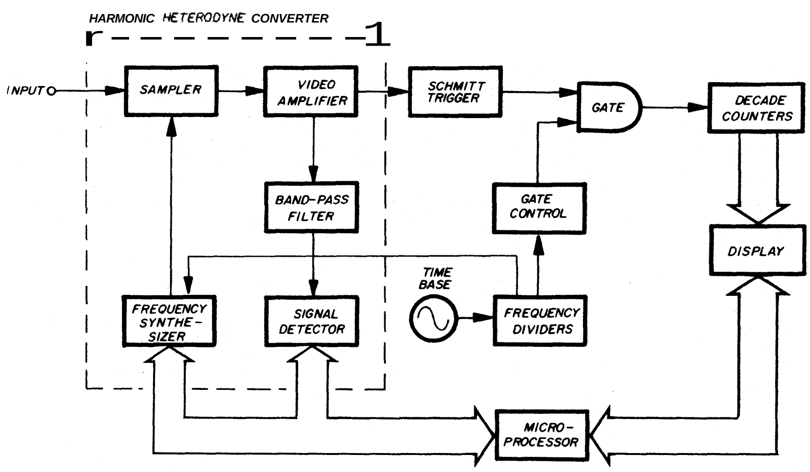


fig. 11. Simplified block diagram of the harmonic heterodyne converter used in the Hewlett-Packard model 5342A Microwave FrequencyCounter.

Since there are now two outputs of known frequencies from the sampler, which result from beating the input signal with N times two known frequencies, the microprocessor is able to perform the simple algebraic computation required to determine N.

The 5342A counter covers a frequency range of 10

Hewlett-Packard's 5342A Microwave Frequency Counter uses a microprocessor-controlled harmonic heterodyne down-converter. When the counter is equipped with its amplitude-measurement option, both frequency and amplitude can be displayed simultaneously. The five left-hand digits are used to display frequency with 1-MHz resolution, and the four right-hand digits display amplitude with 0.1-dBm resolution and a polarity sign *(courtesy Hewlett-Packard Company).*

Hz to 18 MHz, with a recently announced option which extends its upper limit to 24 GHz. Also available as an option is amplitude measurement. This feature allows simultaneous measurement of both the frequency and amplitude of an incoming sine wave. Amplitudes are displayed with a resolution of 0.1 dBm over a dynamic range of -22 to $+20$ dBm. The amplitude-measuring scheme employs a diode detector circuit in conjunction with an internal reference oscillator for level comparison. The amplitude measurement circuit is calibrated during production and, for signals over 500 MHz, error correction values, as a function of frequency, and input level are stored in an amplitude PROM (programmable read-only memory) for use by the microprocessor. This technique ensures an accuracy of ± 1.5 dBm for

The keyboard controls the source-locking circuitry in the EIP model 371 Source Locking Microwave Counter. A separate LED display, located just to the left of the keyboard, displays the desired frequency entered via the keyboard *(courtesy* NP, *Inc.*).

sine-wave input signals within the operational dynamic range.

As can be seen from the photograph of the Hewlett-Packard model 5342A, operation of the counter is controlled by means of a front-panel keyboard. The keyboard provides control of resolution, self-check, automatic or manual modes, amplitude and/or frequency measurements (with the amplitude option installed), frequency and amplitude offset, etc. Such is the power of the microprocessor-controlled instrument. A detailed discussion of the model 5342A appears in reference 4.

A unique instrument manufactured by EIP is their model 371 source-locking microwave counter. This counter is an automatic heterodyning type that covers a range of 20 Hz to 18 GHz, and, in addition, has the ability of locking any signal source between 10 MHz and 18 GHz to the same long-term accuracy and stability as the time-base oscillator in the counter. The only requirements for the signal source are that it have an fm input and that it can be set manually to within 20 MHz of the desired output frequency. A block diagram of the source-locking circuits is shown in **fig. 12.**

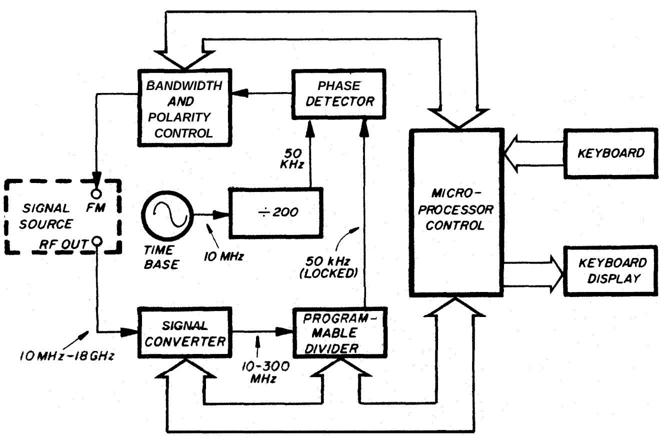


fig. 12. Block diagram of the source-locking section of the EIP model 371 Source Locking Microwave Counter. The frequency of an external signal source can be locked to a preset frequency between 10 MHz and 18 GHz with the same long-term stability and accuracy as the time-base oscillator in the counter.

Source locking is accomplished by converting the input signal to one that is in the 10- 309-MHz range, using heterodyne down-conversion. The microprocessor control then calculates the proper division ratio to produce a 50-kHz output from the programmable divider when the input signal is equal to the desired frequency, which has been entered via the keyboard. (An auxiliary keyboard display on the counter records the frequency which has been keyboarded in). The dc component of the phase detector output, applied to the fm input of the signal source via the bandwidth and polarity control circuit, alters the frequency of the signal source until it is equal to the desired frequency.

The microprocessor controls the overall loop response by systematically varying the bandwidth and polarity parameters until a phase lock is achieved at a nominal bandwidth of 2 kHz. If the loop cannot be locked at this bandwidth, because of inherently low bandwidth in the signal source, the microprocessor repeats the process at a nominal bandwidth of 500 kHz. The automatic bandwidth and polarity control permits the use of the source-locking counter with signal generators and sweepers of different modulation sensitivities and polarities.

summary

This has been a necessarily brief overview of uhf and microwave counters. I have intentionally omitted a comparison of the several down-conversion systems being used today, since such comparisons are often a matter of specsmanship. Readers who are interested in such comparisons, or who want more detailed information on the conversion techniques used by the manufacturers dominant in the field, should consult references 3 through 7.

references

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2. 540B Transfer Oscillator, Operating and Service Manual, Hewlett-Packard Company, Palo Alto, California 94304. pages 3-6 and 3-7.
3. Dick Voyles and Dick Hall, "Microwave Frequency Measurement — A New Approach." Technical Article No. 1412, 1975 IEEE Intercon Conference Record, Trends in Electronic Measuring Instruments.
4. Ali Bologlu and Vernon A. Barber, "Microprocessor-Controlled Harmonic Heterodyne Microwave Counter also Measures Amplitudes," *Hewlett-Packard Journal*, May, 1978, page 2.
5. "Understanding Microwave Frequency Measurements," Application Note 144, Hewlett-Packard Company, Palo Alto, California 94304.
6. "Fundamentals of Microwave Frequency Counters," Application Note 200-1, Hewlett-Packard Company, Palo Alto, California 94304.
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ham radio

vertical antenna
for 40 and 75 meters

Design and construction of a two-band vertical antenna that fits into a modest city lot

Many Amateurs live on small lots in crowded neighborhoods and don't have the space for full-size horizontal antennas for the low-frequency bands. For example, my lot is small and cut into the side of a hill. The house and a swimming pool occupy the only flat area. A few years ago, during the sunspot doldrums, I became interested in working some low-band DX with emphasis on 75 meter SSB. The two-band vertical antenna described here is the result of my experiments.

First I tried a two-band inverted V hung from my beam antenna tower. I had little success competing on 75 SSB, although the antenna worked fairly well on 40 meters. It was also a bother whenever I had to lower the tower because of weather. The next antenna considered was a ground-mounted vertical cut to one-quarter wavelength on 75 meters. This design became very unattractive for a number of reasons. Very little free ground area was available for a good radial system, and the ground is exceedingly hard with low-conductivity soil. The only available site locations were either difficult for running coax or were in locations where a considerable amount of the radiation would be into my house and those of neighbors. About the only place left to consider was the top of the house, which is about 9 x 12 meters (30 x 40 feet).

This led to the design of an inexpensive two-band groundplane vertical antenna for 40 and 75 meters mounted on the tap of the gable roof. The antenna was made from a three-section push-up TV mast, about 8 meters (25 feet) of RG-8/U coax, some galvanized TV guy wire, a TV-type ceramic pot capacitor, a short piece of stair railing dowel, a few insulators, and 61 meters (200 feet) of almost any kind of copper wire for two sets of radials. The antenna is shown in **fig. 1**. A simple fixed-tuned L network was

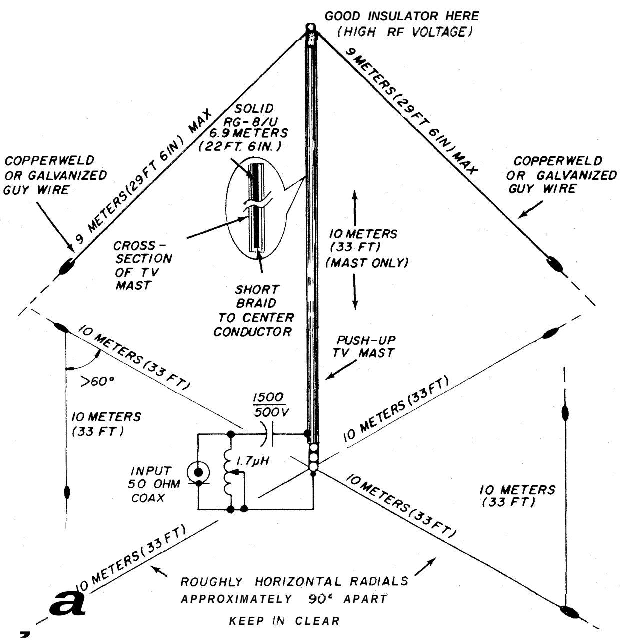

fig. 1. Two-band vertical antenna for 40 and 75 meters.

mounted at the base to obtain a good impedance match on both bands. Two sets of two radials were used, one straight set for 40 meters and a Z configuration for 75 meters.

operating principles

The TV mast **(fig. 1)** is one-quarter wavelength long on 40 meters. From the top a length of RG-8/U

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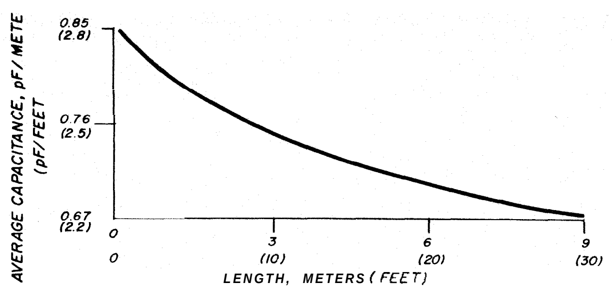

fig. 2. Capacitance of wire in space as a function of wire length. To find total capacitance, multiply length by capacitance per meter (foot). Curve assumes length-to-diameter ratio greater than 50. Curve was used to determine capacitance of the antenna top-hat radials.

coax is dropped, which is shorted at the bottom end. The top outer conductor (shield) is connected to the top of the TV mast. The top center conductor is connected to two slanted radials, which act as guy wires and the capacitive-loading element, or "top hat."

The coax on 40 meters appears as a parallel-resonant circuit and isolates the mast from the top-hat radials. On 75 meters the coax is one-eighth wavelength long and acts as a series inductance of Z_0, or

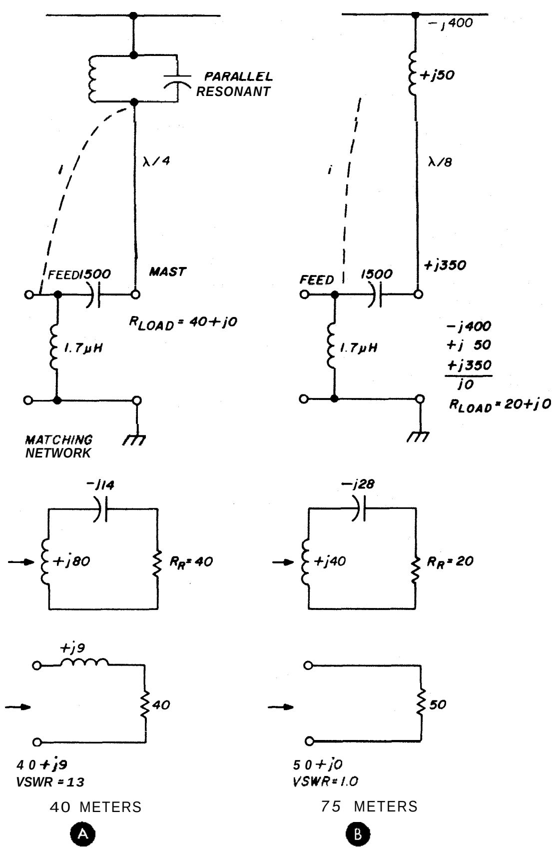

fig. 3. Equivalent circuit of the two-band antenna. The 40-meter version is shown in (A), 75-meter version in (B).

50 ohms. The base section has a characteristic impedance* of about 350 ohms. Accordingly it appears as an inductive reactance on the order of 350 ohms. The top radials act as capacitance loading and have an effective capacitance of about 100 pF with a reactance of about 400 ohms **(fig. 2)**. The mast, coax, and capacitive top hat form a series-resonant circuit on 75 meters, allowing the mast to be an effective one-eighth-wavelength radiator with a fairly flat current distribution. Note from **fig.** 3 that the current

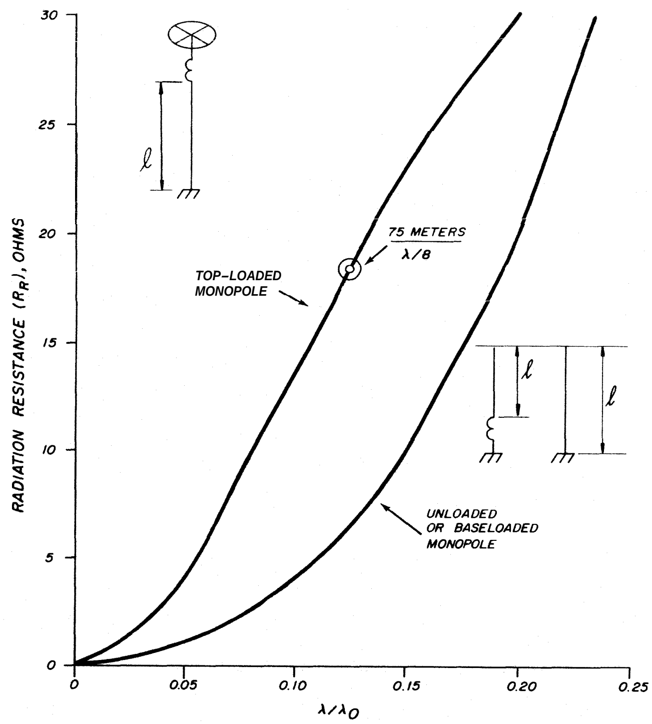

fig. 4. Radiation resistance of short monopole antennas, which was used to derive base resistance of the antenna L matching network.

distribution along the mast on 75 meters is fairly uniform.

Fig. 4 is used to derive the base resistance for the design of the L matching network. On 40 meters the antenna is one-quarter wavelength long and has an input impedance in the order of **40** ohms. On 75 meters, because of top loading, the base impedance is about 18 ohms. If base loading were used, one-eighth wavelength on 75 meters would have a base impedance of only 7 ohms and would be inefficient.

*Characteristic impedance is

$$\sqrt{\frac{\text{inductance per unit length}}{\text{capacitance per unit length}}}$$ and

is approximately $60(\ln \frac{2h}{d} - 1)$, where h is height and d is diameter

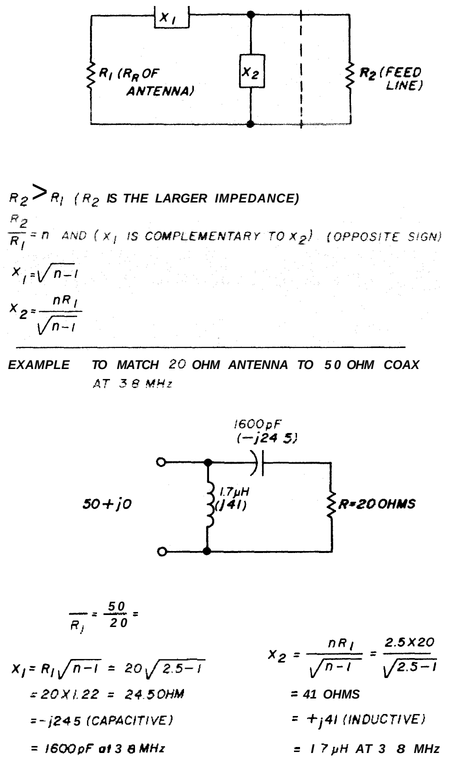

fig. 5. L network matching circuit development.

The L network design is developed in **fig. 5.** The resultant circuit is derived in **fig.** 3. The measured input matching characteristics are shown in **fig. 6.**

adjustment

Adjustment is straightforward. Little interaction occurs between 40- and 75-meter adjustments. First adjust the 40-meter radials, the two straight 10-meter (33-foot) lengths equally until a minimum VSWR is obtained at the desired operating frequency. Next, for 75-meter operation, adjust the Z-configuration radials equally. If this doesn't quite hit the desired frequency a slight adjustment of the top-hat radials may be necessary. These radials don't affect 40-meter operation to any extent. Recheck 40 meter operation. The VSWR on 40 meters will not be unity, because the base impedance is on the order of 40 ohms. The matching network may need a slight inductance change for the best match on 75 meters. This adjustment will have negligible effect on 40 meters.

construction

Details are shown in **figs. 1** and **7.** The top of the mast on 40 meters is at **a** very high voltage, so a good-quality top insulator is needed. **I** used low-sap wood (maple) boiled in wax. The insulator has been reliable over the past four years. Plastic materials such as acrylic are suitable. See **fig.** 7 for top insulator assembly. Each end of the coax cable mounted inside the mast should be dipped in wax or otherwise sealed.

The two top radials are made from galvanized or copperweld guy wire. Soft copper wire was used originally but broke at the top end as a result of wind stress. The top radials are at high voltage on 75 meters but have low current. They are not used on 40 meters. Accordingly, it's not necessary to use high-conductivity wire. The lower end of the radials terminate in small, corrugated antenna insulators. The wire is cut ionger than shown, passed through the insulator, and twisted back on itself so that easy adjustment may be made. The vertical angle of the radialsisn't critical. An anticorona noise loop is formed at the top of the mast either by extending the coax or one of the top radials.

The base-mounting insulator can also be made from wood boiled in wax. This point is at low voltage and insulation isn't critical. **I** used a tilt-over, U-channel TV base for convenience in mounting. The mast should be supported by one set of four insulated guys at the top of the base section of the three-section TV push-up mast. Guy rings are usually supplied with the mast.

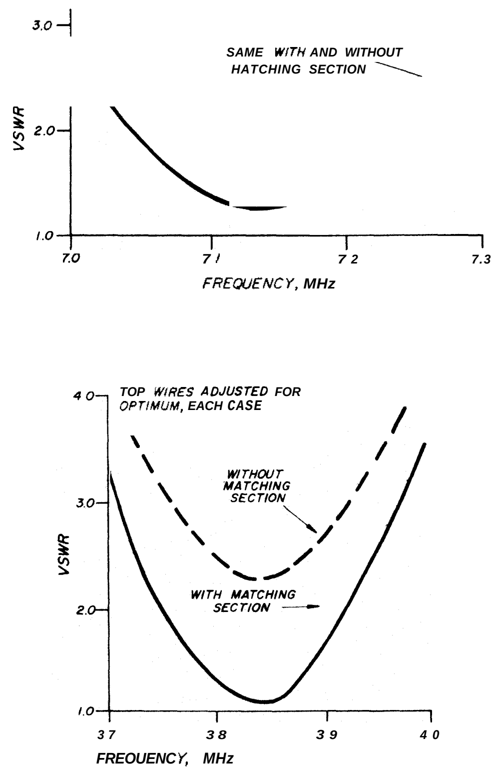

fig. 6. Measured input matching characteristics for the two-band antenna, 40 meters (top) and 75 meters (bottom).

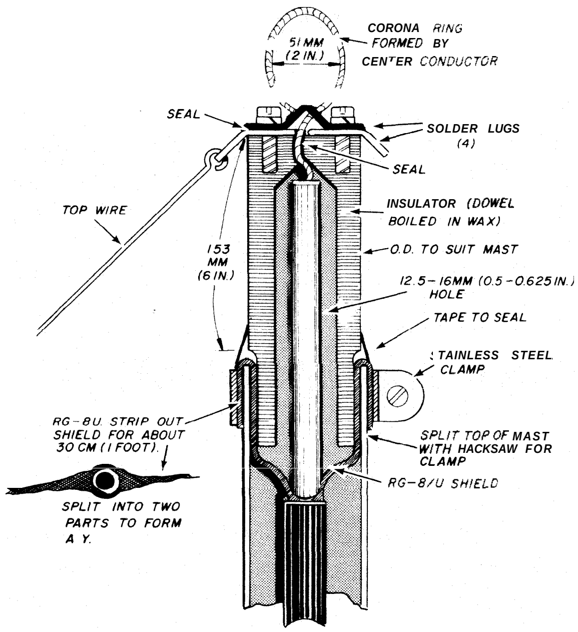

fig. 7. Construction details of the top insulator.

The L matching network is mounted on the base of the mast above the mounting insulator. A heavy, flexible, stranded wire is run from the bottom side of the L network to the center of the groundplane radials. This makes it convenient to tilt the mast without disassembly of the network or feed coax. The 1.7-μH inductor is ten turns of 1.6-mm (no. 14) bare wire, 38 mm (1 1/2 inches) diameter and 35 mm (1 3/8 inches) long. Spacing is six turns per 150 mm (six turns per inch).

The groundplane radials form resonant elements and should be separated from surrounding surfaces except at their center. The vertical angle of the radials is not critical. The ends are supported and terminated by insulators similar to the top radials for convenience of adjustment. Bend the radials to suit the shape of roof. The bend at the 10-meter (33-foot) point can vary slightly. The wire type isn't critical. Anything larger than 1 mm (no. 18) either enameled or covered will suffice.

performance

The antenna has been in use since 1974. Operation has been satisfactory. The only mode used on 75 meters was SSB, with some CW on 40 meters. All continents except Europe have been worked several times on 75 meters, with good reports. If you live in a high noise location, this may not be the antenna for you. If you live in a place where lightning is active, make sure an adequate ground is provided. This antenna makes a dandy lightning rod.

ham radio

updating the Collins KWM-2

Important modifications are described for modernizing the KWM-2 high-frequency transceiver

Introduced to the Amateur world in the fall of 1959, the Collins high-frequency KWM-2 transceiver quickly became the classic, with over 40,000 units in use worldwide by Amateurs, commercial services, and the military of numerous nations. The latest version of this popular rig is the KWM-2A. Time-proven by its robust construction and its long life in these days when circuit-boarded, solid-state gear quickly eliminates obsolete designs, this fine transceiver has more than held its own. Over the years revisions have been made to the original design. This article covers some important modifications to the KWM-2 family and describes how you can incorporate them into your unit to help bring it up to date.

the KWM-2

The Collins KWM-2 high-frequency transceiver is widely recognized as a superior piece of Amateur gear and is continuing a long and useful life. A decade ago a military overview of communications equipment in governmental service praised the KWM-2 for reliability, ruggedness, and ease of repair. Countless thousands of Amateurs agree with this conclusion.*

While the newest KWM-2s retain the original classic appearance, numerous revisions and modifications have been incorporated over the years which make the modern version easier and better to operate than the older sets. Some of the important modifications that can be made by the advanced Amateur with adequate test equipment are described here. For those who don't want to dig into their transceiver, information is furnished on getting the more sophisticated and difficult modifications made by a professional. In any event, before undertaking any revision or modification to your KWM-2, make sure the change has not already been incorporated into your equipment. Many hams own second-hand units, so it's wise to make sure your manual agrees with the particular transceiver you own, at least as far as the schematic and voltage charts are concerned. Be suspicious of an older model KWM-2 that has a new manual. The two may not be in exact agreement.

All modifications should be made with a 40-watt (or smaller) soldering iron, so as to protect the insulation on wires next to the soldering iron. A magnifying glass is helpful, as are needle-nose pliers. You'll be working in an area with a high parts density and you don't want to damage some circuits while you modify others!

the "wing" versus the "meatball"

Around mid 1968, Collins changed their old winged emblem and adopted a new, round escutcheon known as the "meatball." This cosmetic change allows you to determine the approximate age of your

*Gus Browning, W4BPD, tells the story of a KWM-2 he took along on a DXpedition in the Indian Ocean. It was dropped overboard by a crew member during an attempt to land on an obscure island. Native divers finally fished up the KWM-2 and brought it ashore. After flushing with fresh water and drying out for a few hours, the rig was hooked up — and it worked! The only casualty was the meter movement, which had opened up. A local artisan repaired the meter winding and Gus was back on the air.

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KWM-2, as the random serial numbers on the KWM-2 after 1968 no longer date the equipment for the layman. On the used-equipment market, the "meatball" KWM-2 commands a somewhat higher price than the older "wing" model. It's best to buy the KWM-2 on performance and appearance, however, and forget about the emblem. Sometimes you can realize a tidy savings by buying a "wing" model in good condition rather than the newer "meatball" model.

If you do buy a used KWM-2, check it carefully in both transmit and receive modes on all bands before you part with your money. Look under the chassis to make sure the previous owner hasn't made his own unique (and often unworkable) modifications. Many good KWM-2s for sale are showing up in the classified ads, as bedazzled hams trade in their units for the latest solid-state transceiver complete with bells, whistles, and a six-month wait for replacement parts. Good! Their loss is your gain if you want to own a rugged and reliable transceiver that you can service and repair yourself.

minor bugs you may not have observed the first time around

Transmitter instability? Signs of oscillation? Before you tear things apart or attempt reneutralization of the amplifier stage, remove the amplifier-compartment lid and make sure the tube shield of the 6CL6 driver stage (V8), is firmly in place. A loose tube shield can play havoc with transmitter operation!

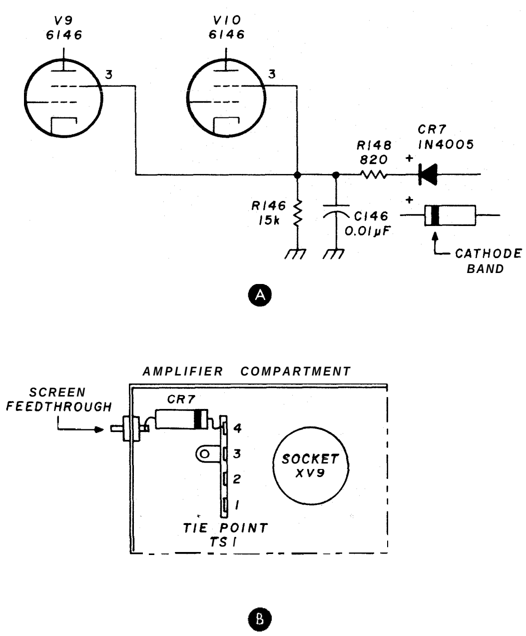


fig. 1. Diode CR7 placed in amplifier screen circuit protects receiver from blocking caused by screen emission of 6146 tubes. (A) circuit modification, (B) diode placement.

Receiver blocking on switch-over? Sometimes you'll notice a delay of up to 30 seconds during which time the receiver is blocked and no signals are heard after switch-over from the transmit mode. This problem is caused by screen emission from the 6146 amplifier tubes (V9), (V10), which paralyzes the receiver agc (automatic gain control) circuit. New 6146* tubes will sometimes cure this annoying problem, but a permanent fix is easily achieved by placing a diode in the amplifier screen power lead, which blocks negative current (fig. 1). This mod is easily and quickly made in the bottom of the amplifier compartment. The diode is substituted for the wire lead between the screen feedthrough terminal in the compartment wall and nearby socket tie-point strip (TS1). The diode anode is connected to the feedthrough terminal. Put insulated sleeving on the diode leads. This mod has no effect on transmitter performance.

ALC meter instability? Does the zero reading of the alc meter float around during transmit, or does it gradually drift up-scale as the KWM-2 warms up? This annoying fault can usually be cured by replacing capacitor C157 (0.01 μF, 200 volts) with a new low-leakage mylar or polypropylene unit. You'll find the old unit attached to pins 1 and 3 of socket XV17A (6BN8).

Equipment runs hot? Short tube life? The 6U8/6U8A and 6AZ8 tubes in the KWM-2 are said to have short lives. The grapevine suggests replacing the 6U8/6U8A with a 6EA8 for longer life. This can be done in most sockets, with no change in performance, except for the 6U8/6U8A used as the audio tone oscillator (V1). Some 6EA8s will not work in this circuit, and others will distort the audio tone signal, which then bleeds into the receiver audio system during CW operation. Stick with the 6U8/6U8A in this socket and look for short-life tube problems elsewhere.

In some KWM-2s the low-voltage dc supply (supposed to be a nominal 275 volts) runs from 300 to over 340 volts when the standard Collins 516-F2 power supply is used.† No wonder some of the small tubes are cooked! Measure your low-voltage supply. It should not run much over 290 volts on receive and

*Folklore has it that either 6146B tubes won't perform properly in the KWM-2, or that 6146Ds are the only tubes to use in the KWM-2. Forget both of these fairy tales. The differences between the 6146. 6146A, and 6146D are minimal (mostly being one-upmanship in advertising policy). All do the job equally well. It's not necessary to match 6146-type tubes, either, although it's suggested that a 6146A not be used with a 6146B.

†Overvoltage is presumed due to various manufacturers having supplied the power transformer and filter chokes. Design and windings of these components seem to vary, especially in the dc resistance of the transformer or choke coils. This could account for the voltage variance.

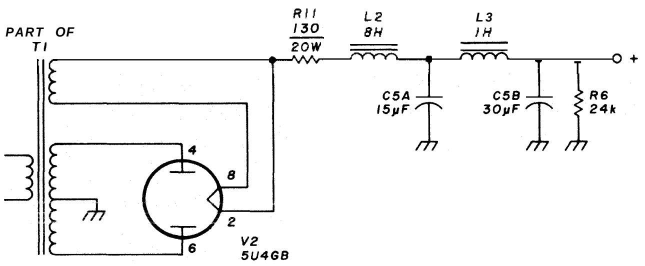


fig. 2. Voltage-dropping resistor R11 added to reduce low voltage to small tubes in KWM-2. A heat-dissipating resistor in a finned housing is recommended. This is bolted to the chassis sidewall near the 5R4GY socket. A Dale type RH-25 resistor or equivalent is suggested.

260 volts on transmit. If the voltage is much higher than these values, add a 75-150 ohm, 25-watt wire-wound dropping resistor, R11, as shown in **fig. 2.** The small tubes in the KWM-2 will run much cooler if you do this. The 5U4GB rectifier in the power supply should be replaced with a solid-state plug-in rectifier, and the resistance value of R11 should be chosen to deliver the correct voltage. Substitution of the rectifier improves regulation and removes 15 watts of filament power from the supply transformer.

You should also replace the 5R4GY high-voltage rectifier with a suitable solid-state plug-in device. This action will remove an additional 10 watts of filament power from the transformer and will increase the B-plus voltage by about 40 volts, providing a few more watts of power output and a cooler-running transformer. This simple substitution also boosts the 6.3-volt filament supply, which is marginal at best. You'll probably have to readjust the amplifier bias control, R9, in the supply for the correct resting plate current of the amplifier tubes after these mods have been made (40 mA for general use or 50 mA when driving a linear amplifier).

Old **filter caps in the power supply?** It's a good idea to replace the high-voltage filter capacitors and the bias filter capacitor in the power supply if the KWM-2 is an older model. The capacitors become leaky with age and the capacitance value drops off at the same time. You can put more microfarads in the same space occupied by the old units and this improves the supply's dynamic stability. When you put the new capacitors into the circuit be sure to observe polarity, for the bias capacitor, which is hooked up "backwards," with the positive terminal grounded. Capacitors C2, C3, and C4 can be replaced with equivalent 80-µF, 450-volt units, and C5A-B can be replaced with a dual 30-µF, 250-volt unit. Capacitors C6, C7 can be replaced with 40-µF, 250 volt units. Unless the shunt capacitor, C1, is defective (a rare occurence), don't bother to replace it.

Dial chatter or backlash? Underneath the VOX plate atop the main tuning dial assembly is a small idler pulley mounted to the front panel to the left of the dial mechanism (as viewed from the front). This pulley holds the two dial plates in alignment as the dial is rotated. Unbolt and lift up the VOX plate; this requires removal of one screw at the top left of the plate and two screws above the panel escutcheon. Now you can see the dial pulley. If it's loose it will rattle, and the dial will show backlash to a greater or lesser degree. The amount of mesh with the dial mechanism is determined by the center screw holding the gear. For a quick fix, loosen the screw and slide the gear into the dial mechanism a very small amount and retighten. Caution! The gear-retaining screw is very short. Don't loosen it too much or it will

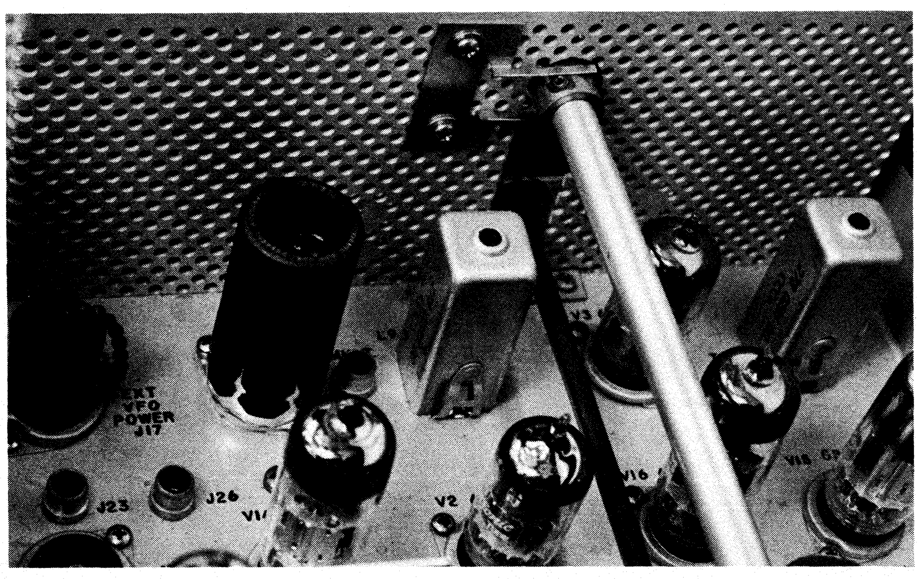


fig. 3. Grounding clip on amplifier loading shaft reduces harmonics escaping from amplifier compartment. Tube shield on V1 (6AZ8) decreases tube temperature. Heat sink shield is used.

fall out and you'll lose dial alignment. However, if you hold the two dial disks together to keep them from losing alignment, you can completely remove the idler gear and coat the gear shaft with silicone grease, which will eliminate dial rattle. Maintain the position of the dial plates so that you don't lose calibration.

TVI on 10 meters? Why do some KWM-2s show bad TVI on 10 meters while others don't? And why does the TVI often worsen when you bring your hand near the final amplifier tuning/loading panel controls? The answer is that these concentric shafts come out of the amplifier compartment and are insulated from the front panel of the KWM-2 by an almost invisible panel bushing. In effect, the shafts act like a radiating antenna for amplifier harmonics that would otherwise remain bottled up in the amplifier compartment. A shaft grounding clip* bolted to the outside of the amplifier enclosure (as shown in **fig. 3**) grounds the outer shaft and reduces the harmonic signal at this escape point to near zero. The grounding clip is held in position with (4-40) hardware.

If your KWM-2 doesn't incorporate a vhf choke (L128) in the power amplifier B-plus lead immediately following plate choke L17, a 120-pH choke should be added to prevent harmonic currents from passing into the power supply **(fig. 4)**.

Receiver i-f tube V1B run hot? Place your hand on V1, the 6AZ8 i-f amplifier tube after the KWM-2 has been running for a few hours. Wow! Hot! No wonder this tube is said to have a very short operating life. And no wonder the S-meter zero-signal reading shifts about on the scale. The latest versions of the KWM-2 have incorporated a protective resistor (R75) in series with pin 3 (cathode) of tube V1 to ground to limit plate current. If you don't have this resistor in the circuit, a 10-ohm, 112-watt resistor placed in series with the ungrounded terminal of the receiver GAIN ADJUST potentiometer, R132, mounted on the VOX plate, will help reduce the tube temperature. In addition a heatsink-style tube shieldt is placed over V1.

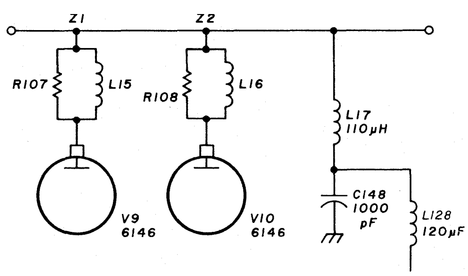


fig. 4. Vhf choke (L128) in B-plus lead to final amplifier helps suppress TVI-causing harmonics. A J.W. Miller 9360-13 choke rated at 400 mA is suggested.

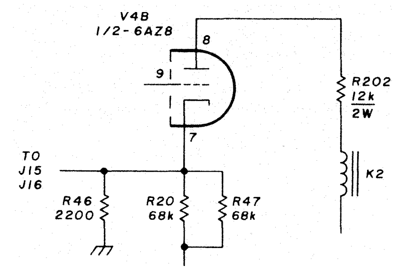


fig. 5. Modified VOX relay control circuit. Resistor R202 is added to reduce current through the relay coil. It may be necessary to reduce the value of the resistor in some cases to provide proper pull-in current. In some KWM-2s resistor R202 is 330 ohms and is located in the cathode circuit between pin 7 of socket XV4 and the circuit to J15 and J16. In this case, no plate resistor is required. In some units resistor R46 is 3.3k. It should be replaced with 2.2k for this rnodification.

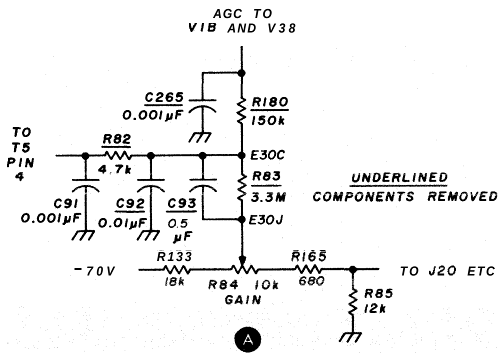
Heat-sink shields are hard to come by, but perhaps your friendly electronics store (or the local flea market) has some. A retainer mounting shell is also required. The shell is mounted to socket XV1 using the existing mounting bolts. You'll probably find (as I did) that the mounting shell has a negative clearance with respect to the socket. The solution is to cut tiny slots around the bottom edge of the shell with metal snips. Cut to a depth of about 1.5 mm (1/16 inch) then bend out the tabs you've made with a pair of long-nose pliers. The shell will then fit snugly over the socket rim. Snap the heat-sink shield over the tube, and longer tube life will be your reward.

Relay problems? Some KWM-2 owners have found to their sorrow that the coil of VOX relay K2 burns out after prolonged use. The popular and expensive solution is to get a "meatball" KWM-2 with plug-in relays. However, a circuit modification somewhere along the long production history of the KWM-2 has solved this vexing problem, even in some of the older models. A 12k, 2-watt safety resistor (R202) is placed in series with the plate of the VOX relay amplifier tube, V4B, **fig. 5.** If your KWM-2 doesn't have this modification it's a good idea to incorporate it, as it might save you a destroyed relay coil. The resistor can be mounted between pin 8 of socket XV4 and a tie-point epoxied to the chassis near the socket.

Lack of receiver sensitivity on some bands? Even after repeated alignment some KWM-2s show

"The Collins part number of the grounding clip is 553-2555-002. You may be able to obtain a clip from Dennis Brothers, WA0CBK, Route 1, Box 1, Potter, Nebraska 69156.

†Suitable heat-dissipating tube shields are manufactured by, and available from, International Electronic Research Corporation, 135 West Magnolia Boulevard, Burbank, California 91502. The shield cools tube bulb temperature to below that of the bare bulb. A type TR6-6020B shield is used for the 6AZ8 or 6U8/6U8A. A TR6-6025B is recommended for use with the 6CL6 driver tube.



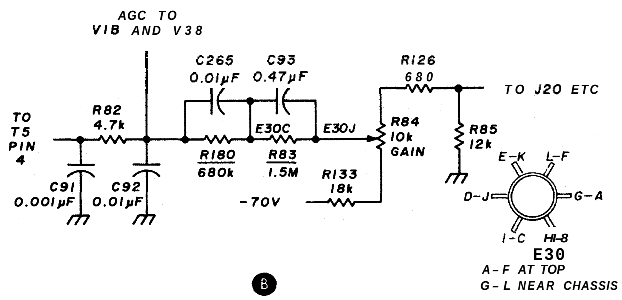


fig. 6. KWM-2 mods to improve agc response. Original circuit is shown in (A). Circuit in (B) is the modified setup as described in the text. Terminal support E30 is adjacent to the audio output transformer T6. Terminal TS8 is equidistant from E30 and T6.

poor sensitivity on some bands, or sensitivity seems to change when sidebands are switched. This problem can be caused by signal overload from low-frequency sideband oscillator V11A, whose signal is coupled into the receiver section wiring harness.

The culprit is rf choke L22 in the crystal-oscillator plate circuit. The choke is part of a tuned circuit and can radiate energy furiously. Radiation from this inductor can get into circuits where it doesn't belong and reduce overall receiver sensitivity. The cure is to remove choke L22 and replace it with a shielded rf choke. A parasitic suppressor (R195, 47 ohms, 1/2 watt) should also be placed in series with the grid, pin 2, of socket XV11A.

Receiver agc pumping and overshoot on noise pulses? Some of the older model KWM-2s use the agc time constant circuit shown in **fig. 6A.** A newer circuit is also shown in this illustration. The components to be changed are on terminal support E30, shown in the technical manual. These are R82 (4.7k), R83 (3.3 meg), C92 (0.01 µF) and C93 (0.05 µF). In addition, R180 (150k) from terminal TS8-1 to E302, and C265 (0.001 µF) from TS8-1 to the power-connector grounding ring are removed. (Note that R180 and C265 are not incorporated in some early models.)

The following components are now added:

1. Connect new R83 (1.5 meg) from E30C to E30J.
2. Connect new C93 (0.47 µF) from E30C to E30J.
3. Connect new R82 (4.7k) from T-5 terminal 4 to TS8-1. Use sleeving on leads and route around E30.
4. Connect new R180 (680k) from TS8-1 to E30C.
5. Connect new C265 (0.01 µF) from TS8-1 to E30C.
6. Connect new C92 (0.01 µF) from TS8-1 to ground ring on power connector J13. Check wiring against **fig. 6B.** Mark the modification in your manual for reference.

Agc overload and audio distortion on strong SSB signals? It's recommended that this useful modification be performed along with the previous one in cases where both arrangements are missing from the transceiver. This modification adds hang agc to the receiver rf amplifier **(fig. 7)** and greatly improves strong-signal reception. Refer to the under-chassis layout of **fig. 2** for placement of parts:

1. Remove screw and lockwasher nearest front panel used to secure audio transformer TE.
2. Install a two-terminal, lug-type strip on T6 using screw and lockwasher.
3. Disconnect the white-green-blue wire at TS8-1, pull it back through the cabling and reconnect it to terminal 1 of the newly installed lug-type strip. Call this new strip TS11.
4. Connect R213 (2.2 meg) from TSII-2 to TS11-1. Use sleeve resistor leads as necessary.
5. Connect diode CR11 (1N458) from TSII-2 (cathode) to TS11-1 (anode). Use sleeve diode leads as necessary.
6. Connect C276 (0.05 µF) from TS11-1 to E30B.
7. Of the two white-green-blue wires connected to E40-I, disconnect and tape the end of the one showing continuity to TS11-1. You'll have to disconnect both wires to make this check, then resolder the wanted wire.
8. Connect an insulated wire from E40-I to TS11-2, routing it along the cabling. Check wiring against **fig. 7.** Mark the modification in your manual for reference.

Audio distortion on strong signals? Aside from the above modification, another cause exists in some KWM-2s for fuzzy audio. Place a 0.01 µF, 600-volt capacitor from the screen of audio output tube (V16B, pin 8) to ground. Also place a 56-ohm, 1-watt

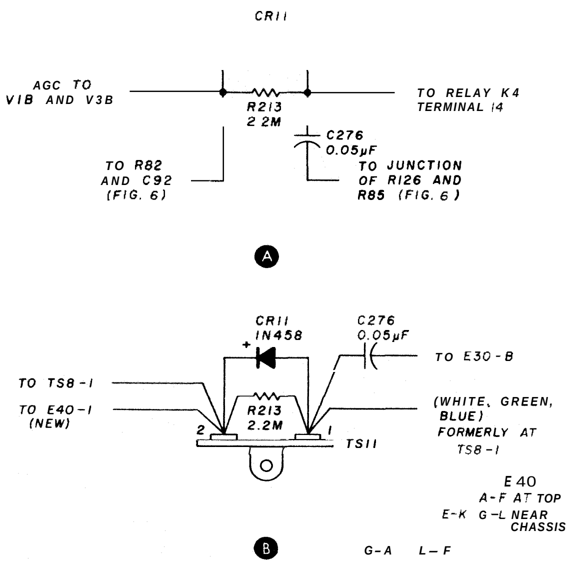


fig. 7. Adding hang agc to your KWM-2, (A). Parts placement is shown in (B). (See also fig. 2.) Terminal support E40 is between socket XV16 and inductor L9. New resistors are ½ watt.

resistor from the yellow (4-ohm) lead of output transformer T6 to ground. These mods will eliminate a weak audio parasitic oscillation sometimes encountered in some receivers.

general modification notes

Modification of the KWM-2 is not recommended for those who have no experience working with small components in cramped spaces. Many KWM-2s are wired with PVC wiring insulation, which melts quickly at the inadvertent touch of a soldering iron. Always check transceiver operation before and after each modification. After your modification, check for wiring errors or shorts and make sure that small specks of solder and wire are blown out of the chassis before power is applied. Also be aware that I've not seen ***all*** existing KWM-2s and that these mods may not work as shown with some transceiver variations. If you don't understand your present circuit wiring or if it doesn't match the schematics, don't attempt the modification!

where to get help

This material has been prepared with the help of Dennis Brothers, WAØCBK, formerly an engineering technician of KWM-2 production at Collins-Rockwell Company. For those not wishing to make these (and other more sophisticated modifications) themselves, I suggest they contact Dennis at Western Nebraska Electronics, Route 1, Box 1, Potter, Nebraska 69156. A self-addressed, stamped envelope for rapid reply is requested.

ham radio

More Details? CHECK—OFF **Page 126**

commutating filters

Discussion of the commutating filter — the application of analog and digital techniques to implement a bandpass filter

The world is becoming increasingly digital. In fact, many engineers and electronic technicians are worried about their positions in a technical scenario wherein the linear art is shrinking as digital techniques take over. The real truth, as I see it, is that digital is *not* going to take over at all, but will provide additional techniques creating circuit solutions where the linear techniques they replace are shaky. In fact, in applying these new digital techniques, the linear circuit area will be even further expanded.

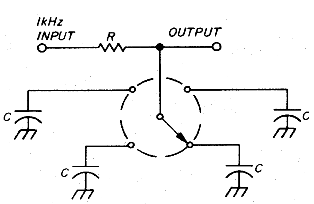


fig. 1. Diagram of a simplified commutating filter. The switch makes one revolution each period of the desired signal frequency to be filtered.

The commutating filter is a good example of how digital techniques provide a simple solution to an analog problem, but which would not work without the addition of some circuitry that is strictly analog. The commutating filter: as presented in **fig. 1**, is designed for 1 kHz; it is a bandpass filter, and its center frequency is dependent only on the frequency with which it is driven. The bandwidth is dependent only on R and C; in fact, this bandwidth is exactly twice the cutoff frequency of the single RC lowpass filter of **fig. 2.** If you look at the voltage on any one capacitor of the filter of **fig. 1,** you'll see a near-dc signal which is the difference-frequency between the drive frequency and the signal frequency. Like the simple RC lowpass of **fig. 2,** it will drop to -3 dB when the signal frequency drops to ½πRC above the drive frequency. It is easily seen why the bandwidth is double the cutoff frequency of a simple RC lowpass filter. Since the output is being commutated sequentially through the four capacitors, it is modulated back up in frequency to that of the input, and phase is preserved.

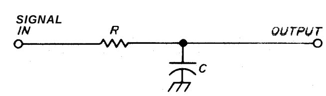


fig. 2. Simple RC lowpass filter from which the commutating filter is derived.

It might sound as if the perfect filter has been

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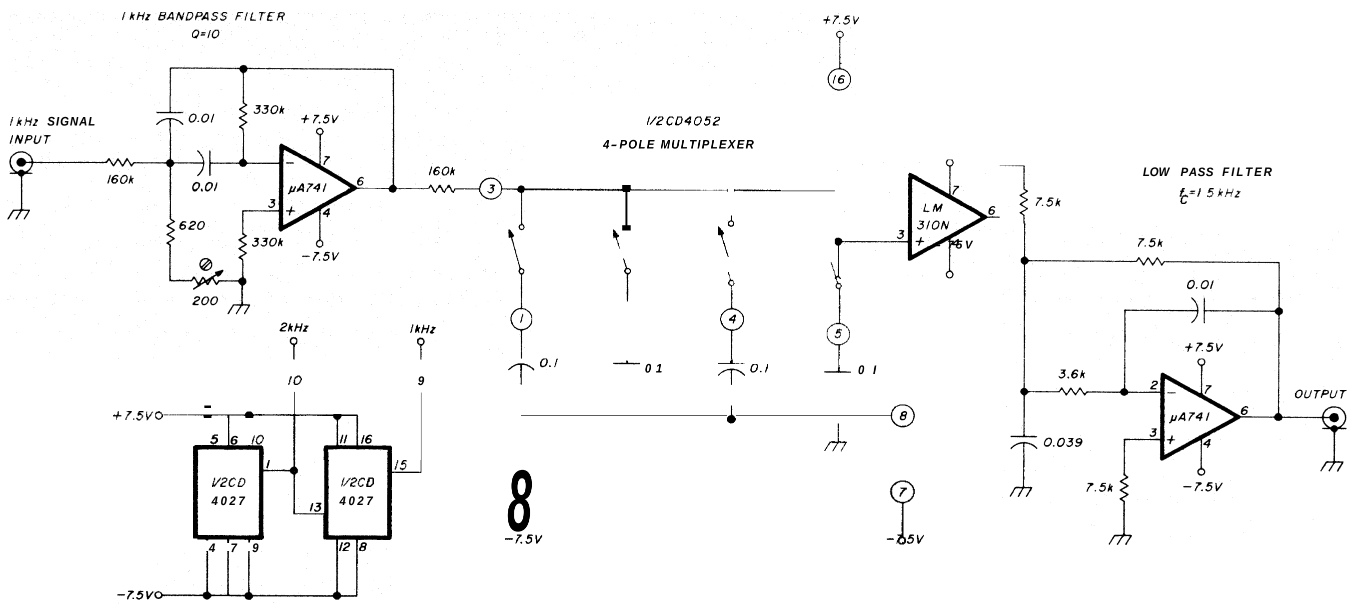

fig. 3. Example of a four-sample-per-cycle commutating filter; circled numbers represent pins of the CD4052 4-pole multiplexer.

achieved with no drawbacks. As usual, there's no free lunch, and you'll find that the commutating filter has some inherent problems. One of these problems is the phenomenon known as "aliasing." Aliasing is the disagreeable habit of filters of this sort to pass not only the same frequency as the drive frequency, but also harmonics of the drive frequency. This can be alleviated by preceding the commutating filter with a simple conventional bandpass filter that attenuates signal frequencies that correspond to harmonics of the drive frequency. This "pre-filter" can, of course, be much broader than the ultimate system bandwidth that our commutating filter provides.

Another drawback of the commutating filter is that

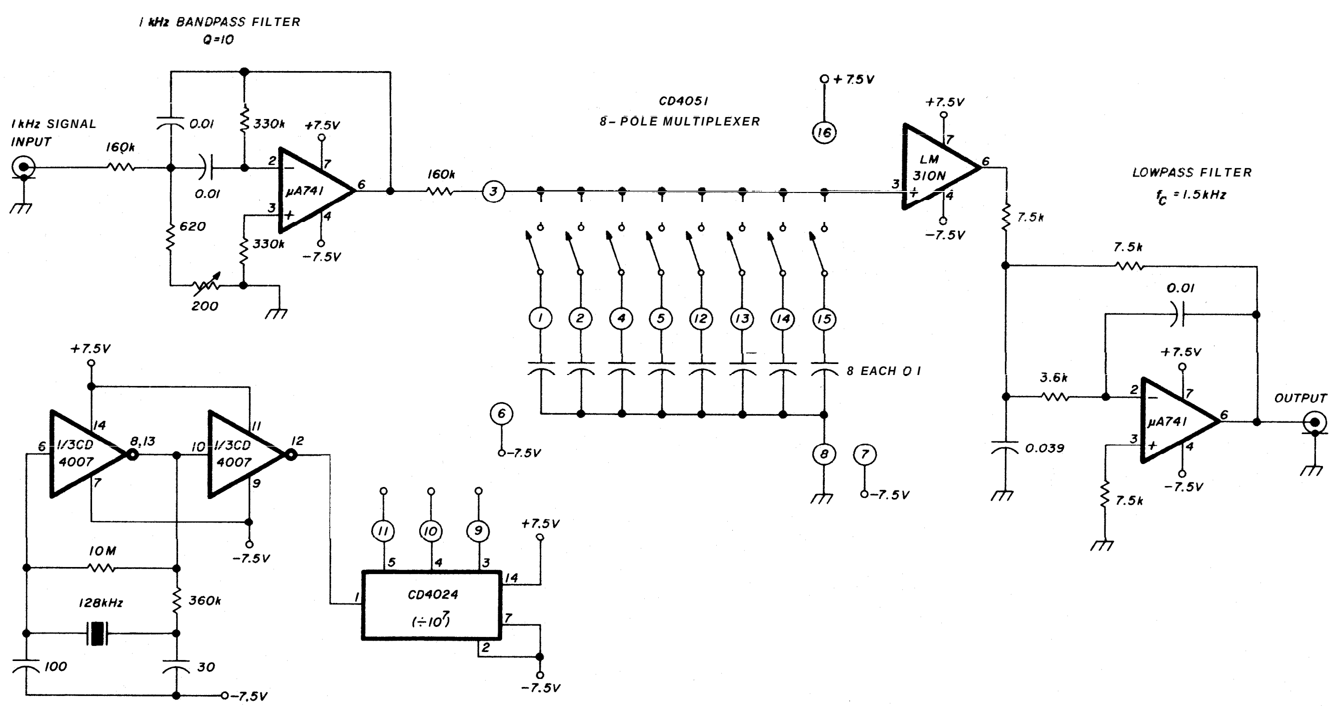

fig. 4. Schematic diagram of an eight-sample-per-cycle commutating filter. In this case, the 4-kHz input has been replaced by a crystal oscillator and a CD4024.

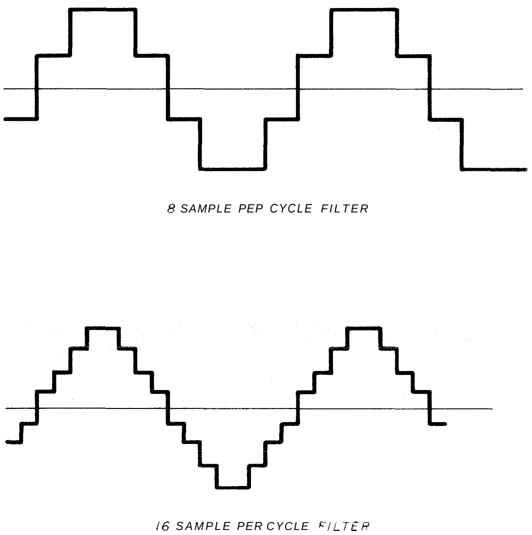

fig. 5. Examples of the idealized waveforms from the 8 and 16 samples per cycle filters. These waveforms are before the post-filter.

it has essentially an infinite output impedance (very much like any simple RC lowpass filter). Practically, however, if the load into which the filter operates is a couple of orders of magnitude higher than R, all is well. The simple solution is to terminate the filter in a high-impedance, noninverting follower.

The last problem of the commutating filter is that it contains harmonics of the drive frequency in its output. These can generally be removed by a lowpass filter, but it is an extra little requirement that must be met if you are to take advantage of the performance of a commutating filter.

As disheartening as all the above restrictions may seem, modern ICs (both linear and digital) come to the rescue to make the commutating filter a fairly simple one. In fig. 3 is shown a complete four-sample-per-cycle commutating filter using an operational amplifier as a pre-filter bandpass, a CD4052 (CMOS multiplexer/demultiplexer) as the switching (and steering) element, an operational amplifier as a noninverting follower, and an operational amplifier as a lowpass post-filter. Since the CD4052 ihalf of it) has built-in decoding (steering), it requires 2 kHz and 1 kHz (two-bit) input. These inputs are derived from a CD4027 dual flip-flop wired as a ripple counter and having a 4-kHz input.

The filter of fig. 3 samples the input signai four times per cycle, and thus the "steps," or discontinuities, in the output (before post-filtering) are relatively large. By going to a filter that takes eight samples per cycle, you decrease the "step" size and ease the post-filter requirements. In fig. **4,** a CD4051 and eight capacitors replace one half of the CD4052 and the four capacitors. This multiplexer is another member of the same CMOS family as the CD4052, but it requires three-bit drive: 4 kHz, 2 kHz, and 1 kHz. To accomplish this drive requirement, another CD4027 flip-flop could be used with an 8-kHz input. Or you could use a single CD4024, which is a seven-stage ripple counter (divide-by-128), and use any convenient three adjacent outputs for the drive. In fig. **4,** the last three outputs (pins 5, 4, and 3) of the CD4024 are used to drive the CD4051, thus requiring a 128-kHz input to pin 1. This 128 kHz is provided by a 128-kHz crystal oscillator made from two-thirds of a CD4007.

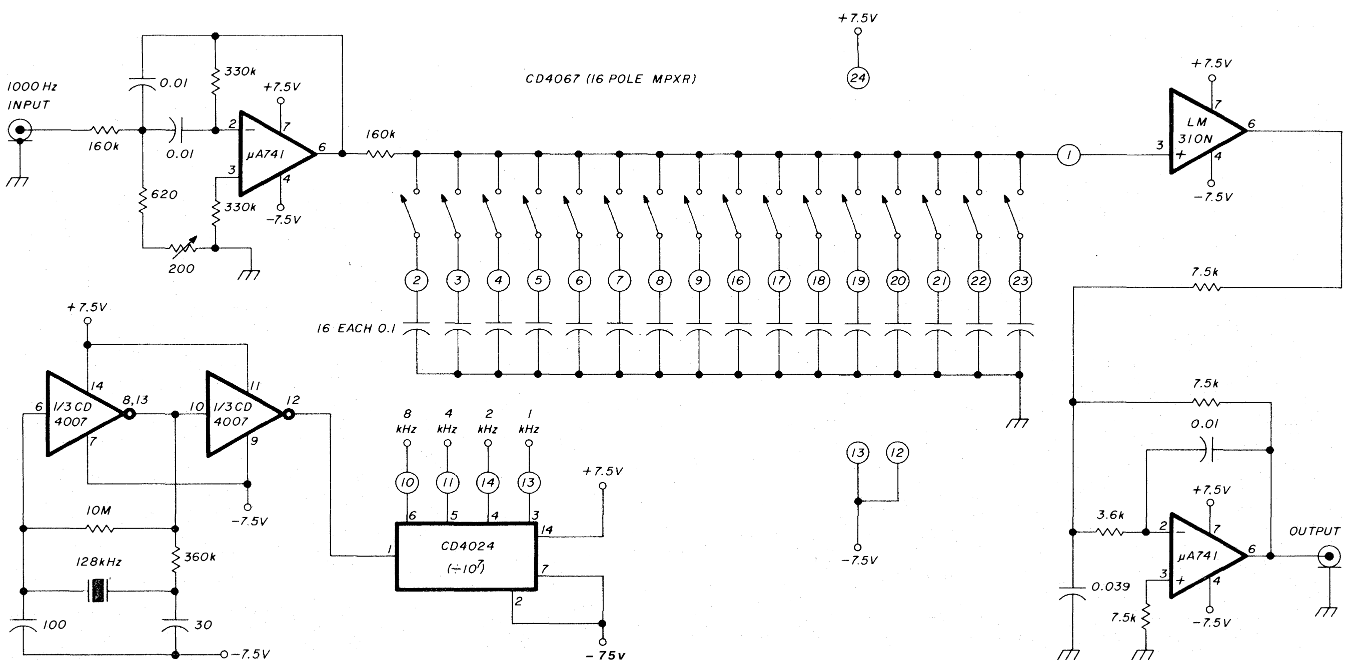

fig. 6. Schematic diagram of a 16-sample-per-cycle commutating filter.

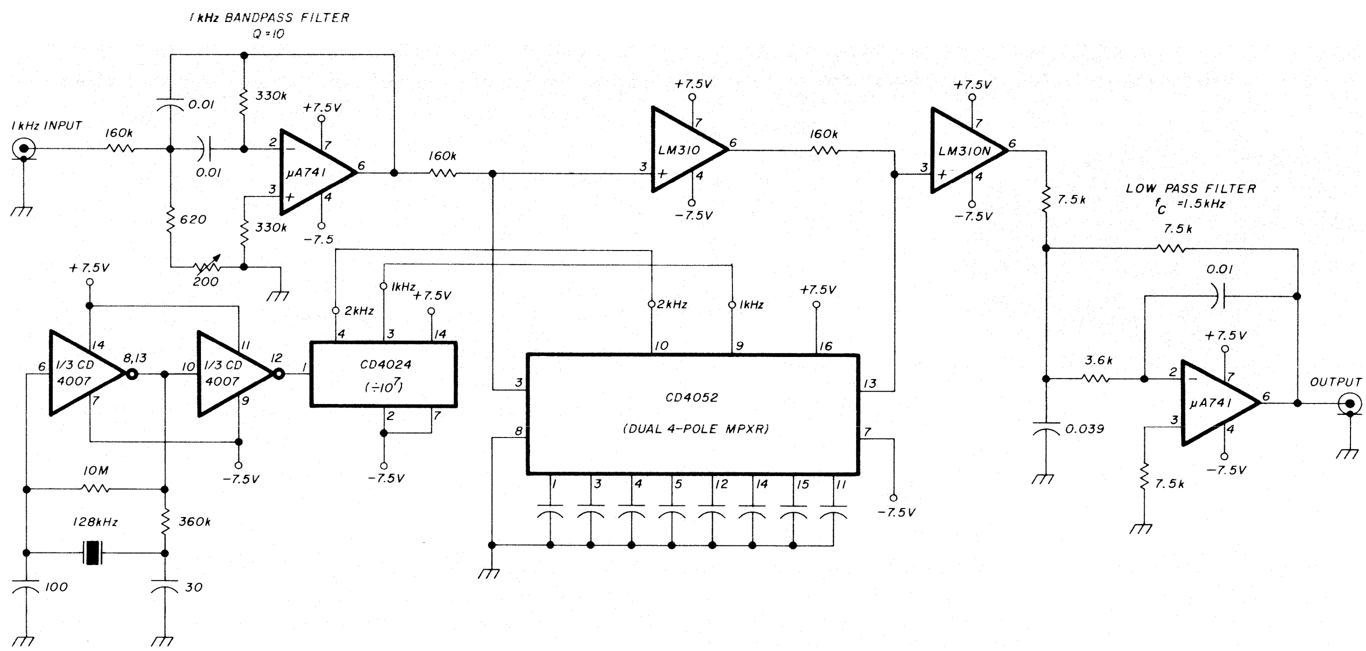

fig. 7. In this case, two stages of four-sample-per-cycle filters have been cascaded. Only one set of pre- and post-filters are necessary, with an impedance follower between the sections to lower the driving impedance to the second section.

Fig. 5 shows the unfiltered outputs of the eight-sample and the sixteen-sample filters. Note how the eight-sample filter has more (and smaller) "steps" in its output, and is thus easier to post-filter. It is even fairly simple to expand the filters of **figs.** 3 and **4** to sixteen samples per cycle, which really cuts down the quantization ripple in the output. Such a circuit is shown in **fig. 6,** using a CD4067 and sixteen capacitors.

It is even possible to cascade commutating filters, and the pre-filter and post-filter need not be replicated. A follower between sections is all that is required for lowering the driving impedance to the second section. An example of a two-section, four-sample-

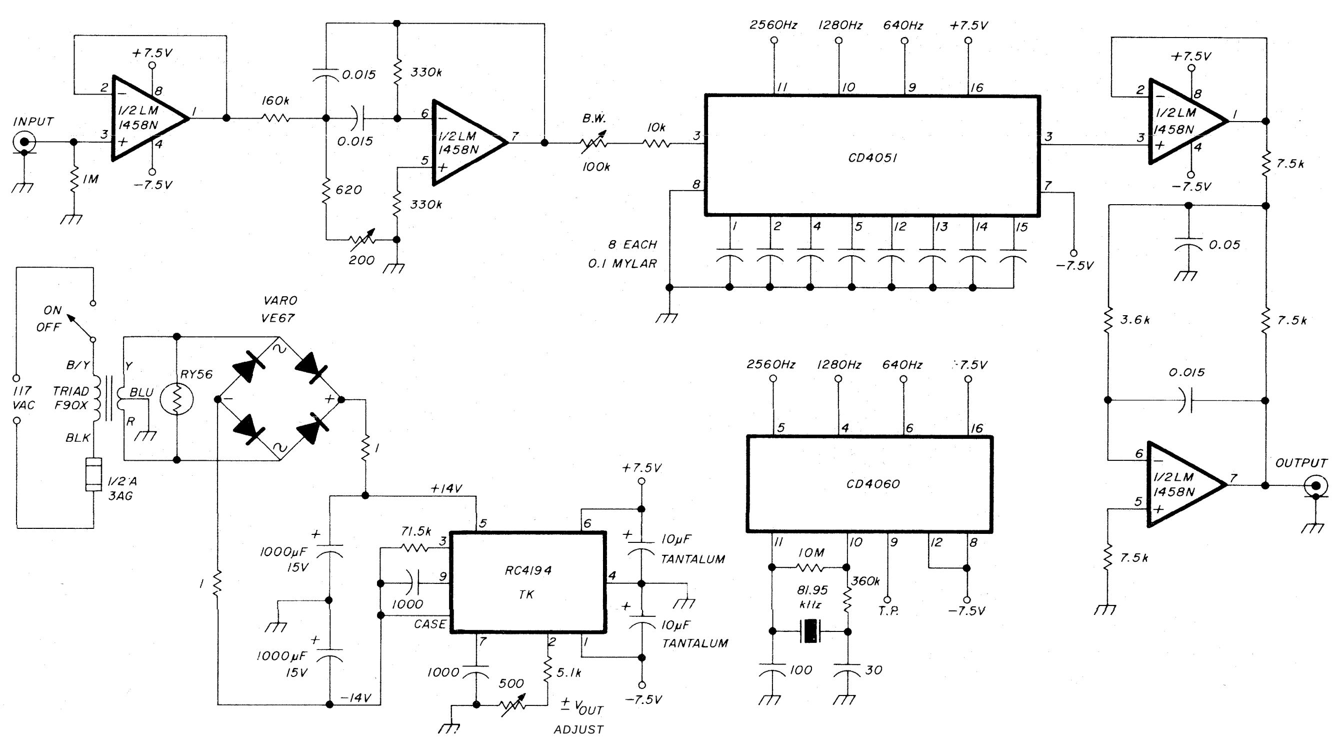

fig. 8. Example of a practical commutating filter. This filter has been set up for an operating frequency of 640 Hz.

per-cycle filter is shown in **fig.** 7. The advantage of cascading is the sharpness of rolloff outside the passband. The rate of rolloff of the equivalent RC lowpass is then 12 dB/octave instead of 6 dB/octave.

Finally, a construction project using a commutating filter in a useful piece of ham equipment is presented in **fig. 8.** The design process was as follows:

1. The operator tuned in a CW signal off-the-air and adjusted the BFO until *the* tone was of an agreeable pitch. This pitch was measured using an oscilloscope to find out what frequency the operator likes to copy. This subjective determination of the operator pitch preference may seem like wasted motion, but many peopie have "holes" in their hearing response (especially older CW operators).

2. Once the desired pitch frequency is determined, multiply it by 128 and get the oscillator frequency to the CD4060. As an example, say that the operator preference turned out to be 640 Hz; then, the input frequency would be 81.92 kHz. If one has a crystal of about that frequency, it can be used directly in fig. **8.** Otherwise, higher frequency crystals could be used, with taps at positions further down the divide-by-two chain. For instance, a 328-kHz crystal could be used, the outputs taken from pins 6, 14, and 13 (still yielding 2560, 1280, and 640 Hz respectively).

3. The pre-filter center frequency is then adjusted to the chosen operating frequency, in this case 640 Hz. The C values scale with frequency so that C is 0.01 μF for 1 kHz and 0.015 μF for 640 Hz.

4. The post-filter cutoff frequency is adjusted to be 1.5 times the bandpass filter center frequency. Again, the capacitor values scale with frequency. Capacitor values of 0.01 and 0.033 μF give a 1.5-kHz cutoff frequency, and 0.015 μF and 0.05 μF give a 960-Hz cutoff frequency.

In **fig. 8,** an input noninverting follower has been added so that the unit may be driven from almost any impedance. An LM1458 dual op amp is used for both the input follower and the pre-filter. Addition of a variable resistance in place of R allows the passband to be varied from 3.0 Hz to 30.0 Hz, continuously. Another LM1458 dual op amp is used in the output section as the noninverting follower and lowpass filter. By using two dual op amps, and using the CD4060 (which combines the crystal oscillator and divider in one IC package) I've reduced the circuit down to four ICs, plus the one IC used as the power supply regulator. The 81.92-kHz crystal was actually an 81.95-kHz unit that is quite common on the surplus market, 81.95 kHz being a standard time base frequency for a variety of distance-measuring devices.

ham radio

accu-keyer speed readout

Another addition to the feature-packed WB4VVF Accu-Keyer — a readout system for code speed

There are thousands of Accu-Keyers[1] already in use, and the appearance of articles[2-8] to add message memories to the basic keyer has undoubtedly resulted in another flurry of Accu-Keyer construction. It is an excellent and highly versatile keyer, and deserves the fine reputation that it has. It might seem that there is little else that one could want from this, or any other, keyer.

There is one useful addition, however. Most of us vaguely know our sending speed. It is true that a speed scale could be put on the front panel behind the control, but the speed vs rotation dependence of most controls is highly nonlinear, especially at the high-speed end of the range where the scale becomes compressed. Any semblance of accuracy is lost in the compressed scale.

A desirable feature, which I have incorporated into the Accu-Keyer system, is a direct words-per-minute speed readout. This is useful for many purposes, and at the least is an interesting conversation piece in the hamshack.

The readout and keyer clock, which I will describe, may be easily used in any Accu-Keyer design, and possibly in other types of keyers as well. The main precaution to be observed with the Accu-Keyer family is to be sure the 5-volt power supply in your keyer is capable of handling the extra current drain, about 370 mA.

I do not consider it feasible to use my readout with a battery-operated keyer,[9,10] but it should be possible to make relatively simple modifications to the circuit and use CMOS integrated circuits. It would be necessary to choose some other type of display, and I would recommend a liquid-crystal type.

principles of operation

A continuous speed readout in wpm ***requires*** a free-running clock. The Accu-Keyer clock, however, is not free running. It starts when either side of the paddle is closed, and is stopped by an inhibit signal from the logic when all characters have been completed. This method has a considerable advantage over a free-running clock, since the operator initiates a character at the time he chooses rather than at the time the clock is finally ready.

This dilemma is easily overcome, and the unit I have developed gives an accurate, continuous readout of the speed without sacrificing the advantages of the operator-started clock. A fringe benefit of the unit is that it does not have the problem, common to some keyer clocks, of a first clock pulse different in duration from the rest of the pulses in the sequence. Because of these features, it may be worthwhile to use the clock portion of this unit, even without the readout.

The speed is variable from five to around fifty wpm, an adequate range for almost anyone from Novice to Extra. The speed display is updated approximately six times each second, whether or not any sending is being done. I incorporated it into the WA9LUD memory version2 of the Accu-Keyer, but of course it can be used with any similar keyer

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design. Different speeds may be selected as you build.

Recent editions of the ARRL *Radio Amateur's Handbook* give the relationship between code speed and keyer clock frequency as:

$$speed\ (wpm) = 1.2 \times f\ (clock\ frequency)$$

Twenty pulses per second of the clock result in a keying speed of 24 wpm. A scheme for reading out this relationship has been described previously,[11] but that system has several disadvantages which are overcome by my circuit.

Suppose you have a high-frequency pulse generator running a 2420 pulses per second. Three decade counters hooked in series would count to 242 if they are allowed to count for exactly 0.1 second. If the least-significant digit (2 in this example) is ignored, it is then possible to display 24 in the readout connected to the digital counters. The reason for this approach will be discussed more fully later.

The high-frequency pulse generator can also be divided down by a decade and a duodecimal (divide by twelve) divider, a total division of 120, to give twenty pulses per second for the keyer clock. If you gate the divide-by-120 divider on and off with the original inhibit line in the Accu-Keyer, the resulting keyer clock line acts much like the operator-started clock, which is the key to the success of the Accu-Keyer design. This scheme allows us to have a free-running clock that can be accessed at the operator's

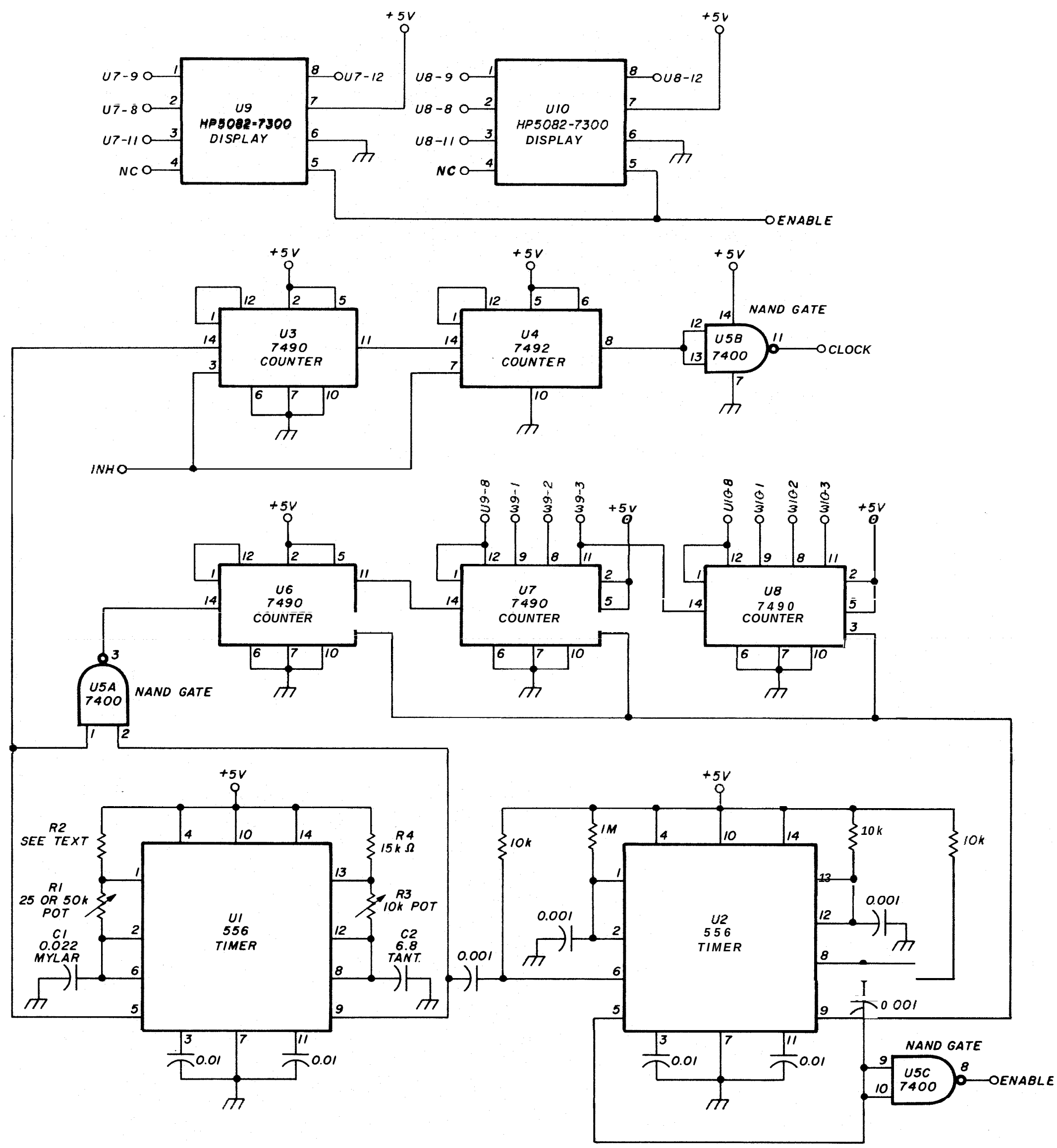


fig. 1. Schematic diagram for the Accu-Keyer speed readout. This circuit incorporates a free-running clock which can be accessed at will by the operator. The frequency of the clock is high enough that the delay between accessing and the first clock pulse is negligible. U9 and U10 are HP 5082-7300 displays that have the latches and display drivers incorporated within the display. C1 and C2 should be of the type indicated to ensure adequate stability.

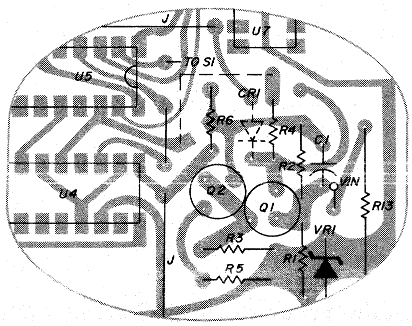


fig. 2. Blowup of the portion of the keyer board which is changed to incorporate the speed readout. CR1 and the original speed control wires must be removed. The foil is cut and new wires attached at the indicated spots.

convenience. I have never been able to detect any delay because of the free-running pulse generator, even at the slowest keying speed.

circuit description

The logic diagram for the clock/readout is given in **fig. 1.** U1 and U2 are 556 dual timers. One half of U1 generates the high-frequency pulses, available from pin 5, that form the basis of the clock/readout. The other half of U1 is the time base for the display counter, with the output on pin 9.

R1 is the speed control and is mounted on the front panel of the keyer. C1 must be a reasonably stable capacitor, *not* one of the ceramic bypass types. C1 and R2 determine the maximum keying speed, and the value of R1 determines the range. The value of R2 will probably be between 6,000 and 22,000 ohms for a 50 wpm maximum, and may be selected for this purpose. If C1 is changed for any reason at some later time, it may be necessary to change R2 to bring the maximum speed back to the one desired.

R3 is mounted on the printed circuit board and is used to adjust the 100-ms time base for the display counter. If it is not possible to adjust the "on" time at pin 9 of U1 to 100 ms, it may be necessary to change the value of R4 to bring the pot within the proper range. C2 is the most critical component in this entire circuit.

U2 is simply a sequential timer. The trailing edge of the 100-ms counter gate triggers a pulse of short duration at pin 5 of U2. This pulse, after inversion by U5C, strobes the count in the decade counters into the display. It also triggers another short pulse, at pin 9 of U2, which is used to reset the counters to zero, preparing them for the next update.

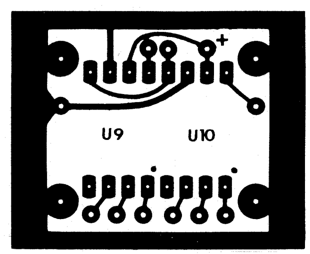
U3, a 7490 decade counter, and U4, a 7492 duodecimal counter, form the divide-by-120 divider that generates the clock pulses for the keyer logic. This divider is gated on and off by the inhibit line from the keyer, with the inhibit signal resetting the divider to zero and holding it there when all keyer action is complete. Inverter U5B ensures that the clock pulses have the right polarity for the Accu-Keyer, and might not be necessary in other keyer designs. This combination forms a keyer clock which is always within 11120th of a dit of starting, a negligible delay at any speed.

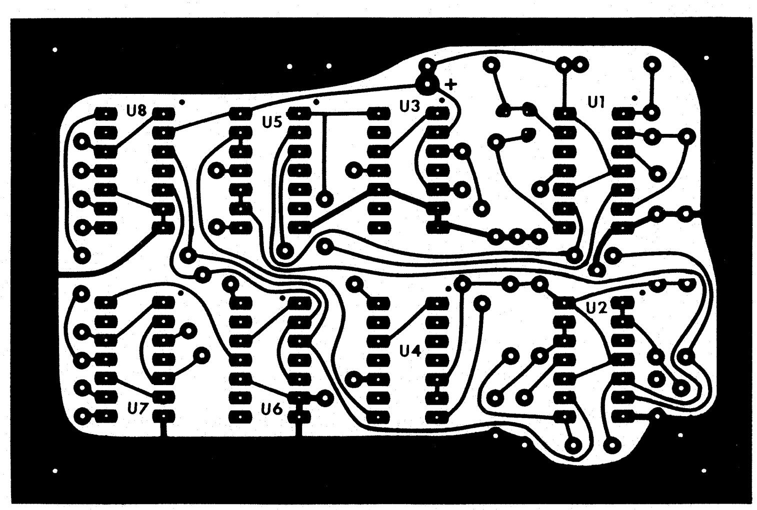
U5A controls the display counting. The pulse generator pulses are fed to the counter only when pin 9 of U1 is high. When it is high for precisely 100 ms, exactly one tenth of the pulse generator frequency is counted. U6, another 7490 counter, is for the least significant digit and, by including it without display, the jitter inherent in this digit is eliminated. This results in a stable display considerably superior to using only two decade counters with a 10-ms time base. U7 and U8, both 7490 counters, are the actual display counters, with U8 serving as the most-significant-digit counter.

The displays themselves, U9 and U10, are easy to use, with an attractive, bright display, although they are a bit expensive. Other displays may be substituted, but it might be necessary to incorporate data-storing latches, which are built into the 5082-7300 displays. A nonblinking display is a necessity, so be sure to add latches if they are not in the displays you choose.

Connection to the Accu-Keyer is really quite simple. CR1 in the original Accu-Keyer clock *must* be

fig. 3. Full-size printed-circuit layout for the Accu-Keyer speed readout. Parts layout is shown in fig. 4.





removed from the circuit. A wire is connected to the vacated hole at the anode end for connection to the inhibit line in the new clock. The foil should be cut as indicated in **fig. 2** and the old speed control wires should be removed. The clock line may then be connected to the vacant hole near the cut in the foil. Connect V_{cc} and ground both the readout and the clock board, and you're in business. You may wish to remove the old clock components from the Accu-Keyer board, but that is not really necessary.

I have not included a power supply, since most will be able to use the supply in the Accu-Keyer. It might be necessary to increase the size of the input capacitor ahead of the regulator to keep the voltage high enough to maintain regulation. If your supply is incapable of providing the necessary current, any standard 5-volt power supply design will be satisfactory.

Full-size board layouts and the component placement diagram are shown in **figs.** 3 and **4.** They are single-sided boards, and should be easy to duplicate by those who wish to roll their own. There is no reason why point-to-point wiring cannot be used, since the layout is not critical.

accuracy and calibration

The key to the accuracy of this unit is how carefully the 100-ms time base for the display counter is calibrated, and how stable it is. It would have been possible, of course, to use a crystal-controlled clock to control this counter, but that seemed quite unnecessary. One half of a 556 timer, with a high-quality, stable capacitor, results in quite adequate performance for this purpose. It saves considerably on circuit complication and expense.

There are three methods of calibration, and they will be described in order of increasing accuracy.

1. Set the keyer to match as closely as possible W1AW's 18-wpm bulletin broadcasts (or better yet their 35-wpm code practice), and adjust R3 until the readout indicates 18 (35).

2. Use a calibrated scope to set the "on" time (output high), as seen at pin 9 of U1, while adjusting R3.

3. Connect a counter with a l-second time base to U1, pin 5, to measure the pulse generator frequency, and adjust the keyer speed control until the counter reads about 4000. Adjust R3 until the display reads 40. This is the method I prefer, and should be used if a counter is available.

My own keyer has been in use for almost three years and seems to be accurate within one wpm at all speeds throughout its range at all temperatures encountered so far in my shack. Accuracy is not a problem if a sufficiently stable capacitor is used for C2.

I'll be happy to answer any correspondence regarding this readout or any modifications people may wish to make. I'll try to furnish circuit board availability information, provided that a self-addressed, stamped envelope is supplied.

It has been a pleasure to use this keyer with its readout. Now, when someone says QRQ by 5 wpm, I can do it quite accurately, depending on my skill of sending, of course!

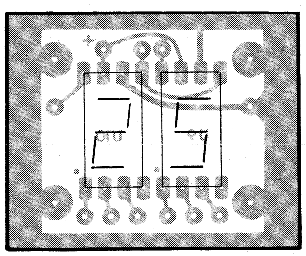
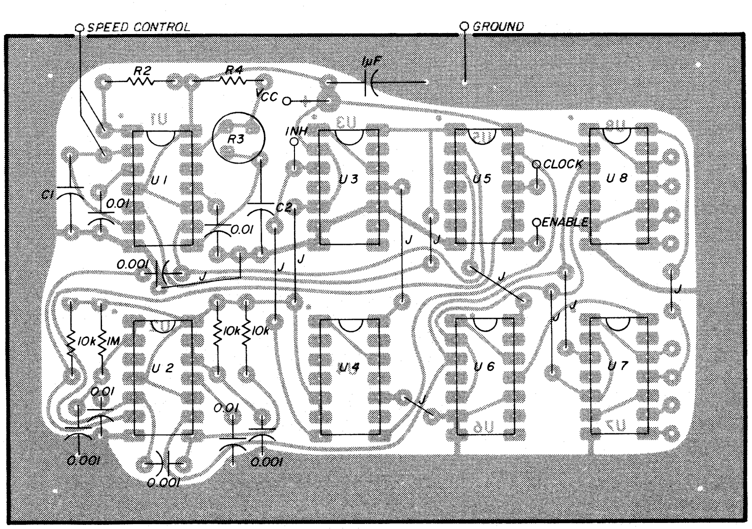


fig. 4. Component placement for the Accu-Keyer speed readout.



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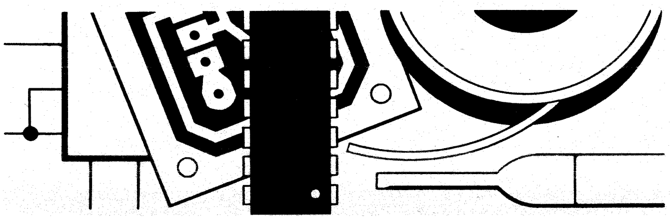
the weekender



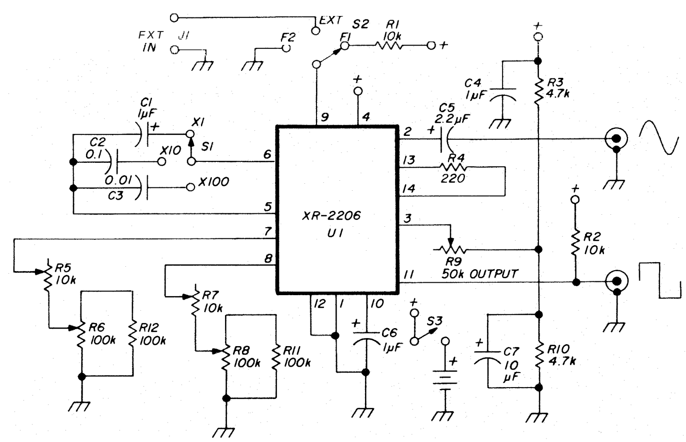
Duplex Audio-Frequency Generator With AFSK Features

Need a stable audio-frequency generator for testing, trouble shooting, and experimenting? Here's an instrument that fulfills these requirements and also provides some extra features for AFSK work. It's a weekend fun project that will reward your efforts with a truly versatile piece of test equipment.

The duplex audio-frequency generator covers the audio-frequency spectrum from 20 Hz-20 kHz, furnishes sine- and square-wave output simultaneously, and is battery operated for portability and interface safety. The generator has two frequency controls that are switch selectable from the front panel, or they can be selected through TTL level applied to the generator. In this manner an AFSK signal, relative to the TTL input, is generated. This electronic switching feature should be useful for the experimenter. The generator is constructed on a single PC board and can be easily completed in a weekend. An etched and drilled board is available for the project **(fig. 1)**,

The duplex audio generator is built into a Mod-U-Box available from Quement Electronics (see text). Controls are f1 (upper left), f2 (upper right), and output level (center).

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



B1	battery, 12.6 V, Mallory 304116 Smoke Detector Battery
C1, C4, C6	1 pf 35 vdc*
C2	.1 pf 35 vdc*
C3	.01 pf 35 vdc*
C5	2.2 pf 35 vdc*
C7	10 μf 16 vdc*
J1, J2, J3	jacks or binding posts of your choice
R1, R2	10K ¼ w*
R3, R10	4.7K ¼ w*
R4	220 ¼ w*
R5, R7	10k trimmer, Radio Shack 271-218*
R6, R8	100k pot, audio taper, Radio Shack 271-1722
R9	50k pot, linear taper, Radio Shack 271-1716
R11, R12	100k ¼ w*
S1, S2	switch, single-pole, three-position
S3	switch, mounts on R9, Radio Shack 271-1740
U1	XR2206, James Electronics XR2206*
Case	Mod-U-Box 3-7-6, Quement Electronics
PC Board	J. Oswald 1006J*

PC board and board-mounted components kit available from J. Oswald, part 1006K (includes parts marked*)

J. Oswald, 1436 Gerhartd Avenue, San Jose, California 95125.
1006J $4.75 PPD. 1006K $17.75 PPD.

James Electronics. 1021 Howard Street, San Carlos, California 94070.

California residents add sales tax.

fig. 1. Schematic of the duplex audio-frequency generator. Design is built around the James Electronics XR-2206 function-generator IC and includes an AFSK signal.

and, because component count is small, cost of the project is minimal.

description

The generator is built around the XR-2206 monolithic function generator IC. This little IC can perform many functions, and in this particular application we're using only a couple of its many features. As seen from the schematic, **fig. 1,** the XR-2206 and a handful of passive components form the entire generator.

The audio-frequency spectrum is covered in three ranges; 20-200 Hz, 200-2000 Hz, and 2 kHz-20 kHz, as selected by the range switch, S1. The range switch is labeled X1, X10, and X100, allowing the use of a single scale on the frequency dials. The specific frequencies desired within these three ranges are selected by the frequency controls, F1, (R6), and F2 (R8). Switch S2 selects the generator output frequency as F1, F2, or EXT. In the F1 position R6 determines the output frequency, while the F2 position allows R8 to control the output frequency. With switch S2 in the EXT position, the output frequency is selected by the signal applied to the external input jacks. Frequency F1 is selected by a high level or

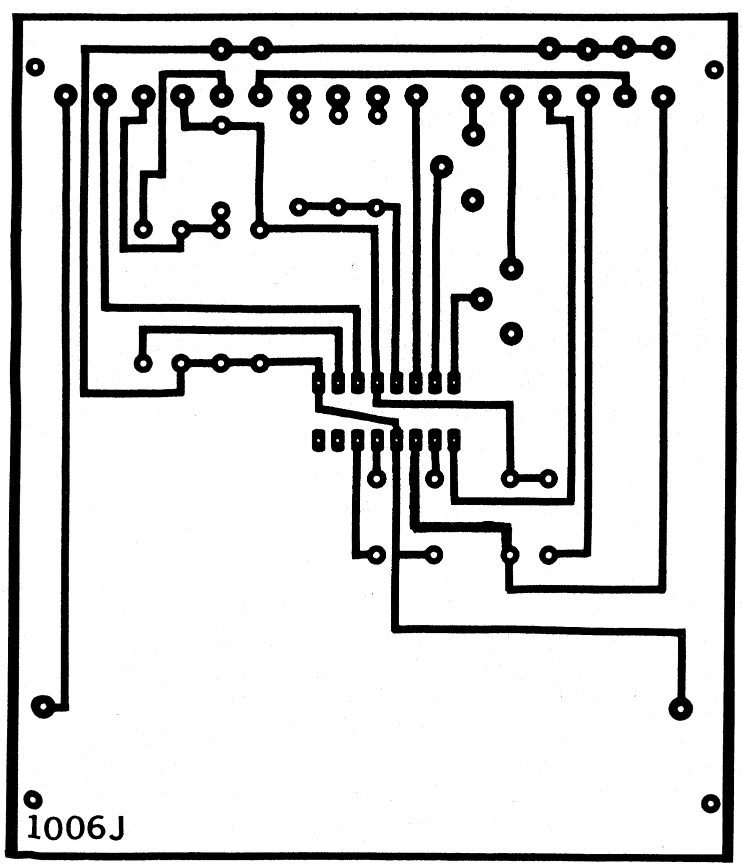
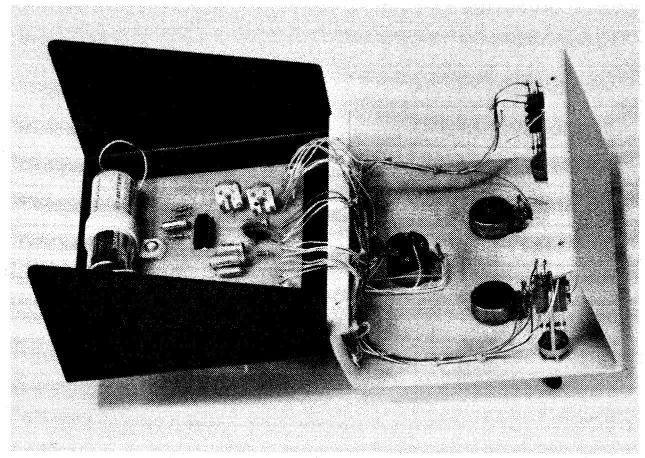


fig. 2. Full-size layout of the PC board (foil side).



Construction of the duplex audio generator showing the two switches and three variable controls on the removable front panel (right), and the printed-circuit board in the base (left). Power is provided by a 12 6-volt battery designed for smoke detectors.

open contact, and F2 is selected by a low level, or closed contact. In this manner an AFSK signal of adjustable frequency, shift, and amplitude is generated.

The sine-wave output of the generator is variable to a maximum of 3 volts peak-to-peak through the output level control, R9. The square-wave output is fixed at a TTL level; and because both outputs are available simultaneously, the square-wave output provides a very handy sync point for scope triggering.

The generator is powered by a 12.6-volt battery. Current consumption is low, so extended battery life can be expected. The basic circuit is not overly critical to voltage changes, and the first indication of battery failure will be flat topping of the sine-wave output at high-amplitude levels. Trimmer resistors, R5 and R7, are used to calibrate the frequency controls, and capacitors C1, C2, and C3 determine the range-multiplier accuracy. Generator output impedance is a nominal 600 ohms and provides a good match to most standard audio equipment.

construction

Virtually any type of construction practice could be used for the audio generator, because the circuit isn't critical the way rf circuits are. PC board construction was chosen for ease of assembly and predictable results. A full-size layout of the foil side of the PC board is shown in **fig. 2.** The component layout as viewed from the top, or component, side of the board is illustrated in **fig.** 3.

A practical approach to construction is to mount and solder all board-mounted components and then add the interconnecting wires to the front edge of the board. Leave these wires about 30 cm (12 inches) long for connection to the front panel after doing the sheet metal work. Drill and deburr all holes for the front panel controls and jacks, as well as the PC-board mounting holes in the case. Next, mount all front-panel components and the PC board, using small standoff spacers to elevate the PC board above the case.

Place the front panel next to the case and wire the interconnecting leads from the PC board to the panel controls, jacks, and switches. As the case goes together, the interconnecting wires should fold over neatly. After you're sure that the wiring doesn't interfere with the case assembly, the wires can be spot-tied to form neat groups and retain their positions. Assemble the case again to insure that everything fits well.

test and calibration

A very simple test of the generator can be made with the aid of a pair of headphones. Set trimmers R5 and R7 to midpoint, connect the phones to the sine-wave output, set S2 to the F1 position, range switch to X1, the frequency controls fully counterclockwise. Advance OUTPUT LEVEL control, R9, until an audio tone is heard in the phones. The tone should be about 250 Hz. Flip the function switch to the F2 position and adjust trimmer R7 until you can flip from F1 to F2 without detecting a change in the tone. Change the range switch to the X10 position and rotate the frequency controls fully clockwise. Again this should yield a tone about 200 Hz.

Flip the range switch to the X100 position. The tone should be approximately 2 kHz. Switch back down to the X10 range, listen to the 200-Hz signal for a few seconds, then move the phones to the square-wave output jack, J3. The signal should be a bit raspy because of the square waveform. Now set the function switch to the EXT position and the F2 control to its midpoint. Short the external input jacks with a jumper, and the 200-Hz signal should shift to approximately 750 Hz, indicating that the AFSK circuitry is functioning. This evaluates all the functions of the audio generator and should make you feel pretty good.

Calibration of the little generator is subject to the equipment you might have available, such as a scope or frequency counter, and also to the degree of accuracy you're trying to attain. The generator is a small package, so there really wasn't much room for large dials that would provide high resolution. With the small dials, accuracy is adequate for general audio work, and if you are going to do anything critical you'd probably use a counter with the generator.

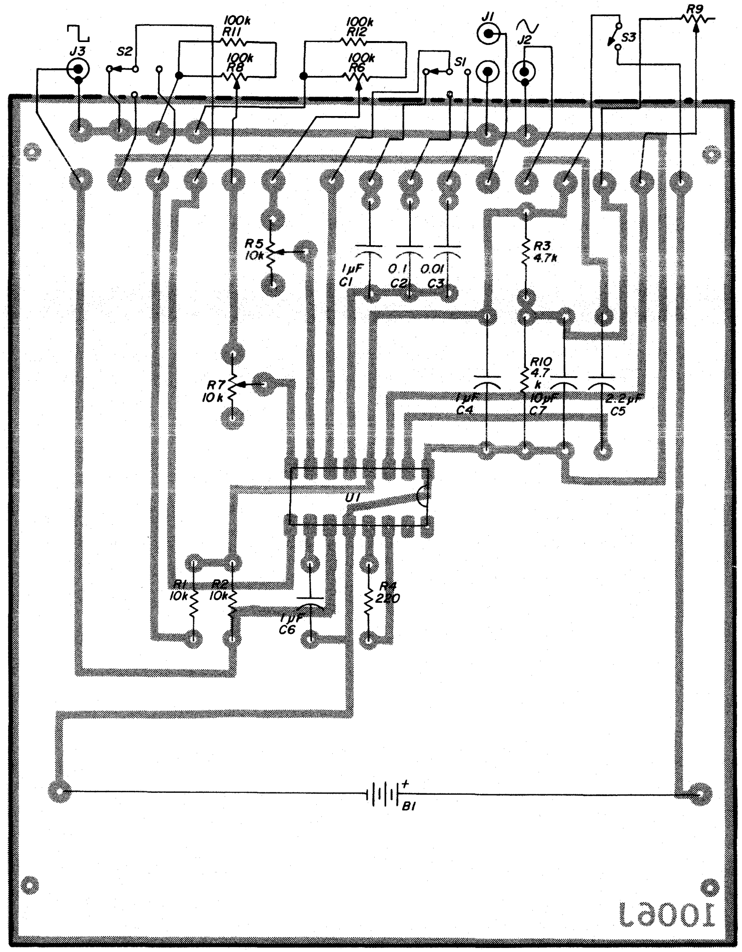


fig. 3. PC-board component layout viewed from the top, or component, side.

The dual outputs make this a very easy thing to do. I used a scope to measure the output pulsewidth and put small pencil marks around the dial area of the frequency controls. Then I removed the controls, switches, and jacks from the front panel and applied the lettering using Datak rub-ons, followed by a light coat of clear Krylon to protect the lettering. This gives the instrument a professional look and provides adequate calibration. Use care in putting the front panel components back in place so that the lettering isn't damaged.

The duplex audio-frequency generator is a very flexible piece of test equipment that can be built at a low cost and is worthy of a place in every experimenter's shop. It can be built by a beginner and will remain useful to him as his interests change to the more complex phases of Amateur Radio. Also, the basic circuitry can be lifted for many other applications, limited only by the creative ability and interests of amateurs and experimenters.

ham radio

GITAL TECHNIQUES

down counters

Most counters can be considered *up* counters. Their binary states usually increase in bit weight. This is sometimes a problem when applying them to a phase-locked loop as the programmable divider. In such an application, the divider is first preset to a particular binary state by the front panel control. Counting then proceeds to an all-ones or all-zeroes state. This state outputs a pulse to the phase detector and presets the counters again. Division ratio is the difference between preset state and end-of-count state. It can be a problem with up counters.

Suppose you want to divide by 888. An up counter must be preset to the nine's complement of each decimai digit (nine minus the desired digit), or decimal **111** in this case. The up counter will then increase through 888 states until an all-ones condition is reached for end-of-count. Confusion arises because the decimal preset is in reverse of the desired decimal division.

A solution is to use a down counter, one whose states decrease with the number of input clocks. Preset and division are now the same number. Motorola makes such a device with the designation MC4016 and it is designed for PLL applications.*

a BCD down counter

The counter portion of the MC4016 is shown in **fig. 1** with waveforms. D flip-flops are used in place of the usual JKs, and all gates are ANDs. G3 is an open-collector AND to sense all-zeroes from the $\overline{Q}$ outputs.

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"The designation was formerly changed to MC74416 but is back to the original number. MC4316 is the military temperature version.

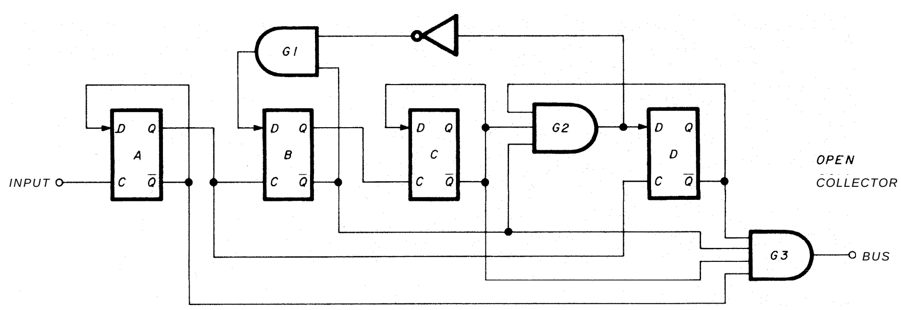
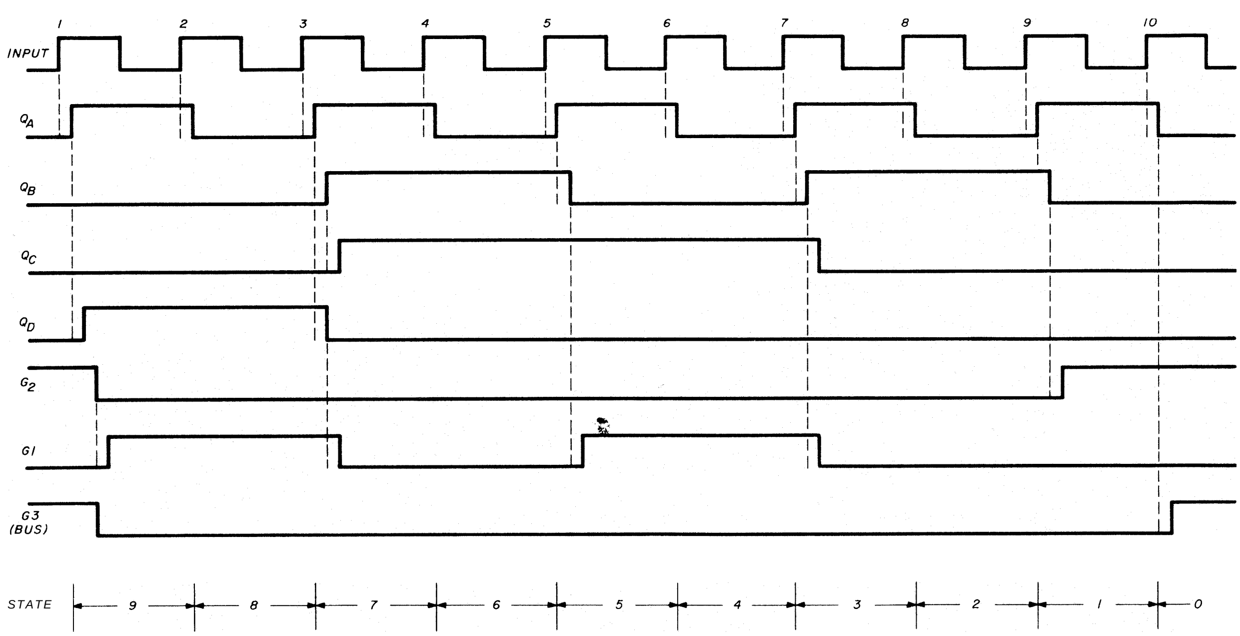
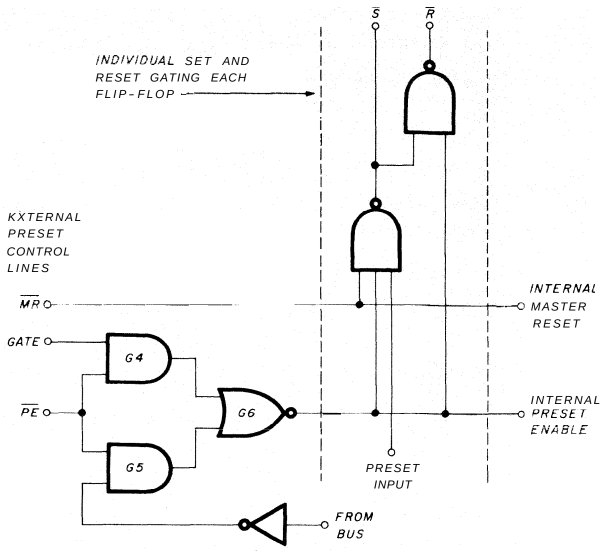


fig. 1. Decade down-counter section of Motorola MC4316/MC4016.





$\overline{MR}$	GATE	$\overline{PE}$	BUS	$\overline{S}$	$\overline{R}$
0	X	0	X	1	0
0	0	X	1	1	0
1	0	0	X	$\overline{P}$	P
1	0	1	1	$\overline{P}$	P
1	1	0	X	$\overline{P}$	P

TRUTH TABLE

fig. 2. Preset control section of the Motorola MC4316/MC4016.

Decreasing BCD states can be seen from the waveforms and the effect of gate output states. As in its up-count version, it's a divide-by-two in cascade with a divide-by-five. Carry out is from Q_D to the next input. A chain of three will go from an initial decimal 000 to a decimal 999 on the first input. Subsequent inputs will change decimal states to 998, 997, 996.

preset control

Each stage has direct set and reset inputs active low. Counting will be overridden when either is low. Two internal buses and three gates provide versatility in preset control. **Fig.** 2 shows the preset control section with a truth table for external control inputs. P indicates the state of the external preset input for each stage. An X is a don't-care state; it may be 1 or 0 without changing a particular state combination.

BUS gate 3 was stated as being open-collector. AND gates with open collectors may be wired-AND just as NAND gates may be wired-OR.[1] The internal connection to the inverter doesn't change the open-collector condition. An internal, separate, pull-up resistor is provided on each package.

Fig. 3 indicates a single package connected for division by eight. Waveforms are expanded to show automatic presetting. Preset inputs are wired for binary 1000 (decimal 8). Control lines $\overline{MR}$ and $\overline{PE}$ are tied high. A preset can occur only when GATE is low and BUS high — the all-zeroes condition.

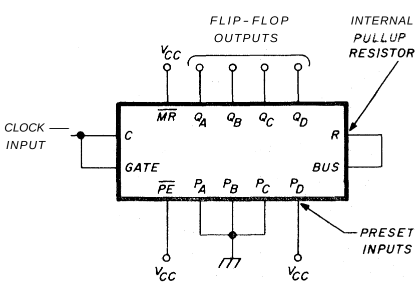
The waveforms assume that six clock inputs have occurred. Counter state is then binary 0010. The next clock (seventh) will make it binary 0001. The eighth clock will cause several actions.

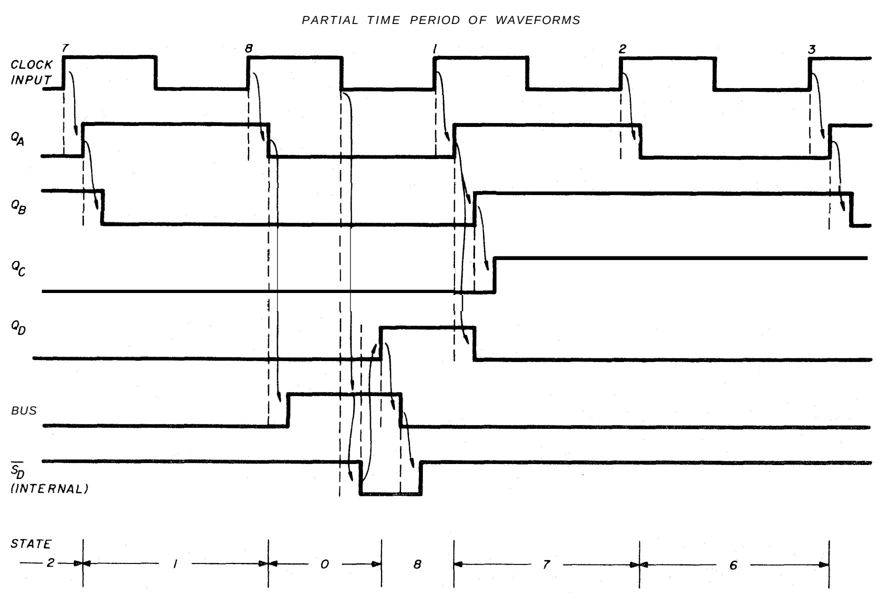
The counter goes first to binary 0000 and the BUS goes high. External control GATE is connected to the clock. It is still high after binary 0000 has been reached, so a preset doesn't begin until the clock goes low. At that time, the internal D stage active low set changes Q_D from 0 to 1. The other three stages reset; it doesn't change anything since they are already 0.

The counter is now set to binary 1000 and BUS is low, but the next positive clock edge will change the counter state to binary 0111. It counts down again until all zeroes are present. The carry out is only the width of the clock low state.

Maximum clock input frequency is limited by three propagation delays: clock positive edge to BUS going

fig. 3. Single MC4016 connected as divide-by-eight.





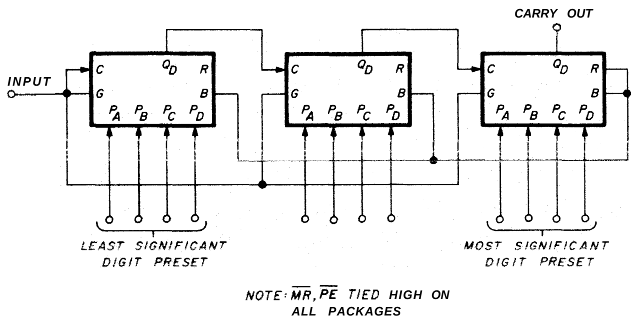


fig. 4. Three MC4016 counters in cascade for variable division.

high (65 ns maximumi, clock negative edge lo any flip-flop set (35 ns), and next-positive clock to a flip-flop toggle (78 ns). Inverse total is 5.6 MHz, but the nominal maximum frequency is 8 MHz.

cascading

Fig. 4 shows the circuit for three packages. It can divide by any number from 1 to 999 depending on the **BCD** input to each counter. An MC4018 can be substituted. It's a hexadecimal (divide-by-sixteen) version, and three packages would yield a maximum count of 2^{12}, or 4096; four-bit binary preset inputs would be required.

All $\overline{\text{MR}}$ and $\overline{\text{PE}}$ control lines are tied high. All GATE inputs are connected to the input clock. All BUS pins are tied together, but only one R or pullup resistor connection is required.

Preset action is the same as in **fig.** 3 and depends on the first, or left-hand, counter for speed. The last, or right-hand, counter will reach all-zeroes first, then the middle. The BUS is almost ready to go high, but the wired-AND connection makes it dependent on the first counter. When the first counter goes zero, preset is enabled to all; the BUS goes high, then the common GATE goes iow.

Carry out may be from the third Q_D, but fast inputs should use the BUS line since it's slightly wider. Speed is limited, but a few extra devices wil! increase this.

increasing speed

Input frequency can be increased to at least 25 MHz by adding a D flip-flop, 5-input gate, and three inverters as in **fig. 5.** Schottky TTL devices are recommended. Note that this version has the first counter's BUS pin grounded and all $\overline{\text{MR}}$ and GATE pins tied high.

Previous connections initiated a preset on the input clock low state. **Fig. 5** allows nearly a full clock period for preset. This is possible by arming the preset when countdown has reached decimal 002 (binary 0010 in the first counter). Presets have been hard wired for 888 division for illustration.

G7 goes low on the 886th input (representing decimal 002). The external flip-flop will toggle on the 887th input. This action initiates a preset by making the common $\overline{\text{PE}}$ control line low. Preset completion will make G7 high but won't change the external flip-flop, because its clock, the 888th, has not yetarrived.

Once the 888th clock arrives, the external flip-flop will toggle, but the counters will not change; $\overline{\text{PE}}$ is

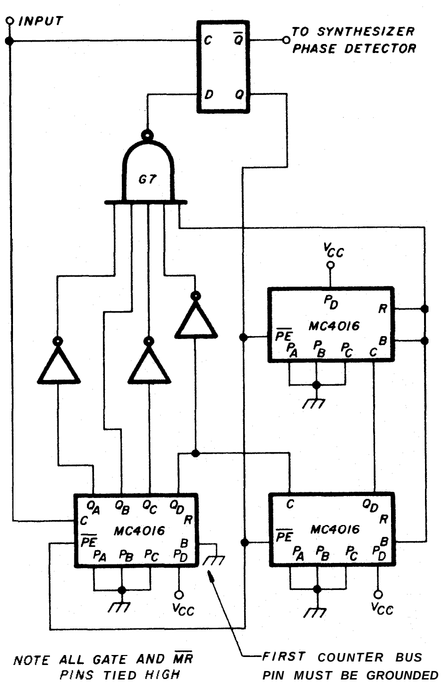
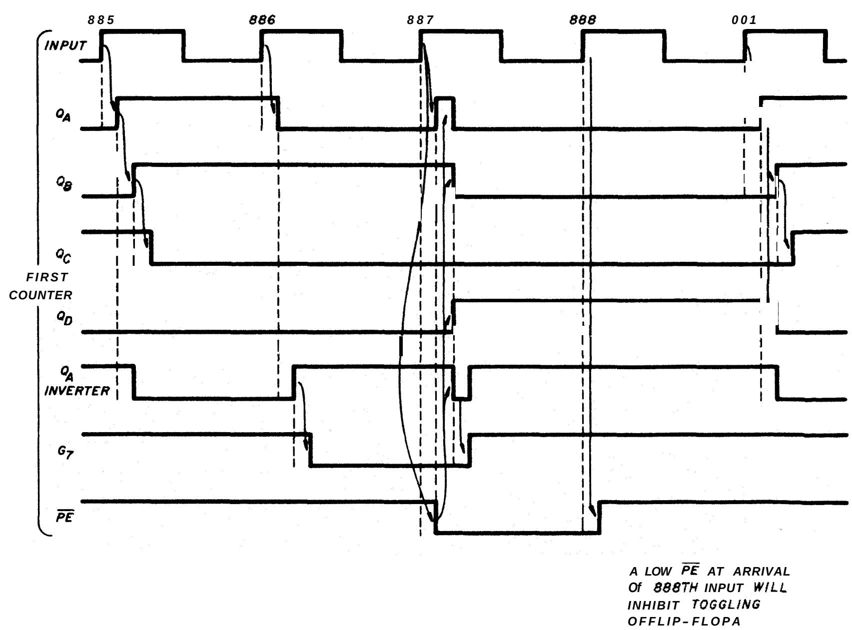


fig. 5. High-speed divide-by-888for use with phase-locked loop.



still held low at clock edge, and all counters remain at preset at that time. (See the third state of the truth table in **fig. 2**.) The first counter essentially ignores the 888th clock.

The external circuitry permits a substantial increase in speed even though the counters are not synchronous. The only disadvantage is slight: division by less than three is not possible.

other packages

Presettable up/down counters are available. These can be connected for down counting only with external circuitry added for similar preset-enable control. The use of synchronous counters and Schottky TTL programmable dividers is possible up to 60 MHz. Great attention must be paid to propagation delay at high speed.

The Motorola device was selected for this example because it contains the essential ingredients of a counter with preset control ability.

reference

1. Leonard H. Anderson, "Digital Techniques: Gate Structure and Logic Families," *ham radio*, February, 1979, page 66.

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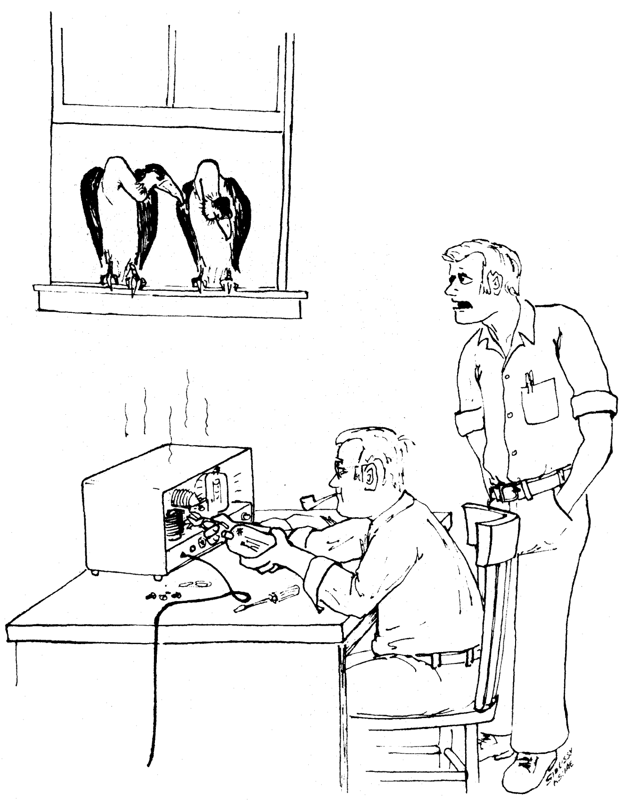

"I think you'd better unplug it, Stan."

More Details? CHECK — OFF Page 126

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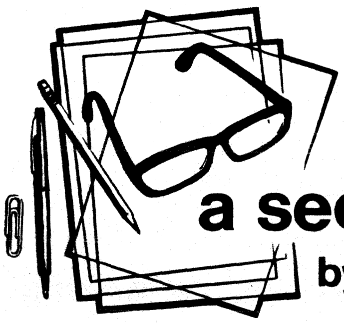


a second look

by Jim Fisk

I've been interested in the history of radio nearly as long as I've been a Radio Amateur, and am continually on the lookout for old wireless equipment to add to my collection. Local ham flea markets and auctions are often the source of unexpected treasures, and I attend as many as I can; not too long ago I *missed* an opportunity of a lifetime when I passed up an auction of surplus electronic equipment at an old, respected New England college. Much of the "surplus" gear dated back to the 1920s and was built into custom-made wooden cases which were then in fashion; most of the buyers, unfortunately, were antique dealers who were interested only in the finely crafted cabinets — the priceless radio equipment inside was destined for the trash heap.

In many respects it was a replay of an event in Oklahoma several years ago: The auction of the electronics equipment and *junk* collection of a prominent local amateur. From all reports, it was quite a collection, filling four large warehouses. Except for the huge volume (and original cost) the collection resembled the typical "hell box" of every amateur who lived through the halcyon days when building your own transmitter was conventional practice and everyone eagerly added to his junk collection at every possible opportunity.

However, there was one big difference in the Oklahoma collection: Where most Radio Amateurs painfully part with dollars, this amateur painlessly parted with thousands of dollars. Think back a few years — what was the most delectable piece of radio gear you remember? It was probably in the Oklahoma collection. And not just one, but several. Parts, radio sets, test equipment, you name it, it was all there in unimaginable profusion. One whole warehouse floor was reportedly crammed full of big transmitters, spark coils, and rotary gaps for 1920-style transmitters, spiderweb coils and thousands of variable capacitors of every possible make and description. A complete inventory would go on for pages.

Now here's the tragedy: These priceless articles, which belong in a museum, were grouped in huge lots with utter junk and sold to junk dealers! The probability that these dealers could differentiate between valuable antiques and valueless junk is frighteningly small. Antique radio equipment that can never be replaced, items not preserved in any museum, were probably bulldozed under at a county landfill dump.

This scene, on a much more modest scale, is probably repeated many times a year all over the country. Without getting morbid, each one of us should realize that we are not immortal. Each of us has a collection of electronic gear that will, if someone doesn't know any better, be bulldozed under with the trash at the city dump when we join the list of *Silent Keys.* Each item, when you acquired it, represented a jewel to be treasured and was carefully put away. If you were ever so careless as to toss out one of these treasures, you almost certainly would have an immediate pressing need for an identical article. I know, because it happens to me every time I clean house!

The point is this: Talk to your heirs. Clue them in as to what items, if any, belong in a museum. Better yet, make arrangements with the executor of your estate to donate certain prized items to a local school or college; antiques should be given to a museum of your choice. The same sort of foresight applies to your newer equipment as well. Give your executor the names of several trusted amateur friends who will help dispose of modern radio gear and test equipment. They will know the fair market value — your executor may not. There have been more than a few cases where an amateur's survivors have been ripped off to the tune of thousands of dollars; don't let it happen to your family.

Jim Fisk, W1HR
editor-in-chief

comments

RTTY demodulator

Dear HR:

I would like to make some comments on the article by KB9AT in the October, 1978, issue: "Digiratt PLL2."

In his third sentence the author states, "The one common drawback to PLL terminal units is that they decode only half the available information present in the RTTY signal." In other words, we could eliminate one of the two tones and still end up with the same results. We could go to a similar signal like CW but Baudot encoded. This would save some bandwidth and let our transmitter finals live longer because of the shorter duty cycle. As far as I know, most if not all TUs use some kind of discriminator, PLL or otherwise, which produces at its output a positive voltage corresponding to one tone and a negative voltage corresponding to the other tone, both with respect to a reference voltage, which might or might not be at ground potential.

These two voltages are then processed through filters and/or a comparator before going to the keyer stage. What if one of these tones is missing? One of these two output voltages is also missing. And the place where the missing tone should be is now filled with noise. When we send that noise through a comparator to the keyer stage, the keyer stage will start switching back and forth on noise pulses. So we see that it's absolutely necessary to have two tones.

My second argument is with the author's poor choice of PLL for his TU. If at any time he had read the data sheets and looked at the graphs that go with this PLL tone decoder, he definitely would have made another choice. First, the *data sheet* gives a simple equation to calculate the bandwidth, which is

$$BW = \frac{V_i}{F \bullet C2}, \qquad (1)$$

and the minimum detectable signal is 20 mW. Inserting this value into **eq. 1** shows that with the author's choice of a filter capacitor, which also determines the bandwidth, we have a bandwidth of 15 Hz. Going to a more normal input voltage of 2 volts we find a bandwidth of 48 Hz. Now, going with these numbers one step further, on the data sheets we find the graph *Greatest number of cycles before output.* We find that as much as 300 Hz occur before the 567 will lock up on that short 2125-Hz tone.

We know that each bit in our Baudot-encoded machines at 60 wpm is only 0.022 second long. This means that we have, from those 2125 Hz, only available per bit 0.022 x 2125 $= 46.75\ Hz$, far less than we need in the worst-case lockup time. So we lose that bit altogether. When we have an R or a Y, where we have six transitions, we switch back and forth between the two PLLs; and every time each PLL must search and try to lock up.

Agreed, the data sheet also says that "assuming random initial input phase, only during 1/6 of the time will there be a lockup time longer than half the worst-case lockup time." This means that we can transmit all bits in a letter such as R or Y. Furthermore, we shouid take into account that most machines will print garble when we have a signal distortion (shortening of the pulse) of 30 per cent on a well-adjusted machine!

Now even with the loop filter C compietely removed, the worst-case lock up time can still be 27 Hz, which is far more than 30 per cent, and is about 58 per cent.

So with all of this in mind we can see that the 567 is a very poor choice for 60 wpm or more on RTTY, and we can see why this chip was never intended to be used for that purpose. There are many other choices that would have been far better, which I've tried in all the years I've been on RTTY. Other chips that are intended for RTTY have a lockup transient only at the very beginning of each transmission, or when the signal appears at the input of the TU. By careful choice of loop filter C, these loops will almost instantaneously follow each tone deviation that might appear within its capture range.

Joe C. Zegers, WB6PMV
Sunnyvale, California

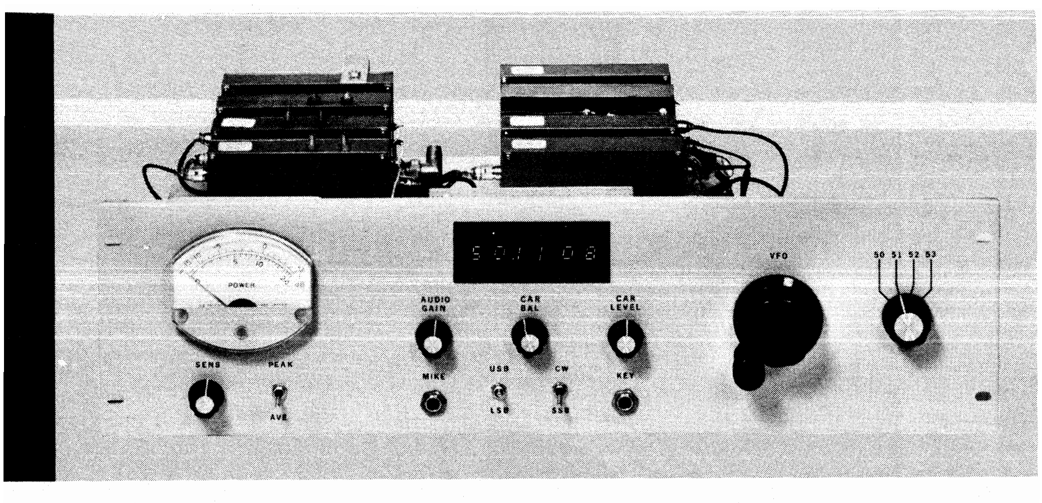
Dear HR:

Reader Zegers has raised some interesting questions in his letter. I wish to reply to his statements with this letter.

I stand by my statement that phase locked loops decode only one frequency, which in RTTY is one half of the information. Actually, a more correct statement would be that they decode only a small range of frequencies within their detection bandwidth.

It is true that PLL circuits (such as the popular 565 for instance) are often connected to an op-amp, and a positive and negative voltage swing obtained for use in keying circuits. However, both the 565 and the 567

(Continued on page 80)



50-MHz SSB exciter

Construction details of a high-performance, 50-MHz exciter which features CW/SSB operation, digital readout, and modular-type construction for circuit improvement and modification

This article describes an SSB/CW exciter with digital frequency readout. It covers ±100 kHz segments centered on 50, 51, 52, and 53 MHz. The power output has been limited to 50 milliwatts because the primary use of the exciter is to drive solid-state transmitting converters. The wide frequency range allows coverage of all present weak-signal band segments and satellite passbands.

A modular approach was used in building the rig. I was convinced of the flexibility of this technique by Joe, W1JR. The rig could very easily have been built with fewer modules and still retained its testability. However, I wanted to be able to change very small pieces of the rig without pulling the whole thing off the air. I also felt that taking modularity to this somewhat sublime limit would make the exciter more reproducible. There is a disadvantage to the approach though; you do not end up with a small box. But if you are a casual builder or apt to change designs in midstream, then this is the only way to go.

The exciter can be broken down into analog and digital sections. **Fig. 1** shows the block diagram of the analog section, which is made up of ten modules. Each module contains a "complete" function. If you start with the audio preamp, the exciter can be built up from it; that is, all existing modules are used in testing the next. I kept the port impedances of the rf modules at 50 ohms to aid in testing. It also allows easy changes to any given module with minimum impact on the whole project.

Most of the analog portion of the rig was stolen in bits and pieces from the many designs that have appeared in various Amateur Radio journals. Wherever possible, I picked the "cheapest" way to go, consistent with getting the rig to fly with minimum

By Rick Commo, K1LOG, 3 Pryor Road, Natick, Massachusetts 01760

problems. I also imposed the constraint that there would be no hot switching of any rf, in order to facilitate a neat and functional panel layout.

The digital section, shown in fig. 2, consists of six modules, or in this case, boards. It was also built from the bottom up starting with the time base. References are included at the end of this article.

The specifications for the exciter, which are tabulated below, are on a par with the majority of commercial low-band rigs advertised in the Amateur journals:

maximum power output	50 mW **CW**
carrier suppression	>45 dB, **with respect to single tone**
IM products	**– 32 dB, with respect to two tones**
harmonics responses	– 50 dB, **with respect to full carrier**
nonharmonic responses	– 45 dB, **with respect to full carrier**

audio module

The audio module (see fig. 3) is not complicated. Two op-amps provide a voltage gain of about 50 and lowpass filtering. An LM308 was used for the first stage because it has somewhat lower noise than most garden-variety op-amps. Just about any op-amp could be used for the second stage, with appropriate changes in (or deletion of) the compensation capacitor.

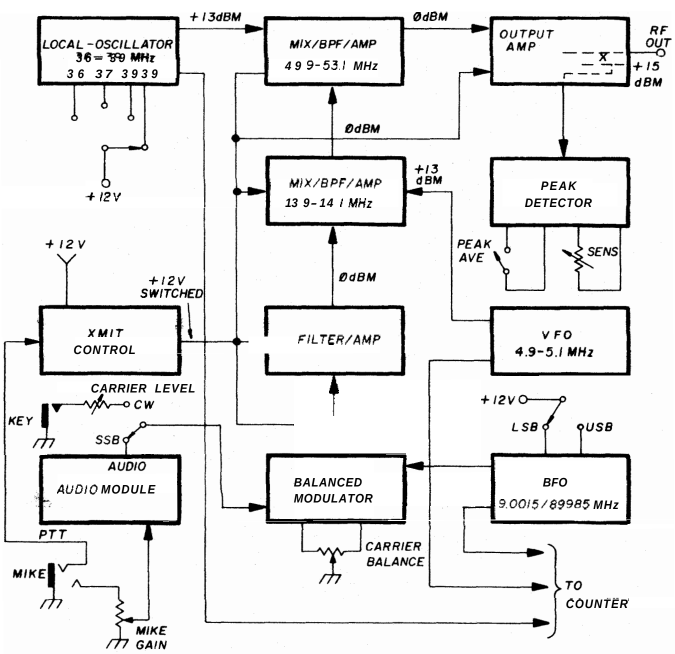


fig. 1. Block diagram of the 50-MHz exciter. To facilitate circuit changes and improvements, each block is built into an individual box, with the input and output impedances set at 50 ohms.

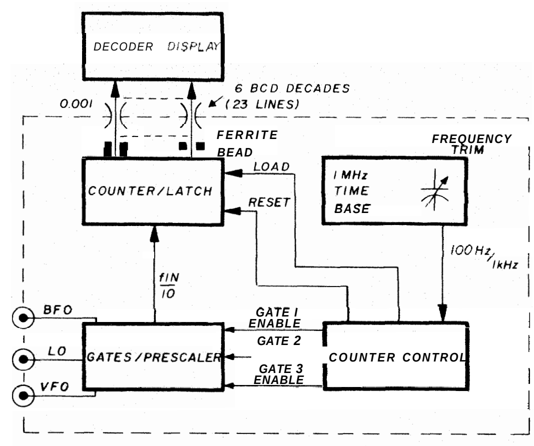


fig. 2. Diagram of the frequency display portion of the exciter. Each oscillator is individually gated through to the main counter, providing a true readout – rather than what would be derived from the less-accurate presetting method.

In this design the first stage has a gain of five and the second a gain of ten. This was more than enough for the microphone that I use. Any gain adjustments the reader feels are necessary in his own case should be done in the first stage. This will preserve the roll-off of the lowpass filter.

Be forewarned: Getting skimpy with audio bypass capacitors is not a good idea, especially in the bias network for the unused inputs. Any hum that appears on these inputs gets amplified right along with the main audio. It doesn't take too much noise to make a poor sounding signal.

BFO module

The BFO module, as shown in fig. 4, contains two oscillators and a buffer. The oscillators are pretty standard, the only thing done special being to match the fets for the approximately equal zero bias drain current (I_{dss}). This was done by shorting the gate to the source and measuring the current with 12 volts applied between the source and drain. The buffer amplifier is described in the ARRL ***Solid-State Design* manual.**[1]

The amplifier was driven directly from the oscillators. This was done to insure a clean waveform. With an input impedance of essentially zero ohms, the amplifier keeps the crystals adequately decoupled from each other.

Two outputs are taken from the amplifier. The high-level output goes to one input of the frequency counter. The low-level output, which is variable (200 mV max) goes to the modulator. The oscillators are adjusted on frequency by C1 and C2. If the counter is built first, it can be used to do the netting. I used the McCoy crystals that came with the filter, although a 9.000-MHz KVG crystal was tried and no problems were encountered.

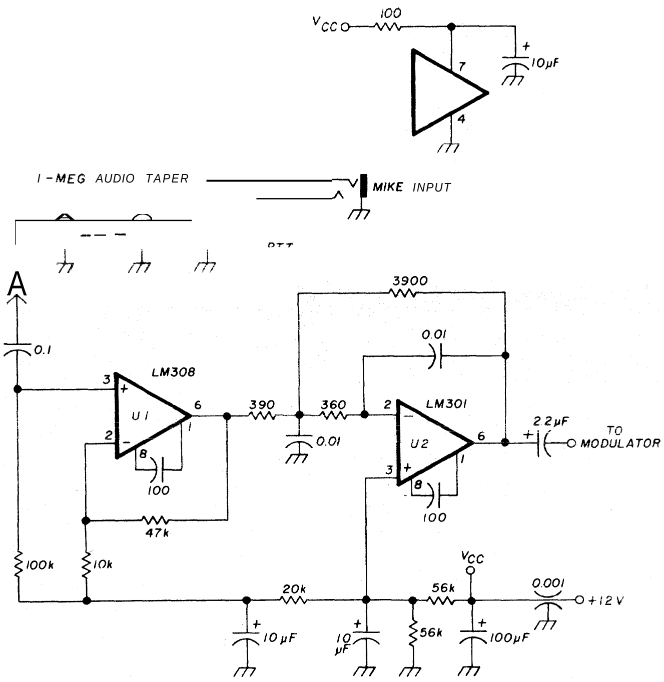


fig. 3. Schematic of the audio module. The filtering in the + 12 volt line helps ensure that no hum will be present on the signal.

balanced modulator

The modulator (fig. 5) uses an MC1496. This circuit has been around a while,[2] but it's easy to get going and has excellent carrier balance. It is fed with about *60* mV of 9-MHz rf and about 600 mV of audio. A balanced circuit was used in the output. This gives a 3-dB increase "for free." A single-ended design will work as well, and has been used in many other designs. The output of the module is about – 10 dBm.

Some thought should be given to the layout of the modulator in order to prevent BFO leak-through and, hence, degraded carrier balance. The initial attempt at building the modulator had all the components stuck in the box. This resulted in some carrier coupling to the audio inputs. Some 200-pF bypass caps cured the problem, but I decided to rebuild the module anyway. The second module was built with many of the modulator pins going to Teflon feed-throughs. In this way, most of the dc and all of the audio was wired on the outside of the box. This resulted in a much cleaner layout and the module worked the first time power was applied.

Care should also be taken to avoid overdriving the modulator with the BFO. Initial efforts with this module were very frustrating because of this problem. The best way to set the modulator up is to null the carrier, then inject an audio tone and reduce the BFO drive level until the output of the transmitter starts to fall off. Leave the BFO set at that level.

The filter-amp module provides gain and matching for the 9-MHz crystal filter (see fig. 6). I used a McCoy Golden Guardian, because I got some at a flea market cheap. Other filters may be used, however, if appropriate changes are made in the matching resistors (560 ohms tor the McCoy). The gain of the module is controlled via the unbypassed emitter and source resistors. The second gate of Q2 can be used as an ALC input if desired.

Attention should be paid to the supply-line filtering to reduce leakage around the filter. While this is not as important in an exciter as it would be in a receiver, it still pays to be careful.

The variable capacitors on the input and output of the filter are there to resonate out the residual inductance of the filter's transformers, thus reducing passband ripple. This is most easily done by padding the LSB crystal to exactly 9.0000 MHz, and then tweaking the two capacitors for maximum output. This will get pretty close to optimum passband ripple.

VFO module

The 5-MHz VFO (fig. 7) is built around an ARC-5 (transmitter) capacitor. This tends to make the VFO quite a bit bigger than it might have been, but it does have the advantage (in VFO design) of mass and built-in gear reduction. The "back wall" of the capacitor was drilled and tapped so that the VFO circuitry could be mounted directly on the capacitor.

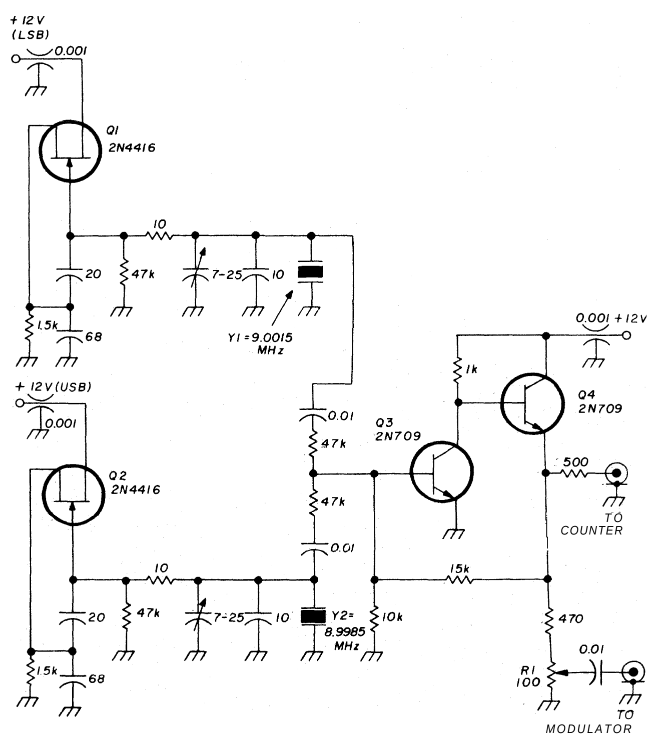


fig. 4. Schematic diagram of the BFO module which supplies the 9-MHz signals to the crystal filter. The level adjust, R1, must be a carbon potentiometer.

The oscillator uses a jfet and is buffered with a dual-gate mosfet. I found that the loading of a mosfet was significantly less than that of a jfet. The output of the buffer goes to an amplifier to raise the output up to a usable level. No compensation has been done on the VFO as yet. However, since the ARC-5 capacitor is made of Invar, the VFO is fairly stable as is. At room temperature, the drift over a two-hour period, starting cold, was on the order of 300 Hz. The frequency variation over wide temperature ranges has not yet been investigated.

Harmonics were a problem with this VFO. Without the half-wave filter, the second and third harmonics were only 15 and 20 dB down, respectively. The third harmonic created a bad spur, since 15 MHz is not that far down on the skirts of the first mixer's filter. Adding the half-wave filter brought the second harmonic down to 30 dB and the third down to 48 dB, at which point the spur problem went away.

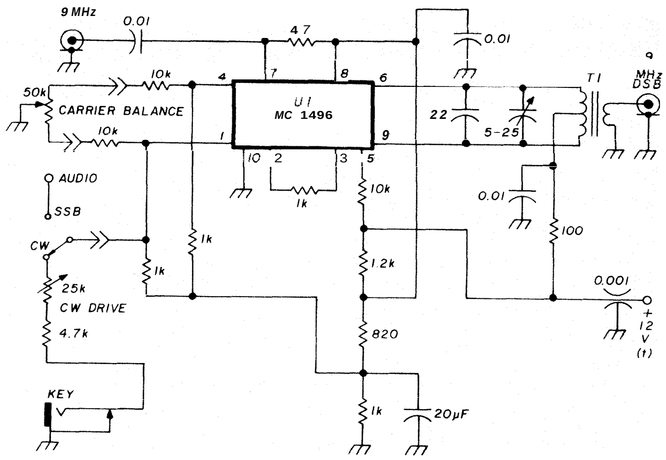


fig. 5. Diagram of the balanced modulator module. C1 is a 5-25 pF trimmer. T1 is wound on a T50-6 core, with the primary twenty turns of no. 24 AWG (0.5 mm) and the secondary five turns of no. 24 AWG (0.5 mm). The two windings are bifilar wound on the core.

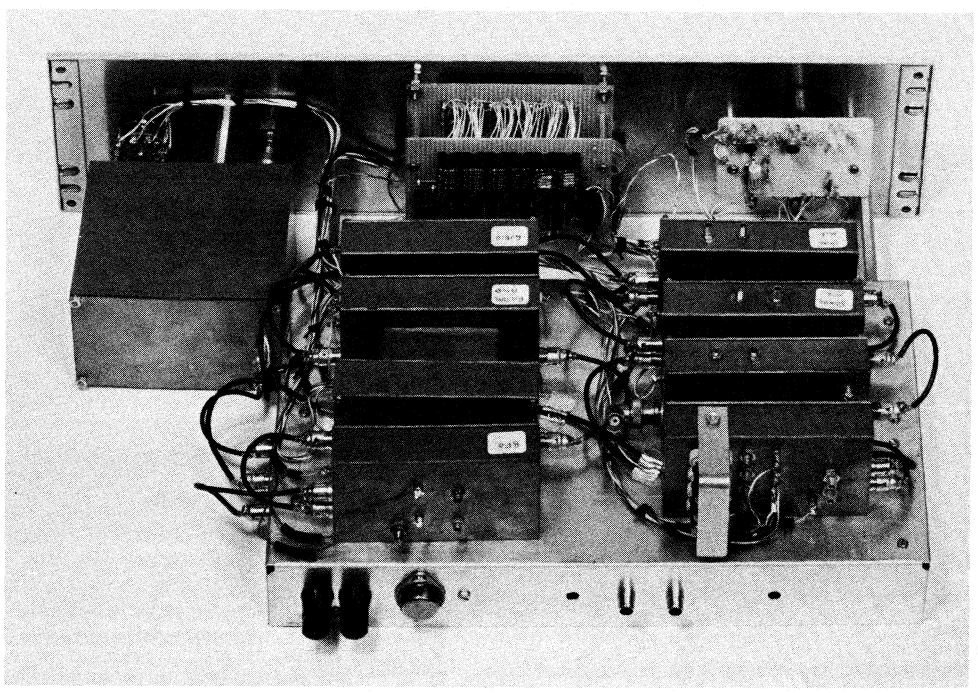
mixer modules

The 14- and 50-MHz mixers shown in fig. 8 are identical, with the exception of the filters. All ports of the double-balanced mixers are padded to insure that they see 50 ohms. If not, the mixer will most likely not perform properly. The output of the filter is also padded for essentially the same reason, to guarantee that the filters are properly terminated at all frequencies. The post-filter amplifier is a broadband design built around the 2N5179. It has more than enough gain to bring the signal back up to a usable level. The i-f signal into the modules is held to 0-dBm max, with an LO level of about+ 13 dBm. This is done to insure that IMD products are at least 30 dB below full output, which runs between 0 and +3 dBm.

The mixer modules turned out to be the weak link in the chain as far as IMD products were concerned. Since access to a spectrum analyzer was limited, further work on the mixers had to be curtailed. I suspect that substituting a 2N3866 for the 2N5179 and increasing the collector current to 30-50 mA would improve the IMD specs considerably.

LO module

The LO module (see **fig. 9**) consists of four oscilla-



Top view of the 50-MHz ssb exciter showing the individual modules used for the analog circuits. Large enclosure at left houses the VFO; modules on the main chassis include (center) audio, filter amp, balanced modulator, and BFO. At right are the 14-MHz mixer/filter, 50-MHz mixer/filter, output amplifier, and local oscillator modules.

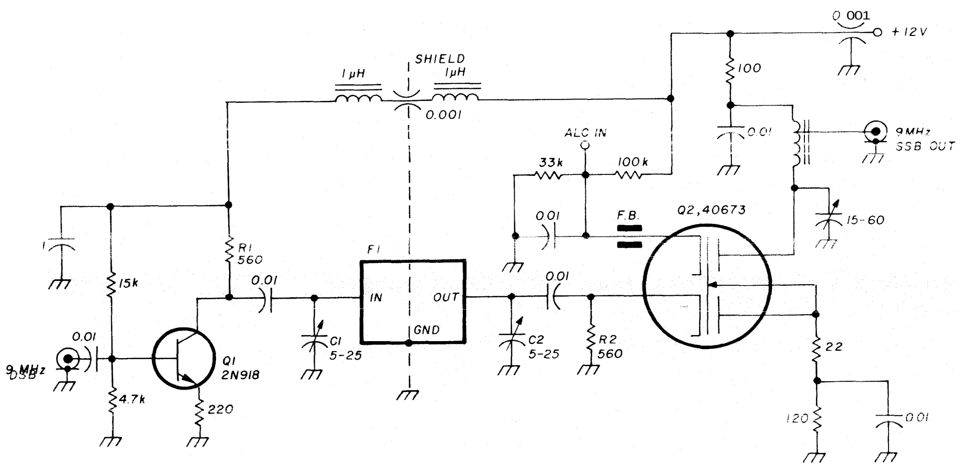


fig. 6. Schematic of the SSB filter and amplifier module. F1 is a KVG 9FA, 9FB. or equivalent, 9-MHz crystal filter. The terminating resistors and trimmers (R1, R2, C1, and C2) must match the filter used. L1 is thirty-eight turns of no 28 AWG (0 3 mm) wound on a T75-2 core The tap is located eight turns from the cold end

tors and an amplifier. I started out by trying to diode switch the crystals but was never really happy with the results. I ended up building four separate oscillators and switching the power to the desired oscillator. The secondaries of the tuned circuits are diode switched so that they would not interact with one another and also to reduce the loading of the active oscillator. While this technique takes more components than other methods of dc switching, it is versatile in that the frequency range of the module is limited only by the bandwidth of the buffer amplifier since no compromises are required on the part of the oscillators. This is significant if you want to steal the circuit for an all-band exciter or receiver where the LO frequencies could vary considerably.

A "broadband" class-A amplifier is used to buffer the oscillators and raise the signal level. The 4:1 balun normally expected in the output has been replaced with a 2-pole filter in order to reduce oscillator harmonics. The harmonics were all better than – 55 dB with respect to the LO signal.

A single stage of amplification was used in the 50-MHz PA module (see **fig. 10).** This was all that I needed to get the signal up to the 50-mW level. It uses a broadband, class-A amplifier into a 2-pole filter. The output is sampled and rectified for use in a peak detector.

I used an MSC CATV transistor because I had some on hand and I preferred to use a stud mount package. A 2N3866 would probably work just as well at a lower cost if you use adequate heat sinking.*

The transistor is biased for approximately 100 mA of collector current. The output filter is similar to the

*MSC transistors are available in stud mount (H002) or T05 package (H003) for $5 plus $1.50 handling for one H002 or two H003s. Send a certified check or money order and your callsign to Ham Trans. Box 383. South Bound Brook, New Jersey 08880, No phone orders.

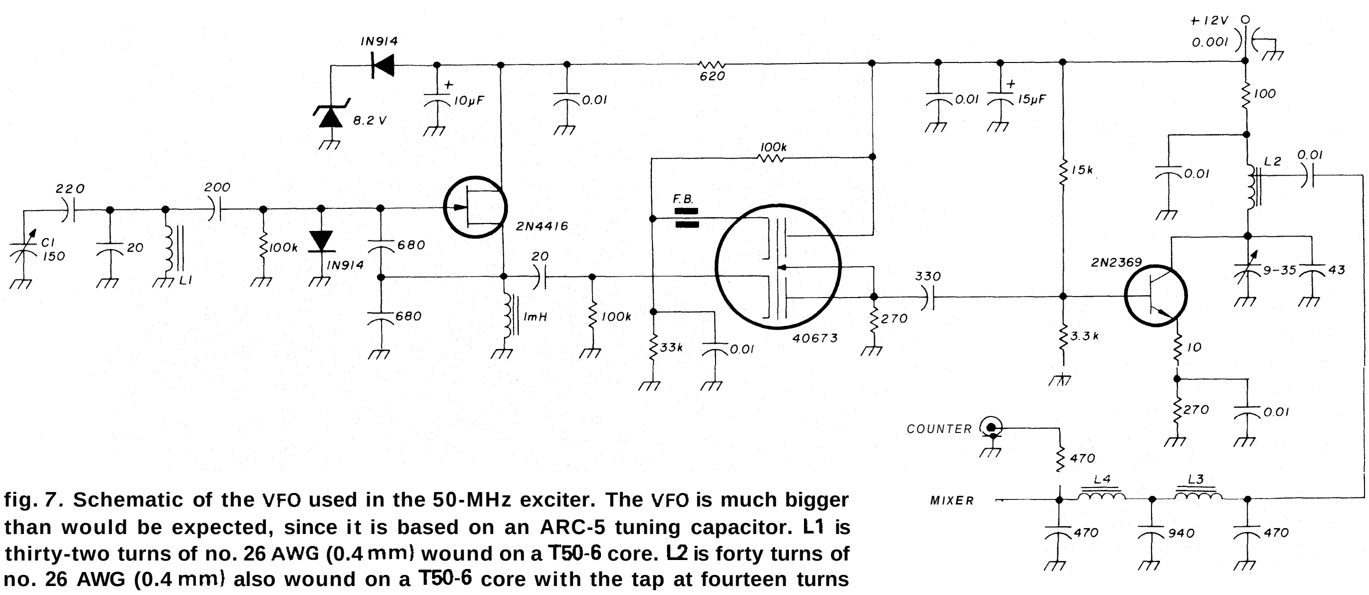


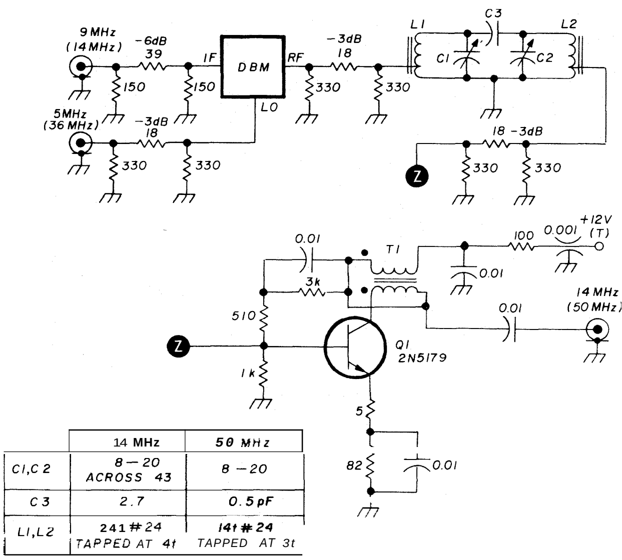
fig. 7. Schematic of the VFO used in the 50-MHz exciter. The VFO is much bigger than would be expected, since it is based on an ARC-5 tuning capacitor. L1 is thirty-two turns of no. 26 AWG (0.4 mm) wound on a T50-6 core. L2 is forty turns of no. 26 AWG (0.4 mm) also wound on a T50-6 core with the tap at fourteen turns from the cold end. The inductors in the half-wave filter, L2 and L3, are both seventeen turns of no. 24 AWG (0.5 mm) wire wound on T50-6 cores.

one used in the second mixer except that the coupling capacitor was increased to reduce losses. It is tuned for maximum signal at 51 MHz.

peak detector

A peak detector is used in place of the usual output detector. This allows the meter to "catch" the voice peaks on sideband so that the operator can tell if he is exceeding the drive-level limit. The output voltage as seen in the schematic, **fig. 11,** from the detector is fed into U1, a unity-gain follower. As the voltage rises, the output of U1 follows and charges up the capacitor through the diode. As the input drops below the voltage across the capacitor, the diode becomes back biased, preserving the voltage on the capacitor. Since the op amps have mosfet inputs, their input resistance is on the order of 10^{12} ohms and the discharge time constant is determined by the resistor across the capacitor and the capacitor's internal leakage. If a Mylar capacitor is used, the time constant can be extended into the minutes range. I used a time constant of 30 seconds because my meter is a heavily damped instrumentation meter and it takes a while to get to full scale. I would recommend a time constant of about 10 seconds (which requires a resistance of 10 megohms). Do not substitute a different op amp for the CA3130s. They were picked because their inputs will operate right to ground and their outputs will go to within millivolts of the supply rails.

In use, the operator tunes up the transmitting chain with the exciter in the CW mode and the peak detector switch set to the "average" position. The meter is set to a convenient reference point with the



	14 MHz	50 MHz
C1, C2	8 – 20 ACROSS 43	8 – 20
C3	2.7	0.5 pF
L1, L2	24t # 24 TAPPED AT 4t	14t # 24 TAPPED AT 3t

fig. 8. Schematic of the mixer and bandpass-filter module. The mixer is an SRA-1. T1 is ten turns of no. 26 AWG (0.4 mm) wire wound on a Ferroxcube 266T125-4C4 core.

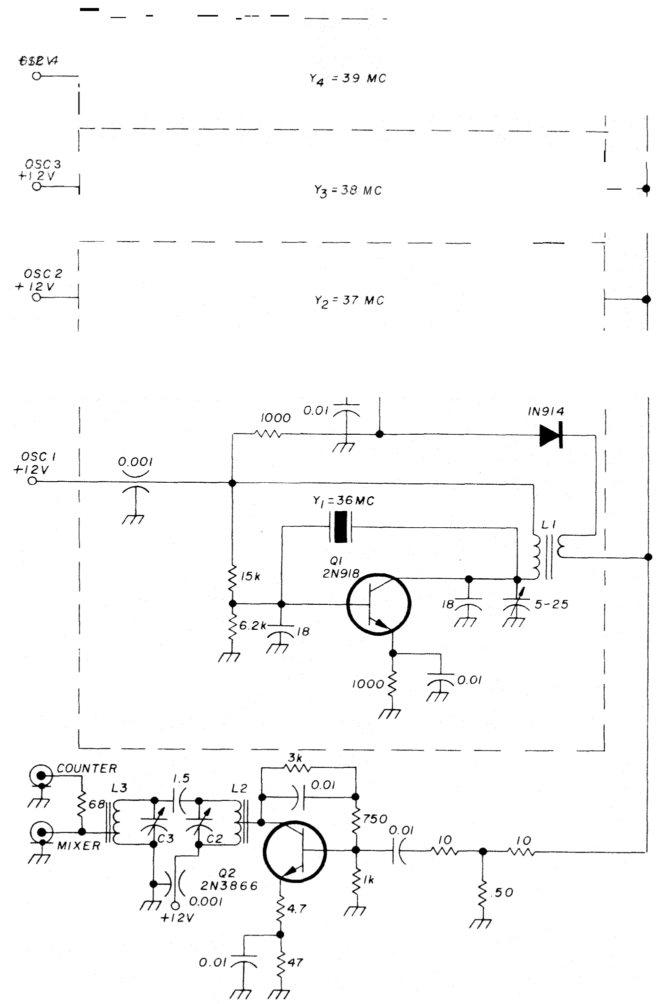


fig. 9. Diagram of the local oscillator module which provides four crystal-controlled frequencies. Each oscillator is tuned by the small trimmer in the collector circuit. L1 is twenty-five turns of no. 24 AWG (0.5 mm) on a T37-10 core. The secondary is two turns wound over the cold end. L2 and L3 are both wound with twenty-five turns of no. 24 AWG (0.5 mm) wound on a T37-10 core. The tap on L2 is eight turns from the cold end, with the tap on L3 at five turns from the cold end. The resistors in the T-type attenuator are selected to provide a 20-mW output.

sensitivity pot and then switched to the "peak" position. The meter will then catch the peaks of the sideband, giving the operator a fairly accurate indication as to whether or not he is exceeding the limit.

PTT control

The exciter is turned on by either the PTT switch on the mike or by a PTT input jack on the back of the chassis (for a foot switch). If either jack shorts to ground, the transmit relay is energized and the supply voltage is applied to the modulator, mixer, and PA modules (see **fig. 12).** I used a relay instead of a solid-state switch for compatibility with other devices in my transmit chain. However, there is certainly no

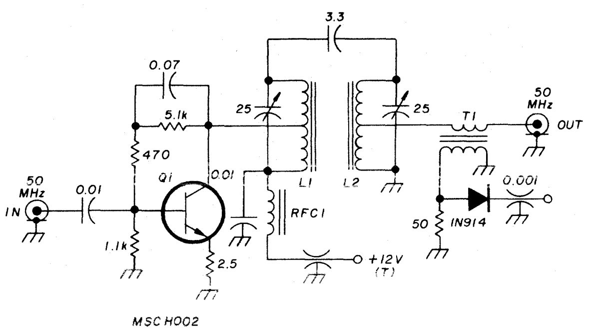


fig. 10. Diagram of the final amplifier transistor and the associated circuitry. L1 and L2 are both fifteen turns of no. 24 AWG (0.4 mm) wire wound on a T50-6 core and tapped three turns from the cold end, T1 is formed by sliding a Ferramic CF101 core over the output lead. The secondary is ten turns of no. 24 AWG (0.4mm) wire.

reason why a power transistor couldn't be used in place of the relay.

counter overview

The digital counter counts each of the oscillators in sequence and then displays their sum. Since the heterodyne scheme for the exciter is all additive, the sum of the oscillators is also the final signal frequency. The counter was broken down into a five functional module as an aid in troubleshooting. All of the modules except the displays and their decoders are mounted under the chassis and all power and signal leads associated with the digital circuitry are well filtered.

A single LM309K is used to regulate the counter. It runs hot and should be well heat sunk. Most of the chips, with the exception of the oscillator and the front-end ICs, could probably be replaced directly with their low-power Schottky (74LS) equivalents without any problems and with a very significant savings in power dissipation. It is one of the things I plan to do in the future.

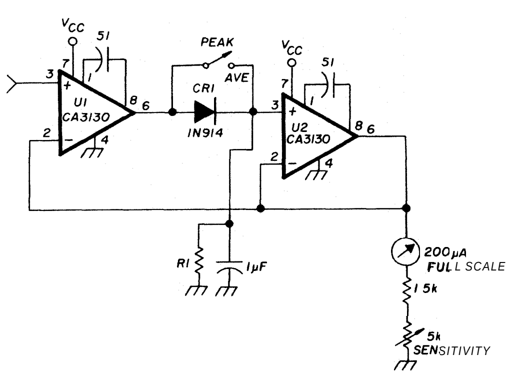


fig. 11. Schematic of the peak detector. C1 should be a Mylar or styrene capacitor. R1 is approximately 1 megohm per second of hold time.

time base

A 1-MHz TTL oscillator is used to drive four decade dividers (see **fig. 13**). The output of either U5 or U6 is selected by the state of TP2. If this point is grounded, the counter will display to the nearest 1 kHz and the update rate will be twenty-five times a second. If this point is high, then the readout will be to the nearest 100 Hz and the update rate will be 2.5 times a second.

The exact frequency of the time-base crystal is set by a trimmer, C1. I had the advantage of owning a precision counter which I used to count the time base frequency at TP1. An alternative approach would be to loosely couple TP1 to your receiver and

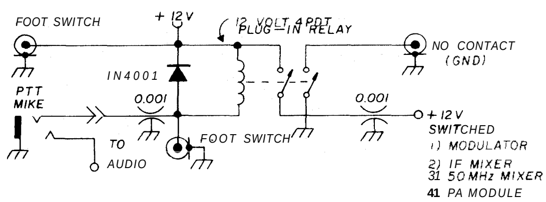


fig. 12. Transmitter control is via a double-pole, double-throw relay which controls 12 volts throughout the exciter.

then beat one of the harmonics to WWV. The crystal I use is one that came from a digital communications board, but it works well. If I were starting from scratch, I would add a divide-by-four stage and use a good quality 4-MHz crystal. For an exciter though, the counter is accurate enough.

control logic

The control logic takes the reference signal from the time-base generator and creates all the control signals for the entire counter. The circuit is more complex than most for two reasons. The first is that the counter must gate in three separate signals and add them together. The second is that I have had some bad experiences using differentiators as sequential pulse generators, as is sometimes done for the load and reset pulses. This type of circuit can often lead to noise problems and I preferred to avoid them. With the illustrated control circuit (see **fig. 14)** there is a finite "quiet" time between all of the gating pulses, the load, and the reset pulses (see **fig. 15**).

The time-base input is inverted by U5 and also divided by two in U2A. Clock 2 is fed to U1, a modulo-six divider. When the fifth state is decoded by U6, pin 9 of U2B goes high, resetting the divider. The next clock 2 pulse causes pin 9 to return low. The result is that pin 8 of U2 creates a gating pulse equal to clock 1 divided by ten with a "quiet" time equal to one clock 2 period.

U3 is wired as a modulo-4 divider. Each of the states of U3 is used for one of the counter functions shown below:

state	U3:3	U3:5	action
1	high	low	G1 high during GE
2	low	high	G2 high during GE
3	high	high	G3 high during GE
0	low	low	load high during (GE-not and CK1-not and CK2-not) reset high during (GE-not and CK1-not and CK2)

U3 is updated to its next state by the clock pulse from U6B. By using a state counter and the four clocking phases, race conditions were avoided and the counter worked the first time it was turned on (wiring errors excepted!).

prescaler

The counter front-end board contains the count enable gates and a prescaler (fig. 16). While it is a simple board, caution should be taken in its layout

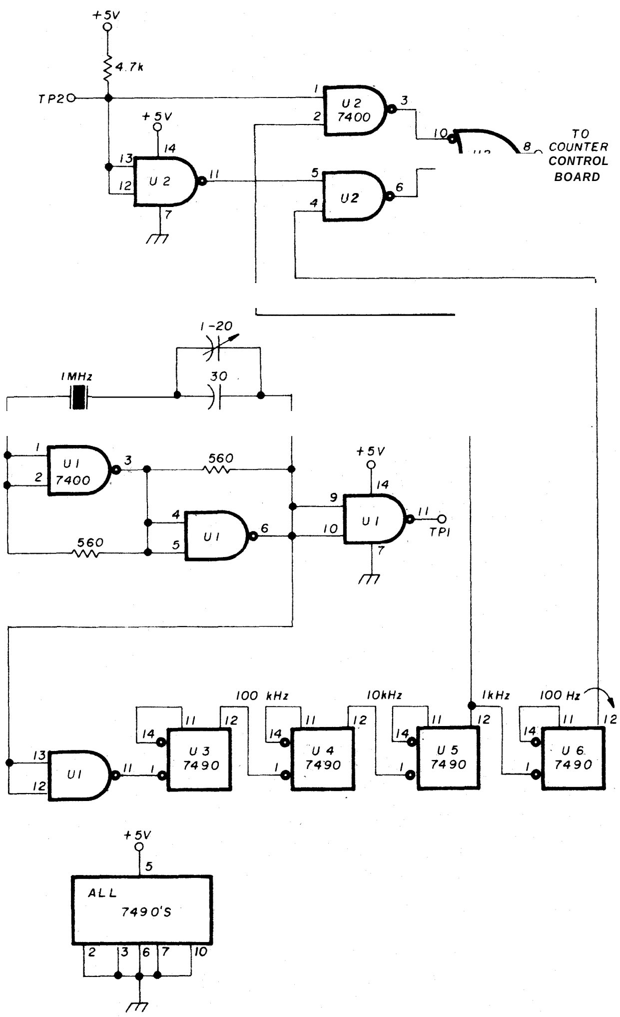


fig. 13. Schematic of the time base oscillator and associated countdown circuitry.

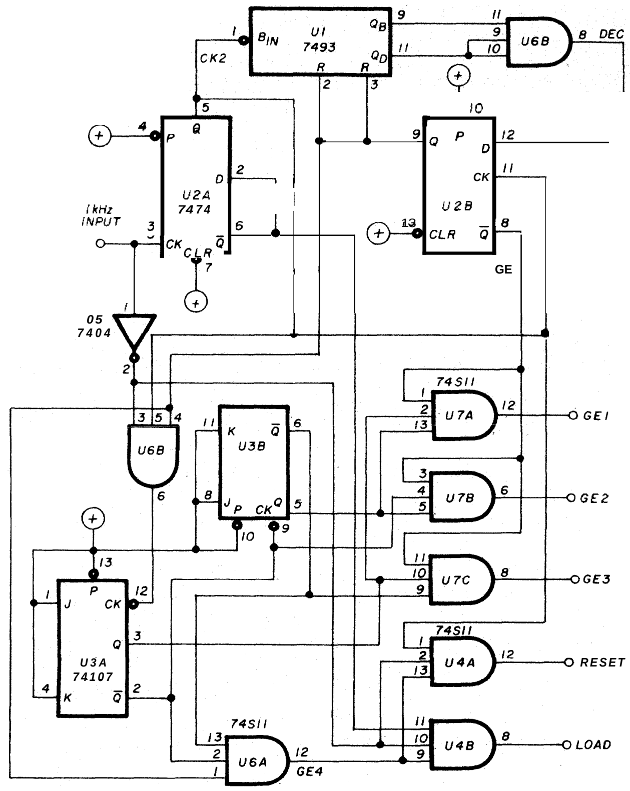


fig. 14. The gating circuitry which generates the pulses and gates for the digital readout. The "plus" sign denotes a common 4.7k pull-up.

and construction. Don't forget, it will have a 39-MHz signal applied to one of the inputs. Each of the oscillators is applied to a section of U1 for squaring up and buffering. The logic signals are then gated and ORed before being prescaled. In order to use the 50-MHz clocking rate of the first stage, U3 is wired divide-by-two, then five. Although pin 12 U3 is no longer symmetrical, it still has a period equal to ten times the input period. That's all that really matters since the 7490s used in the main counter are edge-triggered.

The signals from the oscillators are buffered by sections of U1. These buffers are biased so that their outputs just start to go low. This minimizes the amount of rf voltage needed to toggle them. The analog inputs are decoupled from the digital circuitry by the baluns in order to limit the chance of ground loops reflecting digital noise onto the transmitted signal.

There is some digital noise, however. I believe it is the result of not using high-impedance buffers to drive the timing gates. The noise is way down though; it couldn't be seen on the spectrum analyzer used to measure the transmitter's performance.

main counter

The main counter board, as seen in fig. 17, uses a

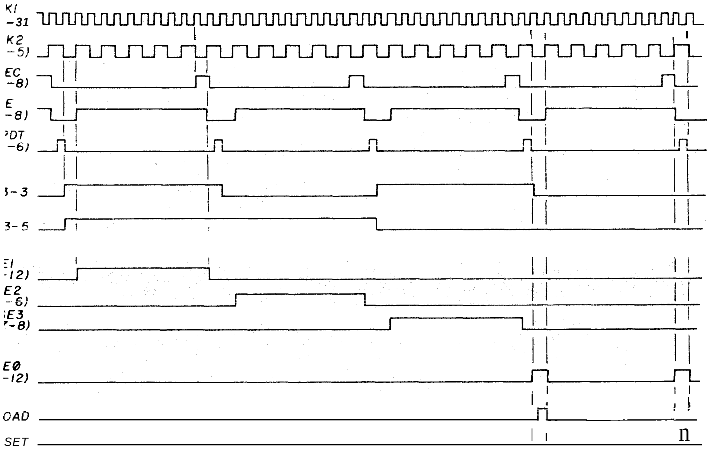


fig. 15. Control logic timing diagram for the circuit shown in fig. 14.

six-digit, latched counter. The 7490s were used because they are cheap and the maximum input frequency of 4 MHz is well below their maximum clock rate. The outputs of the counters are latched to maintain a steady display during each count cycle. Hex latches were used in order to reduce parts count, but any of the 4-bit latches available would do the job with the appropriate wiring changes.

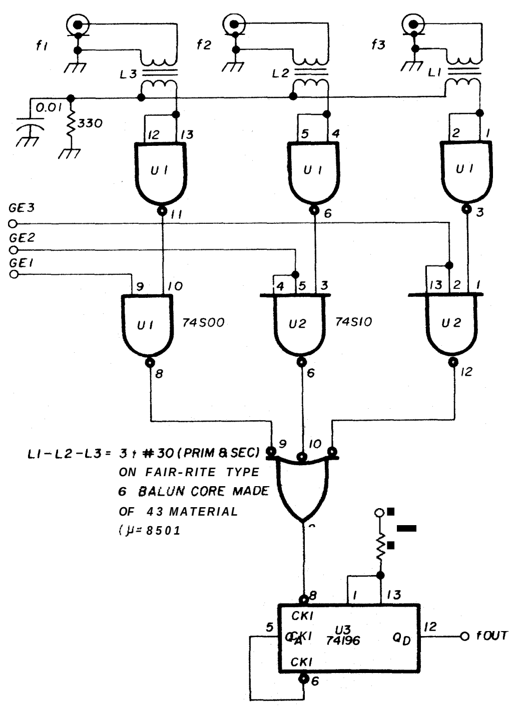


fig. 16. Diagram of the **preamp/scaler** circuit. Individual sections of the **74S00** and **74S10** provide the buffering and gating of each signal. The 74196 acts as a prescaler, dividing each frequency by ten prior to the main counter. L1, L2, and L3 are wound of Fair-Rite type six balun core. The primary and secondary are both three turns of no. 30 AWG (0.25 mm) wire.

display module

This module consists of two boards, one for the displays and one for the decoders and the current limiting resistors (see **fig. 18**). Common-anode displays were used; any of the MAN-1 compatible types should work.

The ripple-blanking feature was used to disable the leading zero when the counter is gated for a 1-kHz

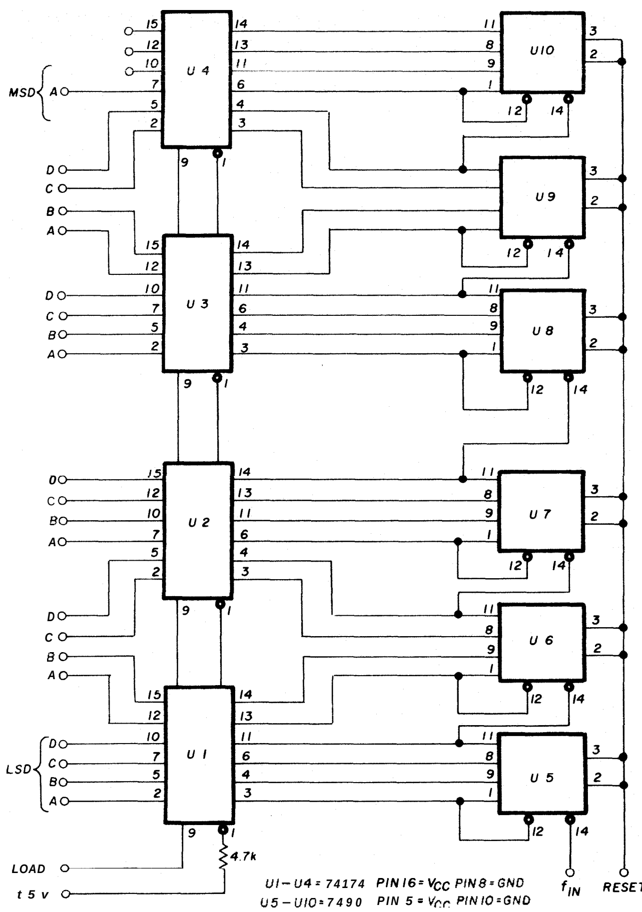


fig. 17. Schematic of the main counter and data latches. Each latch **IC** provides six latches, reducing the number of **ICs** by two.

LSD. I set up the display with two decimals to simulate the commas between the MHz, kHz, and Hz digits. This was done for convenience only, since I always have the display reading out to 100 Hz.

module construction techniques

All the analog modules, except the VFO, are 10 x 6.4 x 2 cm (4 x 2.5 x 0.75 inch); the VFO is 11.5 x **8.9** x 7.6 cm (4.5 x 3.5 x 3 inches). See **fig.** 19. They are all constructed from double-clad circuit board which is soldered into a box. A 4-40 (M3) brass screw (with

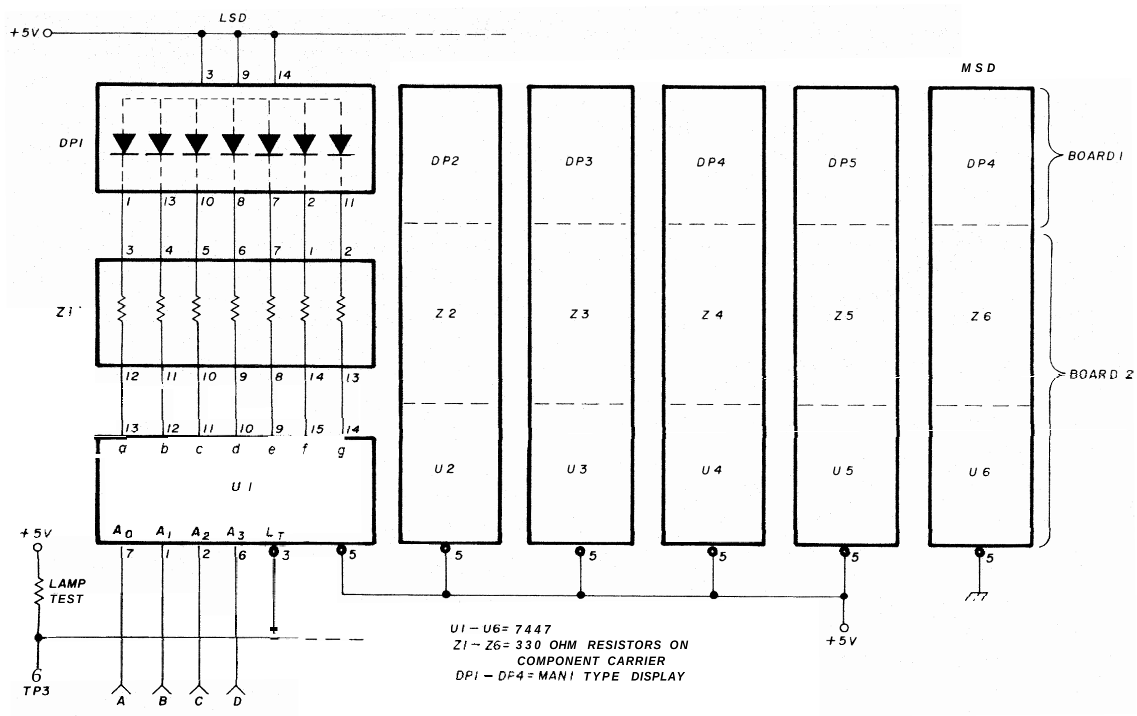
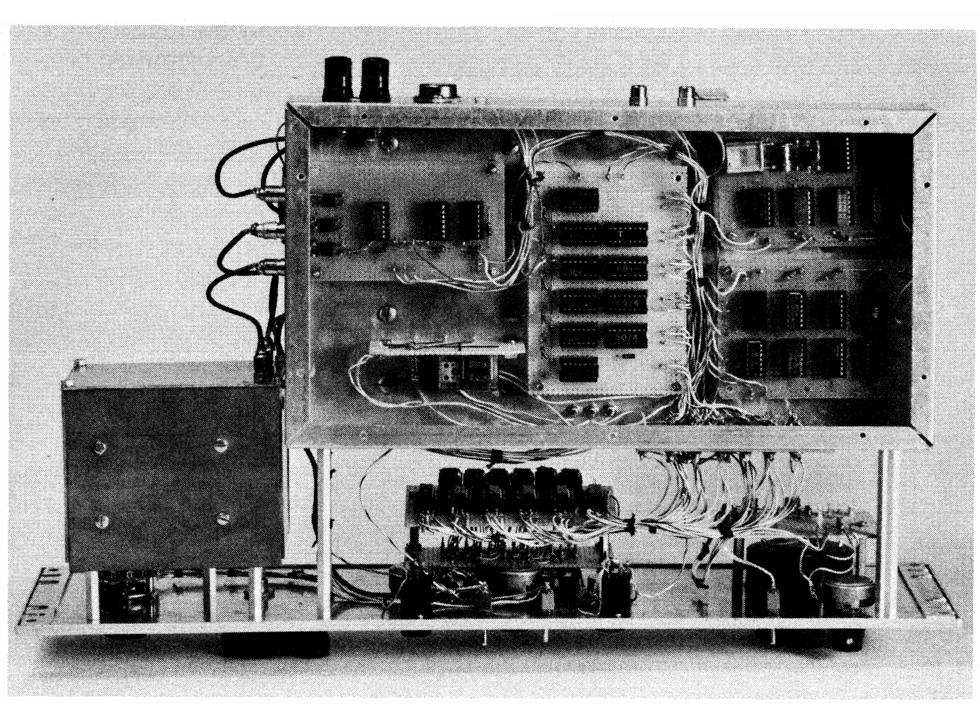


fig. 18. The display module contains the LEDs, limiting resistors. and data decoders. The MAN-1 type LEDs are mounted on a separate board behind the exciter's front panel. The limiting resistors are mounted on component carriers.

the head removed) is soldered into each corner of the box as a means of attaching the cover. For mounting, 8-32 (M4) brass nuts are soldered on the inside (bottom) of the box and the box is then screwed onto the main chassis with 8-32 (M4) screws. I made all the boxes by hand using a hacksaw, a nibbling tool, and a file. It takes about an hour to make up a box. I finally had a friend punch out a bunch of covers and sides for me, which saved a lot of time and made the whole project much more enjoyable.

I use a "standard" bread-boarding chassis which has a row of holes that allows the modules to be



Bottom view of the 6-meter exciter showing construction techniques for the digital counter circuitry.

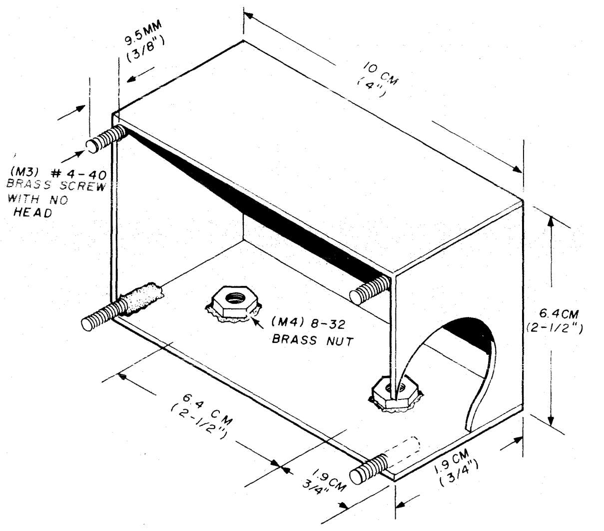


fig. 19. Construction details of the analog modules. Each box is constructed from pieces of circuit board soldered into the desired shape and size.

mounted on 3.8-cm (1.5-inch) centers. The front of the chassis has a piece of angle with 4-40 (M3) clearance holes every 1.3 cm (0.5 inch). Whenever a pot or switch is needed, a small piece of printed circuit board is cut out and attached to the angle to form a "panel." This may seem like a lot of extra work, but it does have the advantage of keeping all the boxes "glued" together. When everything works, the modules are transferred to the final chassis and the breadboarding chassis is ready for the next project.

The modules are interconnected with hook-up wire or RG-174 coax, depending on the type of signal. A couple of ferrite beads were slid over each of the analog power leads. I used phono connectors on

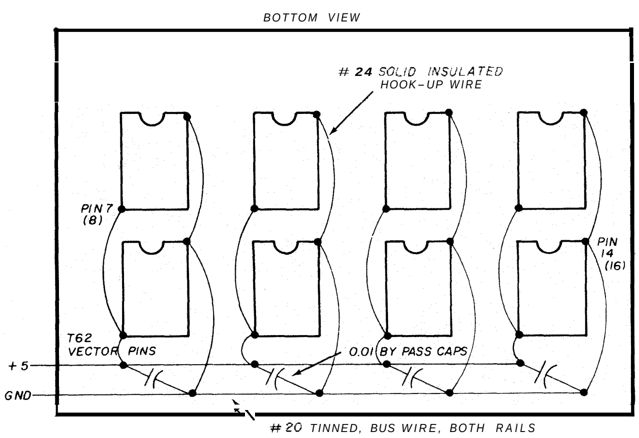


fig. 20. The ICs in the counter circuitry use wire wraps for the point-to-point, with the supply lines bypassed by a 0.01-μF capacitor at each IC. Vector pins are used for the capacitor connection points.

all but the output of the final amplifier module in the interest of cheapness and availability. I used a BNC as the final output connector for compatibility with my switching relays. I believe the phono connectors are fine for low-level work up to about 144 MHz. Beyond that, I prefer to use BNC connectors.

The exciter is built on a 33 x 17.8 x 5 cm (13 x 7 x 2 inch) chassis which is mounted to a 48.3-cm (19-inch) rack panel on 10-cm (4-inch) standoffs. This was done to allow plenty of room for controls. Most of the mechanical work was done with commonly available hand tools. The exception was the red display filter which was machined so that it would look nice when mounted on the chassis.

digital construction techniques

The digital modules were built on perforated G-10 fiberglass boards (fig. 20). Power and ground rails are wired with no. 20 AWG (0.8-mm) wire on the bottom of each board. Individual IC sockets are wired into the rails with no. 24 AWG (0.5-mm) solid hookup wire. All signal wiring was wire wrapped.

The power rail is bypassed to the ground rail at each row of sockets with a 0.01-μF ceramic capacitor. The power and ground rails of all the boards are brought back to a common power and ground point. Interboard wiring is done with hookup wire. The outputs from the counter latches were connected to the decoders via 0.001-μF feedthrough capacitors. Ferrite beads were also used on the latch side of the caps.

The story not usually told in most construction articles is the grief encountered in building the equipment. I have had my share with this exciter. There was a growing pile of copper clad boxes as I tried out different and not so successful ideas. The modular approach saved me a lot of time by allowing the development and comparison of different circuits which performed the same function.

acknowledgments

I would like to acknowlege the helpful comments of Joe Reisert, W1JR, and Hank Cross, W1OOP. Many of their suggestions were incorporated into the final design of the exciter. Thanks also to Javed Bukhari, AP2JB/W1, for manuscript critiques.

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2. Roy Hejhal, K7QWR, "An Integrated-Circuit Balanced Modulator," *ham radio,* September, 1970, page 12.
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ham radio

compact loop antenna

for 80 and 40 meter DX

Quad loops make good low-band DX antennas for those with restricted space — a look at several loop configurations and their advantages

No question about it: the challenge of the five-band DXCC award has created great interest in low-band antennas and has inspired experiments with a wide variety of vertical and horizontal antennas including dipoles, monopoles, delta loops, quad loops, and slopers. Antennas with good performance on 80 meters are typically too large for a city lot. For example, vertical monopoles require extensive ground systems, and horizontal dipoles should be at optimum height. For low-angle radiation, optimum height is on the order of a half wavelength. On 80 meters this is clearly out of the question for most Amateurs.

My home is on a suburban lot where high towers and an effective ground system are impossible. These constraints led to the development of a compact, horizontally polarized loop antenna which has demonstrated competitive performance on both 80 and 40 meters.

loop antenna close to ground

Fortunately, the quad loop offers promise of reasonably good DX performance at low height. Experience has shown that the performance of a Yagi and a quad are roughly the same with respect to height.* There is some evidence, however, that at very low heights the quad may outperform the Yagi.[2] The reason may be that loops contain the electromagnetic field better than dipoles.[3] For whatever reason, the closed loop configuration is relatively immune to nearby objects.[4] The loop configuration also may be relatively immune from some of the effects of nearby ground. For these reasons, the quad loop appears to be a good choice for low band antennas if used with supporting structures within the means of most Amateurs.

L.V. Mayhead, G3AQC, described the performance of loop antennas at low heights in an excellent article, "Loop Aerials Close to Ground," in the RSGB journal, *Radio Communications*.[5] In his article, Mayhead described experiments with a variety of full-wavelength loop antennas close to ground. He modeled the antennas at 470 MHz for radiation-pattern measurements, then constructed two of the best

*Wayne Overbeck, N6NB, in his article "Quads vs Yagis Revisited," (reference 1), states on page 17: "The result was conclusive: the relative performance of the quad and Yagi was exactly the same at both 7.6 meters (25 feet) and 21.3 meters (70 feet)! This was true on zero-angle line-of-sight signals, high-angle 'short skip,' and on long-haul DX signals." *There was no height at which the quad's relationship to the Yagi changed.* (Italics mine). Editor.

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configurations for 75 meters and compared them with a dipole at comparable height. Serious low-band DXers should read this article. It offers further evidence that loops make good low-band DX antennas when height is restricted.

evolution of a two-band loop

An 80-meter quad loop is shown in **fig. 1.** The total distance around the loop is a full wavelength. The feedpoint is at the bottom. When used on 80 meters, the current maxima are at the center of the top and bottom spans of the loop (solid circles!. Maximum radiation occurs from these two in-phase current maxima. The radiation is horizontally polarized.

When the loop in **fig. 1** is operated on 40 meters, four current maxima appear (open circles). The antenna is resonant on the second harmonic; however, the 40-meter current maxima tend to cancel. The radiation pattern deteriorates into many vertically and horizontally polarized lobes. For this reason, two-wavelength closed loops perform poorly, so quad loops are rarely used on the second harmonic.

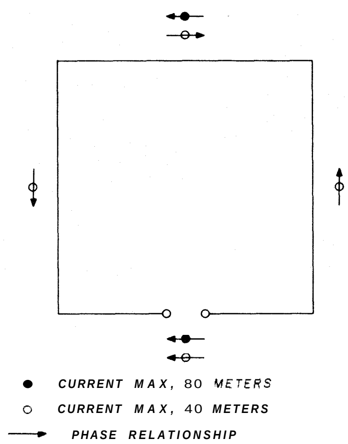


fig. 1. Full-sized 80-meter quad loop showing current maxima for 80 meters (solid circles) and current maxima for 40 meters (open circles). Arrows show relative phase.

Now consider the loop of **fig. 1** distorted into the shape shown in **fig.** 2. The lower left-hand corner folds upward to the right, and the lower right-hand corner folds upward to the left. This forms a half-size loop plus a length of open-wire transmission line. The total length of wire in the loop and the open wire line has not changed. The loop, now reduced in size, presents a capacitive reactance to the open-wire line, which is cancelled by the inductive reactance of the short length of line. Thus the loop and line are still resonant on both 80 and 40 meters.

Fig. 2 shows the positions and relative phase of the current maxima on both bands. Now distort the loop into the familiar delta configurations shown in **fig.** 3. One 80-meter current maximum is still at the top center of the antenna, and on 40 meters all three current maxima add in phase. Radio-frequency current flowing in opposite directions on the open-wire line cancels radiation from the current maximum at the feedpoint on both 80 and 40 meters. For these reasons, a compact loop antenna of the general shape shown in **figs.** 2 or **3** will be a good radiator on two adjacent harmonically related bands.

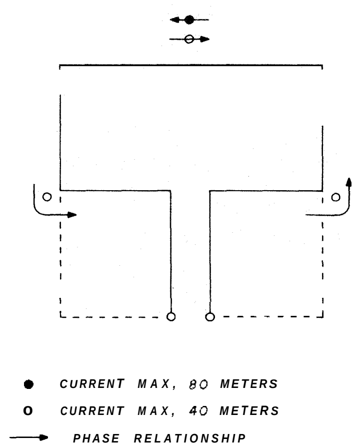


fig. 2. Lower corners of the full-sized loop folded toward the center, forming a 112-area loop plus an open-wire line.

radiation resistance

The area of the loop shown in **fig.** 2 is half the area of a full wavelength 80-meter quad loop. It is twice the area of a 40-meter quad loop. For this reason, on 40 meters, the radiation resistance is high, the efficiency is high, and the bandwidth is broad. The measured impedance at the open-wire line terminals is about 120 ohms. On 80 meters the loop is only half the area of a full-wavelength 80-meter quad loop. Therefore, the radiation resistance is lower and the bandwidth is narrower. On 80 meters the impedance at the open-wire line terminals is about 50 ohms, thus presenting a good match for standard 50-ohm coaxial cable.

Neither the shape nor size of the loop, nor the length of the open wire line, is of particular impor-

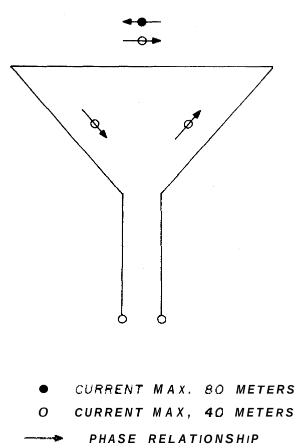


fig. 3. The full-sized loop further distorted into the familiar delta-loop configuration. The 80-meter current maximum is at the top. The 40-meter current maxima add in phase.

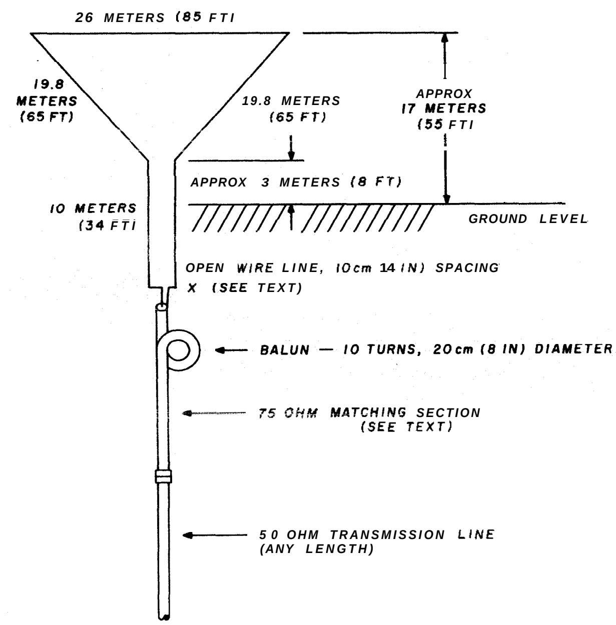


fig. 4. 80/40 meter loop in the delta-loop configuration. Loop can be almost any reasonably open shape. Top should be at least 12 meters (40 feet) high.

tance as long as the loop is reasonably open and the line is between 6 meters (20 feet) and 10.7 meters (35 feet) long. Only the total length of wire, including the loop and the open-wire line, is critical. The total length is best determined experimentally after the antenna is in place; you can resonate the antenna from ground level. Once the antenna is installed, it's unnecessary to take it down again for adjustment.

I built two of these antennas, one at home and the other while on vacation at Lake Tahoe, California. Dick, W3GNQ, also built one at his home in Maryland. All three antennas differ in configuration. The loop antenna at my home is a compromise shape. The top span slopes from an 18-meter (60 foot) tree down to about 12 meters (40 feet). Wires from the

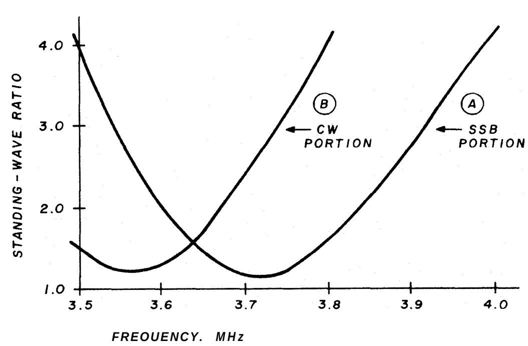


fig. 5. 80/75 meter SWR plots of the delta loop shown in fig. 4. Curve B favoring the CW part of the band was obtained with an extra length of open wire line. See fig. 7.

ends of the top section slant down to the patio overhang and follow the overhang to the feed point. The lengths of wire are in different planes. Last summer, with the beautiful tall pine trees at Lake Tahoe as supports, I built the symmetrical delta loop shown in fig. 4. Figs. 5 and 6 show the SWR plots.

bandwidth

In general loop antennas have more bandwidth than thin dipoles or monopoles. The antenna described here is no exception. The loop is oversize on 40 meters (1½ wavelengths), so its bandwidth is extremely broad. The SWR doesn't exceed 1.3 over

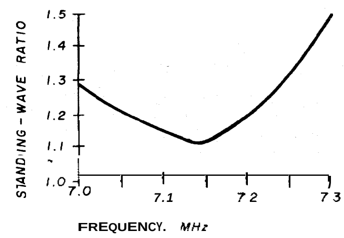


fig. 6. 40-meter SWR plot for delta loop. SWR is less than 1.5 across the band.

most of the band and doesn't exceed 1.5, even at 7.3 MHz (see fig. 6). On 80 meters the loop is only ¾ wavelength. Even with this reduced size, the bandwidth is still satisfactory. The 2.0 SWR bandwidth is 250 kHz, substantially more than the 150 kHz typical of an 80-meter dipole or inverted V. Installing the antenna in a rectangular configuration rather than the delta loop configuration further improves bandwidth.

dual-band feed

A 75-ohm, quarter-wave transformer matches the 50-ohm coaxial transmission line to the 120-ohm input impedance of the loop on 40 meters. On 80 meters, the transformer is only one-eighth wavelength long and doesn't materially affect the 80-meter match. Thus the antenna operates on both bands with a common transmission line.

The two-band feed system (fig. 4) consists of any length of 50-ohm line, a quarter wavelength (on 40 meters) of 75-ohm line, and the open-wire line. A simple balun is included in the 75-ohm matching section.

building the 80140-meter loop

To build the 80140 meter loop, start with a total length of 87 meters (285 feet) of copper wire. The flat top should be roughly 26 meters (85 feet) long. Make the open-wire transmission line between 6 meters (20 feet) and 11 meters (35 feet) long. The wires should be separated by about 10 cm (4 inches). Spreaders can be made by cutting 1.6 mm (1116 inch) fiberglass into 1.3 x 11.5 cm (½ x 4% inch) lengths drilled at each end. Wire these strips onto the open-wire line at roughly one-meter (3-foot) intervals. make the 50-ohm coaxial transmission line from any length of RG-81U or RG-58/U line, depending on the power level. On 80 and 40 meters RG-58/U is adequate up to several hundred watts input. For a full kilowatt use RG-8/U.

Make the matching transformer with RG-11/U or RG-59/U coax cut to 6.95 meters (22 feet 9 inches), the correct length for solid polyethylene insulation. If you use foam polyethylene insulated coax, the correct length is 8.41 meters (27 feet 7 inches).

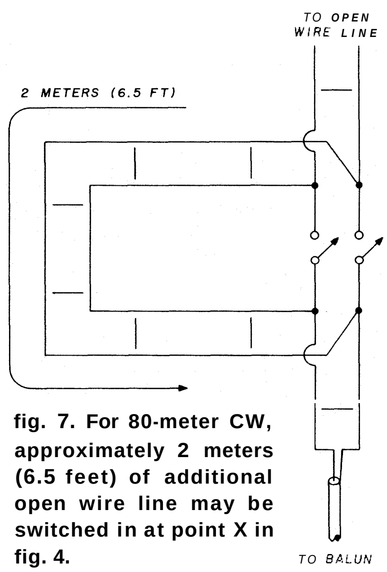


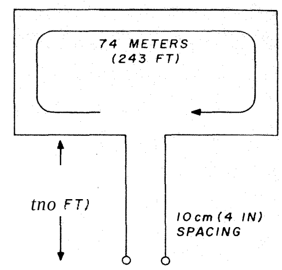
fig. 7. For 80-meter CW, approximately 2 meters (6.5 feet) of additional open wire line may be switched in at point X in fig. 4.

A simple and adequate balun can be added at the loop end of the 75-ohm matching transformer by wrapping ten turns of the coax into a coil. The coil diameter should be ten to fifteen times the outside diameter of the cable used. Place the coil balun as shown in **fig. 4** and use tape to hold it in shape. Connect one end of the quarter-wave transformer to the open-wire transmission line with alligator clips.

Using the alligator clips, adjust the open-wire transmission line length until the **SWR** curve is well centered for 75-meter **SSB.** See curve **A, fig. 5.** Now plot an SWR curve over the 40-meter band. You'll find it broad and well centered (see **fig. 6**). When you're satisfied with the 75- and 40-meter **SWR,** remove the excess wire and solder the 75-ohm coax to the open wire line.

From curve **A** of **fig. 5,** notice that the **SWR** climbs excessively in the CW portion of the 80-meter band. Use the extra length of line and the switch shown in **fig. 7** for 80-meter CW operation and adjust the extra length of line for optimum **SWR** over the CW part of the 80-meter band. (See **fig. 5,** curve **A.**) Add the switch and line close to the coax at X in **fig. 4.** Use a relay or, because point X is near ground level, a switch mounted in a convenient location.

fig. 8. Rectangular loop at W3GNQ. Size and shape of the loop isn't critical. The total length of wire including the loop and the open-wire line must be full-wave resonant on 80 meters. Use two-band feed system shown in fig. 4.



W3GNQ loop

W3GNQ decided to erect his antenna in the general shape of **fig.** 2; his antenna is shown in **fig. 8.** The rectangular shape and short open-wire line result in the SWR curves of **figs. 9** and **10.** The bandwidth of Dick's antenna is broad enough so he can operate on 80, 75, and 40 meters with no switching.

bandwidth and terminal impedance

To gain an understanding of how the input impedance and bandwidth vary with loop size and shape, I made scale models of various loop antennas. All loops were 406 cm (160 inches) in total length including open-wire line. All were resonant between 75 and 90 MHz. I measured terminal impedance and relative bandwidth on a variety of configurations with a Hewlett-Packard RF Vector Impedance Bridge. The curves of **figs. 11** through **14** show the data. **Fig. 11** shows how the terminal impedance and relative bandwidth change with the ratio of square loop size to open-wire line length. When the ratio of the total length of the loop to the length of one side is four, the open-wire line length is zero, and the loop is one wavelength in circumference. The input impedance

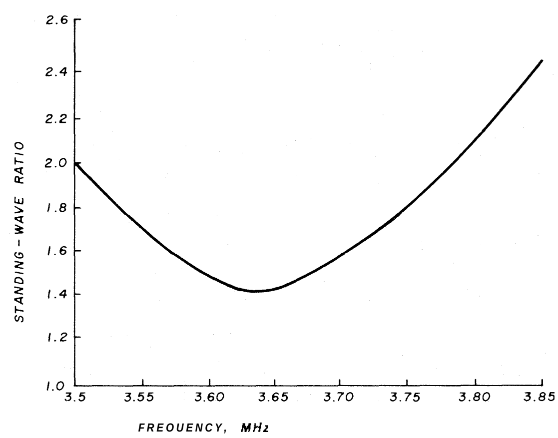


fig. 9. 80-meter SWR plot of quad loop at W3GNQ.

is 145 ohms. As the loop decreases in size, the open-wire line becomes longer and the terminal impedance drops. It reaches 50 ohms at L_t/L_s (total length to side length ratio) between five and six.

I measured the relative bandwidth of the models by recording the impedance-bridge frequency above and below resonance at the +10 and −10 degree points as read on the phase-angle meter. Of course, bandwidth improves as the loop increases in size.

Data taken on a delta loop model is shown in **fig. 12.** Here the open-wire line length was held constant while changing the ratio of top span length to total length. When the terminal impedance is 50 ohms, the bandwidth is the same as that of the square loop when proportioned for the same input impedance **(fig. 11).** The square configuration, however, is capable of greater bandwidth. Furthermore, you can infer from the two sets of curves that a rectangular antenna **(fig. 8)** will have better bandwidth than either a square or triangular loop. The bandwidth

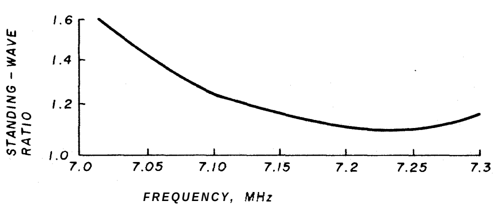


fig. 10. 40-meter SWR plot of the loop shown in fig. 8.

data is only relative and not a measure of the useful bandwidth of the antenna; it's useful only for comparing the scale models.

use on a metal tower

The diamond shape of **fig. 13** is interesting. Because the loop is symmetrical about the vertical axis and is horizontally polarized, it should be a good radiator when supported on a single metal tower. To give some experimental validity to this, during the scale model tests I placed a large metal member near the loop vertical axis, simulating a tower. There was no measurable effect on the terminal impedance, and the resonant frequency changed only one-half per cent. Thus, interaction of a full-scale antenna with a metal tower should be negligible. The inverted delta loop version shown in **fig. 14** also can be used on a metal tower.

diamond vs. square quad loop

On comparing the curves of **figs. 13** and **14,** note the interesting data showing impedance and bandwidth for the diamond quad and the square quad. The full-sized quad loop case is shown on the curves at $L_t/L_s = 4$. The data indicate that a diamond-

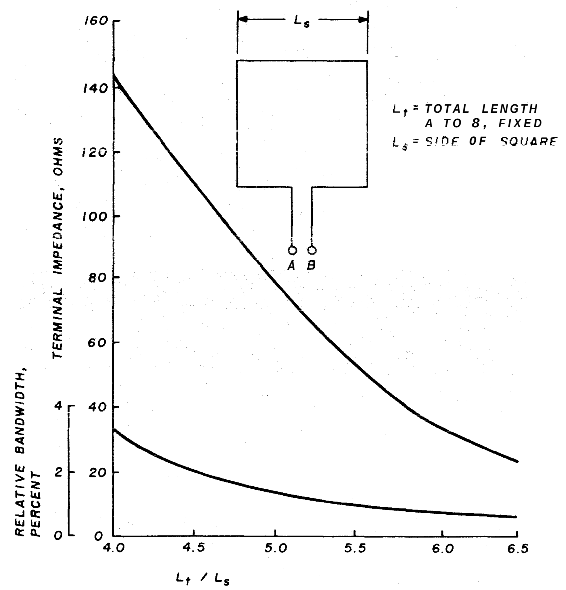


fig. 11. Input terminal impedance as a function of loop side length to total length ratio. Measurements taken on 80-MHz scale model at resonance with an HP Vector Impedance Bridge.

shaped quad is the preferred choice because of higher impedance and greater bandwidth.

Antennas near ground at 3.5 MHz will have a terminal impedance different from that of the scale models because of the effects of nearby ground. I've tried none of the antennas of **figs. 11** through **14** at full scale, so the convenient relationship between 75-

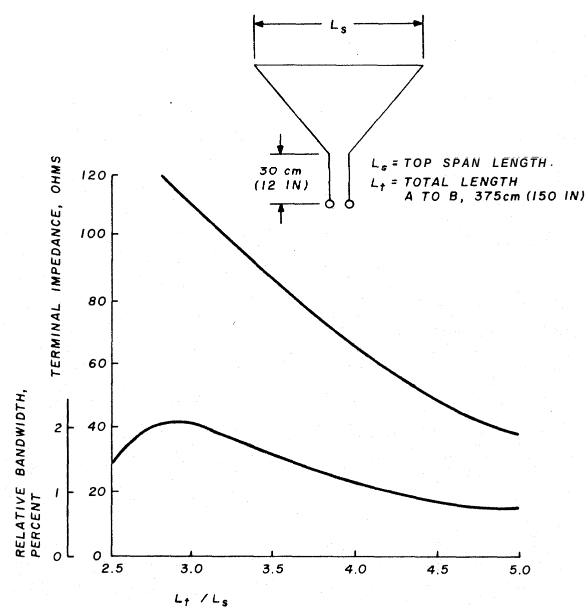


fig. 12. Scale model input terminal impedance as a function of loop top span length for a 375-cm (150-inch) loop including 30-cm (12-inch) open-wire line. Measurements taken at the full-wave resonant frequency near 80 MHz.

and 40-meter resonant length may not hold for these configurations. Therefore, antenna designs based on these curves may require further experimentation.

advantages of the W6TC compact loop antenna

1. The antenna is an excellent radiator on both bands, as theory predicts and as proven in practice.

2. The antenna provides a good match for 50-ohm line on both 80 and 40 meters.

3. The antenna is small compared with other 80 meter antennas. The largest dimension is substantially less than 30 meters (100 feet).

4. The antenna is broadband.

5. The antenna can be supported on relatively low supports and still give reasonably good performance.

6. The shape isn't critical. The antenna can be put in the delta, square, rectangular, or any similar open configuration. It can be installed on a single metal tower in the shape of a diamond or inverted delta.

7. The antenna is horizontally polarized, so a ground system is not required.

8. Only one dimension, the overall length, is critical. This adjustment is made from the ground; it's not necessary to take down the antenna for adjustments.

9. An antenna tuner is not required.

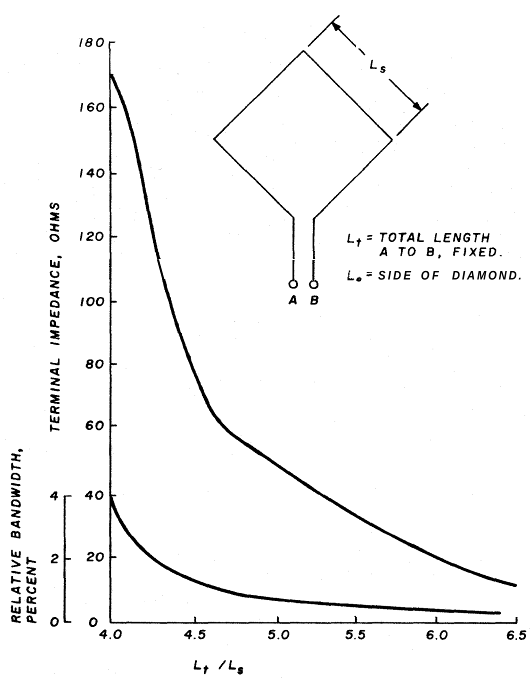


fig. 13. Terminal impedance and relative bandwidth vs diamond side to total wire length ratio. Scale model frequency: 80 MHz. Loop may be mounted on metal tower.

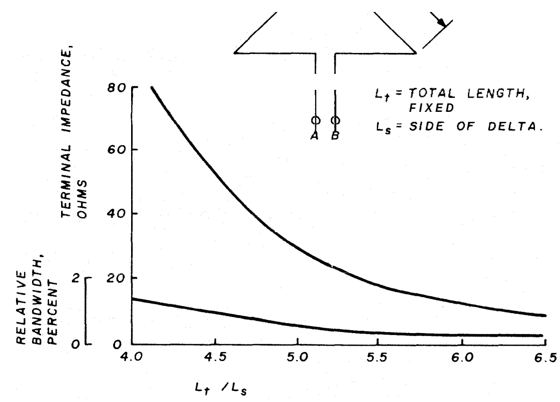


fig. 14. Scale model terminal impedance and relative bandwidth characteristics of inverted delta loop.

10. The center of the long span of the delta-loop version doesn't support a heavy feedline.

conclusion

There's no substitute for full-size antennas supported on tall towers. With low band antennas, "the higher, the better" has always been true. However, if you have room for only one small, low antenna for 80 and 40 meters, this antenna may be your best choice. I hope the examples of full-size antennas and the scale-model data presented here will provide design guidance for your version of the antenna.

Experience with this antenna on the air has been gratifying. While evaluation of low-band antennas is always highly subjective, the W6TC loop appears to be quieter on receive and seems to punch through the pileups better than other low-band antennas tried at W6TC. Working DXCC on 80 meters is difficult from the West Coast. However, after building this antenna, the new countries seemed to come a little easier. Dick, W3GNQ, finds the antenna especially competitive on 75 meters. While direct A-B comparisons were never possible because of space limitations, this antenna, in my opinion, has that extra edge that only on-the-air experience can discern.

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3. H. Leerning, G3LLL, "The Delta Loop Antenna," ***Ham Radio Horizons,*** June, 1979.
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ham radio

the talking clock

An addition to the talking digital frequency readout described in an earlier issue of ham *radio*

It was no time at all after completing the talking frequency readout unit[1] that I recognized the desirability of using its voice for Greenwich Mean Time (or UTC). The investment in the speech synthesis module had already been made and the original readout box was not crowded, so I investigated digital clocks.

clocks

The first clock chip I acquired didn't have BCD output and was meant to drive 7-segment LEDs. Although I located an interesting chip that could have converted the clock data to BCD, I thought its price in unit quantities was too high — above my budget for the whole project. Later I discovered the MM5312 clock chip that sells for $3-$4, which has both BCD and 7-segment data outputs. The MM5312 doesn't read seconds or have all the bells and whistles, but it is a true 24-hour clock on a 24-pin chip. A Radio Shack 5082 miniature 5-digit readout allowed me to get familiar with its operation. (Ignoring the third common cathode makes the middle digit blank.) The visual readout is optional and may be omitted, if desired.

circuit description

The talking clock schematic is shown in fig. 1. I planned to use whatever I could for the original circuit to provide the data sequencing immediately after completion of the previous word, regardless of the word time duration. It would have been possible to obtain both frequency and time in response to each inquiry, because the 74193 can control a sequence of sixteen data positions. However, because of the sentence-shortening features of the frequencv readout, it seemed unnecessariiy complicated to produce a skip past the omitted positions at the end of the sentence. Instead, I chose to provide the clock data in the 0 through 3 (first four) data locations. Stopping after the first four is easiest: merely invert the C address line with a spare gate on the original board and substitute its output for the six- or eight-word stopper.

Two other poles of the frequency/time switch are used to make sure the sentence starts at the 0 address. The switching is ahead of the provisions for deleting the first word on the single-digit MHz bands and ahead of the MHz-delete switch, which delays the start past the first decimal point. Thus in the time mode, all four 74193 preset lines are grounded.

The circuit that causes the voice to say ***point*** in the third and seventh locations (2 and 6) is disabled by switching the fifth data-line input of the voice to ground when in the **time** function; otherwise, the third digit of the time readout would be really strange and come out of the unused vocabulary of the S2A.

switching logic

We must still switch the data itself between the two sources, readout and clock. A 74157 multiplexer must have been designed precisely for this job (fig. 1); it acts as a 4-pole, 2-position switch for digital data in TTL form. When its pin 1 is connected high (to +5 volts), the time data at pins 11, 13, 3, and 6 appears at pins 9, 12, 4, and 7 respectively. When it is low (at ground), the data at pins 10, 14, 2, and 5 from the readout is passed through to the same four output pins. Its power pin, no. 16, must have +5 volts present in either mode. Power may be removed from the 7404, 7400, and 7432 while in the readout mode. Power to the 74193 and other chips on the interface board is required in both modes, since the 74193 and gates are required to step from one digit to the next at its command.

data sequencing

The path is now cleared for clock data to reach the speech module, and four defined sequential slots are reserved for it. As is usually the case, the most inter-

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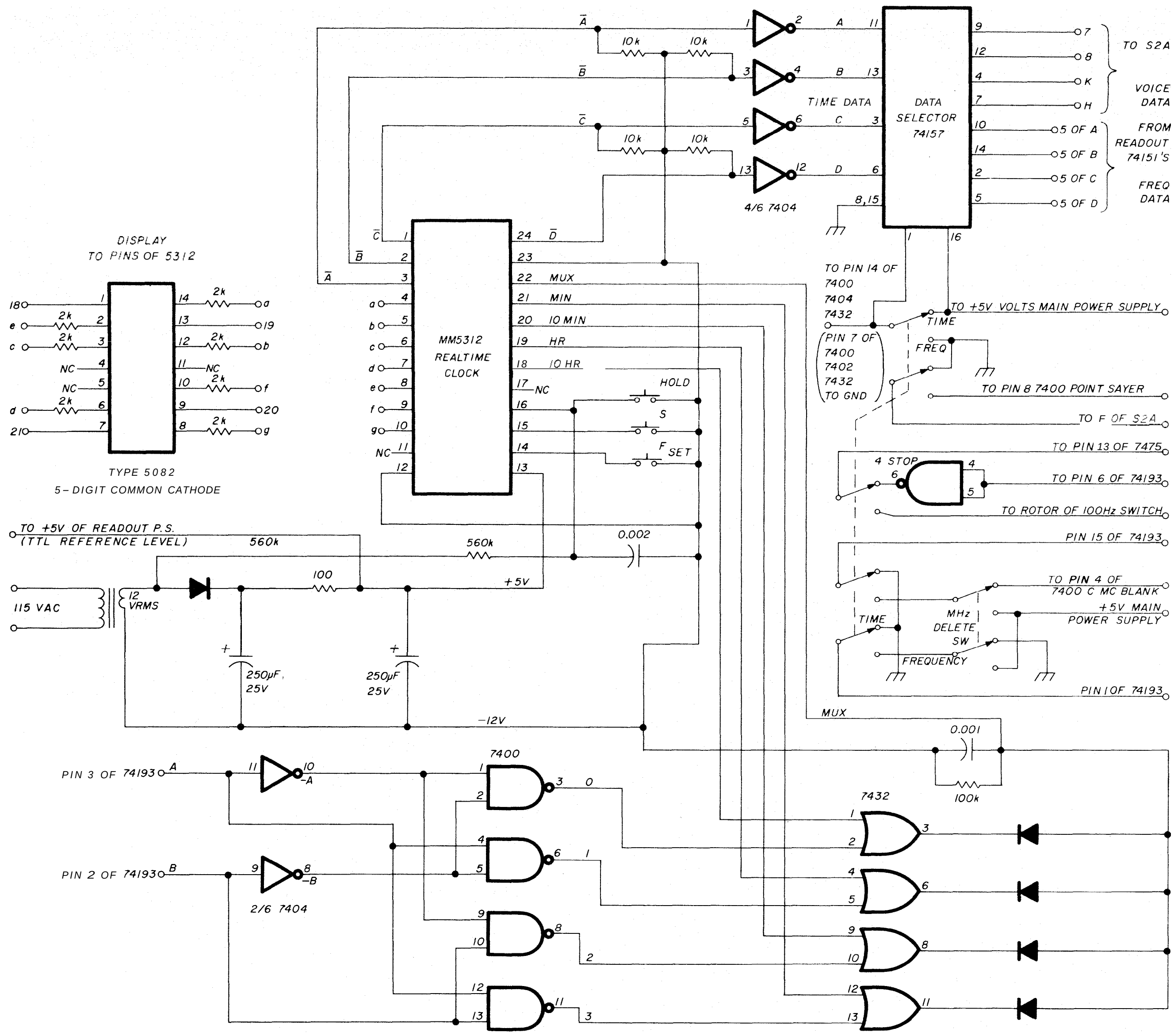

fig. 1. Schematic of the talking clock addition to the talking frequency readout described in an earlier article. It's a useful circuit for the visually handicapped operator.

esting problems come up last. I'd visualized four sets of BCD data available at the same time, which I would have to multiplex, as with the frequency data. But when I examined the 5312 only one set of BCD terminals existed, indicating that the data must somehow be internally multiplexed. Great! That should hold the chip count down. But the problem then was how to control the internal multiplexer. In the normal application, with visual readout, there's no need for this as long as the rate is high enough so that all four LEDs glow simultaneously. But I can't use parallel data for words that have to be said in sequence — and the sequence must be controlled.

The four clock terminals, 18 through 21, go low in sequence so that the appropriate digit LED will glow when the data for that digit is present on the common data lines to all four digits. These four terminals can at least be used to tell which digit's data is appearing on the output BCD data lines at any point in time.

clock mux rate

The clock multiplexing rate is determined by an external RC network connected to pin 22. Looking at this signal on a scope, we see a beautiful repetitive sawtooth waveform. If you use a large capacitor, you'll notice that the four digits light up in sequence, from right to left in backward order. A bit scary. If we ground pin 22, killing the sweep, the readout stops in one of the four positions, more or less at random, but

table 1. Truth table defining multiplexer action.

word sequence to read		first 10 hr	second hours	third 10 min	fourth minutes
address	A		1		1
74193	B			1	1
invert	A	1		1	
7404	– B	1	1		
7400 pin	3	0			
	6		0		
	8			0	
	11				0
5312					
clock pin	18	0			
	19		0		
	20			0	
	21				0
7432					
stop pin	3	0			
	6		0		
	8			0	
	11				0

nevertheless depending on when we do it. A method for controlling the internal multiplexer was evolving. (Maybe I can work out an educated screwdriver to short it out at just the right times.)

A scheme materialized that allowed the multiplexer to scan whenever the clock output data didn't coincide with that requested by the 74193 address position. We want the tens of hours first and the minutes last. When the scanner comes to the correct digit, the multiplexer parks there until it has been pronounced. At the end of the word, the busy line from the voice says, "OK, I'm through," and the 74193 is advanced to the next address. This action releases the scanner lock and it roams (in the wrong direction), looking for the next appropriate digit. It skips over the next two digits presented to it and locks on the third, which is to the right of the last one presented.

Table 1 does a better job of explaining the operation of the circuit. When one of the 7432 positive OR gate outputs goes low, a diode connects it to the scanning waveform and locks it into position. (A positive OR gate has an output low only when both its inputs are low.)

The address is available from the 74193 in BCD form; we need only the A and B lines to describe the first four addresses. A pair of inverters (7404) and four NAND gates (7400) convert this data to decimal form. Since a NAND gate produces a low output only when both inputs are high, we can play the game four ways and come up with combinations minus A minus B for 0, plus **A** minus B for 1, minus A plus B for 2, and plus **A** plus B for 3.

If we apply each output to a separate NOR gate input, then choose the correct line from the four clock terminals, 18 through 21, the 7432 section that will stop the scanner on tens of hours must have the tens of hours terminal (clock pin 18) as one input and the first address, 0, as the other.

It seemed strange to go to all the trouble of sorting things out through two sets of gates only to wind up connecting all four signals to the same place with diodes. But that was pretty much what I started out looking for -- the smart screwdriver. Looking at each 7432 output separately, you should see a square wave that is down one-quarter of the time if you slow it down with a 0.1-μF capacitor across the 0.001-μF.

The visual display stops when the scanner stops, so it's easy to establish what's going on. I used silicon diodes to combine the four signals.

summing up

The gating circuits involving the 7475 latches (reference 1) in the frequency interface board were not modified. I found no error by letting the counter waveform control the start latch, and the reset and address stepping functions were just what I wanted. I added a minimum-design power supply to power the clock chip, which of course is on continuously to maintain time.

Three pushbutton switches permit setting the clock. Upon power interruption, the clock starts running from 0000. When the fast button is pressed, time speeds by at an hour-per-second rate; the slow position advances time at a minute-per-second rate. The hold button permits stopping the clock without resetting.

When listening to WWV, let the clock run to a minute ahead of real time and depress the hold button until the time catches up. On the next minute from WWV, note how many seconds fast you are and hold for a similar period, and you'll find you're right on to the second. The voice repeat switch comes in handy when setting the clock.

Printed circuit boards for the talking readout have been completed and check out. They are available from O.C. Stafford, 427 South Benbow Road, Greensboro, North Carolina 27401. At the time of submission, design of a circuit board for the talking clock addition, incorporating the four chips and diodes shown in the article, has been initiated by O.C. Stafford.

Since the unit now performs two functions rather vital to the visually handicapped Amateur, it's been dubbed the "Second Operator." Let your fingers do the talking.

references

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ham radio

the handi-counter

A high-frequency counter featuring the Intersil 7216-D built into a hand-calculator case

This is a fun construction article. It involves mating Intersil's new ICM-7216-D 10-MHz counter chip into a small hand calculator case. Frequency limit is ± 550 MHz depending on device options.

I purchased one of the first **7216-D** chips. In checking out its many features, I needed an LED display. I had recently purchased two hand calculators in a Poly-Paks penny sale. As I robbed the 9-digit display, the possibility of using the case for a small portable counter evolved into the handi-counter.

design criteria

A small pocket-type counter is designed with sensitivity, frequency range, and accuracy to use for emergency frequency calibration of TR-22 handi-talkies. To meet these requirements, a bare minimum of parts evolved into the circuit of fig. 1 with the following specifications:

Frequency limits: ± 250 MHz

Sensitivity: reads TR-22 at 0.9 meter (3 feet)

Accuracy: ± 100 Hz or ± 10 Hz

Voltage and current: 5 volts at **137** mA

Readout: seven digits at 1.0-second gating;
eight digits at 10.0-second gating

The handi-counter frequency response may be extended to around 600 MHz as shown in fig. **2**,* in which a **74196** chip (U2) is substituted for the **74LS90** (U2 in fig, **1**). The specifications for the circuit in fig. **2** are as follows.

Frequency limits: ± 550 MHz

Sensitivity: 100 mV at 450 MHz
50 mV at 146 MHz

Accuracy: same as the circuit of fig. 1

Voltage and current: 5 volts at 185 mA

Readout: same as the circuit of fig. 1

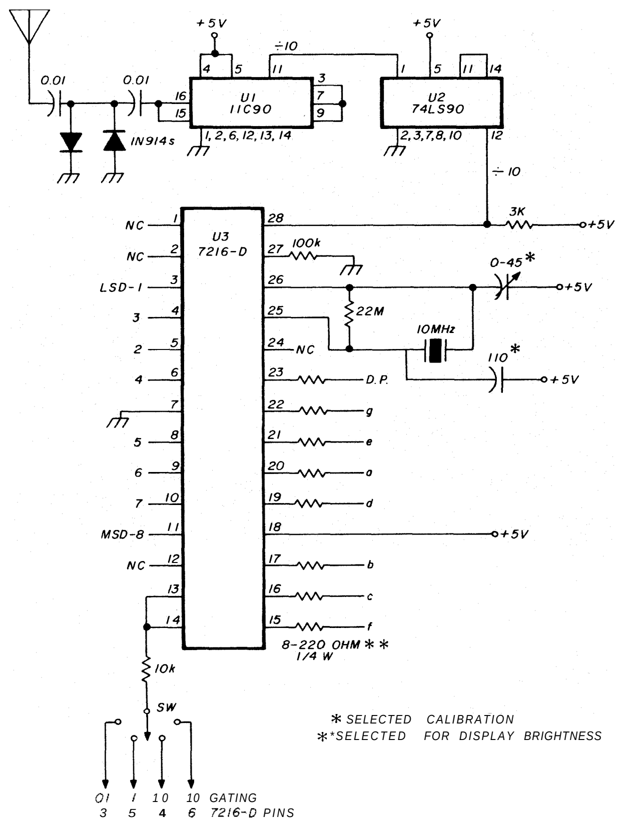


fig. 1. 250-MHz counter using the Intersil ICM-7216-D counter chip.

*A PC-board layout for the 600-MHz version is available from the author. Please send a self-addressed, stamped envelope.

By R. M. Naglee, K5WKQ, 2402 Maxwell, Midland, Texas 79701

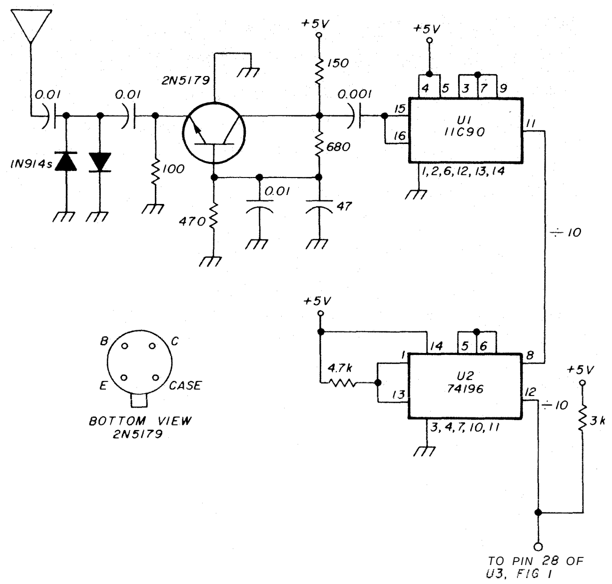


fig. 2. 600-MHz counter, which uses the 74196 chip to replace the 74LS90 divider of fig. 1.

The circuits in **figs. 1** and **2** are the same old standbys that have been used for years, so no credits are given. The 11C90 divides the incoming signal by ten into the 74196 or 74LS90, once more dividing by ten into the 7216-D.

construction

I used the Heathkit laboratory breadboard in the investigation of the many features available in the 7216-D. I used only a few features of the 7216-D (more later about these). Both circuits were thoroughly checked in breadboard fashion. I decided on the circuit of **fig. 1** for my counter; you may wish to do otherwise. I used the small LED readout display from the calculator, **fig.** 3.

The connections for this display were verified in the following manner. I soldered 12.5-mm (0.5-inch) lengths of No. 22 AWG (0.6 mm) wire into each of the pins and plugged the display into the breadboard. I connected a 220-ohm resistor to the 5-volt supply. Then in a trial-and-error method, using a ground wire and the current-limited 5-volt wire, I located each

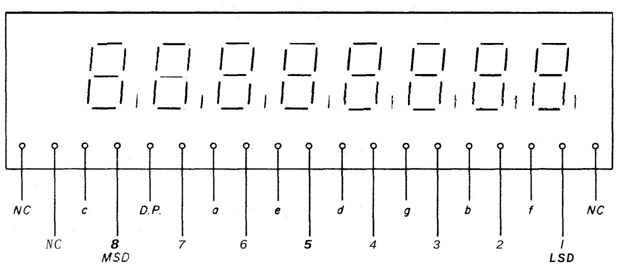
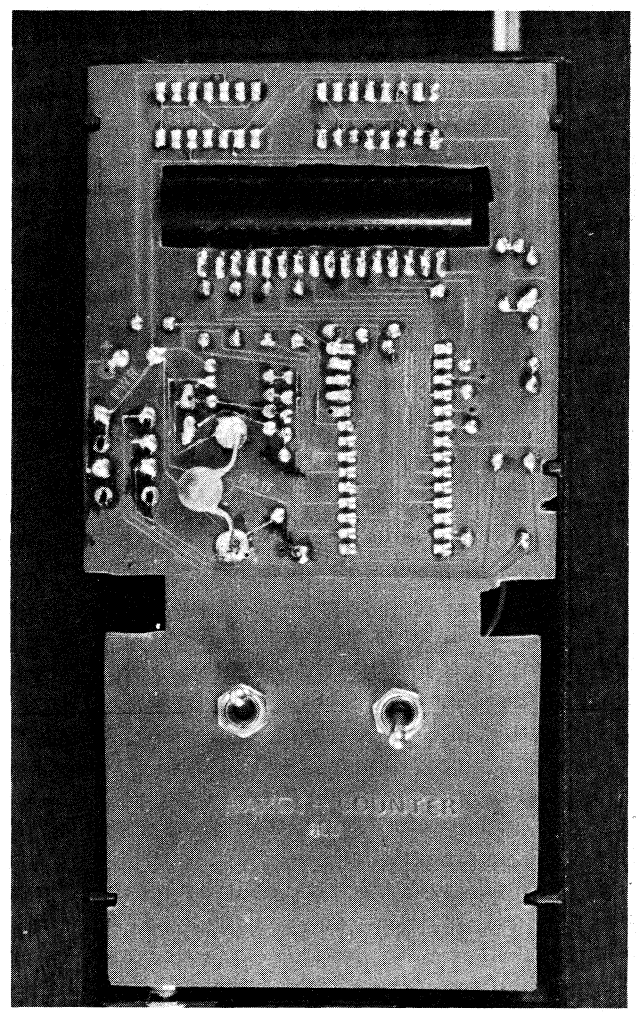


fig. 3. Top view of the 9-digit LED display. (Digit 9 is not used.)

digit and segment. For optimum brilliance and minimum current, 220-ohm current-limiting resistors were used. The frequency-determining capacitors were used for the 10-MHz crystal I had on hand; yours may be different — start out with 39 pF in each.

PC **board.** I used the old PC board with cutout for display from the calculator to get dimensions. I designed the new circuit board in pencil on grid



Handi-counter board for the 250-MHz unit.

paper in reverse. Then, using a piece of clear Mylar as a base, I used rub-on sheets for all ICs and pads, following the pencilled design. I used a Keuffel and Esser (K&E) size 00 LeRoy pen and India ink to connect the pads. This was my first experience using the positive method in PC work, and I was pleasantly surprised at the results.

The next step was to flip the clear Mylar base, placing the image next to the emulsion side of a presensitized fiberglass board. I used a piece of clear

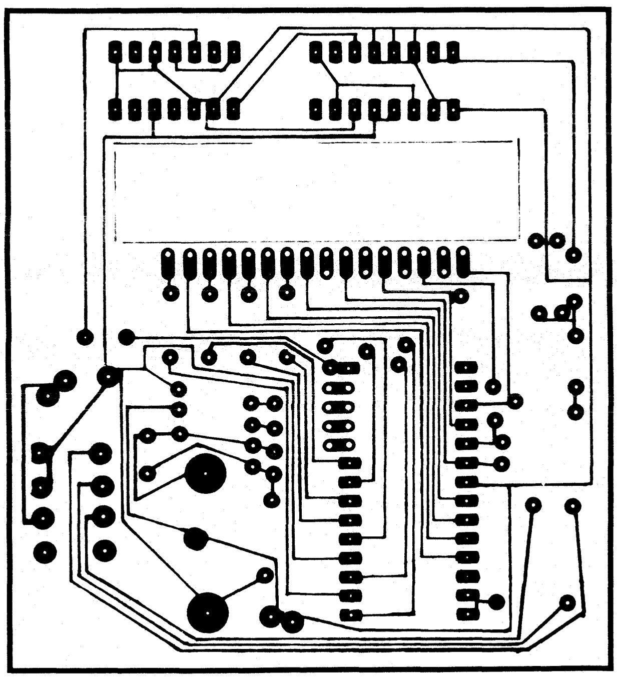


fig. 4. Board layout of the 250-MHz handi-counter. Unit fits nicely into a hand calculator case available from Poly-Paks. A layout drawing is available from the author for the 600-MHz version (see text).

glass about 6.5 mm (0.25 inch) thick to ensure good contact between Mylar and board.

Next was the exposure. I used a 250-watt heat lamp (high infrared) placed about 281 mm (11 inches) away. Exposure time was 5½ minutes.

After development I rinsed and dried the board, then I put it into the etching tank. After etching I washed the board again, dried it, and scrubbed it with steel wool. The board was then ready for cutting and drilling, after which all jumpers and components were connected.

I used no sockets, but they could be used because the breadboard design showed minimum problems with long wire connections. The board shown in **fig. 4** is an example of how I designed around the calculator case and display. Parts placement is shown in **fig. 5;** circuit boards are available.*

Power supply. Four ni-cad AA cells with 450 milliampere-hour ratings were connected to the existing charger plug in the calculator case. I used a small plastic battery holder.

Antenna. The antenna is a telescoping type, which is available from Radio Shack stores. It's bolted to a Bakelite bracket, which is cemented to the case.

"Circuit boards are available from Whitehouse Et Co., Newbury Drive, Amherst, New Hampshire 03031.

ICM 7216-D counter IC

Fig. 1 shows all four available gates. Only two are used in my counter. Pads for additional gates were provided for future use. The pinout of the 7216-D is correct; pins 4 and 5 are digits 3 and 2 as shown. Calibration was for cold starts; that is, no warmup time because for my use the counter is used only for short intervals.

These circuits use only a portion of the capabilities of the 7216-D. During breadboard design all of the features were investigated. The device has such a potential that I've ordered another chip for a counter to be used in my station. It will count between ±0.1 Hz and 600 MHz with a TXCO crysta!. The ICM 7216-D features:

1. Frequency measurement from dc to 10 MHz (mine went to 12.5 MHz)
2. Output drivers will drive large common-cathode LED displays plus overflow LED
3. Selectable decimal points
4. Leading zero blanking
5. Eight-digit multiplexing outputs
6. Single 5-volt supply
7. Either 1- or 10-MHz crystals
8. Selectable 0.01, 0.1, 1.0, or 10.0 second gating times
9. Display hold
10. Display blanking

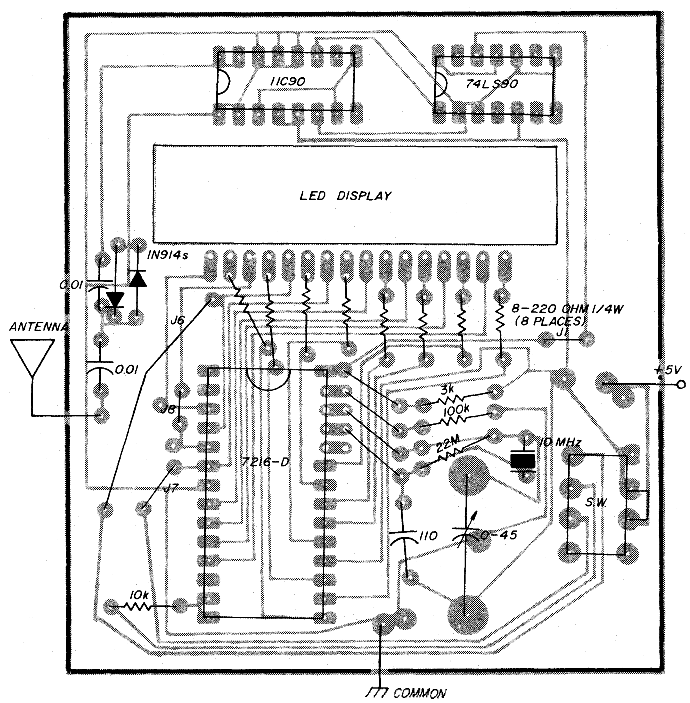
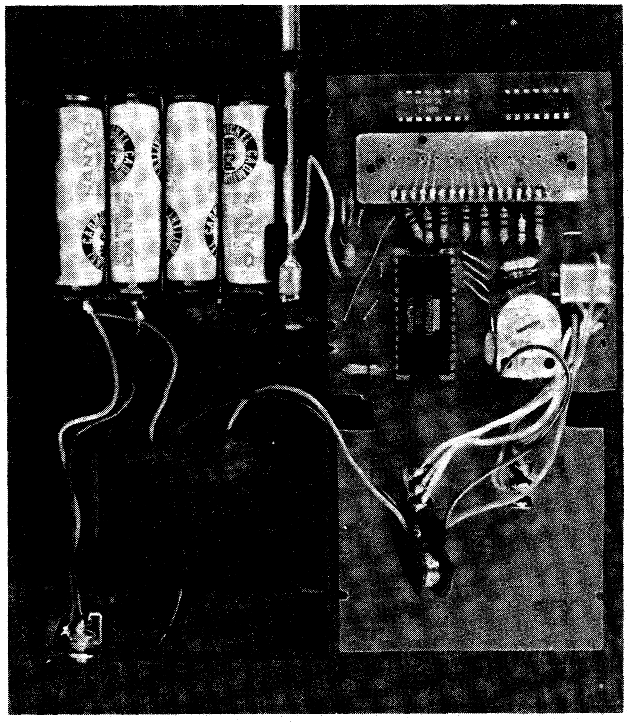


fig. 5. Parts placement for the 250-MHz handi-counter.



250-MHz counter showing component arrangement.

The data sheet accompanying my chip was for the 7216-C, which uses a common anode display. The pinout was *incorrect.* **Fig. 1** pinout is correct.

concluding remarks

I take no credit for any of the designs, only for putting it all together into a small calculator case. I'll be glad to answer all correspondence that includes a self-addressed, stamped envelope. For those interested in cost:

part	source	price
ICM-7216-D	Circuit Specialists	$20.95
11C90	Circuit Specialists	14.95
74LS90	Circuit Specialists	0.40
10 MHz crystal	Circuit Specialists	2.25
Two calculators	Poly-Paks	2.98 each
Four AA nicads		5.95
One battery holder		0.95
One F. G. circuit board		1.95
Miscellaneous resistors and capacitors		1.45
	Total	$51.83.

Watch out for Murphy's Law. The special switch in the calculator board was removed safely but went to pieces. Therefore, two spdt miniature toggle switches, on-off, and 1.0 or 10.0 second gating were needed.

ham radio

repair bench

Joe Carr, K4IPV

building low-voltage dc power supplies

Not many years ago Amateurs and other electronics hobbyists who needed low-voltage dc power supplies used batteries or a simple, unregulated, ac-powered circuit. The most common commercial dc power supplies of that day were unregulated, poorly filtered, and often referred to as "battery eliminators." Such power supplies were used to service car radios and other mobile equipment but were not too good for much else.

Modern solid-state electronic circuits require regulated low-voltage dc power supplies. Many circuits will work best only when the dc power supply voltage is regulated. Oscillators, for example, won't oscillate at the correct frequency unless the power-supply voltage is correct. Even where the absolute operating voltage isn't critical it must remain at the value that existed when the oscillator frequency was adjusted, or the frequency will be different. If the oscillator power-supply voltage changes while the oscillator is running, the frequency will change. Imagine mobile two-way radio if your receiver local oscillator changes frequency enough to move off-channel every time you accelerated away from a traffic light! Sound unreasonable? I've seen many fm broadcast auto radios that shifted 4-6 MHz under acceleration when the internal voltage regulator (8 volts dc) was open.

In digital electronic circuits, especially the very popular TTL, the dc voltage must be regulated. The TTL family of devices will not operate properly or may burn out if the incorrect voltage is applied. TTL devices typically want to see 5 ± 0.25 volts, and in no instance must the voltage rise above 5.6 volts! In reality a narrower range of 5 volts ± 50 mV may be required It seems that some TTL devices, especially complex or multifunction ICs, become flaky at less than 4.9 or 4.95 volts despite the fact that the manufacturer specified a wider range.

The need for regulated, low-voltage, dc-power supplies, then, is well established. What remains is to determine how to go about making such a supply with a minimum of effort and cost and with little or no sacrifice in utility. Before considering voltage-regulator circuits, let's first review the principles of dc power supplies.

transformers

A transformer **(fig. 1A)** scales the 110-volt ac power main's voltage to that required at the power supply output. In the case of our low-voltage dc supply, the transformer must be a step-down type; *i.e.,* the secondary voltage, E_s, is lower than the primary voltage, E_p. Transformers operate only from alternating current (ac) sources and obey the following relationships:

$$\frac{E_p}{E_s} = \frac{I_s}{I_p} \tag{1}$$

Note from **eq. 1** that the current ratio is the inverse of

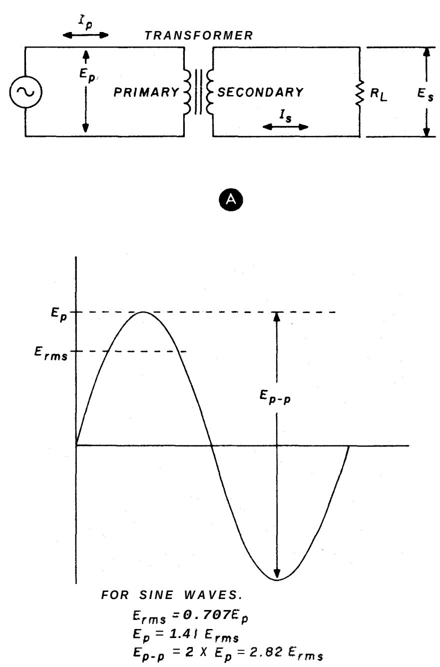


fig. 1. Schematic of a typical dc power supply (A) and the relationships between the ac voltages (B).

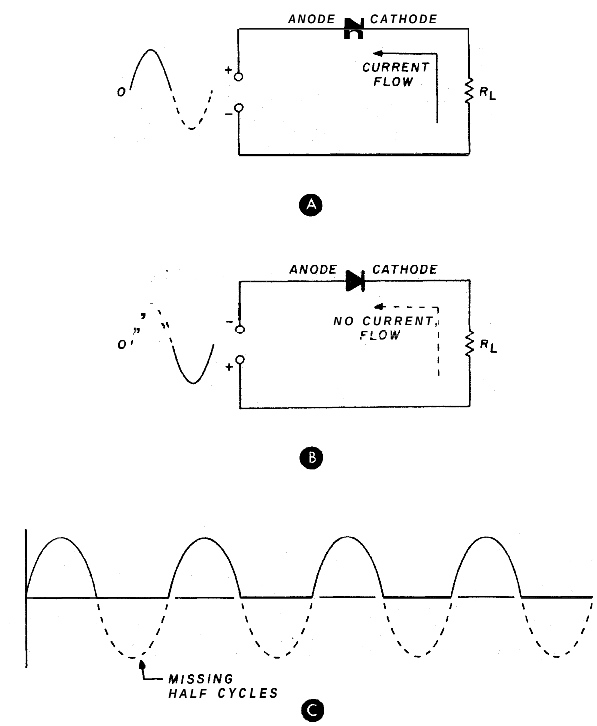


fig. 2. Rectifiers are a necessary ingredient in a dc power supply. (A) and (B) show rectifier current flow in terms of the ac input cycle. The output waveform (C) is unidirectional but pulsating (not pure dc). Because of the missing half cycles, the average voltage at a half-wave rectifier output is only 0.45 times the applied rms voltage, E_{rms}.

the voltage ratio. A transformer that steps voltage down will show an apparently equal step-up of current. This concept can, unfortunately, become confusing at times, because the actual secondary current, I_s, is determined by the secondary voltage, E_s, and the load resistance, R_L, connected across the secondary. The current step-up idea probably arises from the way **eq. 1** is often presented:

$$E_p I_p = E_s I_s \quad \textbf{(2)}$$

Eq. 2 tells us that I_p will vary as changing load conditions cause I_s to vary; as I_s goes up I_p will also go up to keep the equation constant.

Note that **eqs. 1** and 2 contain no loss terms. Real 60-Hz power and filament transformers are efficient devices; efficiency ratings of from 95 to over 99 per cent are common. This fact justifies our use of simplified equations; the error terms are very small.

Transformers are rated by the primary and secondary voltages, the secondary current, and the primary *VA* rating; *i.e.*, volts time amperes. In most cases the primary voltage will be either 110 volts, 220 volts, or will be selectable between 110 and 220 volts. The secondary voltage is determined by the turns ratio between primary and secondary windings. Both voltages are specified as root-mean-square (rms) values, but both the peak and peak-to-peak voltages become important in dc power supply design. The relationships between E_p, $E_{p\text{-}p}$ and E_{rms} are shown in fig. **1B**.

Most transformers are built with the primary winding nearest the core. As a result, the primary is more easily overheated. The primary *VA* rating, then, limits the power available from the transformer. The primary *VA* rating cannot safely be exceeded. Some get away with it but this can't be guaranteed. Only transformers built to military specifications are over-specified sufficiently to make the temptation to overrate reasonable.

rectifiers

Rectifiers (fig. **2**) pass current in only one direction. This features allows them to convert bidirectional ac into unidirectional pulsating dc. Fig. **2A** showsa simple half-wave rectifier circuit. On the positive alternation of the ac cycle, the diode rectifier anode is positive with respect to the cathode, so current will flow. On negative alternations the rectifier is reverse-biased and no current will flow (fig. **2B**). The output waveform (fig. **2C)** although unidirectional is not pure dc; it is pulsating dc. Because of the missing half cycles, the average voltage at a half-wave rectifier output is only 0.45 times the applied rms voltage. To deliver any given level of voltage and current with this circuit requires a transformer with a primary *VA* rating 40 per cent higher.

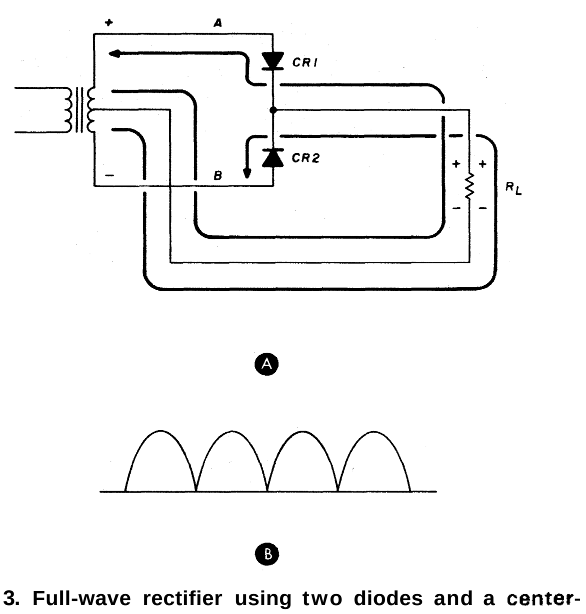


fig. 3. Full-wave rectifier using two diodes and a center-tapped transformer (A) and the double-humped waveform (B), which is characteristic of the full-wave rectifier.

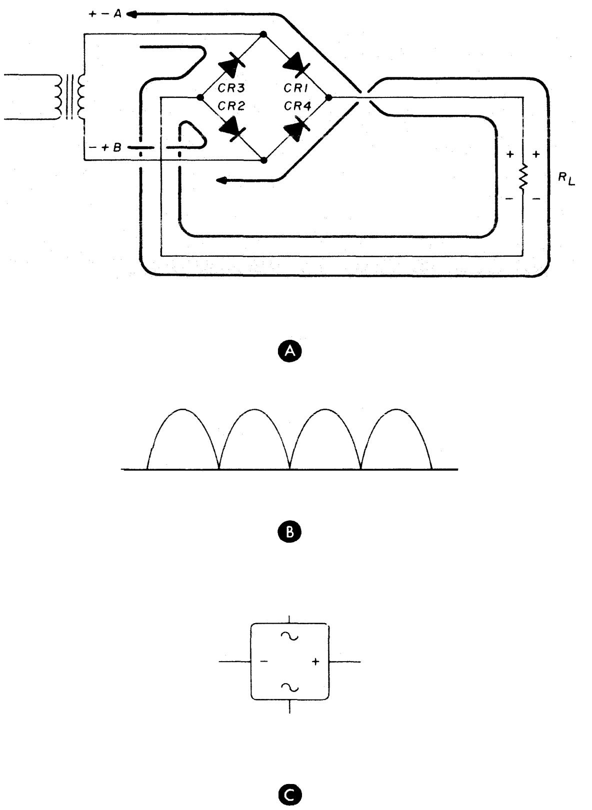


fig. 4. Full-wave bridge rectifier (A). Output waveform is shown in (B) (as in the full-wave rectifier), but average output voltage is 0.9 times E_{rms}. An example of a prepackaged rectifier showing terminal labeling is shown in (C).

The half-wave rectifier, then, is not very efficient. Rectifiers that use the entire ac waveform are called full-wave rectifiers, examples of which are shown in **figs.** 3 and **4.**

Fig. 3A shows a simple full-wave circuit using two rectifier diodes and a transformer that has a center-tapped secondary winding. The centertap is taken as zero reference, so (on any given half cycle) one end of the secondary will be positive while the other is negative. On one-half of the ac cycle, point **A** will be positive and point B will be negative. In this case, diode CR1 is forward biased; diode CR2 is reverse biased. Current flows from the centertap, through load resistor R_L and diode CR1, to point **A.**

On the second half of the applied ac waveform the situation is reversed; point **A** is negative with respect to the centertap and point B is positive. In this case, diode CR1 is reverse biased and diode CR2 is forward biased. Current flows from the center-tap, through load resistor R_L and diode CR2 back to the transformer at point B.

It is important to note that the current flow through the load is in the same direction on both halves of the ac cycle, which produces the double-humped waveform **(fig. 3B)** characteristic of the full-wave rectifier. The average output voltage produced by the full-wave rectifier is 0.9 times the applied rms potential.

A full-wave bridge rectifier is shown in **fig. 4A.** This circuit doesn't need the transformer centertap — but at the cost of two additional rectifier diodes. On one-half of the ac cycle, point **A** wi!! be positive with respect to point B. In this case, diodes CR1 and CR2 are forward biased and CR3, CR4 are reverse biased. Current flows from point B, through CR2, R_L and CR1 to point **A** on the transformer. On the second half of the ac svcle the situation reverses; point **A** is negative with respect to point B. Diodes CR3 and CR4 are forward biased while CR1, CR2 are reverse biased. Current will flow from point **A,** through CR3, R_L and CR4 to point B.

Again we see the current flowing through load resistor R_L in the same direction on both halves of the ac cycle. The output waveform, **fig. 4B,** is the same as in the previous full-wave case. The average output voltage is 0.9 times the applied rms voltage.

The transformer used with the bridge circuit need not be center tapped. The zero potential reference point is designated as the junction of the anodes of CR2 and CR3. This point is labeled –, while the junction of CR1 and CR4 is labeled **+.** Some prepackaged bridge rectifiers **(fig. 4C)** have the dc terminals labeled with the + and – symbols while the other two terminals are labeled **AC** or have the sine wave symbol as shown.

The bridge rectifier will produce an output voltage two times that of the regular full-wave circuit (from the same transformer) because it uses the entire secondary winding on both halves of the ac cycle. But this is not for free, because the primary VA rating must not be exceeded. The secondary current of most center-tapped transformers is rated for the regular full-wave circuit. If a bridge circuit is used, then

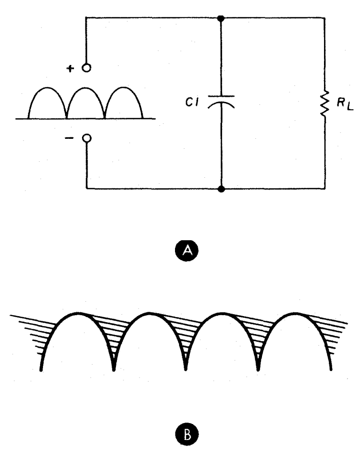
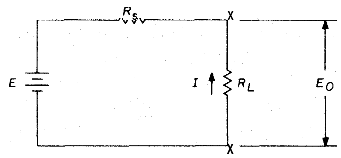


fig. 5. Simple filter circuit (A) and the filter capacitor action on the pulsating dc input (B).

fig. 6. Relationship between R_s (source resistance), R, (load resistance), I (load current), and E_0 (open-terminal voltage). E_0 may be measured by disconnecting R_L and measuring E_0 when no current flows.



the available current is only one-half the rated current or we'll find that the primary VA rating may be exceeded. The full secondary rated current can be obtained only if the transformer is designed for use in full-wave bridge rectifier circuits.

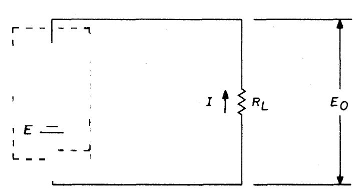
Most electronic circuits can't use pulsating dc but instead require pure dc (or nearly so). The pulsations are called ripple. Half-wave rectifiers produce a 120 per cent ripple component and a ripple frequency of 60 Hz (the ac line frequency). Full-wave rectifiers, on the other hand, produce a ripple component of only 48 per cent and a ripple frequency of twice the ac line frequency, or $(2)(60) = 120\ Hz$, in the *U.S.A.*

filters

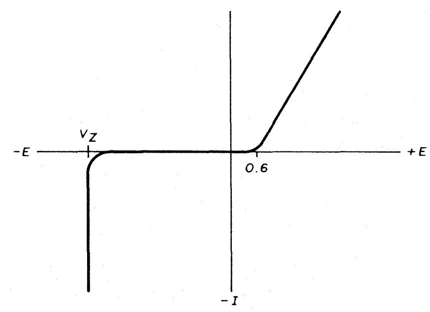
A filter circuit smooths the pulsations to produce nearly pure dc. In the simplest case, **fig. 5A,** the filter capacitor *C1* is across the output in parallel with the load. The action of the filter capacitor is shown in **fig. 5B.** The heavy lines indicate the pulsating dc waveform without the filter, while the light lines show the output with the filter. Capacitor C1 will charge to approximately E_p. But after the peak has passed, the charge will return to the circuit. The effect of returning the charge from C1 to the circuit is to fill in the area between pulses (shaded area, **fig. 5B**). The filter reduces the ripple component to a low percentage.

The value of C1 is critical to the performance of this circuit. In general the minimum required capacitance for C1 is higher in half-wave circuits than in full-wave circuits because of the higher ripple factor. For low-voltage full-wave supplies, the value of C1 should be at least 500 μF in small-current supplies (*i.*e., under 500 mA), and 1000 μF in supplies up to 1 ampere. A general rule of thumb in the over-1-ampere range is to make C1 not less than 1000 μF/ampere, with some authorities calling for not less than 2000 μF/ampere. By the second rule, then, a 4-ampere power supply should have not less than 4 times 2000, or 8000, μF of filter capacitance.

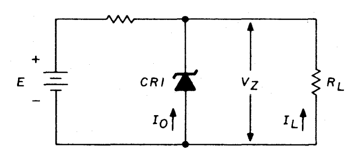
fig. 7. Voltage regulation is a function of output voltage and unloaded and loaded conditions. R_s and R_L form a voltage divider; thus E_0 is only a fraction of the open-terminal voltage.



The circuit in **fig. 5A** is the simplest dc power supply, but even with a high-value filter capacitor it will show some ripple in the output waveform. A voltage regulator will reduce this ripple to almost zero even though the regulator's main function is to maintain the dc output voltage constant. One power-supply manufacturer emphasized the regulator ripple-reduction feature by claiming that the circuit "amplified"



A



B

fig. 8. Zener regulator. (A) shows typical current-voltage curve. A common zener regulator schematic is shown in (B).

the 10,000-μF filter capacitance to make it "equivalent" to a l-farad capacitor! What was meant by this over statement is that it would take a l-farad capacitor to achieve the same ripple reduction obtained from a 10,000-μF capacitor and a voltage regulator.

voltage regulation

All dc power supplies have a certain amount of internal, or source, resistance, R_s. When a load current, I, is drawn from the power supply, a voltage drop, $E = IR_s$, will occur across the source resistance.

The value of the internal resistance can be determined by Ohm's law. The voltage used in this calculation is the power supply open-terminal voltage. This voltage is measured by disconnecting load resistor, R_L, then measuring the power-supply output

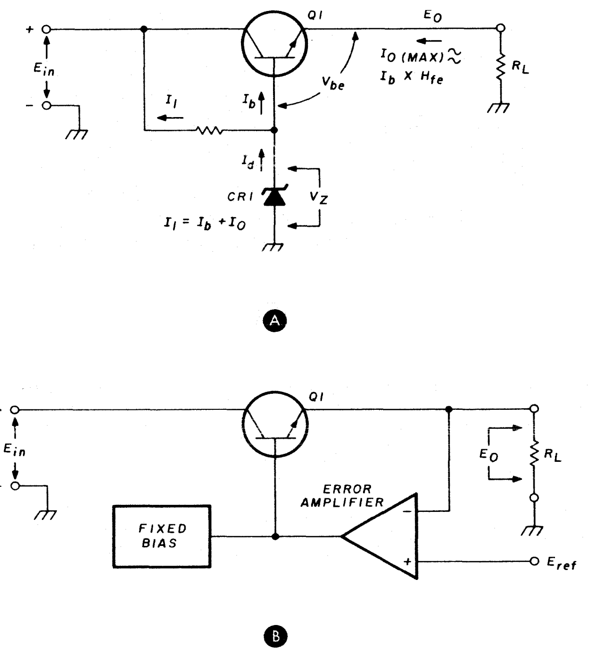


fig. 9. Zener regulator using a series-pass transistor (A). A feedback circuit using an error amplifier IC is shown in (B).

when no current flows (see **fig. 6).** The current used in the calculation is that which will flow when the output terminals are short circuited. Don't try to make this measurement, however! Few real-world power supplies can withstand an output short circuit without damage. An alternative method for determining the internal resistance is this:

$$R_s = \frac{E-E_0}{I} \qquad \textbf{(3)}$$

Where:

R, is the internal resistance of the supply (ohms)
E is the unloaded (*i.e.*, ■ = *0*) output voltage
E_0 is the loaded output voltage (*i.e.*, ■ = *0*)
■ is the output current

Voltage regulation is a measure of how stable the output voltage remains between no load and loaded conditions. **Fig. 7** shows the mechanism that causes the output-voltage change, Internal resistance R_S, is effectively in series with load resistance R_L. The output voltage, then, can only be a fraction of the open-terminal voltage because R_s and R_L form a voltage divider. The output voltage at full load will be:

$$E_0 = \frac{ER_L}{R_s+R_L} \qquad \textbf{(4)}$$

Voltage regulation is usually specified in terms of a percentage, which is calculated from:

$$\textit{per cent regulation} = \left(\frac{E-E_0}{E}\right)(100) \qquad \textbf{(5)}$$

regulator circuits

In this article we'll consider three basic forms of regulator circuit: zener diode, zener-refeienced series-pass, and feedback.

Zeners have a property that allows them to be used as voltage regulators. **Fig 8A** shows the curve for a typical zener. In the +E region the diode is forward biased and behaves exactly like any other silicon diode. But in the – E region, in which the diode is normally reverse biased, the zener behavior is somewhat different from that of the ordinary diode. From zero volts to a point called V_z the zener acts like an ordinary diode; current passed is zero. But at V_z the zener "breaks over" and conducts a reverse current. V_z tends to remain constant and is the voltage to which the zener will regulate the applied voltage.

Fig. 8B shows a typical zener voltage regulator circuit. CR1 is connected in parallel with the load. R1 is a series-current limiter that protects the zener from excess current flow. Without R1 the CR1 would burn out.

The simple zener voltage regulator is used for light-duty work only. A general rule of thumb is that the load current should be held to 10 or 20 per cent of the zener current, hence the limitation to low-current applications.

A better solution is to use the zener as a reference source in a series-pass circuit as in **fig. 9A**. In this circuit Q1 is the control device and CR1 is a reference. The output voltage is given as approximately:

$$E_0 \approx V_z - V_{be} \qquad \textbf{(6)}$$

The maximum output current is approximately the normal load current, I_b, allowed by the zener multi-

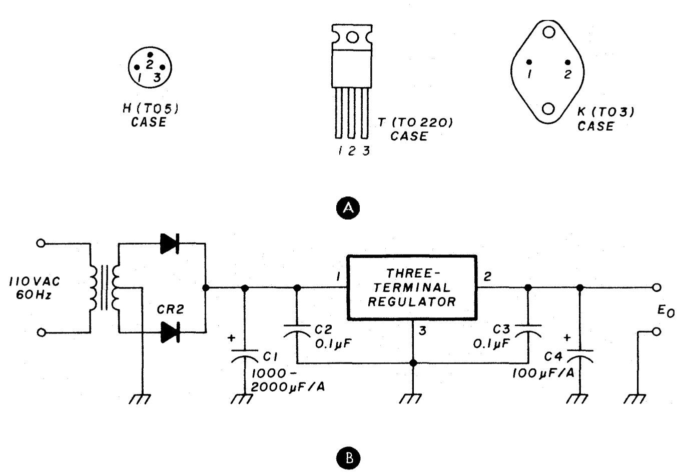


fig. 10. Some case styles for three-terminal voltage regulator devices (A). (B) shows the basic circuit for these regulators. Values for C1 and C4 are minimum (see text).

plied by the beta of Q1; assuming that neither Q1 maximum collector current nor its maximum power dissipation ratings are exceeded.

Another series-pass type regulator is the feedback circuit shown in **fig. 9B**. In this simplified schematic the bias on the series-pass element, 01, is determined by the output of an error amplifier. This amplifier is a differential circuit, meaning that its output is proportional to the difference between two input voltages (*i.e.*, $E_{ref} - E_0$). If E_0 changes from the value set by E_{ref} (when the load current changes), then the amplifier output changes in a direction that corrects the change.

Most readers aren't interested in designing complex dc power supplies but want to know how to make supplies that will meet their needs. The supplies in this section will meet the needs of most Amateur Radio applications for either workbench or project use.

The easiest way to obtain supplies up to 5 amperes capacity at a fixed voltage is to use a three-terminal IC regulator. Such regulators can be obtained in most standard voltages up to 24 Vdc.

Several families of three-terminal regulators are available at current levels of 100 mA, 750-1000 mA, 3 amperes, and 5 amperes. **Fig. 10A** shows several case styles used for these devices. The letters denoting the case style are used as a suffix in the regulator type number. An LM309, for example, is a 5-volt regulator, so an LM309H is a 100-mA device in a TO-5 transistor package, while the LM309K is a l-ampere device in a TO-3 transistor package. The T-style package is generally limited to 750 mA in free air although frequently advertised at 1 ampere.

Often the ratings of these devices are exceeded even though this isn't always good practice. The claim is made that the ratings can be exceeded if proper heat-sinking is applied. I've seen the H-style package devices operated at 150-200 mA, T style at 1 ampere, and the K style at 1.5-2 amperes.

The most common three-terminal devices are the LM340-series and the 78xx-series. For LM340 devices the case style would be denoted by the suffix letter (H, K, or T), while the output voltage is denoted by a hyphen and the voltage; *i.e.*, a 12-volt output 750-mA (T package) device would be given the designation LM340T-12. The 78xx-series devices replace the "xx" in the generic type number with the voltage rating. The 7805, then, is a 5-volt device while the 7812 is a 12-volt device.

The basic circuit for these regulators is shown in **fig. 10B**. T1, diodes CR1 and CR2, and capacitor C1 are the same as in any unregulated supply (a bridge rectifier may be substituted for CR1/CR2). Capacitor C1 is the filter capacitor and should have a value not

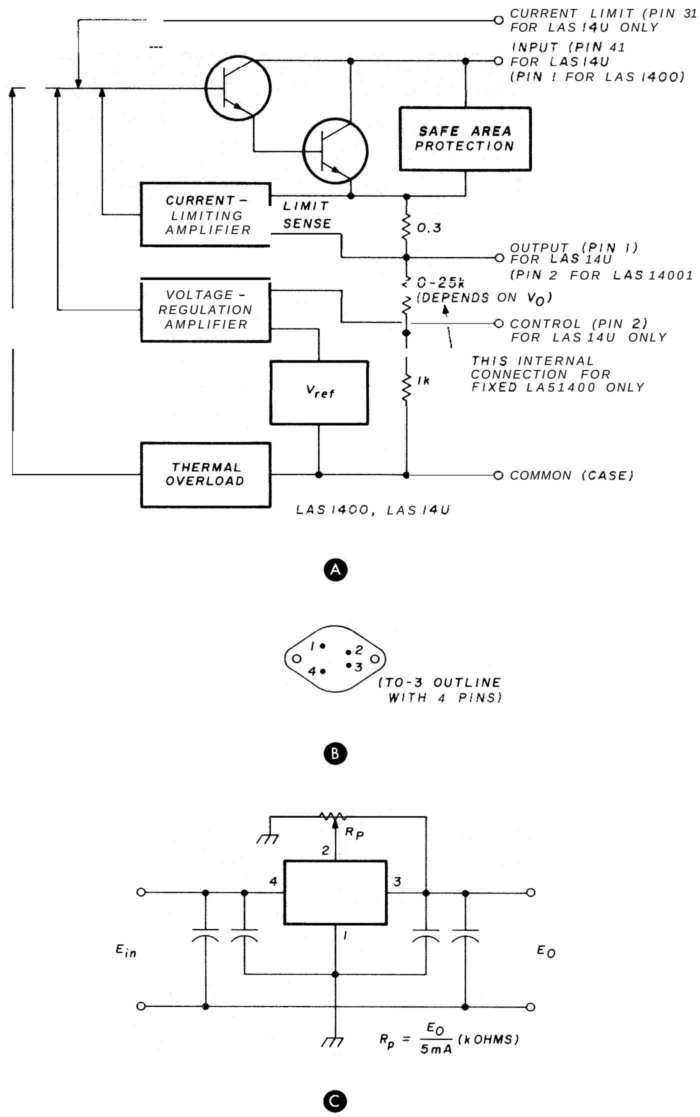


fig. 11. Adjustable four-terminal voltage regulators offered by Lambda Electronics. A simplified circuit is shown in (A). The case (similar to the TO-3) with pinouts is shown in (B). Regulator can be set to output voltages with pot R_p between 4 and 30 Vdc (C).

less than 2000 μF/ampere (I,,,). Capacitors C2 and C3 improve regulator noise immunity and should be mounted as close as possible to the regulator body. Capacitor C4 is optional but improves the circuit transient response. C4 should have a value of not less than 100 μF/ampere.

Until recently there were few three-terminal voltage regulators on the market for current levels over 1 ampere. The LM323, for instance, would pass 3 amperes at a fixed 5 volts. Fairchild Semiconductor now offers several devices in the 5-ampere range. Lambda Electronics* offers three lines of three-termi-

*Lambda Electronics, 515 Broad Hollow Road, Melville, New York 11746 nal devices in TO-3 packages. The LAS-15xx series produce 1.5 amperes, the LAS-14xx series produce 3 amperes, and the LAS-19xx series produce output currents up to 5 amperes. As in the 78xx series, the "xx" is replaced by the output voltage. An LAS-1905, then, is a 5-volt, 5-ampere device, while the LAS-1512 is a 12-volt, 1.5-ampere device.

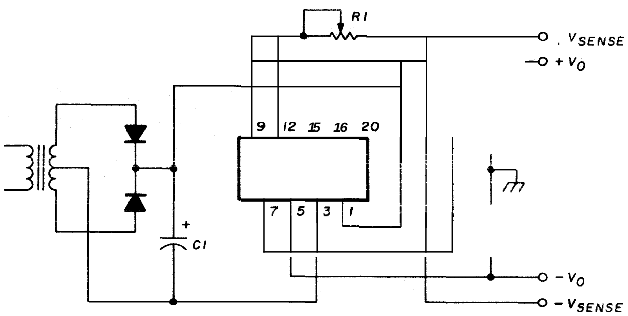


fig. 12. A hybrid voltage regulator offered by Lambda Electronics. R1 minimum value should be 0.25 E_0 **times 1000 ohms per volt except for 5-volt models, in which case 3000 ohms is used.**

An exception to the Lambda numbering scheme is the LAS-19CB, which delivers 13.8 volts dc at the output at up to 5 amperes. This device is designed to power Amateur Radio and CB mobile transceivers on the bench or in a base-station. (Most "12-volt" mobile electrical systems are actually 13.8-volt electrical systems.)

All three-terminal regulators have a certain minimum and maximum input-output differential voltage ($E_{in} - E_0$) rating. A typical number is 2.5 volts for the minimum and a maximum in the 30-40 volt range. The minimum *I/O* differential is the smallest difference that will allow the regulator to operate properly, while the maximum voltage rating is the level that will probably cause device burnout if exceeded.

The ***I/O*** differential, however, can be a trouble area for the unwary. A 5-volt, 5-ampere device may be rated for a maximum input voltage of 35 volts dc and a maximum power dissipation of 50 watts, The power actually dissipated is:

$$P_d = I(E_{in} - E_0)$$

$$P_d = (5)(E_{in} - 5) \qquad \textbf{(7)}$$

So if the maximum input voltage is 35 volts,

$$P_d = (5)(35 - 5) = (5)(30) = 150 \text{ watts} \qquad \textbf{(8)}$$

Clearly, we can't expect to draw the full current at the maximum input voltage without exceeding the device power dissipation rating. By rearranging **eq. 7** to solve for E_{in} when P_d is 50 watts, the maximum input voltage at full-rated output current is 15 volts. Although the 5-volt device is used as an example, the same reasoning applies to other regulators as well.

Lambda Electronics also offers a line of adjustable four-terminal regulators in packages that are similar to TO-3. A simplified internal circuit is shown in **fig. 11A**. Note that it's a feedback device with thermal, safe operating area and current overload protection Overload protection is very important in situations where the device being powered shorts out or where the alligator clips on your bench supply come together! The series-pass element in these four-terminal regulators is a Darlington pair to take advantage of the extremely high gain. The case isimilar to TO-3) and pinouts are shown in **fig. 11B**. Although standard TO-3 heatsinks will work for this device, it's necessary to drill the extra holes.

The Lambda four-terminal adjustable voltage regulators can be set to any output voltage within their specified range using potentiometer R_p **(fig. 11C** is a typical circuit). The range for most models is 4-30 Vdc. The LAS-15U, LAS-14U, and LAS-19U produce output currents of 1.5, 3, and 5 amperes respectively. With the exception of the control pin and the potentiometer, the four-terminal regulator uses the same general circuit as the three-terminal devices.

High-current power supplies (over 5 amperes) are sometimes tricky to design properly. You can't just slap a zener into the base circuit of a series-pass transistor and make it work reliably. Additionally, ready-built, high-current power supplies are often quite costly. A reasonably simple solution to these problems is a Lambda high-current, hybrid-module device:

Lambda series	**ratings**
LAS2000	5 amperes185 watts
LAS3000	10 amperes1140 watts
LAS4000	10 amperes1170 watts
LAS5000	20 amperes1270 watts
LAS7000	30 amperes1400 watts

The LAS5205 is a 20-ampere, 5-volt regulator, while the LAS7215 produces 15 volts at 22 amperes. **Fig. 12** shows an example of a hybrid voltage regulator. The voltage trim pot should have a minimum value of ($0.25E_0 \times 1000$ *ohms/volt*) except for 5-volt models, in which case a value of 3000 ohms is used. In practice these are minimum values and the actual value must be found by experimentation. In the case of my own LAS5205, used in a Digital Group 2-80 computer, the required value was 12 kilohms.

Note that both pins 2 and 20 are plus-input-voltage terminals. This can lead you astray in some cases because pin 1 will operate with only 2.5 volt I/O differential voltage, while pin 20 must have not less than 7.5 volt I/O differential. Pin 1 is the normal high-current terminal leading to the series-pass transistor while pin 20 powers the internal control circuit.

In my computer power supply, E_{in} for the high-current side was 8 volts (obtained from a 6.3-Vac fila-

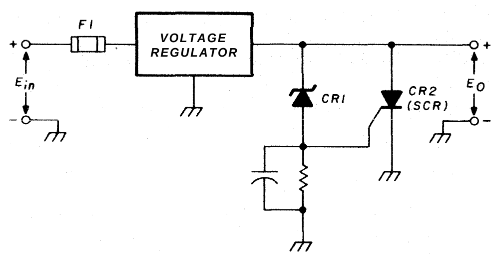


fig. 13. Overvoltage protection scheme using a zener (CR1) and a silicon-controlled rectifier (CR2). The circuit is known as the "SCR crowbar" approach.

ment transformer), while the control circuit input voltage on pin 20 was derived from the +15 volt supply used with a +12 volt regulator in the computer.

overvoltage protection

One aspect of regulated power-supply construction often overlooked by the Amateur is overvoltage protection. Very few electronic circuits can tolerate excessive input voltage for more than a few seconds. TTL IC devices, for example, can burn out if more than 5.6 volts is applied to V_{cc} for an extended period of time. Amateur and CB transceivers that normally operate on 13.8-volt systems will flunk a smoke test when +17 volts is applied, which is the normal voltage obtained from a rectified and heavily filtered 12.6-volt transformer!

The solution is to provide some form of overvoltage protection (OVP); that is, a means of turning off the power supply if the output voltage exceeds a preset limit. **Fig. 13** shows a common approach to OVP design. CR1 is a zener with a breakdown voltage greater than E_0 but less than E_{in}. When E_0 rises to a value higher than the breakover voltage, CR1 conducts and turns on the gate of silicon-controlled rectifier CR2. When CR2 turns on, a short-circuit occurs across the regulator output that blows fuse F1. This a brute-force approach called an **"SCR** crowbar."

Lambda Electronics offers monolithic **SCR** crowbar devices in TO-66 and TO-3 packages as well as certain heavy-current devices in custom packages. Information on these devices may be obtained by writing to Lambda Electronics at the address given previously.

ham radio

active antenna coupler for VLF

A discussion of active low-frequency and very low frequency antenna preamps, with some details for extending the frequency range to 10 MHz and above

The goal of this project has been to have a single, electrically short antenna operate over a wide frequency range with a minimum of interference or operational problems. The biggest problems are those of local noise pickup from 60-Hz harmonics and overload from strong, out-of-band signals (above 300 kHz). In general, any jfet, mosfet, or even CMOS inverters may be used to provide high power gain, but some circuits work better than others.

The primary purpose of a vlf receiving antenna coupler is to convert the low-level signal at the high-impedance pickup point on a short antenna to a low-impedance level for driving the feed cable back to the receiver. The use of bipolar transistors has invariably resulted in problems due to intermodulation distortion or cross modulation from nearby broadcast-band transmitters. A jfet is far less susceptible to this problem over a wide dynamic range. One of the most common jfets available is the MPF102, which I used in the preamplifier circuit shown in **fig. 1.**

circuit description

The input lightning arrester is of obvious value when a good low-impedance ground system is provided at the antenna mounting. The series input resistors, with the capacitance of the neon bulb and trigger diode, serve as double RC filters to help reduce broadcast-band and high-frequency signals. They also provide static discharge protection. The choice of trigger diode for the low-voltage limiter is critical. Some diacs and thyristors are quite nonlinear and have appreciable resistance/capacitance variations. Some General Electric and Japanese diodes, apparently constructed as back-to-back 14-volt zeners, appear to work best and will prevent burnout of the preamplifier in all but the worst-case, direct-hit situation. You should never use parallel, opposed polarity silicon or germanium diodes in place of a trigger diode, because these produce an almost idea! crystal detector for broadcast-band signals with direct audio signals flowing down the cable!

The output transformer in **fig. 1** is operated in a step-down mode from the jfet drain terminal to provide a higher current driver for the cable at a 150- to 350-ohm impedance level. Fifty-ohm cable is not a perfect match, but at these low frequencies the cable looks like a capacitor and there is no VSWR problem because of the very short electrical length of the cable. The transformer is a standard 600-ohm, center-tapped, line-to-line type. Some UTC subouncer models work just as well and will pass frequencies to 300 kHz or more when terminated with a 330-ohm resistor at the receiver end. Power for the preamplifier flows up the cable from the 330-ohm isolating resistor at the receiver end. The cable capacitance helps limit the high-frequency response. In fact, additional capacitance directly in parallel with the coaxial cable may be used to restrict the preamplifier response for use below 100 kHz.

The coupling capacitor to the receiver should be fairly large if you are interested in signals down to the 10-kHz range or below. The preamplifier will drive 50-ohm cables up to 30.5 meters (100 feet) long and still provide adequate response up to 200 kHz. Seventy-five ohm cable can be used to reduce the cable capacitance for longer runs.

The input impedance and sensitivity of the preamplifier are limited by the input RC protective networks and the relatively high-current operating level. It is a good idea to check the current with a meter temporarily connected in series with the 330-ohm power supply isolating load resistor. Current should be about 4.5 mA at +5 Vdc. If you observe a drastically different current, try changing the 330-ohm load resistor, but be sure to check the preamp operation. In testing dozens of MPF102s, a bad one with too

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low an I_{DSS} specification could be found. You should be able to generate a 50-mV rms output signal, with no distortion at the receiver end, when a 50-mV rms, 100-kHz input signal is connected to the antenna terminal through a 100-pF capacitor. A 100-pF antenna input capacitor will roughly simulate a 2-meter (6-foot) whip antenna for testing purposes with a low-impedance signal source.

Limiting will start at about 100-mV rms, which is receivers like the Radio Shack DX-300 which tunes down to 10 kHz.

antenna-mounted preamp

A standard 2-3/4 meter (108-inch) CB whip is used for the antenna, with the preamplifier mounted as shown in **fig.** 3. The antenna should be vertical and in the clear above the immediate terrain if possible. The higher the antenna is mounted, the less will be

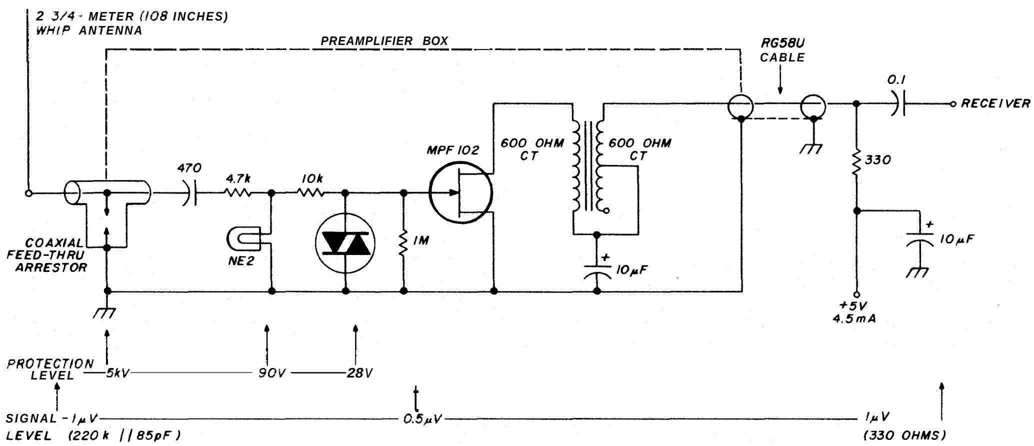


fig. 1. Schematic diagram of one version of a vlf antenna preamplifier. The amplifier's voltage gain is 6 dB, with a noise figure of 3 dB. The − 3dB frequency response, while feeding 15 meters (50 feet) of RG-58U, is 300 Hz to 300 kHz. The maximum signal level without distortion is 100 mV rms. A GE ST-2 or equivalent diode is used as the trigger diode. The coupling transformer is a Mouser TL016.

more than enough range for most receivers. When the input signal is strong enough to start limiting, the output is reasonably symmetric because of the grounded source and high current operation of the MPF102. The noise figure of about 3 dB is adequate for most uses, since the antenna noise is quite high. Typically, in the 100-kHz region, the atmospheric noise level will be over 10 μV/kHz/meter. With a 3-dB noise figure, the preamp generated noise is about 0.1 μV/kHz.

receiver coupling

The simplest way of coupling the preamp to a receiver is shown in **fig. 1**. Other methods using another transformer to drive two receivers or a tuned input circuit are illustrated in **fig.** 2. The wideband, two-receiver circuit is of value in operating a low-frequency, 10.2-kHz Omega receiver in parallel with a wide tuning range receiver with the same antenna.

The tuned-circuit coupling method has been used for a number of experimental receivers with Polyakov's detector[1] and a direct-conversion method with a balanced mixer.[2] These input circuits might also be used with some of the surplus RAK-RBA receivers, the Palomar Engineers vlf converter, the new Elmek LXK 60-kHz receiver, and with various modern

the ac ground noise pickup problems. The arrangement shown is mounted on a cast iron sewer vent pipe which serves as a good low-impedance ground. A small plastic crutch-tip cap at the end of the antenna whip helps reduce corona discharge problems in turbulent weather. I usually seal all joints, including the preamp box, with silicone rubber sealing compound to prevent moisture from entering. However, in the past, water has sometimes entered the box or antenna connectors. A small bleed hole is drilled in the very lowest or bottom part of the box assembly to drain away any moisture that runs into the assembly.

My only bad experience with lightning was when one of the systems failed due to a strike on a tree nearby. The antenna system apparently suffered a side streamer discharge, but the preamplifier itself was not damaged. The coax cable shield was burned to a crisp inside the jacket with not much obvious damage to the plastic outer jacket. There was a burn point where the cable bent over the roof at a well-grounded rain gutter. The receivers connected in the lab were not damaged.

other antennas

The preamplifier of **fig. 1** can be used with hori-

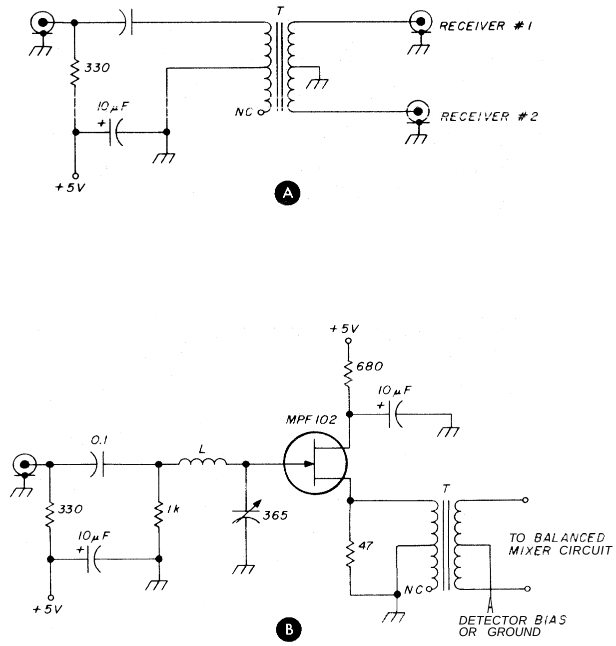


fig. 2. Diagram (A) shows a method for connecting two receivers to a single preamplifier. The transformer is a Mouser TL016. (B) is a tuned input circuit for use with a direct-conversion type of receiver. The inductor is chosen to resonate at the input frequency.

zontal antennas up to 30 meters (100 feet) long, if desired. However, I do not recommend antenna lengths of 150 meters (500 feet), which some DX hunters have used. In many cases, there is not much advantage in long wires over a good vertical radiator. The problem with the long wire is a good ground, underneath the length of the antenna, which will not have fluctuating 60-Hz ground currents. A big problem with long-wire vlf antennas for Amateur users is that the wire picks up 60-Hz harmonics along the entire length of the antenna, which tends to cancel the effectiveness of the length. For a simple installation, a single vertical antenna mounted reasonably in the clear will provide a better signal-to-noise ratio than a long wire strung out over the landscape.

An E-field whip antenna with a wideband preamplifier has one big advantage in that all tuning is done at the receiver end of the circuit. A tuned circuit antenna has to be adjusted for each new frequency range and this becomes a major problem when the antenna is mounted remote from the receiver shack. Loop antennas have a similar restricted bandwidth, compared with this wideband system. H-field loop antennas do have another advantage in that they may be rotated to reduce noise pickup. With an E-field antenna whip, there is no easy way of reducing noise pickup from nearby power lines and varying ground currents except by changing the antenna location, moving it higher, or providing better ground systems directly under the antenna.

audio interference

Audio rectification problems sometimes develop as a result of strong broadcast signals. This does not appear to affect the receiver's signal-to-noise ratio as long as the audio signals do not pass directly into the detector. Some experimental direct conversion receivers have exhibited direct audio feedthrough. This can sometimes be cured with a highpass filter at the receiver input instead of the 0.1-μF coupling capacitor. Tuned transformer or link-coupled input circuits can also be used to reduce direct audio interference feedthrough.

Audio rectification is caused by some nonlinear element or corroded joint and a parallel ground loop where some of the dc current for operating the preamp is being modulated. In one case, on a flat roof building, connecting the antenna mounting to a supposedly conducting member of the roof structure resulted in high broadcast noise pickup at the receiver. The problem was solved by isolating the preamp ground return from the antenna-mount lightning arrester ground such that there is no direct dc connection at the roof. The preamp box "floats" at the end of the coax cable and there is no chance of the roof truss ground providing parallel ground currents along the cable to generate additive or cross modulation effects. This can be done in **fig.** 3 by substituting a tight jam-fit plastic pipe coupling for the

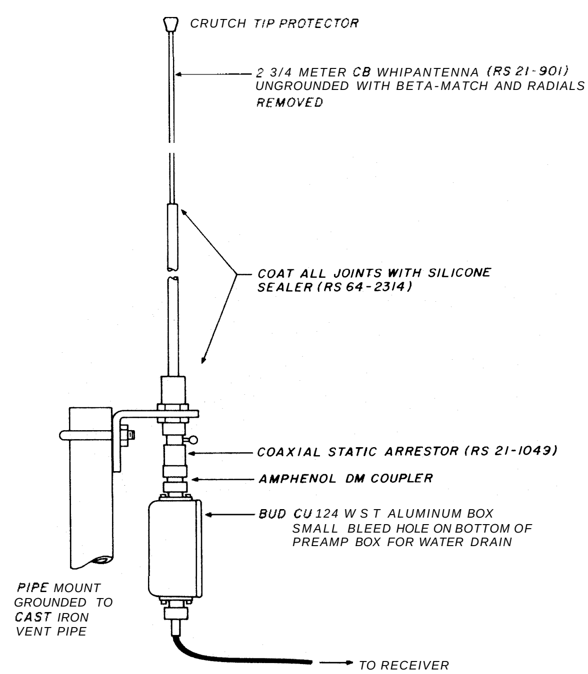


fig. 3. Vlf/low-frequency antenna and preamplifier mounting details.

Amphenol DM (double male) fitting with only the center conductors of the uhf fittings connected with a short length of threaded rod. Thus, the antenna mount is insulated from the preamp and cable with the common ground point at the receiver end of the circuit.

wideband modifications

Attempts to operate these jfet preamps over a wider bandwidth results in many problems. One circuit which has been used up to 10 MHz is illustrated in **fig. 4**. The input protection is reduced to provide less attenuation. The trigger diode will typically contribute 12 to 25 pF of added capacity. At vlf, this is not of much consequence, but at 10 MHz this results in about a 5 dB signal loss. **Fig. 4** uses feedback to provide high current gain and unity voltage gain. The gain is adjusted for the I_{DSS} of a particular 2N5457 by changing the value of the 680 ohm to 1-kilohm resistor. I usually try to adjust for unity voltage gain at 100 kHz. This provides the best linearity and a dynamic range of up to 200,000 μV rms. The output impedance of this amplifier is 80 ohms. There will be a VSWR ripple at the higher frequency range when driving a 50-ohm cable as illustrated in **fig. 4B**. VSWR effects at the high end are noted in both cases. The transformer preamp has a still higher output impedance, but this is of little consequence since I was not interested in performance above 200 kHz. Vlf receivers usually look like a 300- to 600-ohm load to the preamp cable with the preamp of **fig. 1**.

With conventional high-frequency receivers, the input may look like a 50-ohm load, which also produces a VSWR effect. The circuit of **fig. 4** has been used for WWV reception at 10 MHz with performance often better than that obtained using an untuned short length of antenna wire. Thus, while there is an apparent 16 dB voltage loss at 10 MHz, the preamp still has some power gain at this frequency when used with a short antenna. Most high-frequency receivers have excess rf gain and a low enough signal-to-noise performance to be used with this preamplifier up to about 10 MHz. The circuit of **fig. 4** is better matched to 75-ohm or 95-ohm cable providing slightly fewer VSWR effects and somewhat better frequency response.

A number of other variations have been tried, including use of two or more transistors in the preamplifier box configured as a source follower driving an emitter follower. None of these have provided

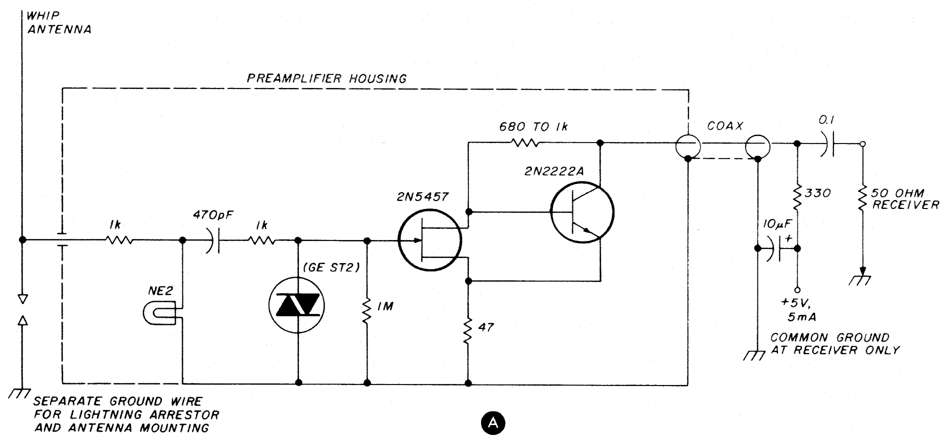
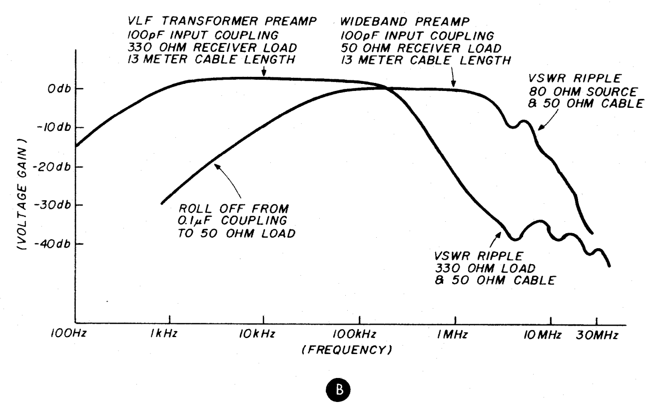


fig. 4. Diagram of a wideband active antenna preamplifier (A). The graph (B) shows the difference between this active preamplifier and that shown in fig. 1.



much better performance than the circuit illustrated in **fig. 4A.** One interesting possibility for those who like to wind their own toroidal transformers would be a bifilar-wound toroid connected as a 4:1 unbalanced step-down transformer from the jfet drain terminal An inductance of about 1 to 4 mH should provide a wideband device over the range from 60 kHz and up. Surplus pulse transformers have been used in this manner with some success, but it's difficult to obtain

receiver threshold properly, hence, some more gain is often desirable. A preamplifier with 20 dB gain over a frequency range of 100 Hz to 1 MHz is shown in **fig. 5.** This circuit uses the same 2N5457 jfet as in **fig. 4**, but is operated as a voltage amplifier instead of a source follower. The emitter follower provides a low output impedance. This preamp will also work up to about 20 MHz, but the gain drops off to 0 dB or unity at this frequency. The circuit is more suscepti-

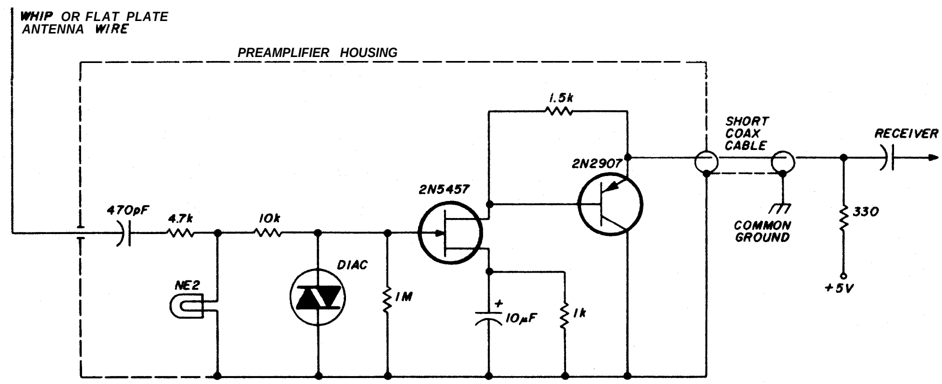


fig. 5. Schematic diagram of a preamp capable of 20 dB gain over the frequency range of 100 Hz to 1 MHz.

a really wideband circuit without reducing all protection at the input to the jfet. There is an inherent problem in that at vlf frequencies you require a very high input impedance for a short whip antenna, but at higher frequencies like 10 MHz, the antenna is a much lower impedance and should be terminated with a lower-impedance circuit.

Still another thought for the experimenter is to consider the methods used in the input circuits of wideband oscilloscopes covering up to 30 MHz. The difficulty here is dynamic range and the circuit complexity, usually requiring a separate dual power supply lead to the preamplifier and perhaps even balanced shielded cable at the output.

High impedance circuits require low capacitance at the input to provide a really high frequency response. One of the most common problems is that of the capacitance to ground of the antenna mount and lightning arrester. This has a big effect on the sensitivity of a short antenna at frequencies like 10 MHz when coupled to a high-Z circuit.

high-gain preamplifier

For very short antenna systems, such as used in mobile or airborne systems, a high-gain preamp is desirable to make up for the low-level signal received on antennas less than 1 meter (3 feet) in effective height. Some vlf receivers are designed such that a low level of 1 μV or so is required for minimum detectable signal. With an electrically short antenna, there may not be enough signal developed to activate the

ble to overload than the previous preamps so some input RC filtering is used to restrict the range.

Other variations are of course possible, including the use of the output transformer coupling method of **fig. 1** and other biasing schemes for operating the jfet at higher gain. In portable/mobile use, there is not so much concern for a low output impedance cable driver since the receiver can be located close to the preamp with a short length of cable. Circuits like **fig. 5** have been used in general aviation aircraft with good results in the 100-kHz Loran-C range. Marine users are cautioned against using these high-gain preamplifiers because of ground-loop interference problems caused by rusty hulls and poor grounding practice in many boats. A conventional preamplifier more like **fig. 1**, mounted up on a mast well away from the superstructure, will usually provide satisfactory performance, particularly when the coax cable is not grounded to the antenna mount, as discussed previously.

results

This antenna and preamplifier system has been used to receive the 10.2-kHz Omega signals on all eight stations including LaReunion Island, halfway around the world. In Ohio, I regularly receive twelve different Loran-C transmissions from the East Coast, Northeast, and Gulf Coast 100-kHz chains. At night, I observe Loran-C skywave signals from the West Coast chain over 4000 km (2500 miles) away. WWVB puts in a strong 150-μV signal in Ohio, but is often interfered with by MSF in England on the same 60-kHz frequency. Other signals noted are the time frequency standard stations in Switzerland on 75 kHz, Japan on 40 kHz, as well as numerous communications and military FSK-type signals in the 14 to 150 kHz range. The system works well in the 1750-meter Amateur band (160 kHz to 190 kHz), but I have not yet made any serious attempt at DX hunting. The most interesting DX received is on 15.625 kHz, part of the USSR Alpha vlf navigation system.

Harmonics of 60 Hz and TVI from harmonics of the 15.75-kHz horizontal oscillator in nearby TV sets are the most common interference observed in urban locations. In my receiving shack, the 60-Hz troubles usually do not start until the mercury vapor arc lights in front of my home start operating, and other 60-Hz uses increase during the prime evening hours.

In some installations, BCI from nearby transmitters can be a problem. These can usually be cured with an additional low-pass filter or trap inserted in the receiver input circuit, with better grounds on the antenna pole mounting, and with proper care in design of the receiver input circuits to minimize cross modulation problems. Common-mode 60-Hz pickup is sometimes a problem caused by combinations of poor ground connections at the antenna coupler box and the receiver location. If this cannot be cured by relocating the antenna, then another method is to use a balanced, twisted-pair, shielded transmission line. The unused half of the preamp output transformer and a similar balanced input transformer at the receiver with a 330-ohm current limiting resistor connected to the center tap can be used to reduce common mode pickup problems. In general, balanced transmission lines have not been used because it is difficult to obtain suitable lines and weatherproof fittings that will pass frequencies to 300 kHz without excessive expense. It is easier to cure these problems by antenna location or ground changes.

acknowledgments

The effort on vlf/low-frequency antenna couplers has been part of a study of low-cost methods for producing receiving systems for the general aviation community, sponsored by the NASA Langley Research Center. This paper is a direct result of that work. The help of student assistant Edwin Jones is appreciated in testing some of the most recent preamplifier systems.

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ham radio

connectors for CATV coax cable

How to beat the high cost of connectors for popular "1-inch" CATV coaxial cable

Large quantities of "l-inch" CATV coax cable have recently become available from various sources. The electrical characteristics of this cable make it attractive for Amateur vhf-uhf applications, particularly for long transmission lines. The cable is offered at modest cost, from tag ends to full-reel lengths of up to 732 meters (2400 feet).

The "l-inch" CATV cable, however, is not without its drawbacks, and this article suggests ways to circumvent these with ordinary construction practices using shop tools and readily available materials.

hardline "1-inch" CATV cable

The CATV cable is shown in **fig. 1. Table 1** gives physical and electrical properties. Perhaps the cable's major disadvantages are 1) its handling properties below room temperature, at which it has the ductility of gas pipe; 2) its relatively large bend radius; and, most of all, 3) the high cost of end connectors.

Fig. 2 shows, from left to right, an expensive commercially available end connector; an Amateur design in the raw; and the Amateur design in a moisture-proof final configuration. Relative costs for the two Amateur designs are quite low.

For the obvious reason of cost, and because of the relative bulk of the commercial connector, I evolved the Amateur design, which can be made with fairly common materials, and, to a large degree, common hand tools.

These were the other major considerations achieved in the Amateur design:

1. No electrolytic action between dissimilar metals

2. Resistance to moisture penetration

3. Minimum discontinuity bumps in line impedance introduced by the connector

materials and tools

Tables 2 and **3** show the materials and tools you'll need for making the connectors. **Fig.** 3 illustrates the raw parts needed for the suggested Amateur design, denoted A, B, and C, D, and G. **Fig.** 3 also suggests not only the assembly steps, but interrelated with the following text, features and mechanics of fabrication, treatment, and parts use to the point of final assembly, as shown in the connector at the right-hand side of **fig. 2.**

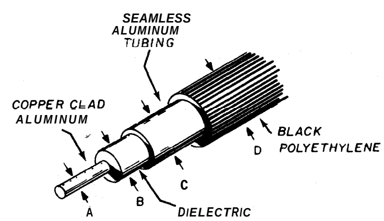

fig. 1. Makeup of typical "l-inch" CATV cable.

The tools should be available to the average builder or his friends, or they can be rented from local outlets.

fabrication

Steps for making your coax connector are given below. Refer to **fig.** 3 as you proceed.

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1. Using a tubing cutter, make a cut through the vinyl cable jacket and seamless-tubing shield about 15-25 cm (6-10 inches) from the cable end.* Then, using a fine-blade hacksaw, cut through the dielectric and inner conductor. Make the cut flush with the outer surface of the cable. Remove aluminum filings from the dielectric surface. E of **fig.** 3 shows what it should look like.

table 1. C A N "one-inch" coax cable characteristics.

C A N cable makeup (fig. 1)

part	material	nominal OD mm	(inch)
A-conductor	solid bare copper-clad aluminum	5.8	(0.227)
B-insulation	extruded foam polystyrene	22.6	(0.9)
C-shield	seamless aluminum tubing	25.4	(1.0)
D-jacket	extruded polyethylene (black)	28.4	(1.1)

physical properties

maximum pulling force	261 kg. (575 lbs.)
minimum bending radius	10 times cable diameter
nominal weight	641 kg./km. (430 lbs./1000 ft.)

electrical characteristics

maximum attenuation, 20C[1]		maximum attenuation, 68F[2]	
frequency MHz	dB/100 meters	frequency MHz	dB/100 feet
5	0.26	5	0.08
30	0.66	30	0.20
50	0.85	50	0.26
216	1.9	216	0.59
240	2.0	240	0.63
260	2.1	260	0.65
270	2.2	270	0.68
300	2.3	300	0.7

1. Attenuation varies ±2 per cent nominal per 10C variance in ambient temperature.
2. Attenuation varies ±1 per cent nominal per 10F variance in ambient temperature.

2. Find the precise center of the copper-pipe-cap outer face. Center punch the cap center, and drill a hole using a sharp no. 42 (2.4-mm) bit. Remove burrs from the pipe cap inside.

3. Force the pipe cap firmly over the square end face of the cable.

table 2. Materials for making the coax-cable connector.

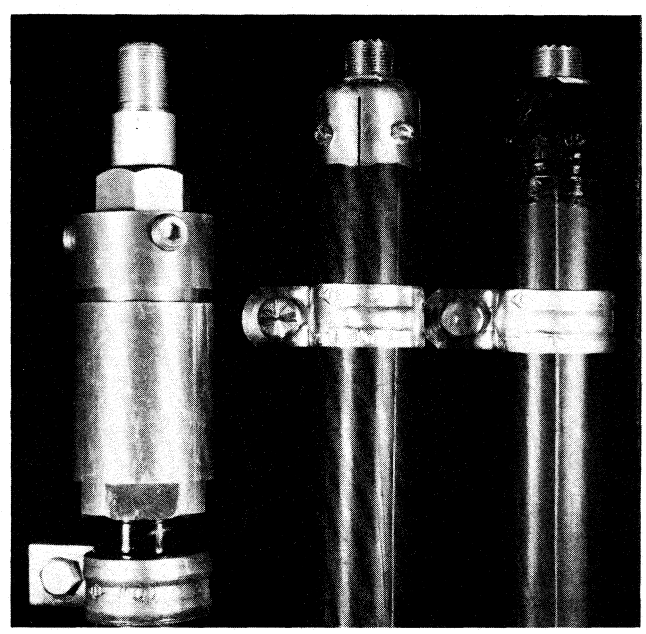
quantity	description
1	4-40 x 112-inch (M3 x 12.5 mm) BH or RH nickel-plated screw
1	83-378 bulkhead vhf female receptacle (Amphenol)
1	copper end cap for 1-inch (25.5-mm) rigid copper water tubing
4	6-32 x 114-inch (M3/5 x 6.5-mm) hex-head sheet-metal screws
1	bottle of Liquid Tape (General Cement)
1	end of CATV cable to attach to connector

4. Clamp the cable vertically in a vise. Using the pipe-cap center hole as a guide, carefully drill a no. 42 (2.4-mm) tap hole ***straight down*** to a depth of 1/2 inch (25.5 mm) into the copper clad center conductor of the cable. For quantity jobs, get a machinist friend to make a drill guide as shown in **fig. 4.†** Then tap the center conductor 4-40 (M3) screw and remove chips from the tap and hole after each turn of the tap.

5. Remove the hex nut and lockwasher from the bulkhead connector (Amphenol 83-879) and discard the lock washer. Mount the connector upright in a vise. Carefully drill down from the top, through the solder tip, with a no. 33 (2.9-mm) drill. Set the connector with its new short tip aside.

6. Reduce the head diameter of a 4-40 x 112-inch (M3 x 12.5-mm) screw to about the same outside diameter (see **A, fig. 3**) as the male tip of a PL-259 connector. This can be done by inserting the screw into a drill and rotating the screw head against a flat fine-

fig. 2. Standard commercial CATV-cable connector, left, and the Amateur design.



*The foam dielectric is not particularly hygroscopic, but some moisture will inevitably accumulate near the exposed end. Remove the section of contaminated cable before installing the connector.

†Use of the drill guide requires temporary cut and removal of about 6.5 mm (1/4 inch) of the black polyethylene jacket to allow the guide to slip down over the aluminum shield (**C** or **fig. 1**).

table 3. Tool requirements.

quantity	description
1	1-inch (25.5-mm) tubing cutter for copper or aluminum
1	fine-tooth hacksaw blade and franie
1	3-inch (77-mm) vise
1	118-inch (3-mm) wide fiat-blade screwdriver
1	fine-tooth flat file and emery cloth
1	150-watt soldering iron or Benzomatic torch
1	center punch
1	electric drill motor with 1/4 or 3/8 inch (6.5-mm or 9.5-mm) chuck
1	4-40 (M3) tap and wrench
1	518-inch (16-mm) OD socket punch
1	1/4-20 (M7) bolt, lockwasher, and nut
1	no. 12 SS compression hose clamp
1	high-speed drill bits, nos. 42, 33, 29, and 1/4 inch (2.4, 2.9, 3.5, 6.5 mm)
1	open-end wrench or socket for 112-inch (12.5-mm) bolt
1	piece of steel wool
1	corrosion-proof rosin-core solder, about 20 inches (50 cm) long.

mesh file. The final outside diameter should be 0.15 inch (3.8 mm).

7. Using another 4-40 (M3) screw to temporarily block the tapped hole of the center conductor, liberally coat and seal the face of the dielectric, shield, and black vinyl jacket of the cable with General Cement liquid tape or equivalent material.

8. Drill out the hole in the face of the pipe cap to 1/4-inch (6.5-mm). Using an upside down 114-inch (6.5-mm) bolt and nut to secure the cap in a vise, make four fine hacksaw cuts 90 degrees apart about 314 inch (19 mm) down from the lip of the cap. (See F, **fig. 3**). Between the slots drill four 0.09-inch (2.4-mm) holes about 1/4 inch (6.5 mm) back from the same lip about 90 degrees apart.

9. Remove the cap from the vise and the superfluous 1/4-inch (6.5-mm) hardware, Insert a 5/8-inch (16-mm) diameter chassis punch. Tighten the bolt on the punch. You should have a clean 5/8-inch (16-mm) hole in the copper cap face.

10. Burnish the cap surface with steel wool. Insert the body of the bulkhead receptacle from the inside of the cap. Tighten the flat nut. (See F, **fig. 3**).

11. With the torch and minimum flux, solder the flat nut onto the cap surface. Use care to fill the void on the flat, keyed side of the receptacle body.

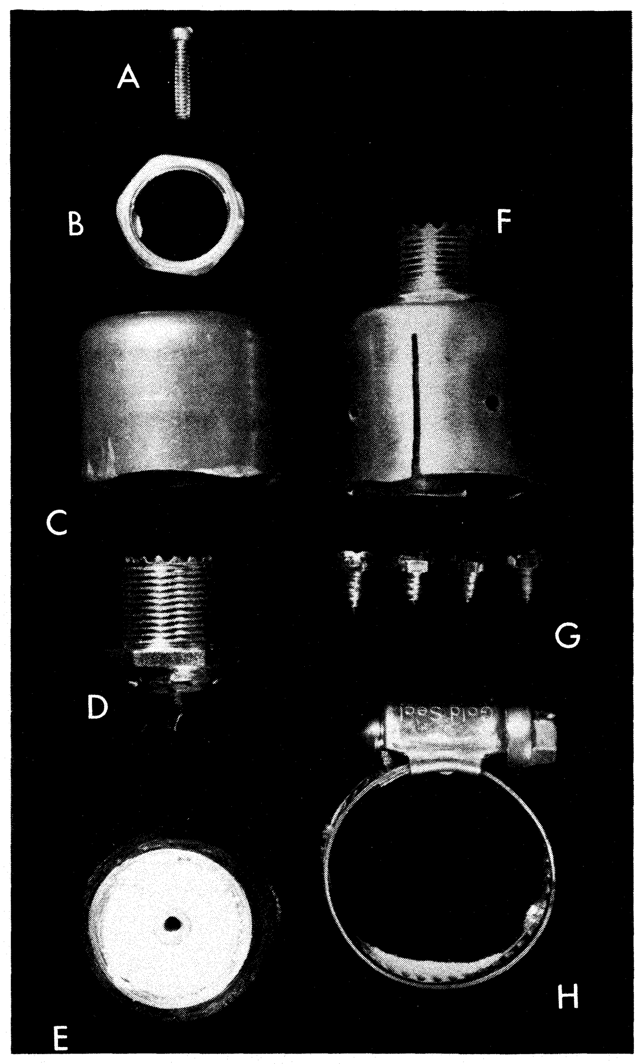
12. Insert the 4-40 (M3) screw, threads down, into the female flange fingers of the receptacle. Push the cap connector assembly down onto the face of the coax, then tighten the screw into the tapped center conductor. Slip the hose clamp over the cap to the cap lip.

13. *Very carefully* drill through the guide holes in a) the copper cap; b) through the black jacket; and c) just deep enough to puncture the outer seamless aluminum jacket of the coax cable. Use the no. 42 (2.4-mm) drill

14. Carefully enlarge the four copper cap holes with a no. 29 (3.5-mm) drill just enough to allow the 6-32 x 114-inch (M3/5 x 6.5-mm) screw shanks (see **G, fig. 3**) to pass through the cap material.

15. Install the sheet-metal screws (four each) through the copper cap, vinyi, and into the cable outer aluminum jacket. Make sure the screws are hand-tool tight, no more.

fig. 3. Raw materials needed for the Amateur design of a coax connector for the hardline CATV cable. Assembly steps are shown for the final connector (extreme right, fig. 2).



16. Remove the hose clamp. Check for short circuits or high leakage between inner and outer conductor. If all is well:

17. Coat the cap body, hex-head screws, and junction of the vinyl jacket with G.E. "Liquid Tape," "Tool Dip," heat-sensitive tape, or 1-1/2 inch (38 mm) heat-shrink tubing treated with the torch. Make sure the connector threads are nice and clean.

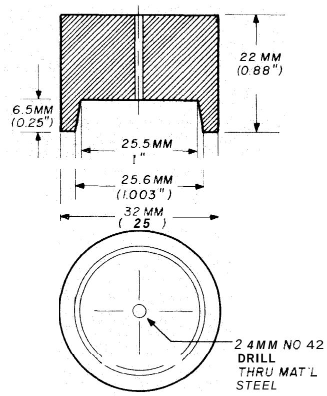


fig 4 Drill guide for quantity production. The guide can be made by a machinist (See step 4 in the fabrication procedure.)

This completes the assembly procedure for the Amateur-design CATV-cable connector.

final notes

Metal-to-metal contacts will be reasonably non-electrolytic; that is, silver-plated connector (inner), to stainless-steel 4-40 (M3) screws, to aluminum inner conductor. Also this applies to the stainless-steel sheet-metal screws, from copper cap to aluminum jacket.

The nominal impedance of the completed connector is about 65 ohms. It will show a small impedance bump at about 146 MHz. For matching considerations using 50- or 75-ohm sources, see references 1-5.

acknowledgments

I wish to express my thanks to the following Amateurs for help on this project: W1SCS, W1WME, K2TJZ, K1MOQ, W1TXK, WA1YRQ, and W1IBF.

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More **Details?** CHECK—OFF **Page** 110

end-of-transmission "K" generator

An easy-to-build circuit for sending the Morse character "K" at the end of a transmission

Ever since the **Apollo** moon shots, Amateurs have been aware of the advantage of a tone at the end of a transmission to indicate that the "over" has finished. Under poor conditions this tone indicates unequivocably to the other station that it's his turn to transmit. There's a growing fashion now in Britain to use not just a single tone but a "dah-di-dah" sequence, making the Morse character "K", meaning "over" in telegraphy. The circuit described here will perform this function using only two standard ICs and a handful of other components.

the "K" generator

The heart of the unit is U2, an IC type CD4017 (**fig. 1**). This device is a decade counter with decoded outputs, which means it has ten outputs: zero, one . . . nine. Initially the zero output is high and all the others are low. When a clock pulse is applied to the clock input, the zero output goes low and the one output goes high. After the next clock pulse, only the two output will be high. This sequence continues until the nine output is high; after that, the zero output goes high again and the sequence repeats.

Also provided on U2 are a reset and a clock inhibit input. When a high is applied to the former, the counter goes straight to the state of having the zero output high. Applying a high to the clock inhibit input freezes the counter in its current state, and further clock pulses have no effect until the clock inhibit input goes low again.

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In this application a free-running clock is connected permanently to the clock input and the clock inhibit is connected to the zero output. Consider the effect of this if the counter is in the zero state: the zero output is high; therefore the clock-inhibit input is high. Thus clock pulses have no effect on the counter and it remains in the zero state, seemingly permanently. However if the clock inhibit is forced briefly low, the counterwill cycle through its outputs, one to nine. When it overflows to zero, the clock inhibit is again high and no further counting occurs.

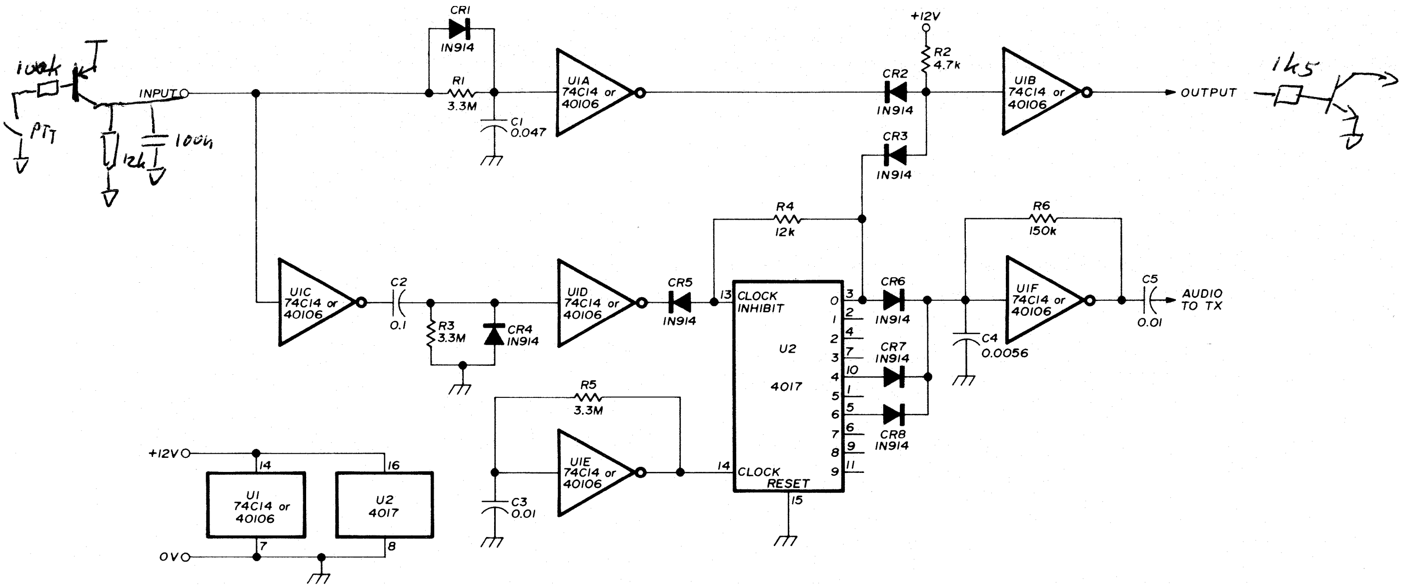

fig. 1. Schematic of the "K" generator. Circuit is designed around U2, a CD4017 decade counter with outputs between zero and nine. Also included in U2 are a clock input, clock-inhibit input, and reset.

We thus have a method of making the counter cycle once through its complete set of outputs. Not counting the zero waiting state, nine counter states remain, and it so happens that the sequence for "K" in Morse is nine dot units long (on-on-on-off-on-off-on-on-on). Arranging the outputs correctly we can make the unit send "K" when wanted.

The other chip used in this project is U1, a 74C14 (40106), which contains six inverting Schmitt triggers. By judicious use of diodes, these triggers are sufficient to complete the required logic functions. The first building block to be constructed from a Schmitt inverter is a relaxation oscillator **(fig. 2)**. Operation is as follows: the capacitor charges through the resistor until the trigger input reaches its upper triggering voltage. The output then goes low, and the capacitor discharges through the resistor until the lower triggering voltage is reached. Then the output goes high, and the capacitor starts to charge again. The output is thus a square wave.

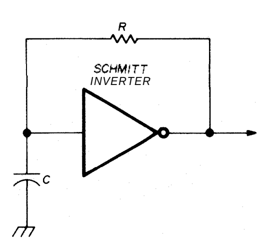

fig. 2. An oscillator may be easily made during a Schmitt inverter IC. This example is seen in U1E, the clock driver for U2 (fig. 1).

In the circuit of **fig. 1,** two of the Schmitt inverters are used as oscillators. U1E is a free-running oscillator, which clocks the counter. U1F is a gated oscillator working in the audio range, which generates the tone. Diode gating is arranged so that the tone is inhibited if any of the diode inputs are high. These diodes are connected to the zero, four, and six outputs of U2 so that the tone is off when the counter is in its resting state and also during the gaps between the dots and dashes.

The input to the unit is in two paths. The first triggers the counter at the end of the transmission. The input signal is inverted and fed to a differentiator C2, R3 **(fig. 1)**. The pulse generated on the leading edge of the input signal is suppressed by CR4, but that generated on the trailing edge is fed to U1D where it's squared. For the pulse duration, the counter clock inhibit input is pulled low through CR5.

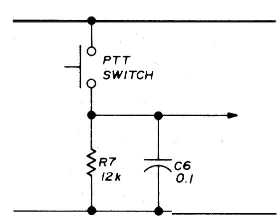

fig. 3. Switch debouncing and interfacing circuit.

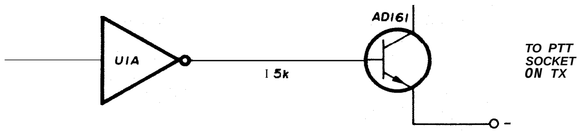

fig. 4. Circuit for interfacing the "K" generator to a transmitter.

Because R4 is relatively high in value, this action will occur regardless of the state of the zero output. Once the clock inhibit input has been pulled low, the counter starts and the zero output goes low, which forces the clock inhibit input low regardless of the U1D output state.

The input voltage is aiso fed to integrating network R1, €1. On the leading edge, C1 charges up quickly through CR1; but on the trailing edge, a delay occurs while C1 discharges through R1. The voltage on C1 is squared and inverted by U1A and fed to the diode AND gate CR2, CR3. Its other input is the counter zero output. If either of these is low, the output is low and U1B output is thus high, feeding the transmit/receive switching.

Thus, as the press-to-talk switch is released, U1A output will be low because of the delaying action of C1. A moment later, the counter zero output will go low, which keeps the transmitter switched on, even though C1 has discharged. Only when the counter has cycled completely will the zero output go high again. Now both inputs to the diode AND gate will be high and the transmitter is switched off.

interface circuits

This completes the description of the logic part of the circuit, but the unit will, at this stage, only work with CMOS input and output levels. Clearly some kind of interfacing is needed in any real situation. For the input side, the circuit of **fig.** 3 will suppress any contact bounce generated by the press-to-talk switch and provide the appropriate input levels to the unit. Because of the time constant of R7, C6 the spikes generated by the contact bounce aren't fed to the input. The value of R7 is smaller than either that of R1 or the on resistance of CR1, so the charging and discharging of C1 is not affected. Clearly, a normally closed switch could be used by interchanging the positions of the switch and R7.

For the output interfacing, the situation is rather more complicated by the variety of circuits with which the unit could be required to operate. I used the arrangement shown in **fig. 4.** The transistor was an old germanium power device, but any variety of NPN transistor with adequate voltage and current ratings will probably work. In my home-brew transmitter, one of the PTT lines was connected to the supply rail so that the transistor operated in common-collector mode. The same circuit will operate just as well in common-emitter mode where one of the PTT lines is at zero potential.

Where the transmitter switching arrangement is unusual, or the unit is to be used with a variety of transmitters with differing switching methods, it's best to isolate the unit by a relay. **Fig.** 5 shows a circuit for driving a reed relay; again the transistor type isn't critical. It is, however, important to make sure that the relay is adequately rated for the current drawn through the PTT line since, if overloaded, the contacts tend to weld together and leave the transmitter permanently on — to the potential embarrassment of the operator!

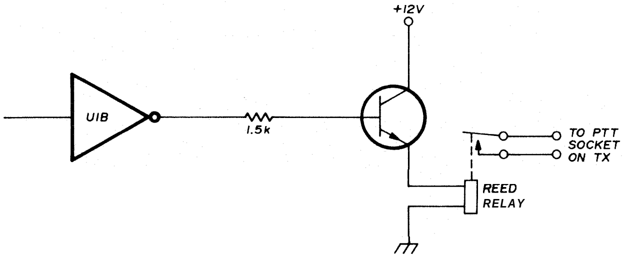

fig. 5. Alternative interfacing arrangement where circuit isolation is essential.

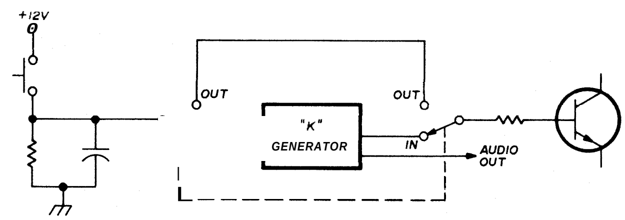

fig. 6. Using a dpdt switch, the "K" generator may be removed from the circuit when desired. The trigger is no longer connected to the circuit so that the audio signal can be retained.

construction

With circuits of this type layout isn't critical. Construction can be in just about any form desired. The circuit can then be either hidden in a spare corner of the rig or built into a separate box. Where the relay driver is not used, the current drain is very low, so a separate battery supply is quite feasible if a suitable voltage isn't available in the transmitter. If you go this route, a common connection must be made to the transmitter chassis, probably through the PTT line, to allow the switching transistor bias current to flow.

Finally, for local contacts, the "K" can be suppressed as shown in **fig. 6** if you wish. Operation is then as normal.

Good luck with your new and improved DX potential from G8KGV, Solihull, England. Dah-di-dah.

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the **weekender**

capacitance measurements with a frequency counter

A digital capacitance meter has been on my shopping list for some time, but it's been difficult to justify the expense of the single-function instruments available. An article in *QST*[1] followed by investigation of the literature on the NE555 led to the design of a frequency counter attachment that allows capacitance to be read from the counter display. Using a 7-digit counter, capacitance values from 1 pF to 1 μF may be read with an accuracy of about 2 per cent ± 2 pF. A range switch isn't required.

design

The NE555, when used as a one-shot, produces a pulse of width

$$T = ARC \tag{1}$$

where k is inherent in the 555, R is the charging resistor, and C is the capacitance being measured. By ANDing an oscillator output with this pulse, a burst of pulses is produced each time the 555 is triggered. The pulse frequency in the burst is that of the oscillator, while the number of pulses in the burst is determined by the length of the pulse produced by the 555. The value of R may be adjusted so that when a 1-pF capacitor is measured, the 555 output causes exactly one oscillator pulse to appear in the burst. Increasing the capacitor to 100 pF will make the 555 output 100 times longer and allow 100 oscillator pulses in the burst. Counting the pulses in one burst will then indicate the capacitance in pF.

To ensure that exactly one burst occurs per frequency-counter gate time, the 555 is triggered by the opening of the frequency counter gate as shown in the block diagram, **fig. 1.**

The effects of stray capacitance are compensated for by providing a third input to the AND gate which forms the burst, nulling the first part of the pulse from the 555. Typical waveforms are shown in **fig.** 2.

circuit description

The design is implemented as shown in **fig.** 3 using a 4009A and a NE555. The functions are shown in the same relative positions in the block diagram and schematic for clarity.

The crystal oscillator is standard and provides a stable square-wave output. The crystal frequency isn't critical but should be greater than 1 MHz to allow measuring 1-μF capacitors using a I-second

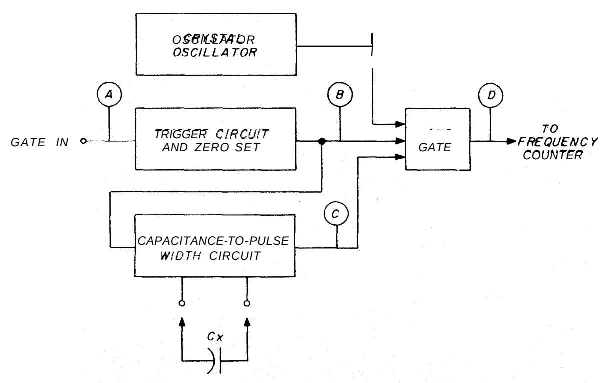

fig. 1. Block diagram of the capacitance measurement attachment for a frequency counter.

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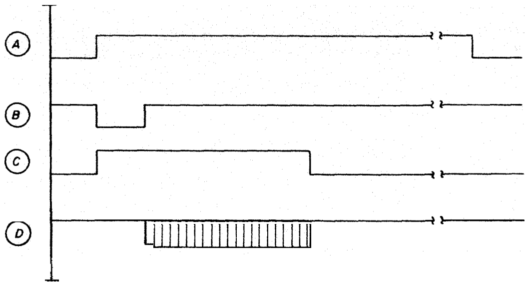

fig. 2. Typical waveforms. Letters at left key the waveforms to both the block diagram and the schematic.

gate time. CMOS speed is restricted when operating from 5 volts, so frequencies above 2 MHz may not work too well.

The trigger circuit consists of a one-shot started by the leading edge of the gate pulse. The output pulse width is adjustable to allow compensation for the effects of stray capacitance, as described earlier. The one-shot is followed by an inverter, which cleans up the pulse shape and provides the proper polarity to trigger the 555 and drive the AND gate. The capacitance-to-pulse width conversion is done by the trusty NE555 in a standard one-shot configuration.

The AND gate is home-grown CMOS/diode logic, which works well and avoids the need for a third IC. The voltage divider on the output reduces amplitude and, more important, the output impedance, thus avoiding stray pickup from the oscillator.

construction and calibration

Perf board construction is adequate if the ICs are bypassed with 0.01-μF capacitors directly at the V_{cc} pins and if the output voltage divider is located a short distance from the crystal.

A metal box should be used to house the unit, since it's sensitive to *60* Hz pickup. This phenomenon is evidenced by a slow variation of about one-half per cent in the reading as the counter gate beats with the line frequency. A variation of over 2 per cent occurs without an enclosure.

The 100-pF capacitor at the oscillator output affects starting and should be adjusted for reliable

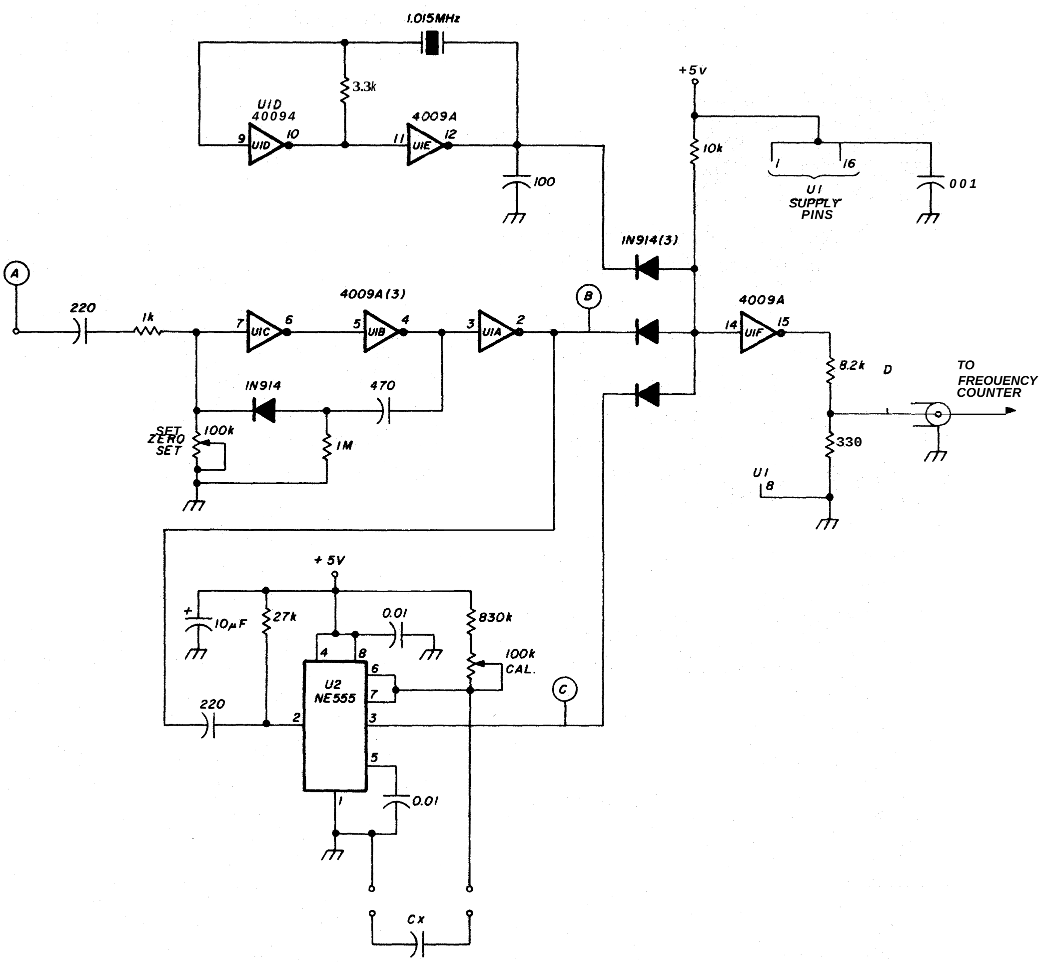

fig. 3. Schematic diagram. Relative position of functional sections matches the block diagram.

operation. The 830k resistor in series with the calibration pot is actually two resistors in series to get a nonstandard value. Using a higher-frequency crystal may require this value to be reduced for calibration.

The frequency-counter gate signal must be made accessibie to allow triggering the unit. I installed two pin jacks in the back of my counter to provide both the gate and +5 volts. I used a 200-pF series capacitor to bring the gate line out, to prevenr damage to the counter in the event that the line is inadvertently shorted.

Calibration is best done using a one per cent capacitor with a value of several thousand pF as a standard. Backing this, several silver-mica capacitors in parallel will probably be adequate, since the deviations from the marked values should average out.

Begin calibration by adjusting the ZERO SET control to the maximum resistance setting. Then, without a capacitor in the measurement socket, adjust ZERO SET until a reading is obtained on the counter. Back off slowly until a zero reading is obtained again. Connect your standard capacitor and adjust the CAL control to obtain the correct reading. Then repeat the procedure to correct for adjustment interaction.

If your counter has 0.1-second and 1-second gate times, be sure to use the 1-second gate when measuring capacitors above 0.01 μF.

variations on the theme

It's possible to add a second range by paralleling the charging resistance (830k + CAL) with values that are a factor of 1000 smaller to allow reading capacitors between 0.001 μF and 1000 μF. Note, however, that capacitor leakage causes artificially high readings, so electrolytics with an internal resistance of 1 megohm or less can't be read accurately.

Much of the circuit is also applicable to using the 555 as a direct-reading ohmmeter covering 1 ohm to 1 megohm in a single range.

concluding remarks

A frequency counter attachment has been described that measures capacitance. My hope is that frequency counters will soon be replaced by multifunction instruments incorporating voltage, current, resistance, capacitance, and frequency-measurement capability, much as the voltmeter was replaced by the VOM. Until that time, the versatility of your frequency counter may be increased with devices such as that described here.

reference

1. D.A. Blakeslee, "An Inexpensive Capacitance Meter," QST, December, 1978.

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More Details? CHECK–OFF Page 110

cross-guide coupler for 10-GHz

Design and construction details of the cross-guide coupler, an easy to build directional coupler for 10 GHz

When it becomes necessary to sample a portion of the transmitted power from a signal source or transmitter, a directional coupler is normally used. A great variety of commercial couplers are available, but they are beyond the means of most Amateurs. Once in a while, surplus dealers have sales of X-band equipment, but these pieces can still be quite expensive. Sometimes at a ham auction or flea market pieces can be purchased quite reasonably. However, directional couplers are hard to come by, so building one is an alternate approach.

The simplest type of directional coupler is the cross-guide design shown in **fig. 1.** Many commercial cross-guide designs have cruciform hole coupling which make the coupling ratio and directivity frequency independent.[1] Broadwall directional couplers also have broadband performance, but it is at the expense of long length and precision machining. The round hole cross-guide coupler has the advantage of short length and good performance over a limited bandwidth. It is the easiest form to design and build.

design

The practical range of coupling in a round hole cross-guide coupler varies between 20 and 45 dB down from the incident power. **Fig.** 2 shows the hole arrangement relative to the coupler ports; also shown is a graph of how the coupling varies as ratio of hole diameter to guide width.[2,3]

The following is an example of a design procedure to build a cross-guide coupler using the graph in **fig. 2.** First, the center of the 10.0 to 10.5 GHz Amateur band is 10.25 GHz. Since the bandwidth of the round hole cross-guide coupler is approximately 10 per cent, the coupling coefficient should not change more than a few per cent in the 10.0 to 10.5 GHz band. The second step is to decide on the desired coupling ratio. Sources such as reflex klystrons and Gunn diodes, which have an output power in the 10 to 20 mW range, require a sampling coupler to have a coupling ratio of 25 dB or greater. This will keep the power level into the detector in the square-law region. Referring to the graph in **fig. 2,** the D/a ratio for 26 dB is 0.335, where a is the broad dimension of the waveguide. For WR-90 waveguide (*i.e.,* X-band guide), the broad dimension, a, is 22.86 mm (0.900 inch). The large hole diameter is:

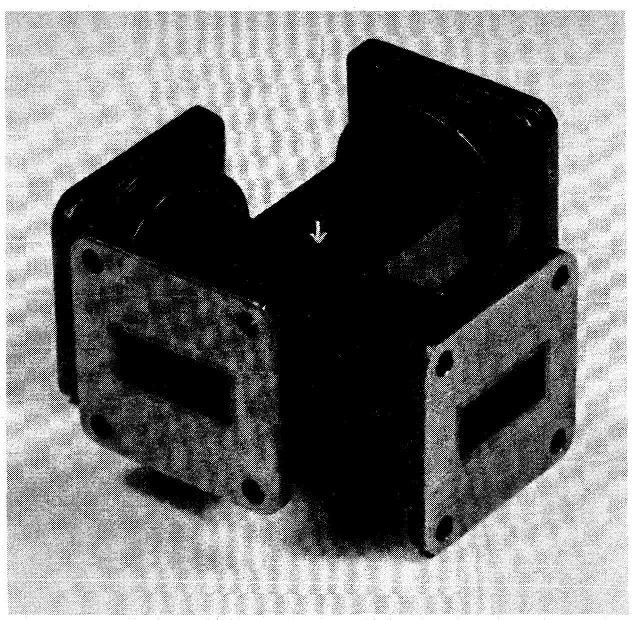
$$D = 0.335\,(22.86)\ \text{mm}\ [0.335(0.9)\ \text{inch}]$$

$$D = 7.66\ \text{mm}\ (0.3015\ \text{inch})$$

The smaller hole is $2D/3$, so:

$$2D/3 = 2(7.66)/3\text{mm}\ [2(0.3015)/3\ \text{inches}]$$
$$= 5.11\ \text{mm}\ (0.201\ \text{inch})$$

Finished directional coupler.



By Richard A. Bitzer, WB2ZKW, 3 Ray Street, Milltown, New Jersey 08850

Now that the hole diameters are known, their distance from the center line of each guide must be $\lambda_g/8$. To calculate λ_g, the quantities λ_o and λ_C must be known.

$$\lambda_o = \frac{C}{f}$$

$$\lambda_o = \frac{3 \times 10^8 m/sec}{10.25 \times 10^9 Hz} \quad \left(\frac{1.18 \times 10^{10} in/sec}{10.25 \times 10^9 Hz}\right)$$

$$= 0.02927 \; meter \; (1.152 \; inches)$$

$$= 29.27 \; mm$$

where λ_o = free space wavelength
C = velocity of light in free space
f = frequency of operation

For TE_{10} propagation mode down WR-90 waveguide, the cut-off wavelength is:

$$\lambda_C = 2a$$
$$= 2\,(22.86)\; mm\; [2(0.9)\; inches]$$
$$= 45.72\; mm\; (1.8\; inches)$$

The guide wavelength, λ_g, for TE_{10} mode is:

$$\lambda_g = \frac{\lambda_o}{\sqrt{1 - \left(\frac{\lambda_o}{\lambda_C}\right)^2}}$$

$$= \frac{29.27}{\sqrt{1 - \left(\frac{29.27}{45.72}\right)^2}} mm \quad \frac{}{\sqrt{1 - \left(\frac{1.152}{1.800}\right)^2}} inches$$

$$= \frac{29.27}{0.768} mm \quad \left(\frac{1.152}{0.768} inches\right)$$

$$= 38.10 \; mm \; (1.5 \; inches)$$

The hoie centers are $\lambda_g/8$ from the center lines of both guides, therefore $\lambda_g/8$ is 4.76 mm (0.0187 inch).

The above example can be repeated for any desired coupling ratio between 20 dB and 45 dB, useful over the bandwidth of the 10-GHz Amateur band.

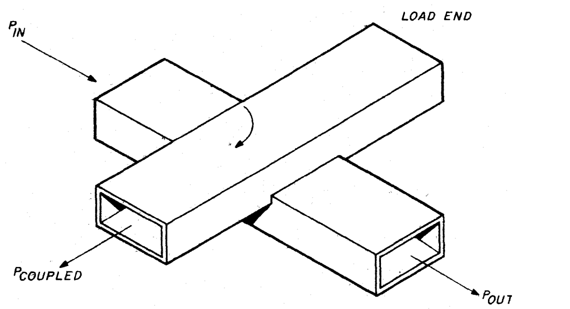


fig. 1. Diagram of the cross-guide coupler configuration. Note that the one end of the coupled arm can be terminated.

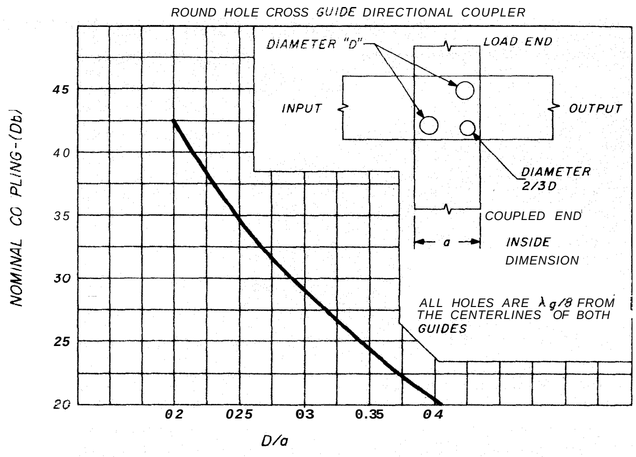


fig. 2. Graph of the relationship between the coupling factor and the physical dimensions of the waveguide and coupling holes.

construction

The directional coupler uses two 76.2 mm (3-inch) pieces of WR-90 waveguide and four UG-39/U cover flanges. **Fig.** 3 shows an exploded view of the coupler. From a length of waveguide, cut two pieces 76.2 mm (3-inches) long and file the ends square and smooth. On piece number 1, scribe and drill the hole locations according to the design data. File off all burrs, since it is important not to have any obstruction inside the waveguide. With piece number 2, a slot is to be cut so that this piece fits centrally over the holes in piece number 1. One way to do this is to cut a piece of wood to fit the inside dimensions of the guide and insert it into the guide. Scribe the slot dimensions on the guide and cut on the inside of the scribe marks with a fine hack saw. Do not cut too deep, for the depth of the cut is the wall thickness dimension. The sawdust from the wood'should tell you when to stop. The two pieces are placed over one another as shown in **fig.** 3 and soldered together. Use a torch rather than a heavy duty soldering gun or iron. Be careful not to let any solder run into either guide. If it does, it can be carefully filed out.

The next step is to solder the flanges onto the waveguide sections. Use a piece of flat aluminum sheet to work on. Place the flange face down. Wrap the guide assembly with a damp cloth to keep the two pieces from coming unsoldered while soldering on the flange. Place the waveguide into the flange, supporting the waveguide while soldering. Repeat this step for the remaining flanges. An alternative procedure in soldering the pieces together is to use a higher temperature solder for the cross pieces and a lower temperature solder for the flanges. To complete the job, use emery cloth to smooth finish the flange faces. For appearance, the coupler can be painted and labeled, giving a commercial looking appearance.

results

The cross-guide coupler was designed to have the coupled power down 26 dB from the incident power. The measured value of the coupled arm is – 26.2 dB. The directivity was measured to be 22.6 dB, about

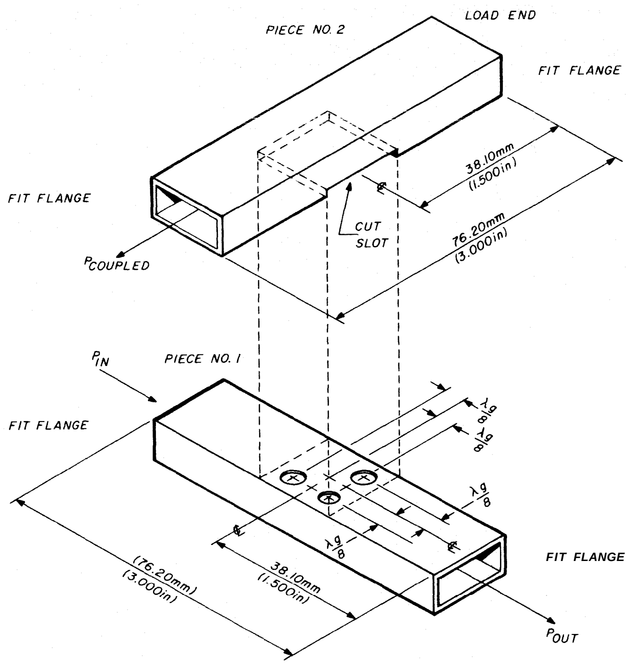


fig. 3. Assembly view of the cross-guide coupler.

average for cross-guide couplers. Referring to **fig. 1,** directivity is the ratio of the power down the coupled arm to that coupled into the load arm from the source. The main arm VSWR is 1.03 to 1, and the insertion loss is 0.05 dB. The coupler was used to sample power from a 10-GHz signal source test bench set up. For this application, the decoupled, or load end, arm has to be terminated with a load, which can be easily made by using a 76.2-mm 13-inch) long guide, flange on one end and a copper sheet across the opposite end. Inside, glue a wedge of 3.16-mm (1/8-inch) thick Masonite coated with Aquadag across the center of the narrow dimension of the guide. The VSWR of such a load measured 1.03:1.

references

1. ***A. F. Harvey, Microwave Engineering, Academic Press, London,*** 1963, ***page*** 107.
2. D. G. ***Fink, Electronic Engineer's Handbook, First Edition, McGraw-Hill,*** New ***York,*** 1975, ***section*** 9-7.
3. ***T. S. Saad, Microwave Engineer's Handbook, Artech House,*** 1971, ***volume 2, section*** 1.

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More Details? CHECK—OFF Page 110

IGITAL TECHNIQUES

gate arrays for pattern generation

Gate array structuring has been the subject of many papers, and a number of methods have been devised to simplify design. All the methods have one thing in common: pattern recognition by examination. This is true whether you have a static or dynamic pattern.

All large gate structures take a bit of time to design regardless of the examination method. Simple methods can be used in most Amateur designs. The example for this part of the series is a digitial **CQ** generator for perfect **CW** keying of those two letters.

organization

The basic time period in **CW** is dot length. Dashes are three dot lengths. Since the letters **CQ** are familiar, spacing can be shortened to three dot lengths with five dot lengths between groups. (It's assumed that a letter group will be repeated a selected number of times.)

Dot length will be the clock for a binary counter that times the pattern. Each *CQ* letter group requires eighteen clocks for key down, fourteen for key up. The total is thirty two, and a five-stage binary counter (2^5) will take care of timing.

Timing generator lines are labeled A through E (**fig. 1**), A being the least-significant bit and the fastest. The first step is to tabulate the timing-generator states and fit them with the desired pattern. Table **1** shows this.

Both letters are nearly the same length. The E timing line can be used as a letter selector. This simplifies state examination to only four timing lines. Note that the first two states in table **1** are not used

The timing generator has not been specified, but some assumptions are made. All zeroes will be in a reset condition and must be key up. A binary 00001 may or may not have the same length from all zero as between other states; this is determined from the case where the timing generator clock is not synchronous with manual pattern start.

pattern breakdown

Table **1** indicates that six states of the first four lines are common to both letters. This simplifies gating, because timing line E doesn't have to be considered for those six states. Only two states are exclusive for letter **C;** four are exclusive for letter **Q.**

Common letter states of lines A through D are all even decimally. A will be low for all six. Concentration now narrows to B, C, and D lines for the common states, which allows a detailed breakdown as shown in table **2.**

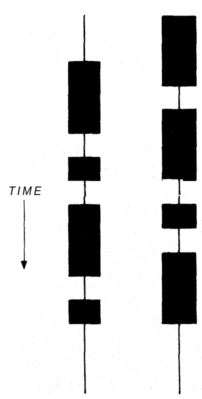
Common states can be reduced to three: CB, $\bar{D}$C, and D$\bar{B}$. The X is a don't-care bit state; that bit can be either 1 or 0. Some further examination of don't-cares will show three other combinations: DB, CB, and D$\bar{C}$.

Either of the three state combinations can be used. It's also possible to set up inhibits from DCB and DCB. The choice depends on the exclusive letter states. Only letter **Q** seems to have don't-ca es, with combinations of **$\bar{D}\bar{C}\bar{B}$** or DBA.

table **1.** Patterns and truth table for digital **CQ** generator.

TABLE I

LETTER C Q



D C B A	DECIMAL STATE	DESIRED STATES BOTH	E=0	E=1
0 0 0 0	0			0
0 0 0 1	1			1
0 0 1 0	2	2		
0 0 1 1	3		3	
0 1 0 0	4	4		
0 1 0 1	5			5
0 1 1 0	6	6		
0 1 1 1	7			
1 0 0 0	8	8		
1 0 0 1	9		9	
1 0 1 0	10	10		
1 0 1 1	11			11
1 1 0 0	12	12		
1 1 0 1	13			
1 1 1 0	14			
1 1 1 1	15			

State combinations should now be tried in a gate array. Remember that inverted, or NOT, bits must have inverters in the array, because the AND operation is possible only with high states. These could

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also come from the timing generator if inverted states are present.

a gate array pattern decoder

Fig. 1 shows one configuration. Common letter states are generated by using the inhibit states from **table** 2 and creating them in G10 and G11. NAND-gate outputs will be low when all inputs are high (the *Nand Rule*).[1] G10 and G11 will thus inhibit G12 on states $\bar{D}\bar{C}\bar{B}$ and DCB; G12 is enabled on the other six states.

Choice of inhibit gating allows using $\bar{D}\bar{C}\bar{B}$ for letter Q. An inverter restores the high state for AND operations. State CBA is common to both letters, as is BA. Use of common partial states allows a reduction in total package count, although it may appear to use more gates. Packages have quadruple 2-input gates and dual 4-input gates; it may be that use of more 2-input gates will result in fewer packages.

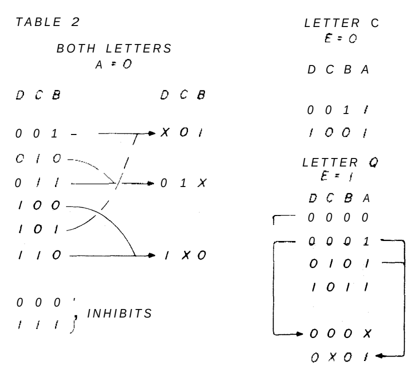
G14 and G15 AND the two states for letter *C*. G19 ANDs the DBA don't-care state from **table 2,** while G20 and G21 AND the remaining two states. G16 and G22 are ORs. G17 and G23 AND each state combination with E and $\bar{E}$ respectively. G24 ORs everything, and key down occurs when the output is high.

Total package count for the circuit of **fig.** 1 is six: one hex inverter, two dual 2-input NANDs, and three triple 3-input NANDs (two gates connected as inverters). This isn't very economical or simple in layout — two other digital devices can help simplify things.

decoders and multiplexers

A decoder (sometimes called demuitipiexer) will provide a single low output on one of two, four, eight, or sixteen output pins depending on one, two,

table 2. Detailed breakdown of pattern states for common and exclusive conditions.



three, or four data select lines. It can be thought of as a rotary switch with a grounded arm and all stator contacts as the outputs. Binary-data-select input will determine switch position.

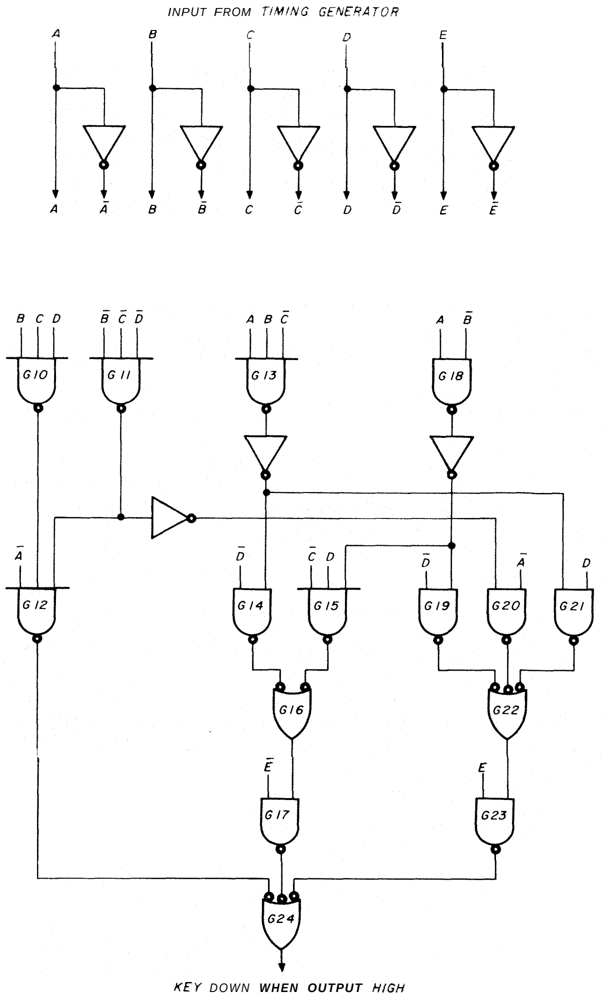


fig. 1. Gate array decoder for *CQ* generator using NAND gates.

A multiplexer will select one of two, four, eight, or sixteen inputs based on binary-data-select lines and transfer this information to a single output. The multiplexer can also be thought of as a switch: all inputs go to stator contacts, and the arm is the output. Position is determined by data-select control inputs.

a simple pattern decoder

Fig. 2 applies a 74L154 TTL decoder package with other low-power TTL NAND gates to the pattern-decoding task. The 74L154 has sixteen outputs active low and fits the sixteen states of **table** 1. Its outputs can be ORed as required.

G30 ORs the two exclusive states for letter C, while G31 ANDs them with $\bar{E}$. G35 is connected as an inverter to generate $\bar{E}$ high. G32 ORs the four exclusive letter *Q* states, ANDing them with E in G33. G34 ORs the six common letter states plus the two ANDed exclusive letter states.

U1 has two chip select pins marked G_1 and G_2.

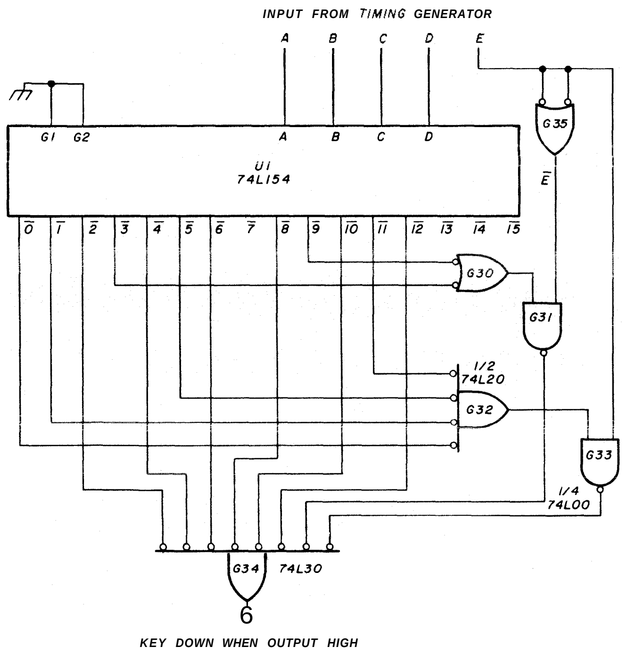


fig. 2. Simpler decoder for digital *CQ* generator.

These pins are used in arrays of decoders for enabling only one out of many. This circuit has only one chip, so it's enabled all the time by tying both pins low.

simplest decoder

This circuit uses a 74150 TTL multiplexer, or mux, and a single inverter as shown in **fig.** 3. The switch analogy describes operation. Timing lines A through D switch the inputs from E0 through E15 to output W. Inputs are high for key down; low for key up.

Common states of key down are wired to. V_{CC}, while unused states (7, 13, 14, 15) are wired to

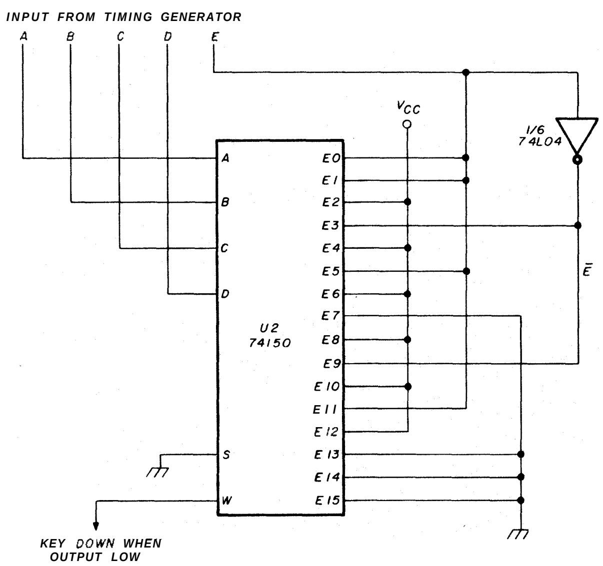


fig. 3. Simplest decoder for the digital *CQ* generator.

ground; if unconnected in TTL, those inputs will go high. Exclusive letter states are obtained from E directly or from the inverter for E.

The S pin is a strobe, similar to a chip select. It is also used for mux array selection. Note that output W is active iow from the inversion bubble **(fig. 3)**.

timing and polyphase clocks

Dot rate is *quite* slow in the *CQ* generator, so small differencein delay throughan array won't be noticed. If the pattern were faster, say for a TV sync generator, the propagation delay differences would be very apparent.

The gate array has propagation delays of three to six gates depending on the path. The decoder delay is slightly shorter but variable. The mux circuit of ***fig.***

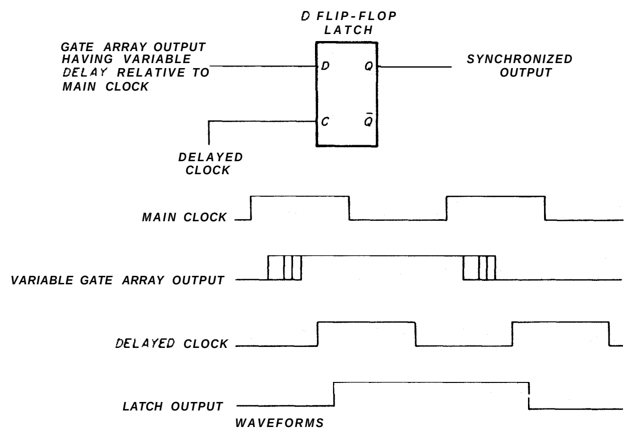


fig. 4. Variable gate array synchronization with delayed clock.

3 can be considered even; it's dependent on the internal structure of U2, which is usually symmetrical.

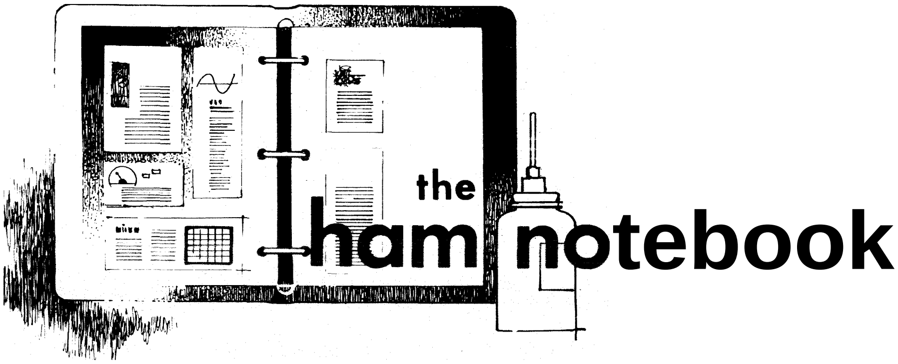
A way to even up differences in delay (some may be from the timing generator) is to use a delayed phase clock line and a D flip-flop for each output. This is shown in **fig. 4** and could be used anywhere in a large array.

Very large arrays can use several clock phases or a polyphase clock generator. Five or more phases are not unusual. In the polyphase clock, the different phases have even increments of time delay in the clock period. Each D flip-flop is a latch for a particular line and is clocked by a phase with a differential equal to the maximum of all gate arrays plus setup time of the D flip-flop itself. Delays may differ in the array, but the latch output is synchronized by the delayed clock.

reference

1. Leonard Anderson, "Digital Techniques: Basic Rules and Gates," *ham radio,* January, 1979, page 77.

ham radio



Drake TR-22C sensitivity improvement

A simple modification to the TR-22C has resulted in a significant improvement to the receiver sensitivity. The modification consists of the removal of the built-in telescopic antenna. The antenna is secured to a small plastic block by a single machine screw. To remove the antenna, simply unscrew the antenna from the plastic block and remove it from the transceiver chassis, along with the protective vinyl sleeving covering the antenna. Leaving the plastic block and coaxial cable intact will permit easy reinstallation of the antenna and sleeving should "over-the-shoulder" operation be desired at a later time.

Following removal of the antenna, perform the rf-alignment procedure as follows:

1. Connect a signal generator set at 146.52 MHz to the rear panel antenna connector. Set the level for approximately half-scale deflection on the signal strength meter. I used a friend's signal on 146.52 MHz in lieu of the signal generator.

2. Adjust Lsl, Ls2, Ls4, Ls9, Ls10, and TCt3 for a maximum signal strength meter reading. Tct3 is located on the transmitter circuit board. The individual trimmers should be peaked several times to ensure optimum performance.

The results of this simple modification have been quite gratifying. Several stations heard prior to the removal of the telescoping antenna were usually in the noise level, if heard at all. After removal of the antenna, these same stations were either full quieting or well above the noise level. It appears that the 0.059-wavelength open-circuited line (*i.e.*, the collapsed telescopic antenna) introduces enough capacitance to the receiver front end to detune it, thereby lowering the sensitivity.

Since I do not have the appropriate test instruments but wanted to verify the much improved receiver performance, I repeated the following tests several times: I contacted several stations using local repeaters or direct simplex operation with both my 5/8-wavelength mobile whip and my J-Pole base station antenna. Signal strength readings were taken before and after the removal of the telescopic antenna (including rf alignment of the receiver). In all tests, the TR-22C showed significant improvement in sensitivity when operated without the built-in antenna.

Further improvementscan be made in the receiver's performance by replacing the jfet, Qs7, with a 2N3823, replacing the cable between terminal "RA" on the transmitter circuit board and terminal "RA" on the receiver board, and making sure to solder the braid and center conductor close to the circuit boards rather than on top of the "pins" serving as circuit-board terminals. Replace the existing coax cables with RG/174/U or equivalent.

***G.* A. Herlich, K7OR**

S-line backup power supply

Recently I acquired a Collins 516F-2 power supply with an open circuit high-voltage winding in the transformer. After the initial enthusiasm of rewinding the transformer* had subsided, I sought other means of providing a backup supply for my 32s-1/S3.

Replacement or rewound transformers are available from independent sources as well as Collins, but even then the cost is approximately $100. A good, used, 516F-2 transformer costs around $150; therefore I decided to homebrew a spare supply using readily available parts.

The price of the transformer used in the backup supply **(fig. 1**) is less than $25 — considerably less than the replacements already mentioned. If a transformer with a higher secondary voltage rating is used (no more than 750 Vac is recommended), a third filter capacitor and bleeder resistor should be added to provide a sufficient safety margin for the plate-supply string. The low-voltage filter as well might be changed from the capacitive input shown to a choke input by eliminating C3 to keep the low voltage supply within the 300-volt range required. A separate 6.3-volt filament transformer connected in reverse supplies the bias voltage. Under full-load conditions, the supply delivers 660 Vdc at 230 mA, 300 Vdc

*A. Wilson, W6NIF, "Repairing High-Voltage Trans formers, *ham radio,* March, 1969, page 66.

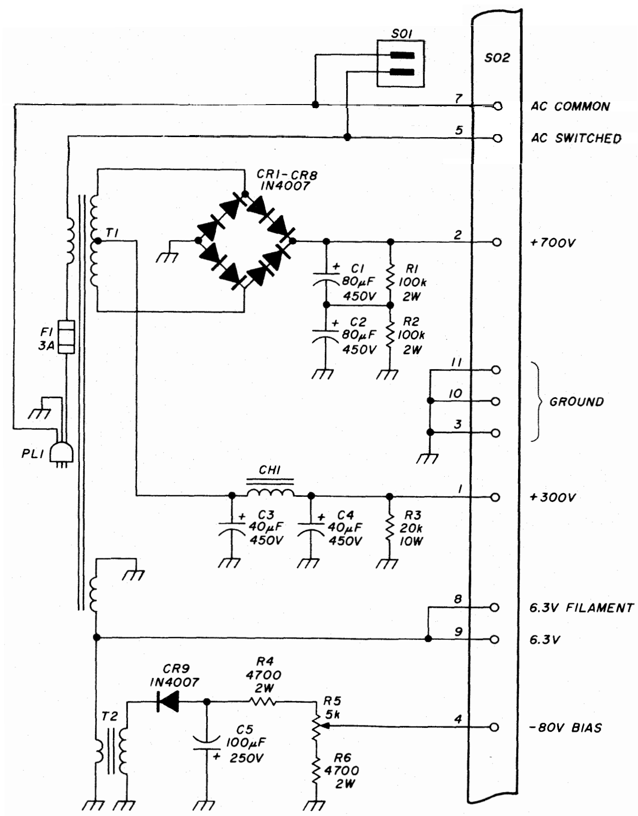
at 175 mA, 6.3 Vac at 8A, and −80 Vdc for bias.

The supply is contained on a 180 x 230 x 51 mm (7 x 9 x 2 inch) aluminum chassis, which is compatible with the 516F-2 cabinet. If you wish to install the supply in the cabinet, the front and rear chassis lips should be reinforced with flat aluminum stock. A paper or cardboard template may be used to transfer the mounting hole locations to the reinforcements.

Socket SO-1 was included on the chassis rear lip to provide a switched ac source to power a cooling fan used on the 32S-( ). Instead of using the standard ac receptacle (Radio Shack 270-642), a 2-pin male cable/female chassis connector (Radio Shack 274-201 and 274-203 respectively) may be used. The female chassis connector requires a 16-mm (5/8-inch) hole. The female cable power socket (SO-2) is an 11-pin Amphenol 78S-11 with a protective cap and cable-strain relief clamp.

The power drawn by the 32S-( ) is well within the transformer's capabilities. After hours of operation on CW, the transformer was only comfortably warm to the touch.

Paul K. Pagel, N1FB



C1,C2	Sprague TVA 1716, 80 MF/450 Vdc	SO-1	Radio Shack 270-642 (or 274-203 – see text)
C3,C4	Sprague TVL 2764, 40 MF/450 Vdc	SO-2	Amphenol 78S-11 with cap and strain relief
C5	Sprague TVA 1718, 100 MF/250 Vdc	T-1	Stancor P-8334, Triad R-77BC or similar
CR1-CR9	1N4007 silicon rectifiers		
CH-1	TV power supply choke, 2-H, 200 mA or similar	T-2	Radio Shack part number 273-050, 6.3V, 1.2 A

fig. 1. S-line **backup power supply**

audio modification for Horizon/2

I have an early-model Standard Horizon/2 two-meter rig. I've noticed a lot of audio distortion when stations with very bassy audio come on. Also, PL tones come right through when a larger external speaker is used. The low-frequency rolloff in the receiver audio section is not adequate to keep the audio amplifier IC from being driven into clipping. By replacing the 1-μF coupling capacitor (C134) with a 0.02-μF capacitor unit, the problem was eliminated.

Roy H. Davis, **WB9RKN**

cliplead carousel

After many frustrating encounters with tangled test leads and several complaints from the YL about "that ball of spaghetti and alligators" hanging from the end of an otherwise neat workbench, I decided there had to be a better way. Encouraged by a sensible system I had admired at the junior college I attend in Rhode Island, I began my improvisation. The method used in school was a long board with slots cut into its edge to accept the ends of the test leads. This was functional, but it took a lot of wall space I didn't have, what with QSL cards, world map, bookshelves, and all. So I "borrowed" a wheel from my son's bicycle parts pile and cut slots into one side of the rim all the way around the wheel. Thinking ahead, I cut some of the slots wider than the rest to accommodate coaxial-type patch cords. The wheel's axle was then secured to the end of a wall bracket with the wheel horizontal at about eye level with the slotted side up. I used a piece of 5-cm (2-inch) aluminum for the bracket, but a standard shelf brace would do as well. Clip leads and patch cords can now be passed up through the inside of the rim and dropped into the slots. This holds them firmly and neatly, easy to see and to get at.

Robert E. Best, WB1AQM

comments

(Continued from **page 6**)

have DTL/TTL compatible outputs and the 567 will sink a 100-mA TTL load at its pin 8 port. This is a low to high logic change, not a positive and negative voltage change in the case of the 567.

Further, Mr. Zegers' statement that "the place where the missing tone should be is now filled with noise" is not supportable by fact, When a PLL, such as the 567, no longer detects a signal within its passband, it simply changes logic states. If this were not so, the Touch-Tone[a] decoder circuit using the 567 would not work.

Perhaps Mr. Zegers should examine my keying circuit in more detail. There is no ± V keying used, and while not Mil spec or RS232, it works quite well.

It is not, therefore, "absolutely necessary" to have two tones decoded from a RTTY signal, but it is desirable and that's why 1 designed the PLL[2].

With regard to the reader's statement indicating I did not read the data sheet, 1 suggest he reread his. The National Semiconductor Corporation and the Signetics Corporation both list the bandwidth equation as:

$$BW = 1070 \sqrt{\frac{V_1}{f_0 \cdot C_2}}$$

Also, they list the minimum detectable signal as 20 mV, not 20 mW as Mr. Zegers does.

Also they specify the measurement at 20 mV/RMS and further state in the National Semiconductor Linear Data Book *that the equation is an* approximation *only usable to ≤200 mV, so therefore, the 2-volt example given is not valid.*

In actuality, the detection bandwidth is 75.7416 percent, *not 15 Hz, at 20 mV/RMS input, which indicates that the reader erroneously assumed the resultant of the above equation to be Hz since he indicated that the bandwidth was 15 Hz. The resultant of the bandwidth equation is the percentage of f_0 that the PLL will detect and lock onto. At 20 mV/RMS, the detection bandwidth is ± 160.8795 Hz, not 5 Hz, and the greatest number of cycles before output at 15.1416 percent bandwidth is 28 Hz not 300 Hz.*

Obviously, since bandwidth will widen out between 20 mV/RMS and the saturation point of 200 mV/RMS, an overlap will result between mark (2125 Hz) and space (2295 Hz), and that is why in the PLL[2] article it was necessary to slew the two PLLs capture frequencies so as to make a "hole" between the two signals. The 2-stage active filter also helped narrow the response envelope to a more acceptable curve.

Finally, I disagree that the 567 was never intended for FSK demodulation. I'm not alone. National Semiconductor agrees with me on page 9-40 of their Linear Data Book, *where they list among other uses for the 567 "wideband FSK demodulation."*

John Loughmiller, KB9AT
Greenville, Indiana

Yagis vs quads

Dear HR:

I imagine you have received a great deal of correspondence concerning N6NB's article in the May issue on the quad vs Yagi controversy. At the risk of adding to the bulk of mail you have already received, I would like to make some comments and offer a suggestion.

NGNB developed an excellent article based upon his experiments using the two portable towers and a triband two-element Yagi as a reference. However, the methods used to provide an accurate standard of comparison were not developed in such a manner that would lead me to conclude that a Yagi is better or worse than a quad. Since NGNB did not isolate and control his variables (such as boom spacing on the two-element quad) he really can't make an objective analysis on the relative performance of each antenna. The question of antenna feedlines for example, was not discussed, yet this variable can be of critical importance in evaluating any antenna array's performance.

While Mr Overbeck has shed some light on the quad-Yagi controversy, additional research is required before one can draw the conclusions developed by NGNB.

I would like to suggest that NGNB and his associates continue these experiments using tighter controls on the experimental process with the objective of publishing a follow-on article; then we might be able to objectively decide whether a quad is better or worse than a Yagi, and more importantly, under what conditions.

Dr. Kenneth Jenkins, WB6MMV
Portland State University
Portland, Oregon

Dear HR:

As Dr. Jenkins points out, there is no question that ignoring feedline loss variables could invalidate comparative antenna gain measurements such as those reported in my article. However, that variable was considered in all of my tests.

Since 1 had described my experimental procedures in an earlier article (which 1 footnoted), I did not repeat the full description in this particular article. I assumed that anyone wishing to critically analyze my procedures would read my previous article on the subject,

Dr. Jenkins is correct in stating that 1 tested only one two-element quad design. It would be interesting to test other two-element quad designs, but 1 did test all of the most popular designs using three or more elements; reporting how those antennas fared in comparison to Yagis of similar size was the main point of the article. Dr. Jenkins has suggested no experimental errors that would invalidate those results.

Wayne Overbeck, PhD, N6NB
Malibu, California

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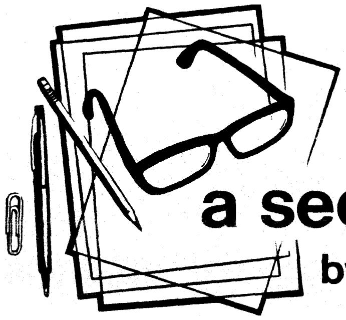


a second look

by Jim Fisk

As i was tuning around 40-meters one evening earlier this fall, I overheard a contact between two old-time Amateurs who obviously felt that Amateur Radio just isn't as exciting, inviting, and mysterious to today's youth as it had been to them when they first got started back in the 1920s and 1930s. But just as the old timer of today would like to return to the homebuilt receivers and transmitters of his youth, the old timer of 1930 probably wished to return to the days of his beloved spark transmitter and galena detector — and the old timer of 2001 will no doubt reminisce about the "good ole days of 1979." The cast of characters is different, but the basic argument never changes: "Modern technology is ruining Amateur Radio." That kind of thinking is as old fashioned and out of place as a-m on 20 meters; if anything, Amateur Radio offers more opportunities now than ever before, and the number and variety of those opportunities increases with each major advance in technology.

As just one example, consider the opportunities available through satellite communications. Rather than wishing for a return to the "good old days," we should appreciate the possibilities of intercontinental communications when we want it, rather than at the whim of the ionosphere. The Radio Amateur's traditional communications expertise, inquisitiveness, patience, and resourcefulness must again come to the fore in the exciting field of satellite communications.

Many old timers also worry that fewer and fewer amateurs now build their own equipment. Although the homebuilt receivers and transmitters of yesteryear have given way to vastly superior (and less expensive) commercial equipment, today's Radio Amateur is still building some of his own gear — speech processors, automatic SWR meters, digital dials, memory keyers — sophisticated accessories that weren't available ten years ago at any price!

There are even those who complain that the thrill of working DX is gone — anybody with enough money and a big antenna can work all the DX he wants. That's always been true, so I guess what they're really saying is that DX is no longer the private province of a small, select group. With the proliferation of high-performance transceivers and high-gain antennas, the competition for rare DX is probably more intensive now than ever before. If that's not challenge enough, there's always the world of QRPp, now growing by leaps and bounds as experienced kilowatt-wielders leave their high-power linears to marshal four or five watts to chase DX around the world.

Modern solid-state technology and manufacturing techniques have provided us with equipment which has fostered the Amateur spirit — perfecting the art of getting the message through in spite of conditions or power limitations. Rather than making more "appliance" operators, high quality commercial amateur equipment offers new challenges and opportunities for fun and training to help Radio Amateurs better serve the public interest. The sophisticated equipment now available also gives us all the ability, and indeed, the *responsibility,* to truly communicate with our fellow Radio Amateurs. And if that still isn't exciting, or challenging, or rewarding, or as new and vital as *today,* then I don't know what is.

Jim Fisk, W1HR
editor-in-chief

comments

biquad band-reject filter

Dear HR:

The article on the biquad bandpass filter for CW in the June, 1979, *ham radio* (page 70) is very interesting; the biquad active filter is a very versatile circuit. With little modification it can be made into a highpass, lowpass, bandpass (as in the article) or band reject (notch) filter. The latter may be of great interest to CW fans who want to try to remove interference near a rare DX signal. The only modifications needed are a 1.5k resistor from the input to the second op-amp inverting input (pin 2) a 1.5k resistor from the input to the third op-amp inverting input (pin 2) and R1 and R2 must be the same value. This gives an out-of-band gain of 1. To vary the Q of the filter, both R1 and R2 must be adjusted. Note that R3 still sets the filter's center frequency. The only problem I've noticed in this circuit is that if R3 is too low, the circuit will oscillate.

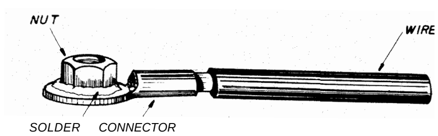
The depth of the notch depends upon how well resistors R1 and R2 and the 1.5k resistors are matched. There is probably no point in being too fussy about this since most audio amplifiers generate some harmonic distortion, so no matter how deep the notch is, some of the harmonics will still come through. Using 10 per cent resistors for R1 and R2 I routinely get rejection notches of about 25 dB; 3 dB bandwidths can be less than 50 Hz.

Dwight Sipler, KB3EH
Pittsburgh, Pennsylvania

electronic paddle

Dear HR:

The "Simple Paddle for Electronic Keyers" in April, 1978, *ham radio* proved to be a very timely article. One small point may help someone else building the paddle. It was a little



difficult starting the nut on the screw inside the plug shell. However, if you solder the nut on a spade lug, the job becomes a snap.

Edward Chromczak, WB2MGY
Somerset, New Jersey

anodizing aluminum

Dear HR:

The article on anodizing aluminum in the January, 1979, issue offers some interesting material. It also proffers some remarks which I feel need some clarification. The statement is made early in the article that the natural surface of aluminum breaks down, causing it to be unsuitable for applications where a long-term, stable surface is needed.

In contrast to this, let me quote a statement by the American Society for Metals:* "Aluminum, a member of Group III of the periodic table, is stable in air because of the presence of an extremely thin, but remarkably tight and adherent, transparent oxide film. Growth of this natural oxide film on aluminum is self-limiting." The two points of view seem to be widely divergent, to say the least.

Other statements in the article refer to ". . . an otherwise easily corrodible metal." Now, all metals are corrodible; in fact, stainless steel depends on a somewhat similar oxide film mechanism to achieve its corrosion resistance. Look around and you'll see bare aluminum performing in such long-term applications as electric transmission lines, culverts, and many others. In most instances the metal chosen for these applications was aluminum because of its ability to resist corrosion.

The photo on page 64 has a caption which speaks of ". . . a carbon speck or other alloying constituent." Granted that the cause of such an anomaly is difficult to ascertain since only the void exists, usually at the time of discovery. But a carbon-alloying constituent it is not. Over seventy-five commercial alloys are presently available in the United States, and carbon is not recognized as an alloying agent in one of them.

On page 66, the author says: "The cathode must be constructed of lead." Cathode materials may be of aluminum, stainless steel, or lead. Some precautions must be considered in the way of cathode placement, but there's a choice of materials to use.

Furthermore, I could not reconcile the current density quoted with that employed by commercial anodizers in this country. Most H_2SO_4 (sulfuric acid) anodizing in the U.S. is done at 12 amperes per square foot. Small pieces may be calculated at 0.0833 amperes per square inch but in either case keep in mind that both sides of the piece are treated at the same time. Thus one square foot of sheet metal will represent two square feet of anodizing surface when calculating current density.

Even thickness should be considered, since it's easy to see that a panel 6 x 12 x 1/16 inch thick will have 2.25 square inches of surface area exposed to the bath along the edges alone. Total surface area then is

$$2(L \times W) + t(2L + 2W)$$

(Continued on page 81)

Aluminum, volume 1, page 22, American Society for Metals.

how to design broadband jfet amplifiers to provide top performance from VLF to over 100 MHz

A discussion of broadband jfet amplifier design, with special emphasis on IMD performance and matching

Broadband rf amplifiers are becoming increasingly useful in hf/vhf receiving applications. A modern wideband upconverting high-frequency receiver, for example, often employs a broadband rf amplifier as the first active stage. The rf stage improves the receiver noise figure and reduces undesirable LO-to-antenna conduction. Broadband amplifiers are also useful in a wide variety of other Amateur applications ranging from antenna preamplifiers to home-constructed test equipment. This article deals with some of the considerations involved in the design and construction of broadband jfet rf amplifiers. The circuits presented can be easily duplicated with readily obtainable components.

Other than to provide selectivity, a broadband rf amplifier must do everything that a narrowband amplifier does. Thus gain, noise figure, stability, and most other parameters must be comparable. There is, however, a major additional requirement. Since the broadband amplifier responds to signals over a very wide bandwidth, it is important that the amplifier have exceptionally high resistance to overload and intermodulation distortion (IMD).

With respect to overload, there are potentially many more signals over the larger bandwidth (in comparison with a narrowband amplifier) that might drive the amplifier into its gain compression (overload) region. With respect to IMD, there are many more combinations of frequencies at which these strong signals can cause the amplifier to produce intermodulation products that could interfere with

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signals we are trying to receive. The intermodulation problem is compounded by the fact that unlike the narrowband amplifier, which is vulnerable only to odd-order IMD, both odd- and even-order IMD can produce interfering intermodulation products in the broadband amplifier. The narrowband amplifier is relatively immune to second-order IMD because its bandwidth is much less than an octave. Within this sub-octave bandwidth, there is no combination of frequencies at which in-band signals can produce second-order intermodulation products that also fall in-band (that is, second-order intermodulation products that are capable of interfering with in-band signals we might be trying to receive).[1,2]

Thus, in broadband receivers, it is very important to employ rf amplifiers that have extremely high resistance to both odd- and even-order IMD. These same considerations also apply to mixers.

broadband jfet amplifier

When considering an active device to be employed as a broadband amplifier, you must look for certain qualities. High transconductance is desirable. Input, output, and feedback capacitances should be low. The device should exhibit good noise performance and high signal handling capability.

High-quality jfets satisfy all of these requirements (with the exception of low feedback capacitance)

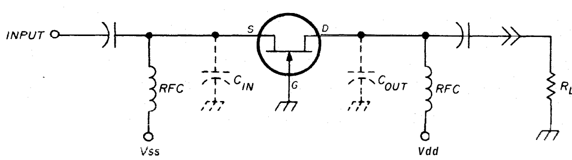

fig. 1. Schematic diagram of the simplified grounded-gate jfet amplifier.

reasonably well. By operating a jfet in the grounded-gate configuration, you can effectively reduce its otherwise high feedback capacitance to a very low level. Earlier, I mentioned that one of the desirable features of rf amplifiers in receivers is reverse isolation, or more specifically, the amplifier's ability to attenuate LO energy from the mixer to the antenna. The grounded-gate/base/grid configuration yields the best reverse isolation of any of the three possible amplifier configurations.

A simplified circuit of a grounded-gate jfet amplifier is shown in **fig. 1**. You can gain some insight into the operation of this circuit by applying some fundamental (and somewhat simplified) relationships. A more general and rigorous presentation of these relationships may be found in reference 3. The first relationship defines the input impedance of the amplifier. Disregarding the imaginary (reactive) component, the approximate input impedance is given by the expression:

$$R_{in} \approx \frac{1}{G_m} \quad \textbf{(1)}$$

where R_{in} is the real component of the input impedance and G_m is the device transconductance. A grounded-gate jfet with a transconductance of 10,000 micromhos, for example, would have an approximate input impedance of 100 ohms.

The second relationship defines the voltage gain of the amplifier. This relationship is:

$$A_v \approx G_m \times R_L \quad \textbf{(2)}$$

where A_v is the voltage gain

G_m is the device transconductance

R_L is the load resistance

Rearranging **eq. 1** to solve for G_m produces:

$$G_M \approx \frac{1}{R_{in}} \quad \textbf{(3)}$$

and substituting this expression of G_m into **eq.** *2* produces:

$$A_v \approx \frac{R_L}{R_{in}} \quad \textbf{(4)}$$

That is, the approximate voltage gain is simply the ratio of the load resistance to the device input impedance.

Referring back to **fig. 1,** observe that the driving source, the jfet, and the load are all in series. Disregarding the input and output capacitances for the moment, it is evident then that the current gain, A_i, of this amplifier must be equal to unity because the current in a series circuit is everywhere the same. Power gain, A_p, is given by the expression

$$A_p = A_v \times A_i \quad \textbf{(5)}$$

Since A_i equals unity in a grounded-gate jfet amplifier, then

$$A_p = A_v \quad \textbf{(6)}$$

In other words, the power gain equals the voltage gain.

It would seem then that to obtain high power gain out of this amplifier, all that is necessary is to make R_L large. This is true up to a point, but even if you disregard the inherent output conductance of the jfet (which could be represented as an equivalent conductance from the drain to the source), you would find that C_{out} (the jfet output capacitance) would limit the output impedance and device gain at high frequencies. Assuming a constant transconductance over the frequency range of interest, and disregarding C_{in},

$$f_{3dB} \approx \frac{1}{2\pi R_L C_{out}} \tag{7}$$

where f_{3dB} is the high-end frequency at which the amplifier gain is 3 dB down from its low frequency value.

To obtain the widest bandwidth, then, eq. **7** indicates that you should employ a jfet with a very low output capacitance driving a low value of R_L. You can arbitrarily select a low value for R_L, but to maintain high power gain. eq. **4** says that you will need a low value of R_{in} (the jfet input impedance). However, since $R_{in} = 1/G_m$ from **eq. 1;** this is actually just another way of saying that you need a jfet with high transconductance.

Given the above, it is evident that a useful figure of merit for a jfet in broadband operation is the ratio of device transconductance to output capacitance. A jfet with an exceptionally high transconductance to output capacitance ratio is the Siliconix U310.4

U310s are rather expensive, but a plastic economy version (the J310) is also available with performance characteristics that are substantially the same. The J310 is manufactured by Siliconix and National Semiconductor. Siliconix also offers a matched pair of J310-type jfets in an epoxy package. This device is the E430.

By the time this article is published, the E430, in all probability will have been phased out in favor of the U430. The U430 will employ the same chip geometry as the E430, but will use an 8-pin metal package similar to the TO-5 package.

The J310 is described in the *Siliconix* FET *Data Books* as a low-noise, wide-dynamic-range device capable of high power gain at frequencies up to at least 450 MHz. The typical transconductance is listed at 12,000 micromhos at a drain current of 10 mA. The amplifiers presented in this article employ J310s and E430s.

The input impedance of a J310 or E430 grounded-gate amplifier can be made close enough to 50 ohms so that a reasonable input VSWR can be achieved without any matching network. The signal can simply be capacitively coupled to the jfet source. The disadvantage of this convenient technique is that the input impedance may not be optimum for best noise figure.

As previously mentioned, the load impedance (R_L) must be high compared with the jfet input impedance to obtain high power gain. Since the required value of R_L is much higher than the assumed 50-ohm load the amplifier is ultimately driving, broadband autotransformers can be employed to convert the 50-ohm load impedance to the higher level of R_L required to achieve reasonable gain in the jfet amplifier. Fifty ohms is selected as the desired ultimate load impedance since this is the nominal impedance level of most broadband mixers that might follow the amplifier.

jfet biasing

In biasing a jfet, there are three general requirements. The first is that the jfet maintain the desired bias current level over the anticipated temperature range. The second requirement is that the biasing circuit should not be device-sensitive. That is, if you design an amplifier employing a J310 biased at 18 mA of drain current, this drain current should be close to 18 mA for *any* J310. The third requirement is that the bias current should be insensitive to changes in supply voltages.

The first requirement is not too difficult to meet. Even with a poor biasing circuit, the bias current will remain fairly constant over a reasonably wide temperature range.

The second requirement can be relaxed somewhat where repeatability isn't so important. Since Amateur home projects are usually built in very small quantities, there is no particular problem with using pot or selecting resistors to achieve the desired bias current (especiallyif doing so permits the use of simpler circuitry or reduces power consumption).

The third requirement is also easy to meet. If the jfet is operated from a single supply voltage (V_{dd}), the inherent constant-current characteristics of the jfet will automatically stabilize the bias current, provided that the drain-to-source voltage is at least 6 volts or so (depending upon the particular jfet). If the jfet is supplied by both positive and negative voltages (V_{dd} and V_{ss} respectively), the bias current may be somewhat sensitive to changes in V_{ss}. However, the situations where dual supply voltages are available will also be the situations where these supplies are most likely to be regulated.

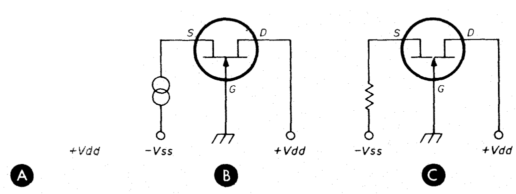

fig. 2. Diagrams of the three jfet biasing configurations with only the relevant dc circuitry shown. Schematic (A) shows the simplest. and poorest, configuration. The circuit is very device-sensitive, in that it is necessary to select the bias resistor for the desired current. Diagram (B) illustrates the use of a constant-current source to bias the jfet. The best overall compromise is shown in (C), where a negative voltage and large-value resistor act as a pseudo constant-current source.

Fig. 2 illustrates three commonly used jfet biasing circuits. For simplicity, only the relevant dc circuitry is shown. **Fig. 2A** shows the most commonly used (and poorest) jfet biasing configuration. Although its performance over a temperature range is adequate in most cases, it tends to be very device-sensitive. It is therefore necessary to select the resistor (or make it variable) to secure the desired bias current. On the plus side, overall power consumption is lower than that of the other two configurations and no negative supply is required. In **fig. 28,** the jfet is biased by a constant-current source. If the constant-current source (usually a bipolar transistor with a temperature compensating diode) is temperature stable, this biasing scheme is nearly impervious to temperature, device, and supply voltage variations, and is thus an excellent biasing configuration.

A compromise configuration is shown in **fig. 2C.** This circuit is very similar to that of **fig. 2A** except that the resistor is larger and is returned to a negative supply. The negative voltage and large resistance act as a pseudo constant-current source. The larger value of the resistor and the magnitude of V_{ss}, the closer this biasing circuit comes to approximating a true constant-current source. If the negative voltage supply is available, this circuit offers the best performance for the number of components required. Temperature stability is very good, and the circuit is reasonably insensitive to device variations. As a brute-force test, this circuit was constructed with a J310 biased at a nominal current level of 18 mA using ±12 volt supplies. Ten different J310s were tried in the circuit. The measured bias currents were all well within a 10 per cent window. Heating the devices for 10 seconds by applying a 25-watt soldering iron of less than 10 per cent. Similarly, chilling the devices with an aerosol spray coolant for a period of 10 seconds also resulted in a bias current change of less than 10 per cent.

Although the amplifiers described in this article all employ the biasing configuration of **fig. 2C,** substan- any of these bias configurations. More detailed information on the subject of jfet biasing can be found in reference (3).

basic jfet broadband amplifier

Fig. 3A shows a simple broadband amplifier. Although this circuit is presented primarily for purposes of illustration, it nonetheless has many practical applications. The circuit is a grounded-gate jfet amplifier employing a single J310 biased at 18 mA of drain current. The output employs a peaking inductor and a 4:1 bifilardwound auto-transformer (detailed winding information is presented later in this arti-

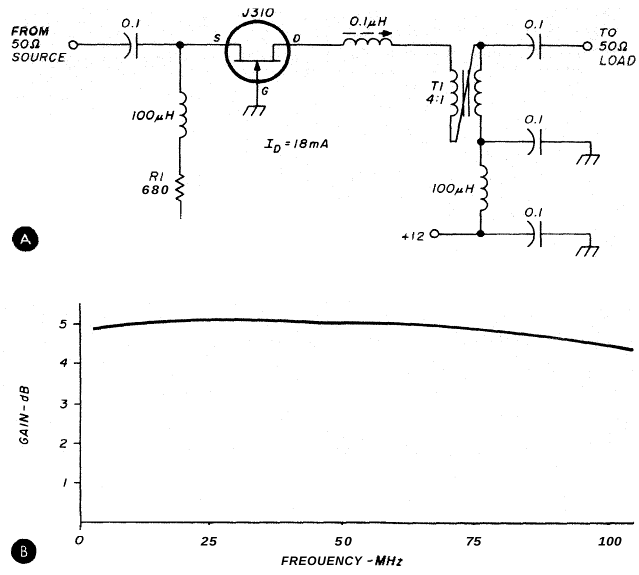

fig. 3. Schematic diagram of the basic broadband jfet amplifier. R1 sets I_D at 18 mA. (B) shows the frequency response of this basic amplifier.

cle). The peaking inductor extends the frequency response. **Fig. 3B** shows the frequency response when L_1 is set for optimum gain flatness with respect to frequency. The setting of L_1 is not particularly critical, although it does substantially affect the high-end frequency response. Other performance characteristics are as follows:

1 dB gain compression level	+13 dBm
2nd-order intercept point	+28 dBm
3rd-order intercept point	+22 dBm
30-MHz noise figure	4.5 dB
input VSWR	1.3:1 from 1.8-100 MHz
reverse isolation	38 dB or better to 30 MHz; 25 dB or better to 200 MHz

The intermodulation and overload specifications for all amplifiers presented in this article are referenced to the amplifier input. The +13 dBm specification for the 1 dB gain compression level, for example, is the *input* (rather than output) level at which 1 dB of gain compression occurs. When evaluating the intermodulation and overload performance of a device, it's very important to know whether the specifications are referenced to the input or output. Unfortunately, many manufacturers specify their devices without providing information as to whether the specification ferred method is to reference the specification to the input.

The output referenced specification is simply the input referenced specification plus the device gain. For example, an amplifier having an input referenced 3rd-order intercept point of +20 dBm and a gain of 10 dB has an output referenced 3rd-order intercept point of +30 dBm.

Some clarification is in order concerning this amplifier. First, the good input VSWR trades off against optimum noise figure. With a 1:1 input VSWR

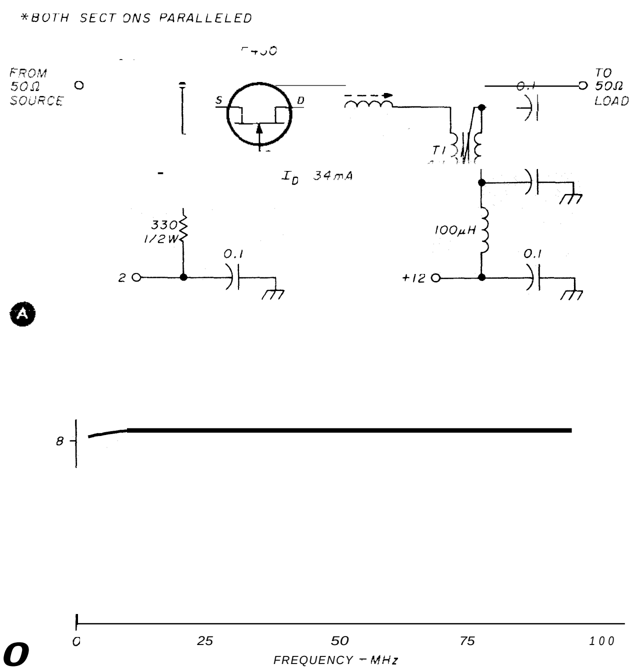

fig. 4. Diagram of the improved broadband jfet amplifier with improved gain, lower noise figure, and better IMD performance. The higher transconductance from the paralleled fets results in a higher gain and better input noise match. Graph (B) shows the overall frequency response of this improved amplifier.

you would have "power match" and best amplifier gain. However, the noise figure in this case could be no better than 3 dB. "Noise match" (optimum drive source impedance for best noise figure) occurs when the driving impedance is considerably higher than the amplifier input impedance. The second point is that although you can specify a 50-ohm output load, the actual output impedance of this amplifier is much higher than 50 ohms (the jfet output is essentially a high-impedance current source). This fact is very important if a filter must follow the amplifier. Where low output VSWR is important, refer to the circuits of **fig. 6** or **7.** All the other circuits in this article have a high output impedance. Finally, the gain of this amplifier is load-sensitive; that is, if the actual value of R_L is greater than 50 ohms, amplifier gain will be higher. Load impedances other than 50 ohms will also require that the value of L_1 be changed for flattest frequency response. If the amplifier is to be used only below 30 MHz, however, L_1 may be omitted entirely. Ampiifier gain may vary somewhat depending upon the characteristics of the particular J310 employed. If a negative supply is not available, return R_1 to ground (instead of -12 volts) and select (or adjust) R_1's value for 18 mA of drain current.

improved jfet broadband amplifier

The amplifier shown in **fig. 4A** is very similar to the one just described, but with higher gain, a lower noise figure, and superior IMD performance. **Fig. 4B** shows typical amplifier gain as a function of frequency. Other performance characteristics are as follows:

1 dB gain compression level	+14 dBm
2nd-order intercept point	+38 dBm
3rd-order intercept point	+29 dBm
30-MHz noise figure	<2.5 dB
input VSWR	1.8:1 from 1.8-200 MHz
reverse isolation	36 dB or better to 30 MHz
	30 dB or better to 175 MHz

This amplifier employs the E430 dual jfet as the active device, with individual sections connected directly in parallel (source 1 tied to source 2, gate 1 tied to gate 2, and drain 1 tied to drain 2) to achieve an equivalent ultra-high transconductance jfet. The higher transconductance (– 36,000 micromhos) accounts for the higher gain of this amplifier as compared with that of **fig.** 3. This higher conductance also causes the input impedance to drop to approximately 28 ohms, which accounts for the 1.8:1 input VSWR. Although you no longer have an optimum input power match you're now much closer to an optimum input noise match, which accounts for the improved noise figure. As far as improved **IMD** performance is concerned, the easiest way to rationalize

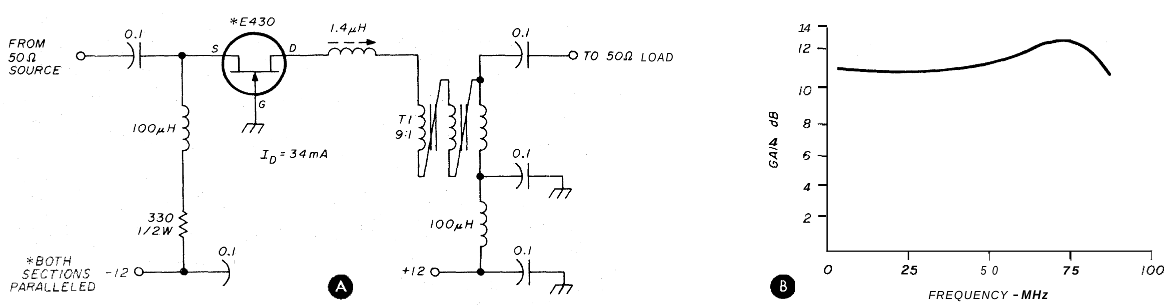

fig. 5. Schematic of a jfet amplifier which exhibits approximately 12 dB gain, though with a narrower bandwidth. In this case, the bandwidth has been sacrificed to produce the higher gain. A plot of the gain vs frequency is shown in (B).

that is to simply say that two devices are carrying the load instead of just one (keeping in mind that you now effectively have two J310s in parallel).

The same considerations with regard to load sensitivity, device variations, and output VSWR are equally applicable to this amplifier as to the one in **fig.** 3. Again, for operation below 30 MHz, peaking inductor L_1 may be omitted. Since the E430 gets quite warm with 34 mA of drain current, it would probably be a good idea to use a heatsink. T_1 is the same 4:1 bifilar-wound transformer as the one shown in **fig.** 3.

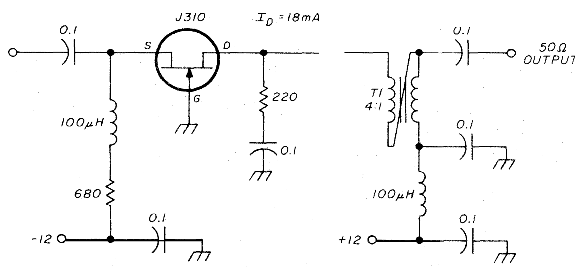

fig. 6. Schematic of a broadband 50-ohm driver where the output impedance is set to **50** ohms. in this case, the **220**-ohm resistor loads the collector, leading to the 50-ohm output through the transformer. Gain in this example is approximately unity.

higher-gain jfet broadband amplifier

From **eq. 4,** you know that amplifier gain can be increased by raising the effective drain load impedance as seen by the jfet. Therefore, if you replace the 4:1 autotransformer of **fig. 4A** with a 9:1 autotransformer, gain should increase. **Eq. 7,** however, tells you that this will also decrease the bandwidth. Since the amplifier of **fig. 4A** has a bandwidth in excess of 100 MHz, you probably can trade off some of this bandwidth for higher gain in many applications. **Fig. 5A** shows the circuit for such an amplifier. **Fig. 5B** shows the gain as a function of frequency. As predicted, gain has increased at the expense of bandwidth. Other performance characteristics are as follows:

1 dB gain compression level	+10 dBm
2nd-order intercept point	+36 dBm
3rd-order intercept point	+24 dBm
30-MHz noise figure	<2.5 dB
input VSWR	1.6:1 from 1.8-100 MHz
reverse isolation	34 dB or better to 100 MHz

Winding details of the 9:1 autotransformer are presented later in this article. L1 cannot be omitted from this circuit for high-frequency operation unless substantial gain roll off (2-3 dB at 30 MHz) can be tolerated.

jfet broadband drivers

There may be occasions where a better-defined amplifier output impedance is required. **Fig. 6** illustrates a broadband driver circuit designed to present nominal 50-ohm impedances to both the source and load. Since this driver produces somewhat less than unity gain, it is intended only to follow one of the previously discussed amplifiers rather than to stand alone. The circuit is nearly identical to the "basic" jfet broadband amplifier of **fig.** 3. Notice, however, that the drain is ac loaded by a 220-ohm resistor to establish the amplifier output impedance (at the autotransformer output) near 50 ohms. This reduces the gain for two reasons. First, you have lowered the impedance as seen by the J310 drain by a factor of two, thus reducing voltage and power gain by the same factor. Additionally, half the output power is now consumed in the 220-ohm resistor. Thus, the power available to the load is cut by a total factor of 4, or 6 dB. The measured gain of this amplifier is –1 to –2 dB from 1.8 to 100 MHz, or *6* to 7 dB lower than that of the "basic" jfet broadband amplifier of **fig.** 3. Other performance characteristics are as follows:

1 dB gain compression level	+13 dBm
2nd-order intercept point	+36 dBm
3rd-order intercept point	+24 dBm
30-MHz noise figure	4-5 dB (estimated)
input VSWR	1.3:1 from 1.8-100 MHz
output VSWR	1.3:1 from 1.8-100 MHz
reverse isolation	35 dB or better to 100 MHz

Fig. *7* shows another broadband driver circuit

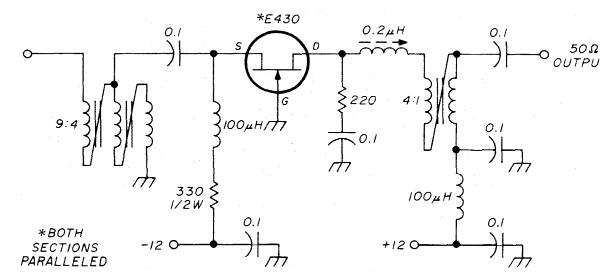

fig. **7.** Schematic diagram of another broadband 50-ohm driver which has the input impedance matched by the **9:4** transformer. The gain, in this case, is **1-2** dB.

which provides some gain rather than loss. It is very similar to the "improved" jfet broadband amplifier of **fig. 4A.** Again, the drain is ac loaded by a 220-ohm resistor to establish the amplifier output impedance. Also, the input signal is impedance matched to the jfet source through a 9:4 autotransformer. This transformer is wound identically to the 9:1 autotransformer employed in the "higher gain" jfet broadband amplifier of **fig. 5A,** but is turned "upside-down" to provide a 9:4 impedance ratio. The gain of this amplifier is 1-2 dB from 1.8-100 MHz. Other performance characteristics are as follows:

1 dB gain compression level	+14 dBm
2nd-order intercept point	+30 dBm
3rd-order intercept point	+29 dBm
30-MHz noise figure	4-5 dB (estimated)
input VSWR	1.3:1 to 30 MHz
	1.8:1 to 100 MHz
output VSWR	1.3:1 to 30 MHz
	7.6:1 to 100 MHz
reverse isolation	40 dB or better to 100 MHz

In both driver circuits, it is important to connect the 220-ohm resistor as closely as possible to the jfet drain with very short lead lengths. If this is not done, oscillations may occur.

winding the autotransformers

The 4:1 autotransformer consists of five turns of bifilar-wound wire on a single-hole ferrite bead. The bifilar wire is made by paralleling and twisting together two dissimilar colored (red and green, for example) of strands of no. 32 AWG (0.2-mm) magnet wire. This is easily done by attaching one end of the paralleled wires in a vice and placing the other end in the chuck of a portable power drill. Maintaining suitable tension on the wires, turn on the drill until the wires have twisted together. Four twists per centimeter (10 twists/inch) is suitable, but this is not at all critical. The only real requirement is that there be enough twists to prevent unraveling but not so many as to cause kinking. The ferrite bead is an FB 43-801. To wind the transformer, wind the bifilar wire through the bead five times, keeping the winding tight to the core. This will result in four strands of bifilar wire against the outside of the bead. Cut off the excess wire, leaving 2-3 cm (approximately 1 inch) or so at each end. Untwist the two ends and bend the green wire of either end and the red wire of the other end toward each other until they meet halfway along the outside wall of the bead, completing the fifth turn. Tin and twist these wires together. Similarly, tin the remaining red and green wire ends. The net result is that the red and green wires are connected *series-aiding,* with their junction being the autotransformer center tap. This 2:1 turns ratio yields a 4:1 impedance ratio. **Fig. 8A** shows an outline drawing of the completed autotransformer along with the corresponding schematic representation.

The 9:1 autotransformer is constructed on the same type of ferrite bead as the 4:1 autotransformer, but is wound with no. 32 AWG (0.2-mm) trifilar wire. The trifilar wire consists of three colored strands, (red, gold, and green, for example) of no. 32 AWG (0.2-mm) wire twisted together in the same manner as the bifilar wire. Wind five turns of the trifilar wire through the ferrite bead and connect the wires series-aiding (see **fig. 8B**). If the input signal and ground connections are reversed, the 9:1 autotransformer then becomes a 9:4 autotransformer.

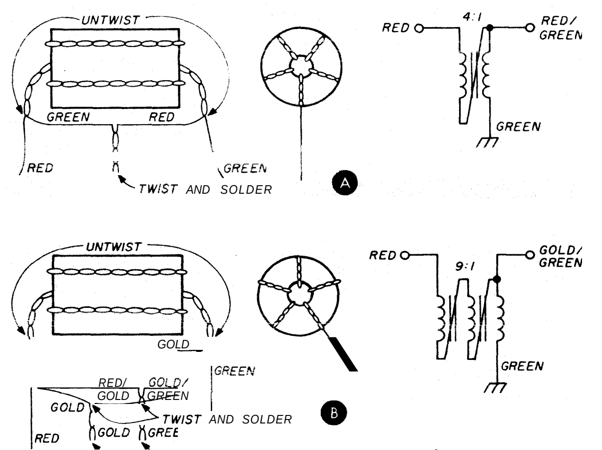

fig. 8. Winding information for the 4:1 and 9:1 transformers. Each transformer uses a ferrite bead, FB 43-801, as the core.

Although the FB 43-801 beads are satisfactory in broadband autotransformer applications, they were selected primarily on the basis of their availability to Amateurs rather than for optimum performance. Two-hole balun cores seem to perform somewhat better.

The FB 43-801 beads may also be used to construct the rf chokes. The 100-μH rf chokes used in the broadband amplifiers may be constructed by winding nine turns of no. 28 AWG (0.3-mm) wire through the beads.

measurement procedures

Swept gain, VSWR, and reverse isolation measurements were made using a Wiltron Model 640 RF analyzer. Noise-figure measurements were made using a calibrated temperature-limited diode-noise generator, a 6-dB pad, a broadband Avantek amplifier, a Heath SB-303 receiver, and an RMS ac VTVM as shown in **fig. 9A**. The noise factor of the pad/amplifierlreceiver combination was first measured. The jfet amplifier under test was then inserted between the noise generator and the 6-dB pad, after which the overall system noise factor was measured. The noise factor of the jfet amplifier alone was then calculated using the well-known gain-noise factor equation in rearranged form:

$$F_1 = F_T - \frac{F_2 - 1}{G_1} \qquad (8)$$

where F_1 is the noise factor of the jfet amplifier
F_2 is the noise factor of the pad/amplifier/receiver combination

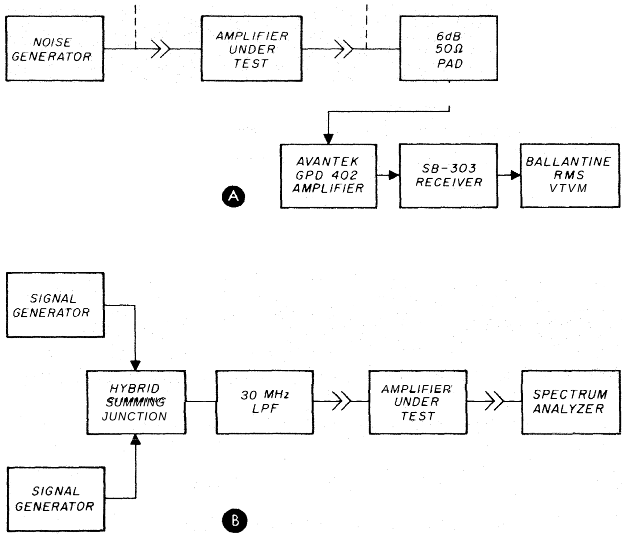

fig. 9. Test setup for the noise-figure and IMD measurements.

G_I is the power gain factor of the jfet amplifier

F_T is the overall system noise factor

The noise *figure* is then simply 10 $\log_{10}$ noisefactor.

Gain compression level measurements were made with an HP-8654B signal generator and an HP-8558B spectrum analyzer, Intercept point measurements were conducted using two HP-8645B signal generators, a 3-dB hybrid junction, a 30-MHz lowpass filter, and an HP-8558B spectrum analyzer. **Fig. 9B** shows the test setup. Second-order intercept point measurements were made by first setting the signal generator outputs to +3.5 dBm (or 0 dBm input to the amplifier after accounting for the 3.5-dB loss at the hybrid junction) at frequencies of 14 and 15 MHz. The difference in amplitude between the 14/15 MHz signal levels and the 29-MHz sum product as observed on the spectrum analyzer was then added to the 0 dBm 14/15 MHz amplifier input signal level to compute the sum product second-order intercept point. The signal generators were then tuned to 27/30 MHz, and the second-order intercept point was again calculated, this time for the 3-MHz difference product. The amplifier second-order intercept point (referenced to the amplifier input) was then taken as the lesser of the two measurements.

Third-order intercept measurements were accomplished by again setting the signal generators to 14/15 MHz at 0 dBm input levels to the amplifier under test. Third-order intermodulation products appeared at 13 and 16 MHz. The difference in amplitude between the 14/15 MHz signals and the greater of the intermodulation products was divided by two and added to 0 dBm to arrive at the third-order intercept point (referenced to the amplifier input).

summary and conclusion

Broadband amplifiers for receiving applications require superior odd- and even-order intermadulation performance due to their greater bandwidths. Jfets make excellent low-noise broadband amplifiers in the high frequency and low vhf range, providing moderate gain and unsurpassed third-order intermodulation performance for the amount of current drawn. Second-order intermodulation performance is good, but may not be as good as that of certain bipolar transistors, particularly when these bipolar transistors are connected in push-pull.[6]

Other devices for consideration as low-noise broadbaad high-intercept point amplifiers include the Siliconix VMOS[7] and the Signetics DMOS fets.[8] The VMOS fets are capable of performance superior to that of E430s in terms of gain, bandwidth, and dynamic range. To secure maximum gain and linearity, however, it is necessary to run hundreds of milliamperes of current through the device, impractical for most receiving applications. A test of a Siliconix VN33AK VMOS fet at 50 mA of drain current (in the device square-law region) resulted in significantly poorer bandwidth and dynamic range than that of an E430 running at 34 mA. A test of the Signetics SD202 DMOS fet at 20 mA of drain current resulted in a gain somewhat greater than that of a J310 and a third-order intercept point comparable to that of an E430. The extremely low output capacitance of the SD202 resulted in improved bandwidth as well.

Both VMOS and DMOS fets characterized for rf applications are still rather expensive, but both technologies are rapidly advancing in terms of performance and man, facturability. As a result, prices are certain'to come down while performance improves (some Siliconix VMOS fets characterized for switching applications already seii for under one dollar in large quantities). In the months ahead, we undoubtedly can look forward to exciting developments in both of these expanding technologies.

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ham radio

the Hallicrafters story

A fascinating account of one of Amateur Radio's great names. How it grew, what it produced, and its demise in today's Amateur market. But is Hallicrafters really dead? Of course not

By William I. Orr, W6SAI, 48 Campbell Lane, Menlo Park, California 94025

prologue

The unmarked car pulled up in front of the small factory. It was a sultry, humid day in Chicago in August of 1941. The driver got out of the vehicle and strode purposefully into the office of the little company. He displayed his credentials and, after a short pause, was ushered into the office of the president.

The president, young Bill Halligan (W9WZE, now W9AC), greeted his visitor and quickly found what was wanted.

"Mr. Halligan," the visitor said in self-assured voice, "we need an HT-4 transmitter."

Bill Halligan shook his head. "I'm sorry, we haven't one in the place. And we don't even have one in production now."

The visitor looked around the office and glanced out the door to the small production line.

"How about a used one, or one that you keep for test purposes?"

"None available at all, sorry."

The visitor leaned forward across Bill's desk. "Do any of your dealers have one? I mean anywhere in the U.S.? How about sending some telegrams?"

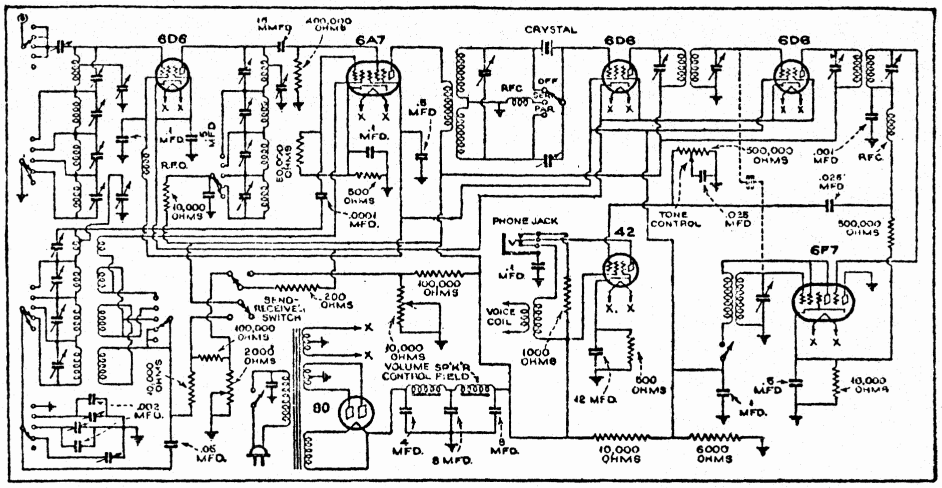
Bill Halligan produced a wad of letters and telegrams from his desk. "It's the other way around," he replied. "They're all cleaned out and are asking for more HT-4 transmitters."

"You don't seem to understand," said the visitor. "This is important and urgent. We must have a transmitter. At once. It is of the highest priority!"

Bill sighed. "Well, how about my HT-4 transmitter? It's at home in my shack . . ."

The transmitter was immediately brought to the factory, an old red brick building on Indiana Avenue. That weekend Bill and an engineer went over the HT-4, checking it out, and on Sunday an Army bomber roared away from Chicago Municipal airport with W9WZE's supreme sacrifice aboard – his own personal HT-4 transmitter. Not even the persistent, close-lipped visitor knew of its ultimate destination, but it was bound for a rendezvous with history (see epilogue).

Bill Halligan received an early start in the world of wireless. As a high-school student before World War I, his homemade spark transmitter cut a broad swath in the 200-meter band, causing great anguish to the Navy radio operators in the Boston area. And after marine and Navy radio work in the Great War, he immediately went into radio as a profession in the Boston/New York area. But by 1931 he was a manufacturer's agent in Chicago, just in time for the depths of the depression. It was a challenging time for an energetic fellow who wanted to build high-quality ham equipment – something he'd never done but always wanted to do. Perhaps now was the time. What did he have to lose?



Schematic of the Super-Skyrider, the first Hallicrafters short-wave superheterodyne receiver.

"I've just the name for you," said Lloyd Back, Bill's good friend in a local advertising agency. "Why don't you call your new company The Hallicrafters? I got the idea from an outfit called the Roycrafters, a printing company in New York that produces fine printing under the leadership of Elbert Hubbard. You'll be the artisans of the new science of radio communications. How do you like that?"

the coveted RCA license

The whole idea caught fire like wet leaves. It was 1933, and hams were broke too. No one seemed interested in the simple ***Sky Rider*** receiver that Bill had designed. Another serious problem was also at hand: The little company was unable to secure a license to manufacture sets under patents held by the Radio Corporation of America. RCA was virtually a patent pool, holding patents on almost every basic radio circuit that existed. True, a licensing system existed for those well-heeled outfits that could pay the price. But it was too high for Hallicrafters. The company went into low gear, and Bill Halligan had to go to Silver-Marshall Company – a licensed competitor – to produce sets of his own design.

It was a tough road. Bill would take a briefcase full of wiring diagrams, drawings, and photographs to ham dealers and secure orders for nonexistant receivers. When he had 50 or 100 orders, he'd deposit the orders with Silver-Marshall and the sets would be produced – for cash only.

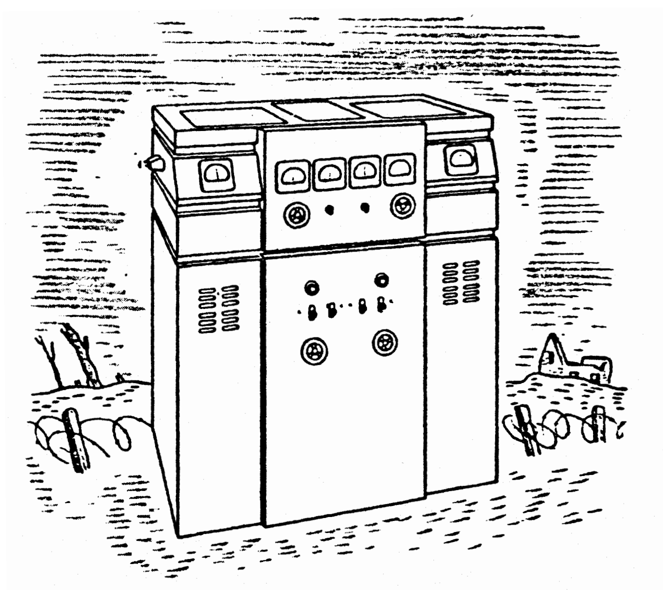
"I've got to get an RCA license of my own," Bill told his wife. He knew that security and a future depended upon an RCA license. And he was going to get one.

harsh days in Chicago

Difficult days were ahead for Bill Halligan. He finally heard of a new, ultra-modern radio factory in Marion, Indiana, which was closing for lack of business. It had an RCA license! Hurrying to Marion, he sold the Hallicrafters' idea to the owner. Shortly thereafter the Hallicrafters' name appeared above the factory entrance. But the two partners lacked the money to produce anything, and the grim specter of bankruptcy loomed on the horizon. Both partners were deep in debt. Bill now had the RCA license, but he could do nothing with it!

By one of those fascinating turns of fate, Bill Haligan chanced to meet Ray Durst. Ray was the credit manager for the Ecophone Company in Chicago, which has a large account with the plant in Marion. Ray felt that if he could collect the debt Marion owed to Ecophone, he might not be fired by that company.

Bill and Ray got together to see if they could solve their immediate problems. The upshot of these discussions was that Bill and his partner agreed to dis-

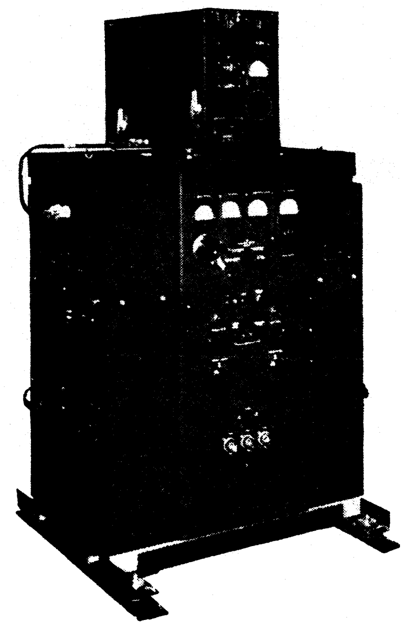


The Hallicrafters HT-4 transmitter.

HALLICRAFTERS BC-610 Xmtr

450 WATTS C.W.
325 WATTS PHONE

now available to the amateur!



NET PRICE complete........ $500.00

This high-power transmitter, famed for its performance in the SCR-299 mobile radio station, is ready now for YOU. Includes all regular features of the familiar HT-4E . . . plus battle-tested iniprovenients that make it better than ever. Furnished complete with speech amplifier, tubes, 3 sets of coils (1.5 to 18 mc.), and simple modification instructions for operation on 10 meters. Like new — used only slightly. Fully guaranteed

The BC-610 was a three-stage transmitter using plug-in coils. Tube line up was: 6V6 crystal oscillator, parallel 807s buffer, and 250TH final amplifier. The modulator used p-p 100THs. A separate speech amplifier and antenna tuner were a part of the transmitter package. Pre-war ham price was about $750. Post-war surplus price was $500. The BC-610 was a popular ham transmitter up to 1948 when it was killed off because of excessive TVI. Tuning units and coils for this famous transmitter are occasionally found on the surplus market today.

solve their contract. Bill and Ray then went to Chicago. There Bill met Clem Wade, the inventor of the ***Eskimo Pie*** ice cream stick and sole owner of the Ecophone Company's equally frozen assets. Would it be possible that some kind of agreement could be reached whereby Bill could make use of the facilities at Ecophone? Perhaps some kind of cross-licensing?

Yes, it was possible. After a tangle of lawsuits with creditors, contracts, licensing agreements, incorporations, and expensive paperwork, Bill Halligan and Ray Durst emerged with a company. Now, perhaps, the Hallicrafters dream would come true.

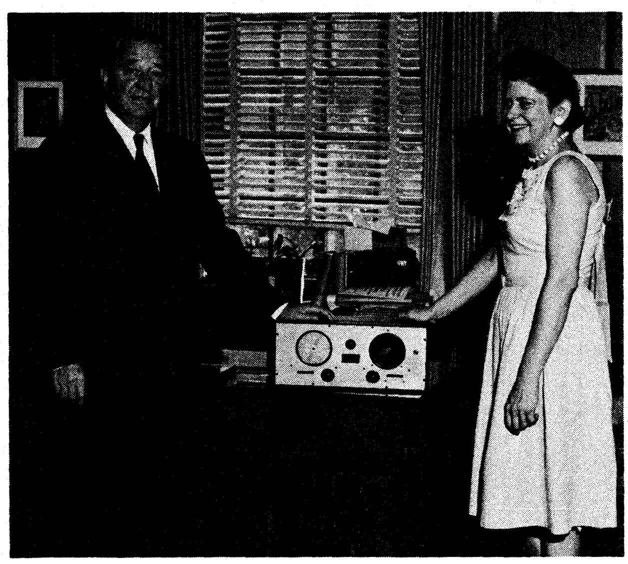
before Pearl Harbor

Hallicrafters was in business. It made radio receivers and phonograph combinations for Capehart, Magnavox, and other Chicago radio companies. And it wasn't long before handmade Hallicrafters shortwave receivers were being made. At long lasr, Bill Halligan had his factory and was in the ham radio business.

Hallicrafters entered the ham market like a rocket. The ***Super Sky Rider*** receiver was announced in early 1935. Modern marketing techniques and joint advertising with distributors was tried — a new approach to the staid market of Amateur Radio. Bill Halligan's earlier work as a manufacturer's agent began to pay off. Within months Hallicrafters was a household word in Amateur Radio. Hallicrafters fielded a whole series of radio receivers and probably had more models on the market than all the other receiver manufacturers combined. Yearly model changes were made, high-style cabinets and attractive panels were used, and the massive volume of receiver production held the costs low — a Hallicrafters receiver existed for every purse.*

the Hallicrafters HT-4 transmitter

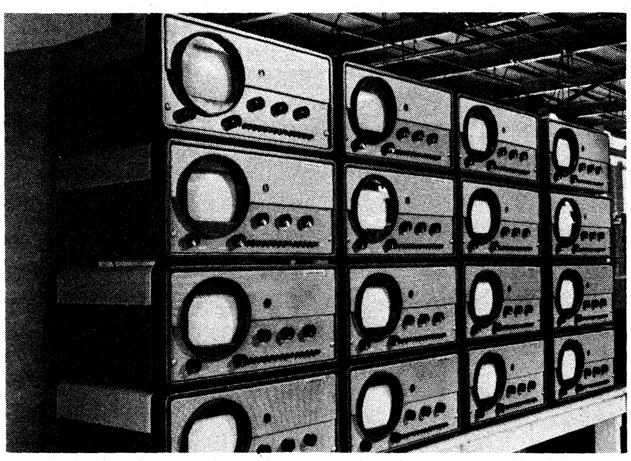
"Why not build a modern ham transmitter," Bill Halligan asked himself one day. With the aid of Bob Samuelson, the design for a powerful ham transmitter was drawn up — target price: less than ***$800.00.***



Bill Halligan and wife, Katie, with the very first model Sky Rider receiver of long ago. Receiver was a trf job, soon to be supplanted by the Super Skyrider. a superheterodyne.

Bill hoped that the Radio Amateur, now accustomed to buying a receiver rather than building it himself, to buying a receiver rather than building it himself, might be ready for a factory-assembled rig.

The ham transmitter of 1938 was a sight to behold. The ham kilowatt then was a massive and awesome contrivance that towered in one or more relay racks to a height of six or seven feet — and with all the mobility of a grand piano.



Early Hallicrafters television receivers are a collector's item now. This model had a push-button tuner that included TV channel 1. now the Amateur 6-meter band.

Starting with a typical high-power transmitter design, Bob Sarnuelson and his engineers spent months over wiring diagrams and layouts, reducing the design to a height of 37 inches and decreasing the conventional seven power supplies to three. Complex switching circuits were simplified. Plug-in tuning assemblies were designed. The heavy steel relay rack construction was scrapped for a light steel frame using the "stressed skin" techniaue develooed for airplane fuselages. Transmitter weight was reduced from over a ton to about 500 pounds!

The final design was a 450-watt phone and CW transmitter that contained no aluminum. It was made of automobile sheet steel, and, with the exception of the heavy transformers, it was easy and inexpensive to produce.

mass production — a real problem

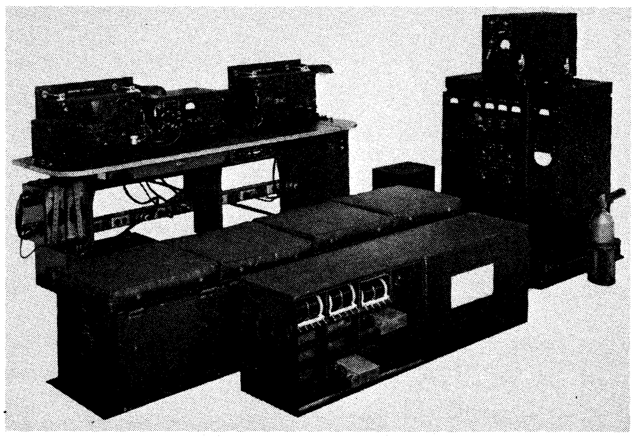
Bill Halligan had no way of knowing that a war was about to start and that his HT-4 transmitter would be a mainstay for World War II communications. After all, he had built only twenty of them by the fall of 1940. He was soon to be surprised. Suppose America got into the war? Would anyone be interested in the HT-4? Could production be stepped up to the unheard-of rate of one a day? What if someone wanted more?

Less than a year later these questions became real when a delegation of French and British purchasing commissioners ordered more transmitters than Bill Halligan had built in the eight years his company had been in existence. Deliveries were a matter of life and death. And how about SX-28 receivers? Did he need money? Would cash in advance help speed things up? The heat was on. It was quite a shock to realize that Hallicrafters had a sold-out factory. Now, instead of selling people on the idea of buying Hallicrafters radio equipment, Bill had to sell other people on making them!

Hallicrafters was well on the road by the time of Pearl Harbor. Production was the all-absorb,, .g passion. The factory was geared to devise faster, more efficient assembly techniques. Three shifts were started, and by the fall of 1941, a steady stream of radio equipment was being shipped to the British (who had taken over the French commitments). On the morning of December 7, 1941, Colonel George H. Sparhawk of the Army Air Force telephoned Bill Halligan and told him that the Armed Forces were commandeering the British transmitters and that they were to be packed and out of Chicago by air that very night.

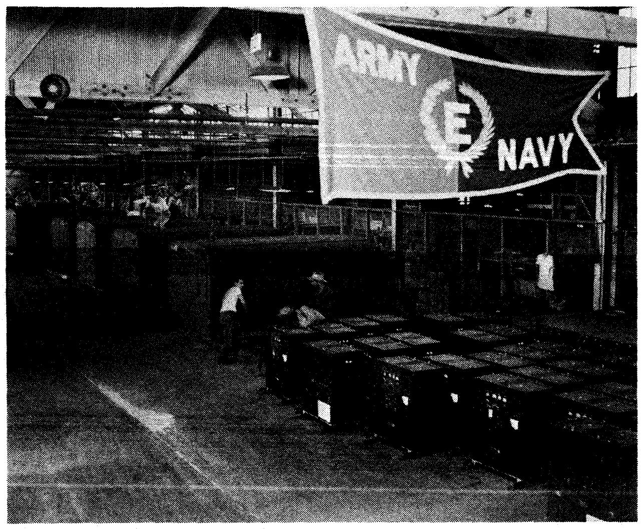
the war years the SCR-299 is born

Hitler's stunning victories in Poland, France, and the low countries revealed his secret weapons: speed, fluidity of attack, and ***instantaneous commu-***



Radio equipment installed in the famous SCR-299 or SCR-399 communication truck. At the right is the BC-610 transmitter and antenna tuner. At rear is the operating desk with two BC-342 short-wave receivers, speech amplifier, and control panel for the transmitter. In the foreground are the chests that contained extra transmission coils, tuning boxes, and spare parts.

***I still have my Hallicrafters S20R, purchased with paper-route money in the 1940s. Wouldn't sell it for any price. Editor, W6NIF**



Hallicrafters won five Army-Navy E awards during World War II for excellence in production. This photo shows BC-610 transmitters in final inspection before they were mounted in the communications vans in the background.

nications. The U.S. military service saw the need for a mobile radio station capable of communications over a wide range of frequencies, even while en route over rough terrain. The rough design for such a communications center was worked out at Fort Monmouth, New Jersey, under the direction of Colonel Roger B. Colton. Known as the SCR-299, the system consisted of a 1½-ton panel body truck with four-wheel drive coupled to a heavy-duty, two-wheel trailer. In the truck body was a complete short wave receivers (BC-342s). In the trailer was a gasoline-driven generator that was used to supply power for the mobile station. A rugged whip antenna was mounted onto the truck.

After long deliberation and tests, the ham-type HT-4 transmitter was selected for the SCR-299 communications truck. It was chosen because of its simplicity, small size, light weight, and ruggedness. And it had good, clear audio. Special military tuning units and handles were added to ensure quick and easy equipment removal from the truck. With some other minor changes, the ham rig was just right for military service.

the SCR-299 in military service

Production was now rolling. The SCR-299 was in service on every Allied battlefront from Alaska to China. SCR-299s were airlifted to Guadalcanal, North Africa, and Sicily. Among the first units to roll ashore at Normandy on invasion day was the SCR-299. Each SCR-299 carried a mile of telephone cable on a drum, permitting telephone lines to be set up for communications with distant areas. This enhanced the value of the SCR-299 where mobility was limited.

the Sebold incident

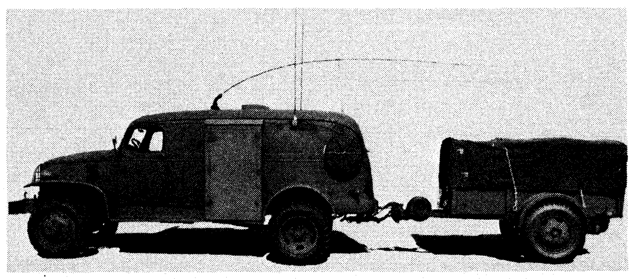
Finally, there was the celebrated case in which a German-American citizen, William Sebold, helped the FBI to apprehend a ring of Nazi spies. Sebold had returned to Germany to visit his family in 1939. Through coercion and intimidation, Sebold was forced to go through the great spy schoo! in Hamburg. When he left Germany in 1940, Sebold had instructions and authority to establish a short wave radio station in the United States for direct communications to Hamburg.

Upon arriving in New York, Sebold went straight to the FBI, which helped him establish a secret station on Long Island, equipped with a Hallicrafters HT-4 transmitter. He did a fine job of putting carefully doctored misinformation into the hands of the spy chiefs at Hamburg.

after the war

By 1946 the war was over, and Hallicrafters turned to the future. The Hallicrafters transmitter (known by the military label BC-610, part of SCR-299) was being released on the surplus market for as little as $500, complete with tubes, antenna tuner, and speech amplifier – only slightly more than the manufacturing cost. As far as Hallicrafters was concerned, there was no ham future in the HT-4 design.

To get peacetime production rolling, Bill Halligan brought out the S-38 and S-40 receivers (revamped



Always on the go! The SCR-399 communications truck was used to negotiate the surrender of Rommel, the "Desert Fox," in north Africa. The SCR-399 was present at the D-day invasion of Europe and accompanied General McArthur during his historic conquest of Leyte Island in the Philippines. Gasoline-powered generator is in trailer. Truck was supplied with three whip antennas for transmitting and receiving while in motion.

versions of pre-war models) and started to plan ahead. By late 1946 he proudly announced the SX-42, an advanced superheterodyne receiver that tuned from 540 kHz to 110 MHz in six bands.

The famous Hallicrafters SX-28 receiver. Over 50,000 were produced for the Allied Forces. High styling and good performance made the receiver a favorite among pre-war Amateurs. Receiver incorporated two rf stages and tuned up to 43 MHz. In 1946 it was replaced by the SX-42.

A simple television receiver was also in the works. Hallicrafters had heard about single sideband transmission. Yes, the future looked bright indeed in 1947 as Hallicrafters settled down to profitable, post-war production of modern radio equipment. They had the know-how.

More than fifty thousand SX-28 receivers had been built, and more than eighteen thousand SCR-299 (BC-610) transmitters had been built, as well as nearly ten thousand S-29 receivers. And now single sideband, television, and high-fidelity fm lay ahead, as well as CB radio on the distant horizon. The world looked good to Bill Halligan in 1947. He had really arrived in style!

The problem was conversion from a war-time to a peace-time industry. Hallicrafters was spread over fourteen plants, some as small as a garage. The first matter of business was to consolidate operations. A new facility was built, and two other locations were retained for growth.

Returning GIs remembered the name Hallicrafters. The company decided that, in addition to building ham gear, they would expand into the entertainment field. Television receivers, hi-fi equipment, and the Lowery organ were built by Hallicrafters. The company continued to build military electronic gear, particularly counter-measure equipment.

But all was not smooth sailing. The home radio and television market was cutthroat. By 1956 Bill Halligan was fed up. He couldn't compete and retain the high standards he'd set for Hallicrafters. The crisis of this uncomfortable situation was reached in the fall of 1957, when Bill Halligan sold out to the Penn-Texas Company. While Bill gave up ownership of Hallicrafters, he still retained management. This operation was even less satisfactory, and late the following year Bill bought the company back from Penn-Texas. Management of Hallicrafters was turned over to Bill's son, Bob Halligan. Bill remained chairman of the board.

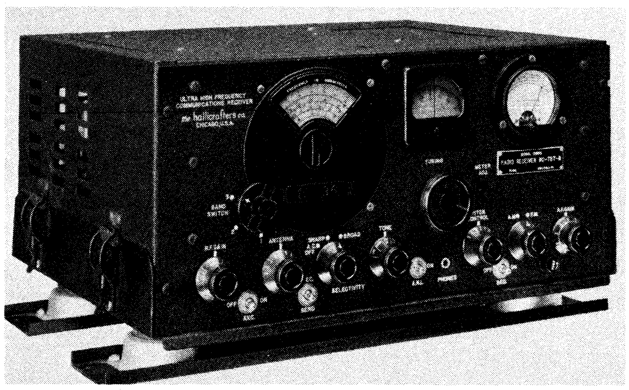
epilogue

Over the years, Hallicrafters contributed many firsts to the electronics industry and to Amateur Radio. Some of these were the first use of silk-screened panels in place of expensive engraving, the use of smooth paint in place of black-crackle paint, the calibrated S meter, the dual-diversity receiver, the automatic noise limiter, the ternperattire-coinpensated, high-frequency oscillator, the battery-portable, all-wave receiver, the dual AVG system, the bridge-T notch filter, and commercial production of an electronic keyer.

W9AC is still active on the air, mostly on 7 and 14 MHz CW. On occasion Bill can be found on 20 meter SSB. During the summer months he's on the air from W4AK in Florida.

Keep your ears open for this pioneering Amateur! Bill's interest in Amateur Radio is as keen as ever. Even though he's not manufacturing ham radio equipment, he can still use it along with the best of today's operators.

In 1950, J. Edgar Hoover of the FBI told Bill Halligan that the HT-4 transmitter Bill had sold to the FBI agent in August, 1941, had been flown to Pearl Harbor and that it had been installed in the hills above Honolulu. It was the only active communications link to the mainland during the Japanese attack, because saboteurs had cut telephone lines in Honolulu. The old W9WZE transmitter served its country well during those painful hours of need.



Large numbers of the Hallicrafters S-36 (BC-787B) receiver were sold during wartime. Tuning range was from 27 MHz to 140 MHz, which included vhf communications ranges of Allied. German, and Japanese ground and air forces. Receiverworked on both a-m and fm.

ham radio

phaselocked up-converter

Last of a three-part series on a frequency synthesized local-oscillator system for the high-frequency Amateur bands

In the first article of this series, I described the basic VCO synthesizer, or "first loop," which covers 100-1600 kHz in 10-Hz steps.[1] Part 2 covered the phaselocked 9-MHz BFO system.[2] This article describes the phase locked up-converter, which translates the 100-1600-kHz output of the first loop to the LO frequencies required for each of the Amateur bands between 160 and 10 meters.*

*Parts kits and circuit boards will be available from the author if there is sufficient interest. Send a self-addressed, stamped envelope for information.

For coverage of the high-frequency Amateur bands in 500-kHz segments, the VFO signal must be translated to a higher frequency for each band without degrading the phase noise performance. The output frequency is the sum of the first loop[1] output frequency and that of a stabilized crystal oscillator. This frequency plan is shown in **table 1. Fig. 1** is a block diagram of the up-converter. The output loop works by first mixing the VCO signal with that of the crystal oscillator (XO) then phase comparing the 1100-1600-kHz result with the first loop output to produce the VCO control voltage. The presteering system ensures that the loop acquires lock and remains locked despite large temporary differences between the VCO and XO frequencies caused by bandswitching or sudden large frequency changes. Crystal-oscillator frequency accuracy is guaranteed by a simple divide-by-*n* loop, with the phase detector operating at 50 kHz.

Crystal-oscillator section (fig. 2). Each 500-kHz band segment is selected by activating a crystal oscillator operating at a frequency 1100 kHz lower than the bottom end of the required LO frequency range. The collectors and the control voltage inputs of the nine oscillators are connected in parallel. The bandswitch, S1, applies base bias voltage to the selected oscillator, leaving the others cut off. Common-base

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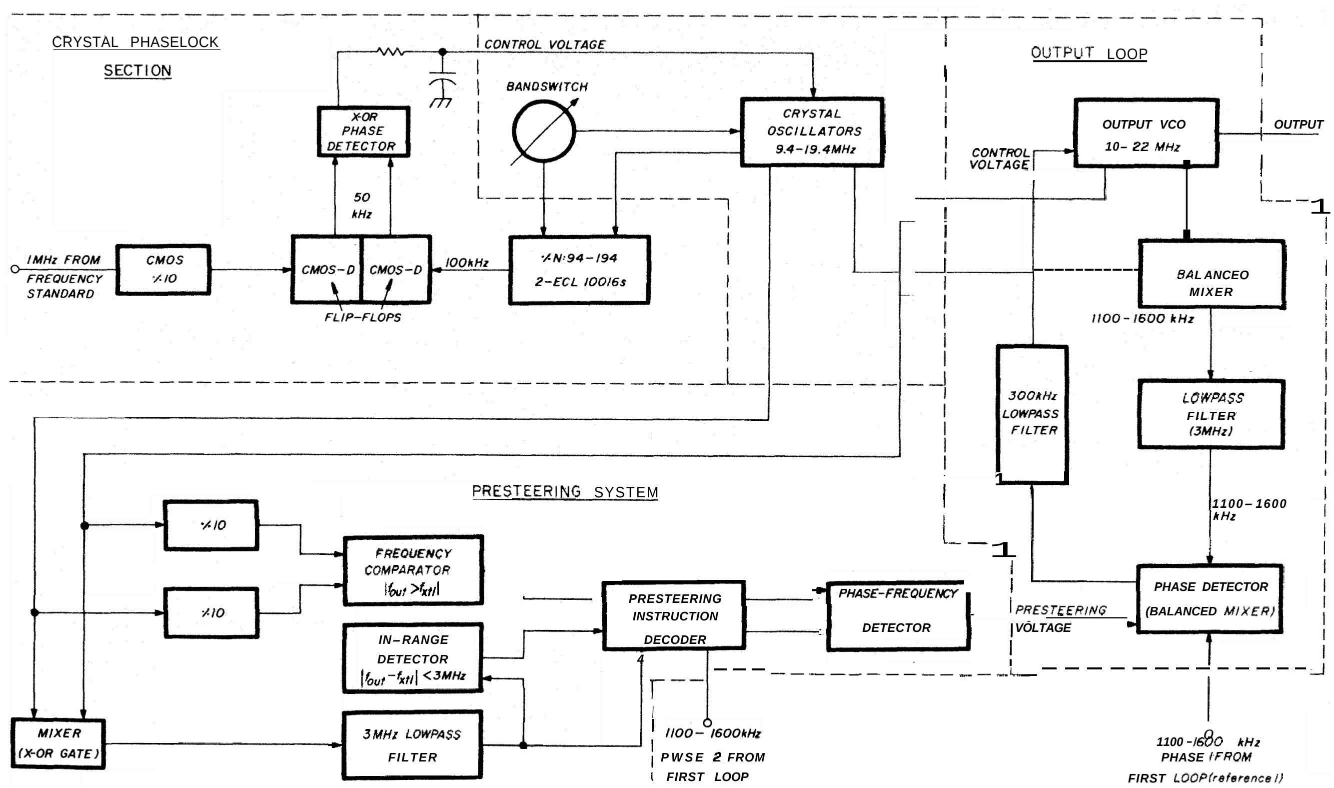


fig. 1. Up-converter block diagram.

buffer Q2 provides about 50 dB isolation between the *rf* output and the inputs to U2. The additional isolation provided by U2 and the PNP differential pair, Q3-Q4, effectively eliminates reverse signal-flow from the ECL and TTL circuits they drive.

Phaselock section (fig. 3). The ECL output of the crystal oscillator section is divided to 100 kHz by 8-bit programmable binary counter U6-U8. The diode matrix controlled by S1 sets the division ratio; the desired division ratio plus the binary number preset by the diode matrix must always equal 256. Because the phase detector requires symmetrical square-wave inputs for best performance, the 100-kHz outputs of the counter and reference divider U3 are first divided by two in U5. The combination of very low VCO gain in the crystal oscillators and the high reference frequency used in the phase detector make it possible to suppress reference-frequency modulation of the VCOs by more than 100 dB.

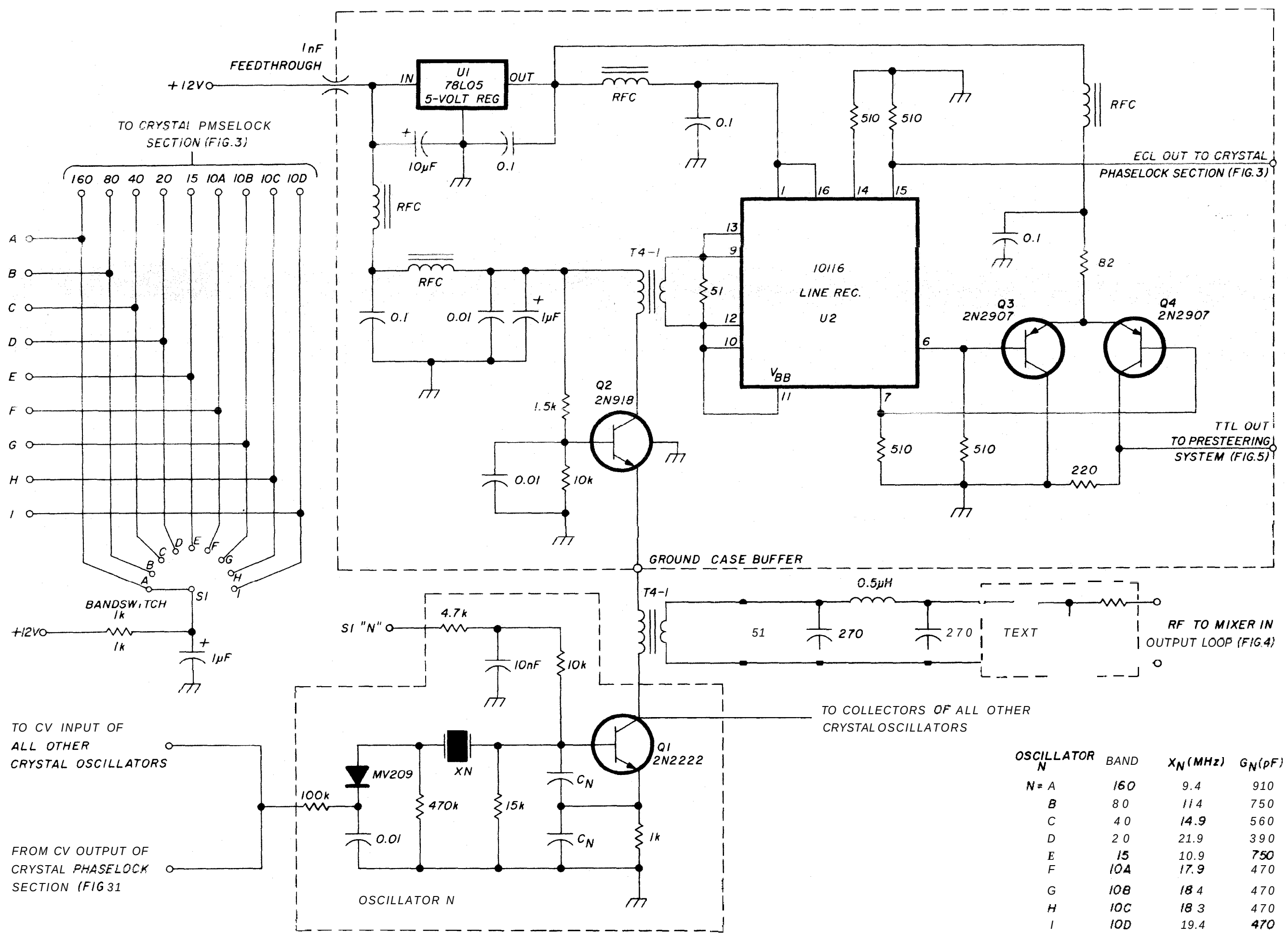
Output loop (fig. 4). With care and good layout, the loop of **fig. 4** will suppress the reference-frequency sidebands by at least 100 dB and reproduce the superb phase-noise performance of the first loop within about 6 dB.

Accompanying these advantages are some special problems. The capture range of the loop is about 250 kHz, and there are two frequencies where the unassisted loop could lock for each bandswitch setting and first-loop output frequency.

Assume the first loop has an output of exactly 1500 kHz and the bandswitch is set for 40 meters. Then the output of the down-mixer can be exactly 1500 kHz when output oscillator Q7 is set for 16.4 MHz (desired frequency) or 13.4 MHz (image). The phase detector alone produces a maximum of about ±100 millivolts. This is good for only about ±200 kHz of VCO tuning range. Because the i-f port of the double-balanced mixer is dc isolated, a presteering voltage can be added to the phase-detector output voltage to put the output oscillator frequency within correct locking-frequency capture range.

table 1. Frequency relationships for the frequency synthesized LO system.

crystal-oscillator frequency (MHz)		system output frequency (MHz)	±i-f (MHz)	frequency band covered (MHz)
9.4	+ 1100-1600 kHz =	10.5-11.0	−9	1.5- 2.0
11.4		12.5-13.0	−9	3.5- 4.0
14.9		16.0-16.5	−9	7.0- 7.5
21.9		23.0-23.5	−9	14.0-14.5
10.9		12.0-12.5	+9	21.0-21.5
17.9		19.0-19.5	+9	28.0-28.5
18.4		19.5-20.0	+9	28.5-29.0
18.9		20.0-20.5	+9	29.0-29.5
19.4		20.5-21.5	+9	29.5-30.0



OSCILLATOR N	BAND	X_N (MHz)	C_N (pF)
N = A	160	9.4	910
B	80	11.4	750
C	40	14.9	560
D	20	21.9	390
E	15	10.9	750
F	10A	17.9	470
G	10B	18.4	470
H	10C	18.3	470
I	10D	19.4	470

fig. 2. Crystal-oscillator section schematic.

Presteering system (fig. **5).** Two conditions must be met to guarantee correct lock. The first is that output oscillator Q7 frequency must be greater than that of the crystal oscillator. Call this condition **B**. This ensures that the system will not lock onto the image frequency. The second condition is that the difference between the two frequencies must be less than the 3-MHz cutoff of the lowpass filter following the mixer (condition A). A phase-frequency discriminator can take over at this point and bring output oscillator Q7 frequency to within capture range.

The presteering system is basically a crude downmixing loop (such as the output loop), except that it can detect when either – or both – of the above conditions are not being met, and will steer the output frequency in the proper direction until these conditions are met. U9-U12 test for condition B. The 4044 phase-frequency detector doesn't work reliably aboveafew MHz, so both inputs must first be divided by ten. Condition **A** is tested by the lowpass filter, ac peak detector, time-delay filter, and U17. If both conditions are met, the down-mixed signal from U16 and the first-loop signal through U20 are passed to U19 by instruction decoder U18. Phase-frequency discriminator U19 and amplifier U21 bring the system into lock at exactly the correct frequency.

If condition A is not met and condition B is met, the output oscillator frequency is more than 3 MHz above that of the crystal oscillator. U18 interrupts the normal signal flow and places U20 output on the R input of U19, causing U21 to pump down the presteering-voltage U17 signals.

If condition B *is not met, regardless of condition* **A,** the output oscillator frequency will be below that of the crystal oscillator. Thus U18 interrupts the signal from the downmixer while leaving the first-loop signal intact. This action causes the presteering output to pump up. Since, in this case, the output-oscillator frequency is below that of the crystal oscillator, these signals must pass through zerobeat in the downmixer before correct lock is possible.

But a zerobeat condition will give no ac signal for

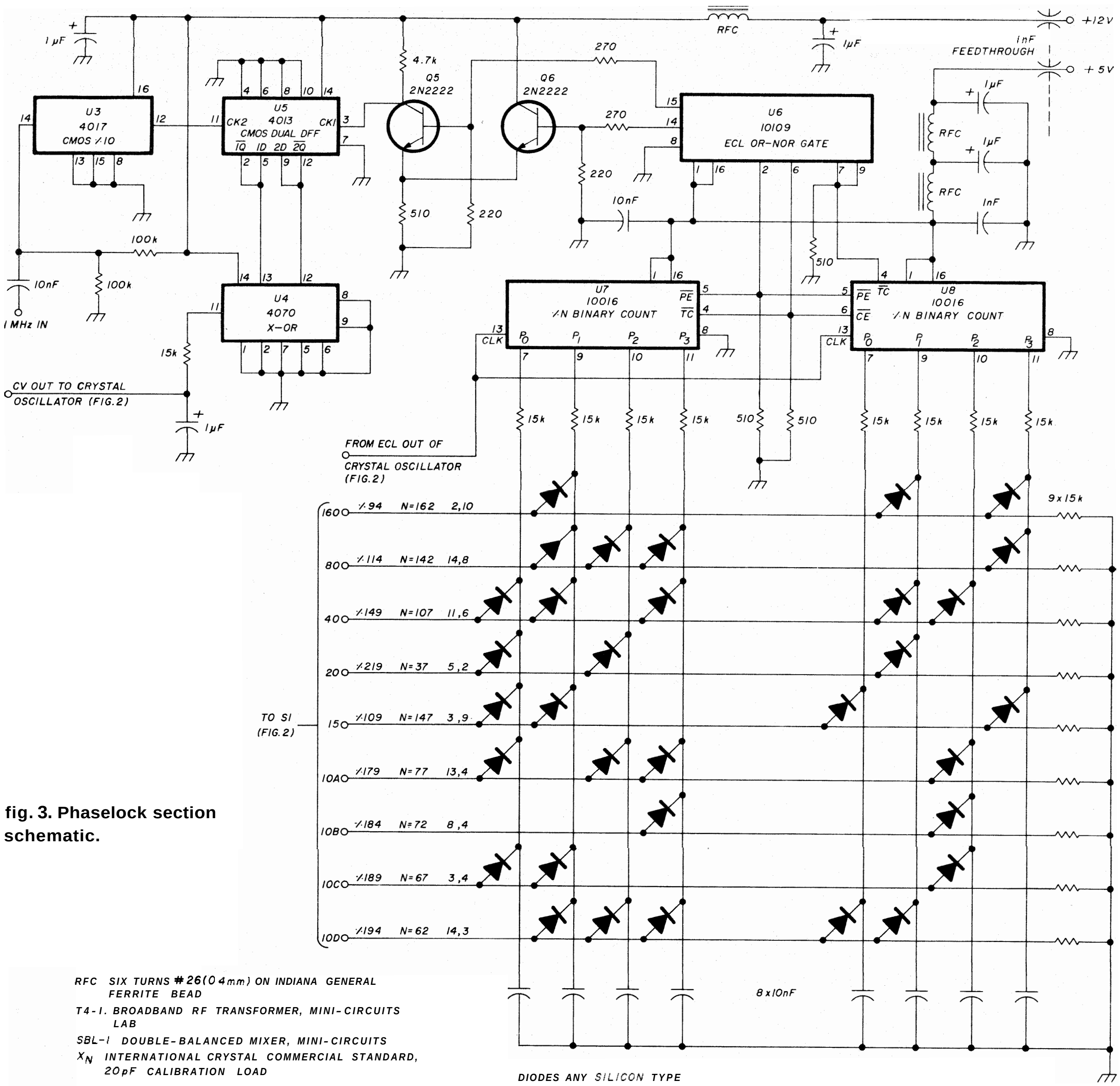


fig. 3. Phaselock section schematic.

the peak detector. This action would give a false alarm on condition **A,** except that the time-delay filter at U17 input keeps the input high until the beat returns. (Without this provision, the presteering system goes into a "dither" condition, which holds the output oscillator frequency close to that of the crystal oscillator.)

test procedures

For simplicity it's desirable to check out each section before connecting them. Sections already checked out can be used as test generators for successive sections.

Crystal oscillator. Set bandswitch S1 to 160 meters. The dc voltage at the base of Q-1 (the 9.4-MHz oscillator) should be about 6 volts. Check for the 9.4-MHz signal with a scope or counter connected to the collector. With a potentiometer, vary the control voltage to check that it oscillates over the entire range and that it is exactly 9.4000 000 at some setting near midrange. Check the ECL and TTL outputs for the 9.4-MHz signal. Repeat this procedure for each oscillator.

Crystal phaselock section. Connect the oscillator ECL output to the clock input of the programmable

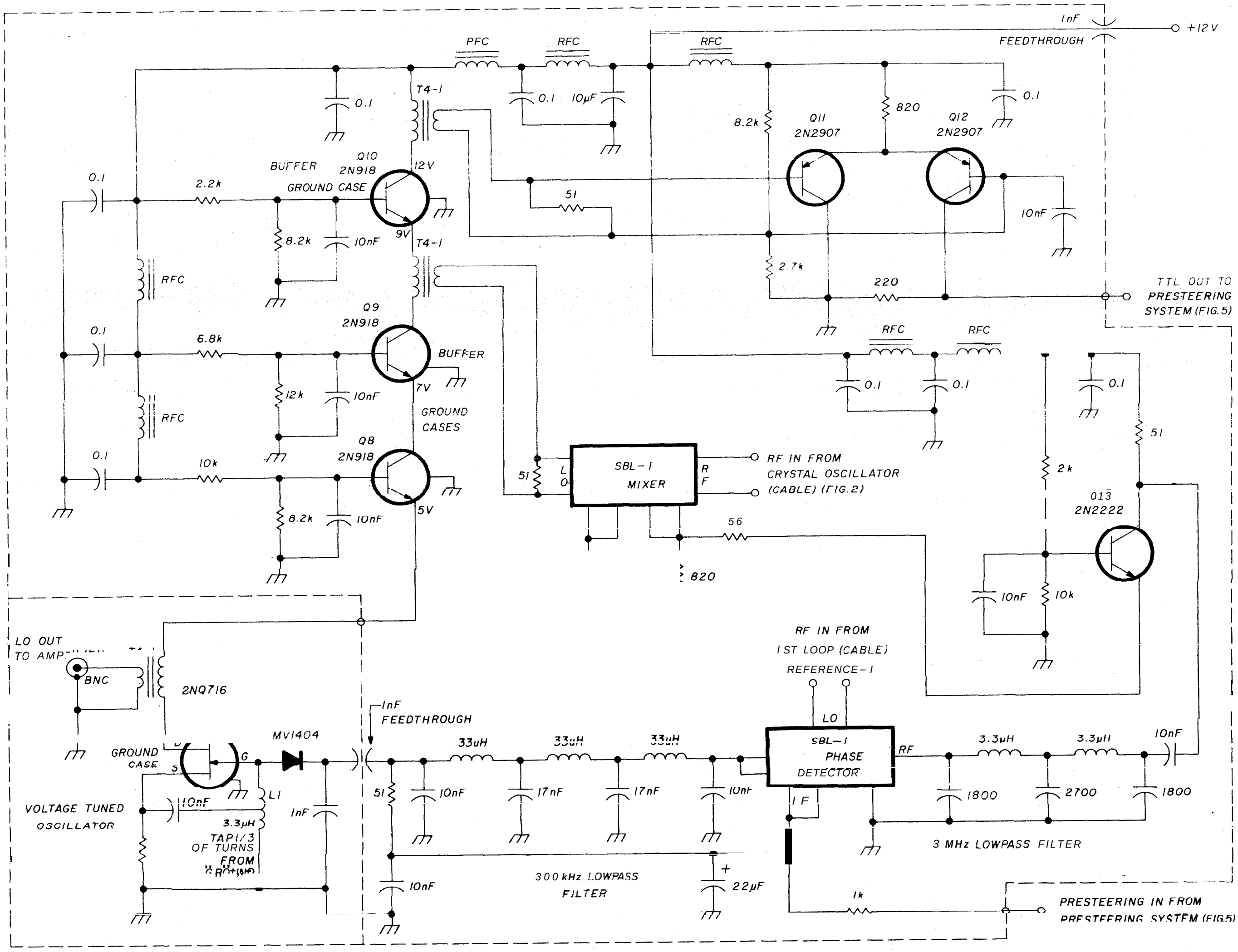


fig. 4. Output loop schematic.

divider, and make the connections to S1. Connect a counter to the collector of 05. For each setting of the bandswitch, it should be possible to adjust the control voltage (as before) so that the counter reads exactly 100 kHz. This checks the diode matrix and programmable counter. Connect a 1-MHz frequency standard to the reference input. U3 pin 12 should show 100 kHz. U4 pin 12 should be exactly 50 kHz, and U4 pin 13 should be very close to 50 kHz.

Check the CV output with a scope. As the CV input of the oscillator is varied, the phaselock section CV output should be a very-low-frequency triangle waveform, which goes through zero beat as the CV input voltage is brought through the middle of its range. Now close the loop by connecting the CV terminals of the two sections and removing the potentiometer. Put the counter on the oscillator-section TTL output and keep the scope on the (connected) CV terminals. Switch through all bands and check that the counter reads exactly the intended frequency and that the control voltage quickly settles to a constant dc level.

Output loop. Connect only the voltage-varying potentiometer to the presteering input and the counter to the rf output. Adjust L1 to obtain a frequency of approximately 15 MHz when the presteering voltage is 6 volts. Check for the TTL-level signal at the same frequency from the TTL output. Connect the crystal-oscillator section rf output to the downmixer rf port input. This signal level should be approximately 200 mV rrns. A 50-ohm T-pad on the oscillator output should be used to reduce the output to this level and provide a suitable termination for the mixer.

Set the bandswitch for 40 meters. Connect a scope to the 3-MHz lowpass filter output, which drives the phase detector. Vary the presteering voltage to observe the difference signal (zero to about 3 MHz) at a level of about 100 mV rms. Connect the first-loop rf output to the phase detector and set

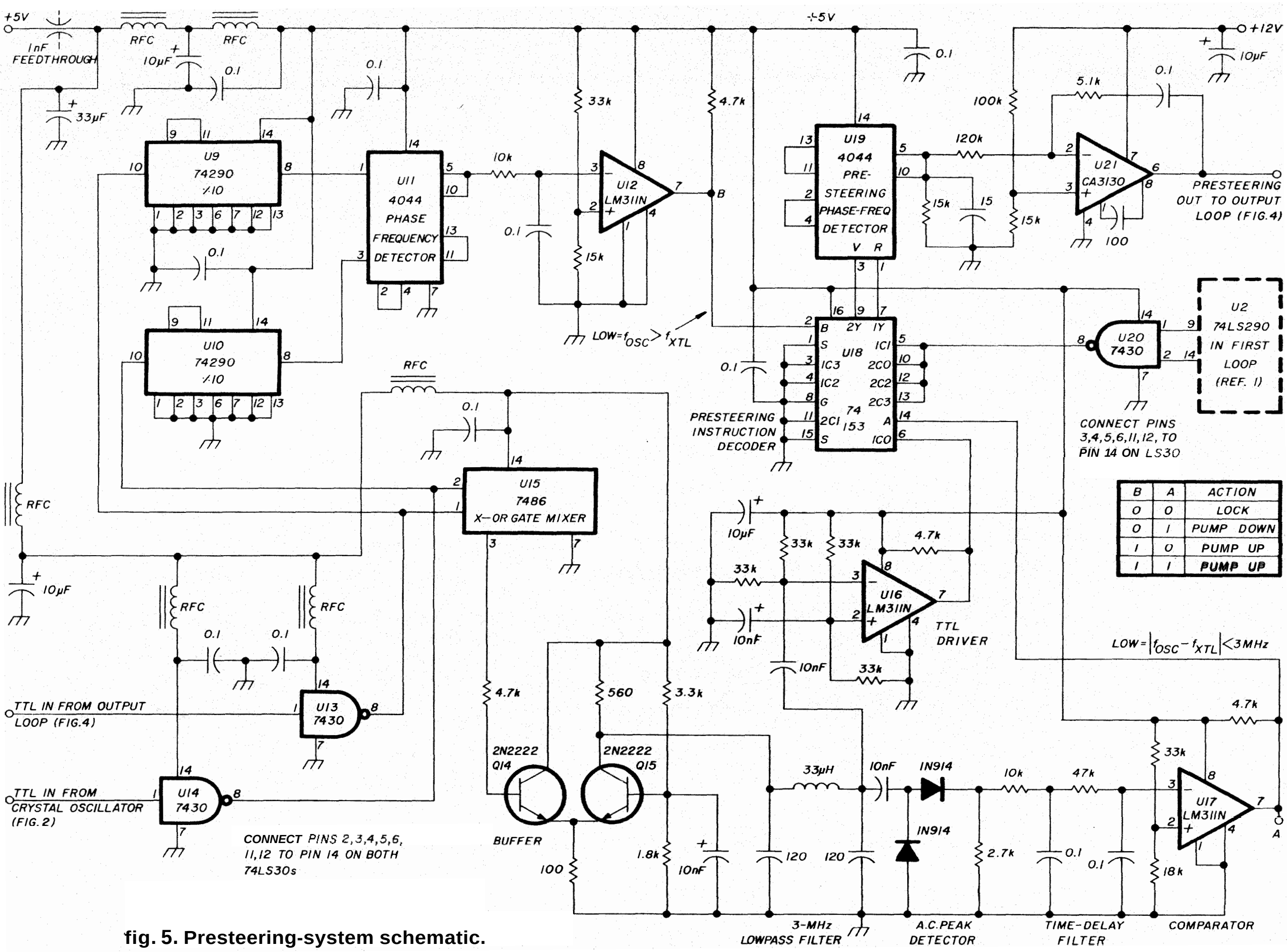


fig. 5. Presteering-system schematic.

the first loop for 1100 kHz (all switches zero). Adjust the presteering voltage so that the frequency is 16 MHz. For a small range of presteering voltage variation, the output frequency should remain at exactly 16 MHz.

Presteering system. Connect everything except the presteering output. Keep the potentiometer on the output loop presteering input. As you vary the output oscillator frequency slightly above and below 14.9 MHz, voltage B (U12, pin 7) should switch between zero and 5 volts. The output should be near zero volts when the output VCO is above 14.9 MHz.

Put the counter on U16, pin 7. It should show a square wave at the difference frequency. As this difference frequency goes above 3 MHz, approximately, voltage **A** (U17, pin 7) should jump from zero to 5 volts.

Now connect a voltmeter to the presteering output. When the VCO frequency is below 14.9 MHz, this output should increase to nearly 12 volts. When the VCO frequency is above 18 MHz it should drop to near zero volts. Remove the test instruments, connect the presteering output to the presteering input of the output loop, and then operate the system.

note on phase detectors

The phase detector in **fig. 4** yields its best performance when the two input signals are more than 30 degrees out of phase. Phase-frequency discriminator U19 in the presteering system requires that the two signals be *exactly* in phase. If both are operating from the same set of input signals, they work against each other, producing chaos. This may be eliminated by driving U19 from a separate phase-shifted output of the first loop. Thus, while U19 sees a zero phase difference, the SLB-1 mixer sees its inputs shifted by at least 60 degrees, resulting in stable operation. U20 delivers the required shifted output.

references

1. Raymond C. Petit, W7GHM, "Frequency Synthesized Local-Oscillator System for the High-Frequency Amateur Bands," *ham radio, October,* 1978, pages *60-65.*

2. Raymond C. Petit, W7GHM, "Phase-Locked 9-MHz BFO," *ham radio,* November?1978, pages 49-51.

ham radio

wideband amplifier summary

A powerful tool for designing wideband amplifiers using transformer feedback — features include low intermod distortion and low noise figure

Feedback amplifiers have been used in solid-state circuits for many years. Applications include wideband amplifiers for undersea cables, instrumentation amplifiers, and antenna amplifiers. For receiver front ends it's essential to combine good input impedance matching, low noise figure, sufficient gain, and a high intercept point. Good linearity can be achieved by using resistive feedback while sacrificing noise. A new circuit is presented combining all these advantages. It can be produced at very low cost.

feedback intermod, and noise figure

Bipolar transistors have a number of inherent nonlinearities:

1. Exponential base-emitter diode characteristics

2. Current-dependent diffusion-layer input capacitance

3. Voltage-dependent depletion-layer output capacitance

The distortion is highly dependent on the generator source impedance. If the generator source impedance is very small compared with the transistor1 input impedance, the input voltage will be directly converted into an output current. This exponential transfer characteristic is responsible for all current nonlinearities.

The only cure for this type of distortion is current feedback. But current feedback provides two unpleasant side effects: it increases the device input and output impedance and therefore creates a mismatch. If the transistor is driven by a current-source generator of infinite impedance, distortion will depend mainly on the current-gain linearity. Since the output is converted into a voltage gain, voltage distortion will result. The only cure for voltage distortion is voltage feedback. Voltage feedback has the disadvantage of reducing the device input and output impedance.

The standard technique for current feedback is to use an unbypassed emitter resistor. This emitter resistor adds a significant noise contribution to the circuit, which is not phase correlated to the transistor internal noise sources. The resulting noise figure is typically between 6 and 10 dB. An amplifier with such performance cannot be considered a low-noise circuit.

Voltage feedback is accomplished by using a resistor feeding voltage from the collector to the base. Again we find resistive losses resulting in noise,

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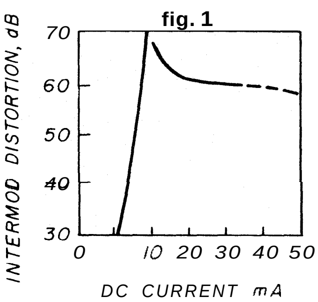
which appears amplified as a product of the transistor gain at the output and which further reduces amplifier performance. Typically the noise figure of a wideband amplifier using this type of feedback as the sole source is at best 4 dB when ultra-low-noise transistors such as the Siemens BFT66 are used. The noise figure of the same circuit, at the same dc operating point without any feedback, is about 1 dB. An attempt should therefore be made to maintain the 1-dB noise figure while increasing the dynamic performance.

The intermodulation distortion as well as the noise figure is dependent on the emitter current. **Fig. 1** shows the two-tone test performance for two carriers of zero dBm at the input as a function of the dc current. It's obvious that the performance is not improved above 10 mA, which also indicates that the cutoff frequency peaks around 10-15 mA. This performance will vary from transistor to transistor. A



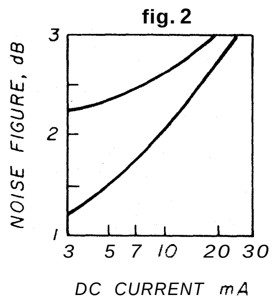


fig. 1. Two-tone test results (left) for two carriers of zero dBm at the input of a wideband amplifier as a function of dc emitter current. Improvement is limited above 10 mA.

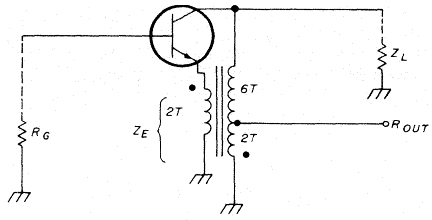
fig. 2. Noise figure (right) of a wideband CATV transistor amplifier versus emitter current. The transistor is a BFT65 used in a recent design for wideband antenna amplifiers.

typical *CATV* transistor, such as the 2N5109, has a flat curve of constant intermodulation distortion produced between 20 and 80 mA.

Fig. 2 shows the noise figure of a wideband *CATV* transistor as a function of dc current. This transistor (a BFT65) is a recent design for wideband antenna amplifiers. However, similar performance can be achieved with the less-expensive 2N5109. It's obvious that until the amplifier is driven to a level that the output voltage swing gets close to the collector dc voltage, the distortion and intermodulation is caused by the *exponential transfer function.*

noiseless feedback

If we use a feedback system as shown in **fig.** 3, in which the collector resistor is transformed back in series with the emitter by a transformer, the emitter will be grounded through a resistance depending upon the collector load. It's apparent that the voltage



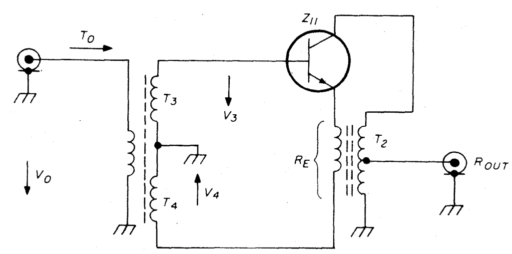
$$\frac{Z_L}{R_E} = \frac{Z_L}{Z_E} = \left(\frac{N_1}{1}\right)^2 \frac{Z_L}{R_{OUT}} = \left(\frac{1}{N_3}\right)^2$$

fig. 3. Feedback circuit in which the collector resistor is transformed back in series with the emitter by a transformer. Transistor emitter will be grounded through a resistance depending on the collector load. thus voltage gain is independent of load changes. The core is a Siemens B62152-A0004-X001.

gain is independent of any load changes, which results from the definition that the voltage gain is equal to the collector load resistor values divided by that of the unbypassed emitter resistor. Since this emitter resistor is derived by feedback, it adds no noise contribution. Therefore, this type of feedback is called "noiseless feedback."

Let's assume that the transformer collector-to-emitter turns ratio is three to one. Then the impedance ratio is nine to one, and the 450-ohm collector load will result in a 50-ohm emitter-current feedback impedance. Ideally this impedance is resistive. If the transistor is operated at 20 mA, the differential output impedance from emitter to ground will be about 1 ohm, assuming a current gain of fifty and a generator impedance for the transistor stage of 50 ohms. Therefore, we have a voltage division whereby the input voltage between base and ground is divided by forty-nine parts across the emitter impedance and one part across the base-emitter junction. The amount of current feedback is fifty and the linearity improvement is also roughly fifty, or 33 dB. The third-order intercept point, as shown previously without feedback, was about 30 dBm and has now been increased by 33 dB, resulting in about 63 dB.

The power gain of this stage can be calculated



$I_4 = I_2 = I_1; V_4 = -V_3 N_2; V_3 = I_1 \cdot Z_{11} - I_4 R_e + V_4$

$I_4 = -I_3 N_2; V_0 = V_3; I_0 = I_1 + I_3; I_2 = -I_1 B^{-jF/F_T}$

fig. 4. Schematic diagram of a bridge-type circuit which adds voltage feedback in order to lower the circuit's input impedance.

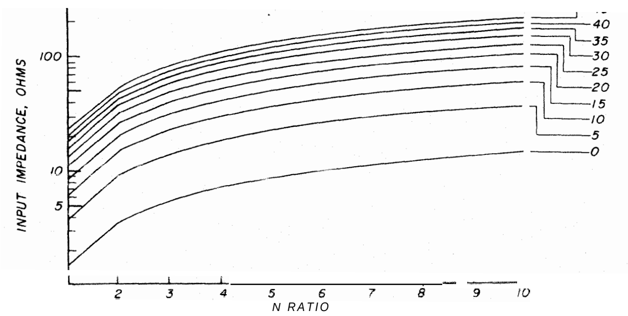


fig. 5. Voltage and current feedback amplifier using a 2N5109. Shown is input impedance (ohms) as a function of transformer turns ratio with R_E as a parameter.

from the voltage gain, which would be ten divided by the impedance scaiing at the output, from 450 ohms to 50 ohms. Therefore a voltage gain of ten divided by three results in a power gain of about three. This power gain is defined by the input and output, whereby the transistor input impedance has become fairly high in value.

As stated earlier, the second source of distortion is voltage distortion. We will now apply voltage feedback to decrease the input impedance to a suitable value, such as 50 ohms.

voltage feedback using a transformer

Fig. 3 shows a bridge circuit, which transforms the transmitter emitter-to-ground impedance to a value determined by the bridge transformer turns ratio and puts it in parallel between base and ground. This feedback reduces the output impedance and is therefore counteractive to the current feedback. For developing the mathematical equations, R_E is the emitter and bypassed resistor value, which has been obtained by using a collector transformer. The various feedback network voltage and currents are included. While the mathematical derivation of this circuit takes a few minutes, only the results are shown.

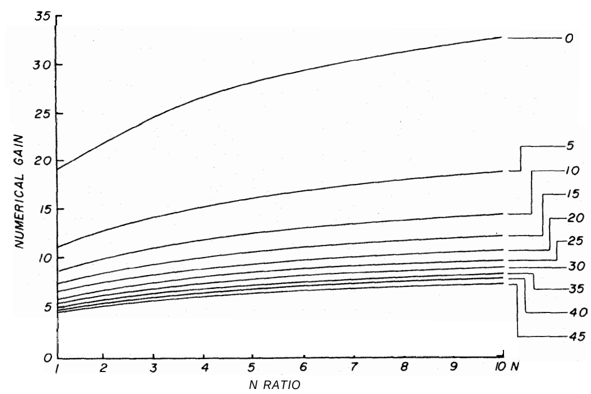


fig. 6. Same amplifier as in fig. 5. Gain as a function of transformer turns ratio is shown with R_E as a parameter.

The circuit input impedance is:

$$Z_{in} = \frac{Z_{11} + R_E \beta C^{-jF/F_T}}{\left(1+\frac{1}{n_2}\right)\left(1+\frac{1}{n_2}\beta e^{-jF/F_T}\right)}$$

$$= \left(\frac{n_2}{\left(1+\frac{1}{n_2}\right)}\right)\left(R_E + \frac{26\,mV}{I_e}\right) \quad \textbf{(1)}$$

and the output impedance is:

$$Z_{out} = \left(\frac{R_G + Z_{11}}{\beta}\right)\left(\frac{n_1}{n_3}\right)^2$$

$$\approx \left(3\Omega + \frac{26\,mV}{I_e}\right)\left(\frac{n_1}{n_3}\right)^2 \quad \textbf{(2)}$$

The power gain has been determined as:

$$P_G = \left(\sqrt{\frac{Z_L}{Z_{in}}}\right)\left(\frac{\left(1+\frac{1}{n_2}\right) Z_L}{\left(R_E + \frac{26\,mV}{I_e}\right)}\right) \quad \textbf{(3)}$$

$$\text{and } R_E = Z_L \left(\frac{1}{n_1}\right)^2$$

$$Z_L \approx 500\Omega; \textbf{from } Z_L = \frac{(V_{\beta AT} - V_{SAT})^2}{2P_{OUT}}$$

test data

Figs. 5 and **6** are the results of computer runs showing test data obtained with an experimental wideband amplifier using a 2N5109 transistor. **Fig. 5** shows input impedance (ohms) as a function of transformer turns ratio, N, with R_E as a parameter (ohms). ***Fig.* 6** shows amplifier numerical gain as a function of transformer turns ratio, N, again with R_E as a parameter. To obtain a 50-ohm input impedance with $R_E = 10$ *ohms*, for example, a transformer turns ratio, N, of about eight is required. For $R_E = 50$ *ohms*, the required turns ratio is about two.

conclusion

It's apparent that this circuit in its final form, providing an intercept point for third-order distortion of more than 70 dBm, is a very powerful tool in designing new wideband amplifiers. Because of transformer feedback, the noise figure is only about 2 dB. In a pushpull version, this circuit has shown a second-order intercept point of more than 120 dBm. These numbers appear to be much better than those for previously published amplifiers.

ham radio

improved GaAs fet preamp for 144-432 MHz

Experiments with a new mesfet from NEC featuring simplified bias and amplifier circuits

Since its publication in the April, 1978, issue of ham *radio,* my article describing a 432-MHz low-noise preamplifier using a NE24406 GaAs fet[1] has resulted in many inquiries from readers. These inquiries have prompted some experiments with a new device, the NE24483 GaAs mesfet (metal semiconductor fet). The NE24483 has characteristics identical to those of the NE24406, but it costs less. This article presents the results of my experiments with the new device, which include

1. bias-circuit simplifications for the 432-MHz preamp;

2. applications of the NE24483 to 144-MHz amplifiers; and

3, circuit simplifications for the 432-MHz preamp.

simplified bias circuit for 432-MHz preamp

Fig. 1 shows the preamp circuit in reference 1. Separate power sources are arranged for the gate circuit (minus voltage) and the drain circuit (plus voltage). (This circuit is discussed later.) When EME communications (*i.e.*, high-power systems) are considered, this is the safest bias circuit for fets. However, with GaAs fets, if the drain voltage is applied first, a current will flow that reaches I_{DSS} (saturation current when the gate voltage is zero); in some low-noise transistors this current may reach 100 mA. Therefore, it's desirable that a method be used that always switches on the minus voltage to the gate. However, it's difficult to provide such a minus voltage.

Fet bias circuits. Fig. 2 shows five methods for supplying bias to fets.[2] Each method has its advantage, and no method can be said to be the best; however, the method easiest to use (considering component mounting and operation) has been employed.

Although the method shown in **A** of **fig. 2** is a bother to implement, it's a superior bias method for fets at extremely high frequencies, as in an 18-GHz amplifier. This is because, with this method, the source can be directly grounded, and the grounding inductance can be maintained smaller than with any other method. So this method will be significant when high gain at the high-frequency bands, or a low-noise amplifier, is desired.

In all methods other than **A,** a bypass capacitor is inserted in the source. Of these methods, D and E require only one power source. If a sudden increase in supply voltage occurs, series resistor R_S is connected to the source (D and E, **fig. 2**) so that a voltage, E, will appear across the R_S terminals and will automatically suppress the voltage increase between drain and source:

$$E = (R_S)(\Delta I_D) \quad (1)$$

where ΔI_D is the increment of drain current caused by a sudden increase of supply voltage

Transistor protection is thus automatic.

Although circuit E **(fig. 2)** has this feature, circuit

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E is inconvenient to implement because a minus power supply must be used. In general, circuit D should be used.

Design. Generally, about 15 per cent of I_{DSS} is optimum as the recommended biasing point of these transistors for low-noise applications. Characteristics are shown in table **1.** Table **1** shows 60 mA as a typical value for I_{DSS}, so the bias circuit design is based on this value. A value of 20 milliohms is typical for the gm (transconductance) when $V_{DS} = 3$ *volts* and $I_D = 10\ mA$.

To determine R_S when $I_D = (0.15)\ (I_{DSS}) = 9\ mA$:

$$R_S = \left(\frac{I_{DSS}}{I_D} - 1\right) /gm \qquad \textbf{(2)}$$

R_S is determined as:

$$R_S = \left(\frac{60\ mA}{9\ mA} - 1\right) /20\ mmho$$
$$= 5.67/0.02$$
$$= 283 \text{ ohms}$$

If a drain current, I_D, of 9 mA flows when $R_S = 280$ ohms, the voltage across R_S will be:

$$R_S I_S = (280)\ (0.009) = 2.52 \textit{ volts} \qquad \textbf{(3)}$$

If the voltage, V_{DS}, between drain and source is set to 3 volts, it will be sufficient for the bias if the power supply delivers $V_{DS} + (R_S I_D)$:

$$V_{DS} + (R_S I_D) = 5.52 \text{ volts}$$
$$= 3 + [(280)\ (0.009)] = 5.52 \text{ volts}$$

The bias circuit is now complete; its design is shown in fig. **3.** Fig. **4** shows the fet amplifier of reference 1 in which the bias circuit has been arranged to that described above.

Practical considerations. If a 5-volt, three-terminal voltage regulator is used in the power supply, further protection against damage is provided. Needless to say, in this case the voltage between drain and source will be somewhat lower than 3 volts, but absolutely no change will be noticed in actual use.

Attention should be paid to the value of R_S when the fet gm, V_p (pinch-off voltage), or I_{DSS} is irregular. R_S should be set to a value between 210 and 280 ohms at which the drain current, I_D, is the specified value.

With a preamplifier employing this biasing method (self-bias), supplying bias will be very easy, even when the amplifier is mounted directly under the antenna.

144-MHz band preamplifier employing GaAs fets

The impedance characteristics of the NE24406 in the 435-MHz band were shown in reference 1. These characteristics are shown in fig. **5.**

Γ_{FOPT} indicates the impedance at which the noise figure, NF, becomes minimum when the transistor input circuit is matched to this impedance. Theoretically, this value has the following meaning. It shows what noise figure, NF, will be obtained when a certain impedance is connected externally to an element having an intrinsic minimum noise figure of F_0.

$$NF = F_0 + \frac{R_N}{G_S}\left[(G_S - G_0)^2 + (B_S - B_0)^2\right] \qquad \textbf{(4)}$$

Here, G_0 and B_0 are the conductance and susceptance, respectively, when the noise figure is minimum. They have the following relationship:

$$\Gamma_{FOPT} = G_0 + jB_0 \qquad \textbf{(5)}$$

From eq. **4** it can be seen that the minimum noise figure will occur when $G_S + jB_S = Z_S$ becomes:

$$\Gamma_{FOPT}\ (G_0 = G_X;\ B_0 = B_S) \qquad \textbf{(6)}$$

When any other impedance is connected, a noise figure is obtainable that's always worse than the case where $Z_S = \Gamma_{FOPT}$.

table 1. Typical electrical characteristics of the NE24406 and NE24483. Ta = **25C** *(32F)*

	symbol	conditions		min.	typ.	max.	unit
drain current	I_{DSS}	V_{DS} = 3.0 V,	V_{GS} = 0	30	60	100	mA
pinch-off voltage	V_P	V_{DS} = 3.0 V,	I = 100 μA	– 1.5	– 3.5		V
maximum oscillation frequency	f_{max}	V_{DS} = 3.0 V,	I = 30mA		55		GHz
transconductance	Gm	V_{DS} = 3.0 V,	I, = 10 mA		20		mV
maximum available power gain	MAG	V_{DS} = 3.0 V, I = 30 mA	f = 4.0 GHz		17		dB
			f = 8.0 GHz	10	12		dB
			f = 12.0 GHz		9		dB
noise figure	NF	V_{DS} = 3.0 V, I_D = 10 mA	f = 4.0 GHz		1.5		dB
			f = 8.0 GHz		2.7	3.8	dB
			f = 12.0 GHz		3.7		dB

Now take another look at **fig. 5.** Γ_{FOPT} is at the edge of the Smith chart in a position where it is a predicted value for the 144-MHz band. In any case, a matching circuit for this value seems feasible.

The experimental 144-MHz GaAs fet schematic is shown in **fig.** 6. The pi network in the output is a 3-dB attenuator. (When these transistors are used in the 144-MHz band they may oscillate.)

A Johanson air trimmer is in series with the input, since only 4 to 5 pF is required. An inexpensive Philips trimmer capacitor could also be used. **Fig. 7** shows assembly and simple structural drawings of the preamplifier.

Incidentally, an $NF \leq 0.7$ dB and a $gain \geq 22$ dB were obtained with this amplifier. These component values are approximately the same as those of the 430-MHz preamplifier previously described. They show that, at frequencies in this range, no improvement in noise figure occurs as a result of lower frequency, and the noise figure has a flat characteristic.

BNC connectors have been used in the input and output, but type N connectors would probably be better. The preamp shown in **fig.** 6 employs self-bias. The bias circuit described previously can be used.

further simplification of the 432-MHz system

At the beginning of this article I described a simplified method for supplying bias **(fig. 4)**. However this circuit employs four expensive air trimmer caps. I've received some comment about the difficulty of adjusting these air trimmers, so further simplification is in order.

The idea was to match the input circuit using a single fixed capacitor for $C1$ instead of the air trimmer **(figs. 4** and **6).** Theoretically it should be possible to replace the variable cap with a fixed cap of the correct value to obtain minimum noise figure.

Take another look at **fig. 5.** For minimum noise figure in the 430-MHz band, the impedance should be $50 + j400$. This means that a $j400$ reactance should

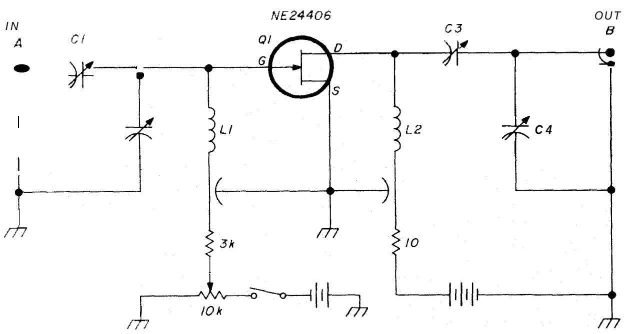


fig. 1. The low-noise 432-MHz preamp described in reference 1. An improved self-bias circuit has been designed (fig. 4).

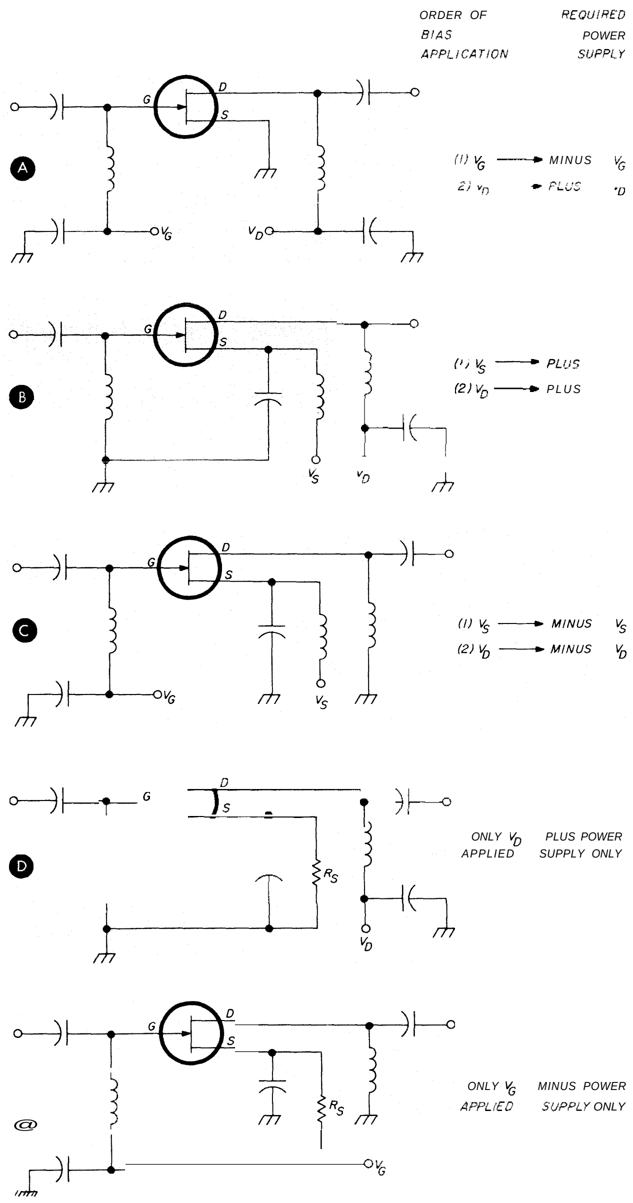


fig. 2. Five methods for supplying bias to GaAs fets (from reference 2).

be connected to the 50-ohm line. Close examination of **fig. 5** shows that an impedance of $50 + j400$ is situated on the 50-ohm impedance line at a point in a counterclockwise direction when seen from the center of the Smith chart. A reactance element that gives a trace moving in a counterclockwise direction on the impedance line is a series capacitance. With this information, it can be seen how the GaAs fet preamplifier is designed.

The $j400$ impedance is expressed by:

$$\frac{1}{2\pi f C} = j400 \qquad \textbf{(7)}$$

where $f = 435\ MHz$

$$C = \frac{1}{400 \times 2\pi f}$$

$$= \frac{1}{400 \times 2\pi \times 10^6}$$

$$= 0.915 \times 10^{-13}$$

$$= 0.915\ pF$$

Therefore, it can be seen that, matching for a minimum noise figure, NF, can be accomplished by employing a 0.915-pF series capacitor.

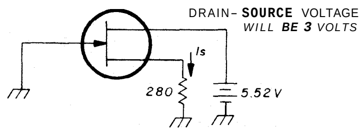


fig. 3. Self-bias circuit for the NE24406/NE24483.

In practice, the circuit will be affected by the circuit series inductance. However, since a series inductance produces a trace which moves in a clockwise direction on the impedance line, the capacitance of the series capacitor must be increased to compensate this inductance.

As seen from the size of the chassis used, the inductance of the capacitor leads can be estimated to be several tens of nH. About 1.2 pF can be considered optimum. Therefore, it will be ideal if the air trimmer in the input circuit and the single-turn coil resonate at the desired frequency in the 432-MHz band and they are employed only as an infinitely large impedance.

The circuit is shown in fig. **8.** In this circuit the number of air trimmers has been reduced by one.

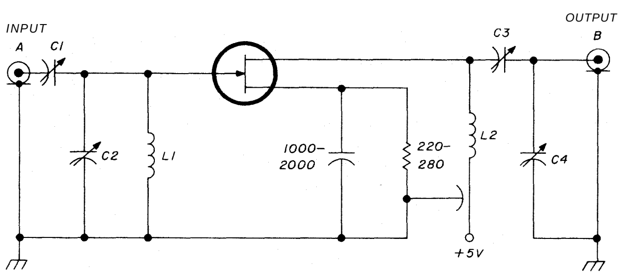


fig. 4. GaAs fet preamp using self-bias. Reference 1 shows component values.

With the method of connecting a 1.2-pF fixed capacitor in series with the input, a noise figure of $NF \leq 0.7\ dB$ was obtained, as were characteristics identical to those of the preamplifier presented before.

system sensitivity

All these preamps have a noise figure, NF, less than 1 db; thus they are candidates for receiving systems of tremendous sensitivity, but this isn't easy to attain. In a high-sensitivity receiving system, thought must be given to the system *as a whole,* including the antenna and coaxial cable.

Noise at receiver terminals. Here the relationship between antenna and receiver sensitivity is discussed. In fig. **9,** a "no-loss" bandpass filter with a bandwidth of B is assumed. A load resistor is connected across the output terminals, and a resistor equal to the filter input impedance, within the passband range, is connected to the filter input terminals.

In this state, when the resistor connected to the filter input is maintained at absolute temperature, T *(degrees Kelvin),* thermal noise will be generated by

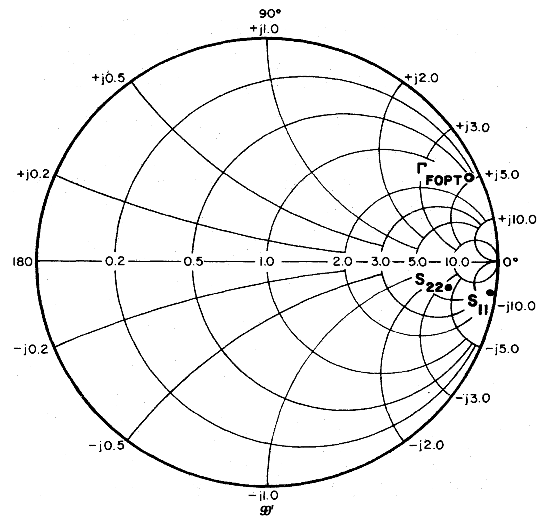


fig. 5. NE24406/NR24483 impedance characteristics.

this resistor and will flow into the filter output load.

The noise power, N, flowing into the load resistor is:

$$N = \frac{hfB}{e^{hf/kT} - 1} = kTB\ (hf << kT) \qquad \textbf{(8)}$$

where

h = Planck's constant (6.62×10^{-34} *joules/sec.)*
k = Boltzmann's constant (1.38×10^{23} *joules/deg.)*

Eq. 8 shows that the noise generated by the resistor is proportional to the *absolute* temperature.

Accordingly, this absolute temperature is called the noise temperature. The magnitude of the noise can be expressed by the noise temperature, T; the noise power, N, can be expressed as $N = kTB$.

In the example above, I've shown the results of thermal noise generated within a typical receiver input circuit. The resistor connected at the filter input

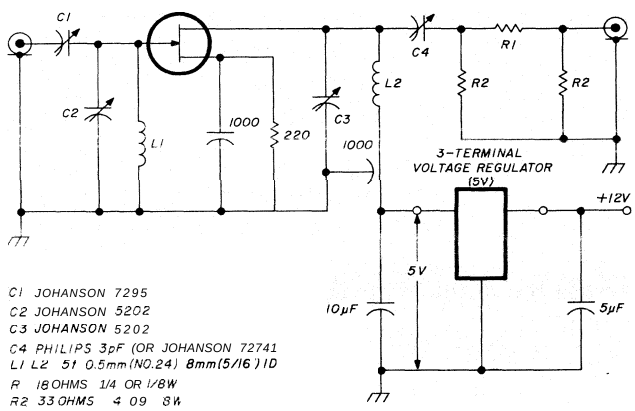


fig. 6. GaAs fet preamp for 144 MHz using self-bias.

represents this noise. But what about other noise, such as that entering the receiving antenna from space?

In this case, the receiving-antenna output terminals are connected to terminals 1 and 1' (fig. **9**) instead of the resistor.

Noise coming from space or artificial (manmade) noise will appear at the antenna output. Connect a noise-power meter of N watts to the bandpass-filter output. Determine:

$$Ta = \frac{N}{kB} \qquad \textbf{(9)}$$

which is the antenna noise temperature. This temperature is the same as that of the resistor in fig. **9,** whose thermal noise exactly replaces the noise coming into the antenna.

Receiving-system noise characteristics. As shown in fig. **10,** the antenna is connected to the receiver through a transmission line (coaxial cable). In this case, a) the noise coming into the antenna, b) the thermal noise generated by the resistive loss of the transmission line, and c) the noise generated inside the receiver are compounded. Let's convert all of these noise powers to their equivalent at the receiver input terminals.

Antenna noise. If the antenna noise temperature is assumed to be Ta, then, as described before, the noise power flowing into the transmission line from the antenna output will be $kTaB$. Now, if the transmission-line insertion loss (coaxial cable) is taken as $10 \log_{10} L$ (dB), the antenna noise power flowing into the receiver input terminals will be:

$$\frac{kTaB}{L} \qquad \textbf{(10)}$$

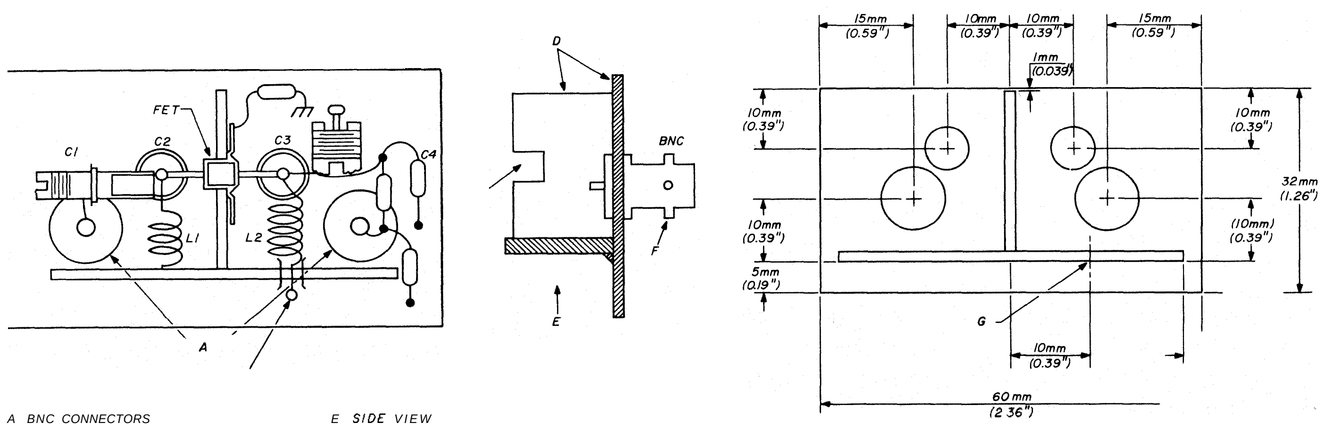
where L is the cable insertion loss.

Transmission line noise. The transmission-line absolute temperature is assumed to be T_0K, and a matching load resistor is connected in place of the antenna in fig. **10.** Then a load resistance is connected, and the load resistance and transmission line are maintained at a temperature of T_0K.

The noise power occurring at the receiver input terminals will be kT_0B. Of this noise power, the portion generated by the matching load resistor, which appears at the receiver input terminals, will be kT_0B/L. Therefore, the actual thermal noise power generated by the transmission line will be:

$$kT_0B\left(1 - \frac{1}{L}\right) \qquad \textbf{(11)}$$

If the line has no loss, L will become 1; therefore, from eq. **11,** noise generated by the line will become zero. This is a natural result, considering the principle that thermal noise is produced by resistance.



A BNC CONNECTORS
B 3-TERMINAL REGULATOR
VOLE FOR PLACING TRANSISTOR
MADE OF DUPLEX GLASS-FACED EPOXY
E SIDE VIEW
F BNC CONNECTOR
G PARTITION BOARD. BOTH SIDES PC BOARD, HEIGHT 16 mm (0.625")

fig. 7. Parts layout and assembly drawings for the 144-MHz preamplifier.

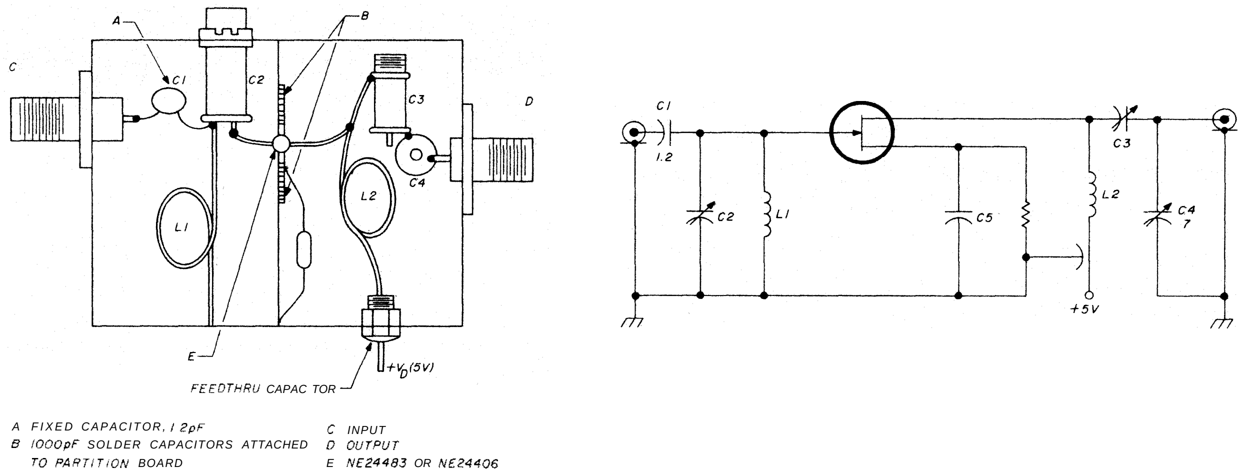


fig. 8. Simplified 144-MHz preamp using a fixed capacitor for the input circuit. A noise figure of 0.7 dB was obtained, as well as characteristics identical to those of the preamp in fig. 6.

Receiver-generated noise. Whatever noise figure an amplifier has, when an attempt is made to obtain gain, the noise figure will always be degraded compared with the $SN(S_{in}/N_{in})$ of the input signal.

When a preamplifier having a noise figure NF_1 and gain G_1 is connected in front of a receiver having a noise figure NF_2, the overall system noise figure is:

$$NF = NF_1 + \frac{NF_2 - 1}{G_1} \qquad (12)$$

In this case, if $NF_1 << NF_2$ and $G_1 >> 1$, the receiver noise figure will be improved.

As previously mentioned, noise power can be converted into temperature; this relationship is:

$$T_e(290)\,[(NF-1)] \qquad (13)$$

When considered in terms of power:

$$N = kT_eB \qquad (14)$$

and this amount of noise power will appear at the receiver output.

Over-all noise characteristics. The sum of **eqs. 10, 11,** and **12** is the overall noise power at the receiver input terminals. When this is taken as N:

$$N = kB \quad \frac{T_a}{L} + T_0 - \frac{T_0}{1} + T_e \quad = kTB \qquad (15)$$

Therefore:

$$T = \frac{T_a}{L} + T_0\left(1 - \frac{1}{L}\right) + T_e \qquad (16)$$

T is the receiving system over-all noise temperature.

The smaller the value of T, the better. (Note thar the coaxial-cable loss, L, always has a value larger than 1.) The noise at the receiver output is always the sum of a) the noise coming into the antenna, b) the noise generated in the coaxial cable, and c) the noise generated in the receiver. So the receiving system should be constructed with the distribution of these noises in mind.

When the receiver noise, including the coaxial cable loss, is higher than the antenna noise temperature, it will be necessary to obtain a signal that will override this noise. In such cases a preamplifier ahead of the receiver will be effective. But since the over-all noise temperature won't become lower than the antenna noise temperature there may not be much effect, even when the receiver noise temperature is extremely low compared with the antenna noise temperature.

Preamplifiers and receiving systems in practice. The relationship between the equivalent temperature of natural noise and frequency is shown in **fig. 11.** Using this relationship as a datum, let's discuss receiving systems using GaAs fet preamplifiers. (Understanding will be made easier if actual numerical values are inserted into **eq. 16.)**

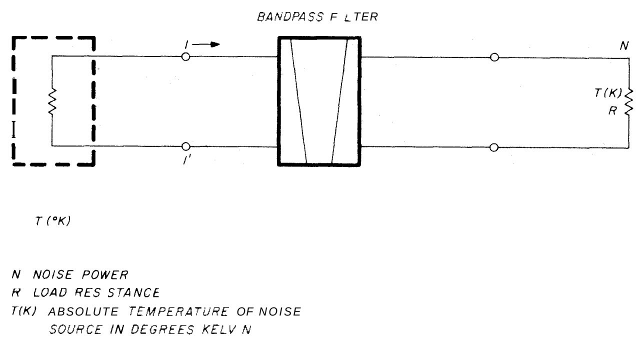


fig. 9. Explanatory diagram for noise-temperature discussion.

An ideal antenna is assumed, whose noise temperature in the 432-MHz band is determined solely by natural noise. Then, from **fig. 11**:

$$T_a = 48K \text{ at } 435 \text{ MHz}$$

A system as shown in **fig. 12** is used as an example. Let's consider a state when the temperature is 17C:

loss of coaxial cable, $L = 3\ dB$

noise figure of receiver $= 1\ dB$

The over-all noise temperature, T, for this case is determined from **eq.** 16. First, the loss, L, of the cable is converted into an antilog:

$$log^{-1}\ 3\ dB = 1.9 \quad \textbf{(17)}$$

Next, the $NF = 1\ dB$ of the receiver is converted to noise temperature, T_e, using **eq. 13**. For this, $NF = 1\ dB$ is converted into an antilog:

$$log^{-1}\ 1\ dB = 1.3 \quad \textbf{(18)}$$

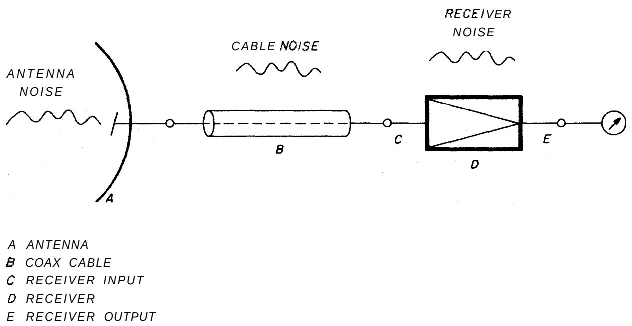


fig. 10. Noise-generating points in a receiving system.

Substituting this value into **eq.** 13 yields:

$$T_e = 290 \times (1.3 - 1) = 87K \quad \textbf{(19)}$$

$$T_0 = 273 \text{ degrees} + 17 = 290K \quad \textbf{(20)}$$

Calculation of **eq. 16** yields:

$$T = \frac{48K}{1.9} + \left[290K\left(1 - \frac{1}{1.9}\right)\right] + 87 = 249K \quad \textbf{(21)}$$

If a coax cable with absolutely no loss can be used, L in **eq. 16** will be $L = 1$ and will be sufficient if:

$$T = T_a + T_e \quad \textbf{(22)}$$

is calculated.

In this case:

$$T = 48 + 87 = 135K \quad \textbf{(23)}$$

will be obtained.

Let's take another look at **eq.** 16, which shows the over-all noise temperature of the receiving system. Suppose you have a receiver with a noise figure of

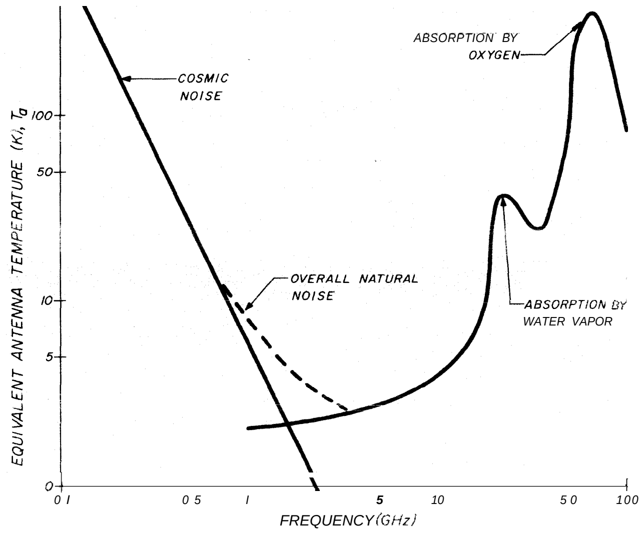


fig. 11. Antenna equivalent temperature as a function of frequency with natural noise as a parameter.

$NF = 1\ dB$ connected to a coax cable with a 3-dB loss. According to **eq.** 21, the over-all noise figure, NF, will be:

$$NF = 10\ log_{10}\left(1 + \frac{249}{290}\right) = 2.7\ dB \quad \textbf{(24)}$$

This can be explained as follows. An antenna having a noise temperature of $T_a = 0K$, and a receiver having a noise temperature of $T_e = 0$ are assumed. When a 3-dB attenuator is connected to the receiver input, what will become of receiving-system noise figure? The noise figure should be 3 dB. But according to **eq.** 16 the noise temperature is about 145K, and the noise figure, NF, will be 1.8 dB. When thinking about a receiver, the signal-to-noise ratio should be considered. When a 3-dB attenuator is connected

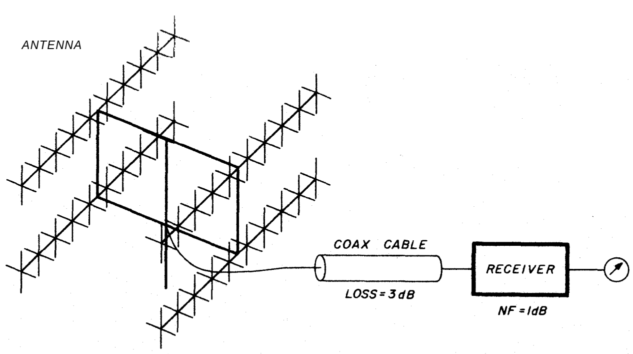


fig. 12. 432-MHz receiving system used as an example for showing the relationship between natural noise and frequency. Antenna noise temperature, T_a (from fig. 11), is 48K at 432 MHz.

to a receiver having 0-dB noise figure, the noise figure will be 3 dB (noise temperature 290K).

signal-to-noise ratio

With regard to the signal entering the receiver input, the transmitting-antenna gain is expressed by G_T; the power output by P_T, and the receiving-antenna gain by G_R. A loss occurs between the transmitting point and receiving point, which is expressed as L_S. Furthermore, when the coaxial-cable loss from the antenna is expressed as L, the strength of the signal entering the receiver will be:

$$S = P_G G_T G_R \frac{1}{L_S L} \tag{25}$$

To calculate the signal-to-noise ratio, eqs. **15** and **16** are added to eq. **25.** Then:

$$SN = P_T G_T G_R \frac{1}{L_S L} \quad /kTB$$

$$= \frac{P_T G_T G_R}{L_S} \times \frac{1}{[T_a + T_0(L-1) + LT_0]\, kB} \tag{26}$$

Only the second term of eq. **26** is the portion in which the receiving-system sensitivity is shown. Therefore, from **eq. 26,** it can be understood that, when a 3-dB attenuator is connected to an amplifier of $NF = 0\,dB$, the system noise figure will be degraded by 3 dB, as well as the signal-to-noise ratio, SN.

$$T_r = T_a + T_0(L-1) + LT_e \tag{27}$$

is defined as the noise temperature of the receiving system including the coaxial cable. When the above example is calculated again using **eq. 27:**

$$T = 48 + [290(1.9-1)] + [1.9 \times 871 = 474K \tag{28}$$

This value has the surprising amount of 339K difference compared with the value when the coaxial-cable loss isn't considered.

This difference is more than 3 dB, so if an amplifier or preamplifier of good noise figure and ample gain is placed directly after the antenna, without any coaxial cable, the noise figure may be improved more than the loss (3 dB, in this example) of the coaxial cable. This occurs because the coaxial cable has resistance and generates some noise.

points for EMErs

To perform EME in the 432-MHz band in Japan, a maximum output of 500 watts is sometimes permitted. However, in such cases, when transmitting, some power may detour and damage the GaAs fet amplifier.

Consider **fig. 13.** What degree of isolation should

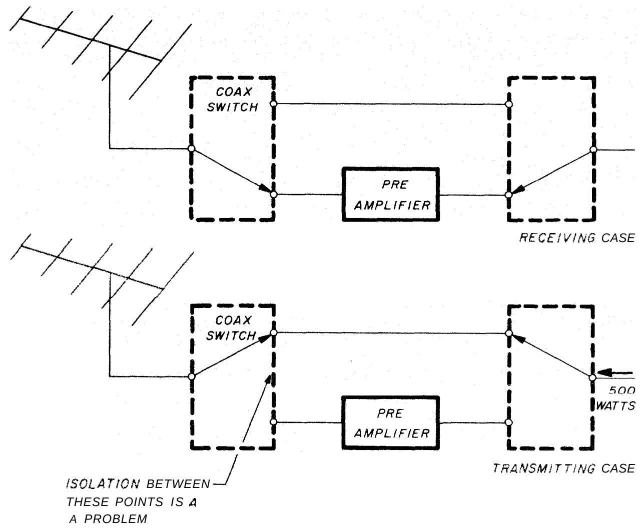


fig. 13. When using high power, isolation in the coax switch is important to protect the fets in the preamp.

the coaxial switch have on the input side of the preamplifier to be adequate?

Here, a transmitting power output of 500 watts is expressed as 57 dBm. If the coaxial-switch isolation is 30 dB in the 432-MHz band, the power that will detour to the preamplifier input during transmission will be 27 dBm (500 mW). Will the fet be protected at this level?

Table 2 shows the power level at which the GaAs fet approaches breakdown in the 432-MHz band when the input power is gradually increased. These are values determined by my experiments in the 432-MHz band. Answers to such questions as "How long will the fets withstand a level 1 dB lower than these values?" are, I regret to say, not yet available.

The power differences in **table 2** occur because of the difference in the biasing methods, and are of great interest. Let's consider why this difference occurs.

Fig. 14 shows two biasing methods. When the input power is inceased in method **A,** at a certain

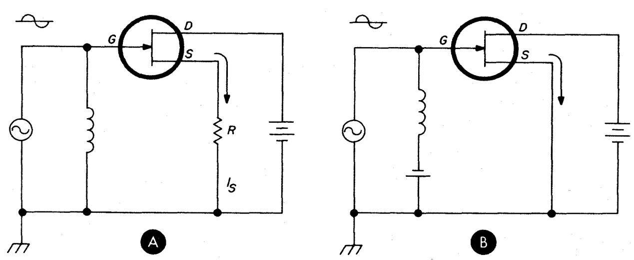


fig. 14. Allowable maximum input power is smaller for the self-bias method (see text).

moment the gate-source span will become forward biased, the current I_S will increase. However, when current I_S increases, a potential drop $I_S \times R$ is created; and in time, the source potential will rise, impeding current flow in a forward direction in the fet. Then, only a voltage in the reverse direction will be applied between gate and source; and when this voltage exceeds the reverse breakdown voltage between gate and source, it will bring about fet breakdown.

However, in the case of **B** of **fig.** 14, the gate voltage is always maintained lower than the source potential by the application of a constant potential. This constant potential prevents the gate voltage from being more greatly negatively biased (reverse direction! through the rising of the source potential by the input voltage. From this, it can be seen that method B is strong against breakdown.

From the standpoint of construction and adjustment, method **A** of **fig. 14** is very stable. However, there is the contradiction that this stability is hard to obtain under actual operating conditions. It's imperative that the maximum input power of the preamplifiers described in this report be designed to have a value 3 dB *lower* than the values shown in table 2.

table 2. Allowable maximum input of GaAs fet preamplifiers.

fet	CW	ssb
with two bias circuits:		
NE24483 NE24406	23 dBm (100 mW)	28 dBm (158 mW)
with self-bias circuit:		
NE24483 NE24406	20 dBm	25 dBm

Therefore, when a self-bias circuit is employed, an isolation of 40 dB will be required to transmit at 500 watts (7 dBm). Furthermore, although these values have been determined experimentally, I've found that when input power is applied without applying bias, breakdown will occur at values 1-2 dB lower than those shown. Thus, when high-power operation is attempted, it will be safer to keep the bias applied during transmission.

Finally, it's desirable to consider using a delay circuit that will ensure transmitting power is always cut off before the coaxial switch is moved to the receiving position. This will ensure further safety of operation.

references

1. S. Sando, JHIBRY, "Very Low-Noise GaAs fet preamp for 432 MHz," ***ham radio,*** April, 1978, pages 22-27.
2. G.D. Vendelin, "Five Basic Designs for GaAs Amplifiers," ***Microwaves.*** February, 1978.

ham radio

diversity reception:

an answer to high frequency signal fading

Diversity reception techniques are discussed with ideas on how they can be implemented with today's equipment

Signal fading is one of the principal problems confronting Amateurs in the high-frequency or short-wave bands. This seems strange, because fading was one of the earliest high frequency problems to be investigated. A 1927 QST article[1] shows that a worthwhile reduction in the adverse effects of fading can be obtained by using diversity reception.

What is diversity reception? With diversity reception, two or more different, or diverse, antenna/receiver combinations are used to receive the same signal. A two-channel system is known as dual diversity; a three-channel system, triple diversity. Diversity reception is widely and effectively used in commercial high frequency installations but has never been popular with Amateurs. One wonders why. Considering what the development of stereo did for the hi-fi industry, I'm surprised that the receiver manufacturers didn't push diversity reception years ago.

In this article I discuss fading and explain how diversity reception can minimize signal loss due to fading. I then discuss equipment considerations for a diversity reception system.

diversity reception

Although it's not apparent to a listener with one receiver and one antenna, fading is not uniform over the surface of the earth. If the listener had several antennas separated by between two and ten wavelengths, with each antenna connected to its own receiver, he'd find that the signal received by the various antennas faded more or less independently of one another. The probability of all receivers being in a fade at the same time is very small. So, if several receivers are connected so that the receiver with the strongest signal can be chosen, the effect of fading can be greatly reduced.

Fig. 1 is a strip-chart recording of a CW signal received on 19 MHz using a triple space-diversity system. The first three rows show each channel individually, whiie the bottom row shows the combined signal. Note the reduction in fading of the combined output.

The fading characteristics of the two signals of a dual-diversity system may be described mathematically by what is known as the correlation coefficient of fading, R. This coefficient may have any value between -1 and $+1$. When $R = +1$, the two signals will vary in the same direction; *i.e., both signal will be either above or below a reference "minimum usable signal level" (MUSL) at the same time, In this case, diversity operation will obviously not provide any improvement.*

When $P = -1$, the two signals will always fade in opposite directions; when one signal is above the MUSL, the other will always be below it. In this situation, diversity operation will provide fade-free reception, since one of the signals will always be above the MUSL. Unfortunately, negative correlation factors are seldom found in practice.

When $R = 0$, the two signals will fade completely independently of each other. In this case, the proportion of time that both signals spend below the MUSL simultaneously is equal to the product of the proportion of time that each signal will be below that MUSL individually.

The advantage of diversity reception is measured by what is called "diversity gain" and is given in dB. Diversity gain is the increase in average signal level obtained from a diversity receiving system compared with the level obtained from a single-channel receiver averaged over a period of time, usually 5 to 10 minutes. Diversity gains of between 3 and 20 dB are typical in commercial practice, and gains approaching these values are probably obtainable in Amateur practice, a worthwhile improvement in average received signal level.

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Commercial stations using high frequency propagation commonly use three antenna/receiver combinations; the law of diminishing returns applies for more than three. A substantial improvement can be obtained, however, using only two receiving systems, and it's doubtful if more than two channels are justified for Amateur applications.

fading

To understand how diversity reception improves reliability, it's necessary to understand the fading phenomenon. Fading in the high frequency, or shortwave, bands is basically of two types: path failure and multipath.

Path failure occurs when the ionosphere can no longer reflect the transmitted frequency back to earth. A good example of this is the way signals on 10, 15, and 20 meters fade out at night: The signals just gradually disappear into the noise. This type of fading is also known as "flat fading," since all frequencies over the usual information bandwidths fade together. Nothing can be done to overcome this type of fading except to change frequency; if the signal is not there, two receivers are not going to hear it any better than one. Path failure isn't a serious problem anyway, because propagation forecasts can generally predict which frequencies will fade and when, so that you can arrange your operating plans accordingly.

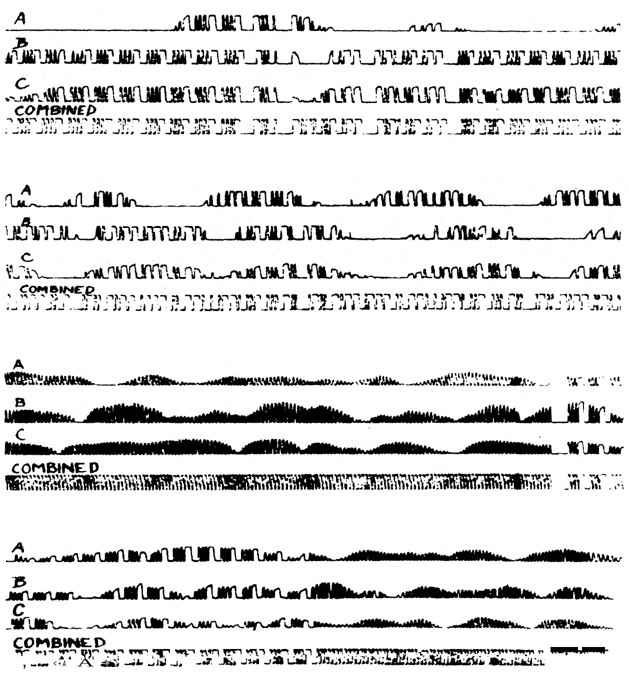

fig. 1. Strip chart recording of keyed CW signals from the output of the three separate channels of a triple-diversity receiver is shown in rows A, B, and C. The combined output is shown in the bottom row, illustrating the reduction in fading possible with this type of system (from reference 2, page 543).

Multipath fading is much more annoying and is the result of two or more waves from the same transmitter traveling over different paths and arriving at the receiver with different phase relationships. If the length of these paths differs by an odd multiple of a half wavelength, which is only about 10 meters (35 feet) at 14 MHz, the two waves will arrive out of phase, and a fade will occur at that frequency. If their path lengths differ by one wavelength, the two waves will arrive in phase, and a "fade-up" will occur.

The distance from the East Coast to the West Coast of the United States is about 5000 km (3000 miles), and the radio path is slightly longer because of its round trip to the ionosphere. A path difference of only 10 meters (35 feet) represents a very small percentage difference between the two, so it isn't any wonder that multipath fading occurs and creates the problem it does.

types of diversity reception

The ionosphere is not stationary but dynamic — more so at some times than others. Paths are constantly changing in both number and length. Therefore signals fade in and out randomly at different locations, at different times, and on different frequencies, all depending on the signal polarity and its angle of arrival. This phenomenon gives rise to five different types of diversity reception: space, polarization, angle of arrival, time, and frequency.

Space diversity. The most common form of diversity reception used by commercial high-frequency stations is space diversity. In commercial practice triple diversity is usually used, with the three antennas spaced at the corners of an isosceles triangle measuring two to ten wavelengths on a side.2 Increasing the spacing beyond this amount doesn't materially improve reception, nor does using more than three antennas. Many experimenters, including Amateurs, have found that a worthwhile diversity gain can be obtained on the 20-meter band by using only two antennas spaced about 15 meters (50 feet) apart. Therefore, space diversity can be practical for Amateur stations restricted to a modest suburban lot. With correlation coefficients of fading as high as 0.6, space diversity can still provide a significant diversity gain.

Polarization diversity. Where space is a limiting factor, as it is at many Amateur locations, a considerable reduction in the effects of fading can be obtained from polarization diversity; that is, using one horizontal antenna and one vertical antenna, each connected to its own receiver. The same tower that supports the horizontal antenna, or one end of it, can also act as the vertical antenna. Polarization diversity is possible because the vertical and horizontal components of the received signal do not usually fade simultaneously, even at the same location.

Some Amateurs report an unusual effect when using polarization diversity: The ionosphere gets hung up on one polarization for extended periods of time, often several days. When this happens, a single-receiver channel using the wrong polarization would report that conditions were bad, whereas a polarization diversity system would report good conditions.

The advantages of space over polarization diversity, if any, are not clear. Grisdale et *al.*[3] report more diversity gain with space than with polarization diversity under some conditions, and vice versa under other conditions; the differences are too detailed to list here. In any event, significant diversity gains are obtainable with either type of diversity, with the difference between the two usually limited to 2.5-3 dB.

Angle-of-arrival diversity. This method uses one or more antennas with lobes at various vertical angles of arrival. Experiments have shown that waves arriving at vertical angles differing by as little as two degrees will give significant diversity gains. Close control of the vertical radiation pattern requires a vertical antenna many wavelengths tall; therefore this type of diversity system doesn't appear to be practical for most Amateurs.

Frequency diversity. Two separate frequencies are used to transmit the same message, because different frequencies don't necessarily fade at the same time. By transmitting the same message simultaneously on different frequencies and listening to the stronger of the two, circuit reliability can be improved. Frequency separations as small as 400 Hz will give considerable improvement on long-haul, high-frequency paths. It therefore appears possible to receive the two sidebands of an a-m or DSB signal, demodulating each sideband separately with two different SSB receivers, thus receiving frequency diversity. (I will discuss this later.)

Time diversity. Time diversity uses two channels, usually with the same transmitter, antennas, and receiver. Imagine a transmitter capable of transmitting two teletype signals simultaneously. Start a message on channel A and a minute or so later restart the same message on channel B. At the receiving end, match the messages received on the two channels. The probability of the circuit fading out during the same portion of each message is much lower than the probability of a fade on only one channel, so that an improvement in circuit reliability can be obtained. Delay times of between 0.05 and 95 seconds, depending on conditions, have been found to give improvement.

Note that both frequency and time diversity improve reliability by sacrificing channel capacity, *i.e.,* by halving the number of messages that can be transmitted over the circuits in a given period of time. If there were **no** fading, two different messages could be transmitted over the same two circuits. Space, angle of arrival, or polarization diversity, however, don't reduce the channel capacity.

diversity transmission[7]

Diversity reception has been shown to increase the average signal level, so one might reasonably ask **if** additional improvement could be obtained by transmitting over two or more antennas. The answer is no. Because fading is caused by multipath, using two transmitting antennas with either space or polarization diversity would double the number of possible

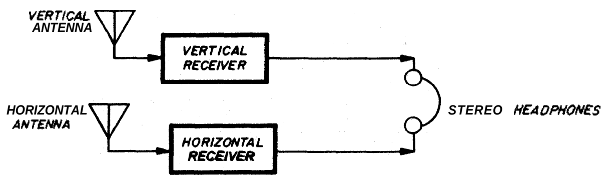

fig. 2. A simple polarization diversity receiving system.

different signals reaching the receiver, thereby increasing the possibility of multipath fading.

The best thing that the transmitter can do is to concentrate its available energy in as small a beam as possible, *i.e.,* use an antenna with as much gain as is practicable. This is standard practice anyway, so no changes are necessary at the transmitter.

diversity receiving techniques

If we assume polarization diversity, which appears to be the most practical for Amateur use, the simplest form of a diversity receiving system consists of two separate receivers, one connected to a vertical antenna and the other to a horizontal antenna. The output of each receiver is connected to separate headphones, such as are commonly sold for stereo use; see fig. **2.** This is a simple and effective method, but it has the disadvantage that the receiver whose input signal is "down" generates noise, making it difficult for the operator, since the noise changes from ear-to-ear.

This problem can be easily corrected by tying together the agc circuits of the two receivers. In this way the agc of the "up" receiver will tend to mute the "down" receiver, minimizing the noise in the down channel; see the sketch in **fig.** 3.

The next obvious step is to combine the audio output of the two receivers in a common amplifier, as shown in **fig. 4.** However, this technique can only be used for phone reception, a-m or SSB; not for CW. The reason is that for CW the audio-tone output of each receiver has a phase that depends upon the phase of the rf signal received by the respective antenna. As the relative phases of each rf signal vary in a random manner because of multipath effects, the phase of each audio tone will vary randomly, too, and there will be times when the audio tones will be 180 degrees out of phase. A fade will then occur in the receiver combined output, even though the signal in each receiver is strong. This is just what we are trying to avoid!

There are two solutions to this problem. The first, used by the commercials, is to take the second-detector output as a dc pulse and add the pulses in a simple summing network. The resulting pulses will be relatively fade-free and are used to key an audio oscillator, which the operator hears in his headset.

The second technique uses what is called a "heterotone"[4] oscillator. which is simply a multivibrator operating at about 400 Hz generating two square waves 180 degrees out of phase. These are used to alternately gate each diversity i-f channel. This signal modulates the CW signal at the intermediate frequency.

Tuning a CW signal using a heterotone oscillator is definitely different from tuning one with a heterodyne oscillator, or BFO; with the heterotone no change occurs in pitch as you tune through the signal.

early diversity receivers

Considering all the advantages of diversity reception, there have been surprisingly few attempts to develop diversity techniques for Amateur use. The earliest attempt of which I am aware was in 1936 by Carll Roland,[5] who used two antennas 183 meters (600 feet) apart connected to two short wave broadcast receivers. Even with such primitive equipment and many trials and errors, Roland's results were good. The final sentence in his article reads: "If the

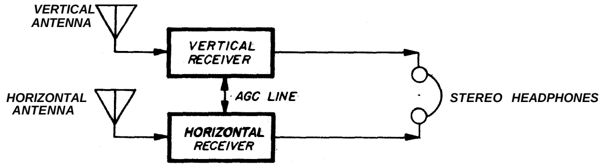

fig. 3. A polarization diversity receiving system with agc muting of the "down" receiver.

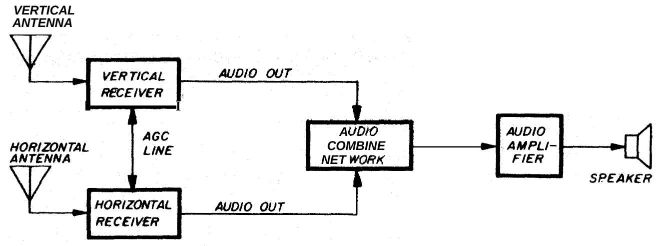

fig. 4. A full dual-polarization diversity receiver for phone work.

broadcast listeners had not wanted their receivers back, we would have kept on using diversity reception."

The second step was taken by James J. Lamb and J. L. A. McLaughlin,[6] who designed what is probably the first single-tuning-control diversity receiver. They developed this receiver specifically for Dr. James M. B. Hard, an American who will be remembered by old timers as XE1G in Mexico City.

QST for December of 1937 describes the third step in an article by J. L. A. McLaughlin and Karl W. Miles.[7] They refer to the May, 1936, receiver (my reference 6) and say, in part:

It has conclusively demonstrated the practicability and desirability of diversity reception for amateur and experimental communications work. Even with two antennas spaced but 50 feet apart, good diversity action has been obtained, especially on the 14-Mc band. Dr. Hard reports that many times when fading conditions and heterodyne interference became so bad as to make his other single receivers useless, the dual diversity still brings in an intelligible signal.

This receiver was considerably improved over the earlier version, mostly in a simplified mechanical design and an improved i-f amplifier. Apparently it was also built specifically for Dr. Hard and became the prototype of the Hallicrafters dual diversity receiver model DD-1. It contains many unique and advanced engineering features, even by today's standards. I'll not go into detail now; see the photograph of my model in **fig. 5.** It's a very impressive piece of equipment!

With diversity reception it's not necessary to use specially made receivers or even identical receivers. Taylor8 describes a 10-meter diversity system using a Hallicrafters SX-17 and a Skyrider 5-10 receiver. His antennas were a horizontal 10-meter dipole and one-half of a vertical 5-meter beam. One end of the 10-meter dipole was attached to the pole that held the 5-meter beam.

As an example of his results, Taylor describes the 10-meter reception of a GM6 late one afternoon:

. . . most of the Britishers had already passed out of the picture. With a single receiver and antenna his signal was so hashed up by a fast fade from S9-plus down into the mud that only about one word out of five was understandable. On switching in the other half of the diversity combination his signal was brought up and smoothed off at a level which rarely fell below S-8; a solid and completely intelligible signal . . .. We will guarantee a thrill the first time you see one of the "S" meters . . . drop down to the bottom of the scale with the signal still pouring out of the 'phones in fine style.

A slightly different approach to diversity reception has been suggested by Bartlett.[9] Bartiett connects each antenna to a separate rf preamplifier; the output of each of the two preamplifiers is connected in parallel to a single receiver of conventional design; this would be the normal station receiver. The preamplifier stages are switched on and off, 180 degrees out of phase, at an audio rate usually between 300 and 1000 Hz. A block diagram is shown in fig. 6. In this manner only one antenna at a time is connected to the receiver so that phase relationships between the two antennas are not important. The receiver output is proportional to the strongest signal present in either antenna at any instant of time.

Because the incoming signal is modulated at the switching frequency this method is useful only for CW. This method also has an unusual effect on the receiver output. If the signal in one antenna is up and the other completely down, the signal reaching the receiver is modulated at the switching frequency. If both antennas receive equal signal levels, the signal reaching the receiver is modulated at twice the switching frequency.

If one antenna has a strong signal with the signal from the other antenna fading in and out, the effect on the output is a changing tone that depends on the strength of the fading signal. Bartlett claims this effect is "very pleasing" to most CW operators; I haven't tried it myself. It sticks in my memory that a device similar to this was advertised in QST right after World War II, but I've not been able to find the advertisement in my old magazines.

fig. 5. A Hallicrafters dual-diversity receiver, Model DD-1. This is the only commercially made receiver intended for diversity reception, circa 1939.

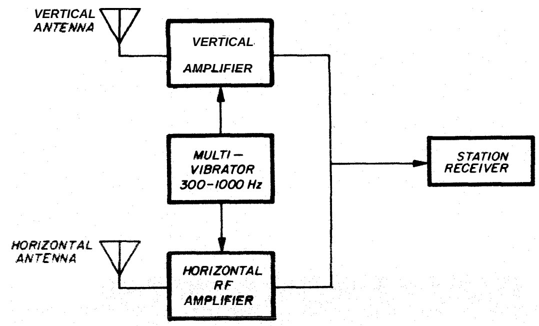

fig. 6. Block diagram of Bartlett's dual-diversity preselector/adapter for CW work.

Bartlett[9] also states that when using polarization diversity, there may be "days at a time" when the vertical signal is 10 to 15 dB lower than that of the horizontal signal. I don't know if this is true in general, or if it results from the use of a smaller vertical antenna than horizontal antenna.

equipment considerations – receivers

By now you may be wondering what changes are needed to equipment designs to make diversity reception practical. The design of a diversity receiving system is not that difficult. At one time a diversity receiver was a truly substantial piece of equipment in both size and cost; fortunately, the development of modern semiconductor devices has reduced both the size and the cost of receiver components. And, since the second receiver will be a duplicate of the first, there will be no additional engineering costs.

As the details of different receivers vary considerably, and as each receiver designerlbuilder has his own ideas as to what a good receiver should be, I'm not going to discuss a detailed receiver design. The characteristics required of a good diversity receiver are the same as those needed for a good single-channel receiver: sensitivity, stability, low noise figure, low intermodulation response, good shielding, and so on. The only difference is that you build it twice! Numerous articles on this subject have been published by *ham radio;* I need not repeat that information.

Good shielding is very important. First, it's necessary to keep the horizontal and vertical channels electrically separate. Leak-through from one to the other before final detection will cause a loss in diversity effectiveness. With two separate receivers, using separate local oscillators, leak-through of one oscillatar to the other mixer will cause birdies, because the two oscillators will, in general, not be on exactly the same frequency.

There are several ways of minimizing the oscillator leak-through problem other than the use of shielding.

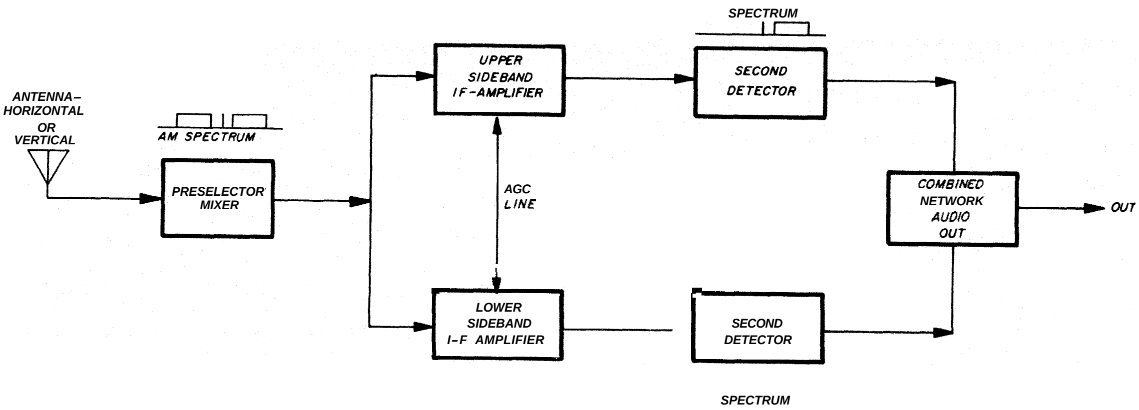

fig. 7. Simple frequency diversity receiver for independent demodulation of the upper and lower sidebands of an a-m signal.

One is to use receivers with different i-fs. Taylor[8] used a Hallicrafters SX-17 with a 465-kHz i-f and a Skyrider 5-10 with a 1600-kHz i-f with good results.

If receivers with the same i-f are used, one oscillator can be realigned to put it on the high side, with the other oscillator on the low side, of the signal. (This may have an adverse effect on tracking in the modified receiver.)

Probably the best arrangement is to use the same oscillator for both channels. Even here, though, considerable care must be used in mixer design to ensure that the received signal from one channel doesn't leak through the common oscillator bus into the other channel. What has been said concerning the local oscillator applies equally well, of course, to all local oscillators in a multiple conversion or SSB receiver.

adapting current transceiver designs

Because the current trend in Amateur equipment design is toward the transceiver, I'll present some general ideas on adapting current transceiver designs to diversity reception. As pointed out earlier, there's nothing the transmitter can do to improve diversity reception, thus the transmitter portion of a transceiver will remain unchanged. Most of the bulk, weight, and cost of a modern high frequency transceiver is in the transmitter section, the transmitter power supply, and the frequency control unit (synthesizer); the receiver itself is very small. And this is the only portion of the transceiver that must be duplicated.

Probably the single most important thing that transceiver manufacturers can do to aid in diversity reception is to make the various oscillator injection voltages and agc bus available, suitably buffered, on the *back apron of the transceiver.* This will permit an external adapter, either commercially manufactured or homemade, to be easily attached. It will then be practical to add an external diversity adapter containing the rf, i-f, audio, and combining circuits necessary to complete the diversity receiving system.

In the preceding material I've assumed a simple summing network for combining the output of the two receivers, as this appears to be the simplest and most appropriate for Amateur use. Actually, the subject of an optimum combining law for two (or more) signals has occupied many, many pages in various journals.

Combining laws can vary from a hard-switching law (*i.e.*, switching to the receiver with the strongest signal) to more sophisticated and beneficial laws. Leonard R. Kahn[10] has asked this question: "For a given ratio of diversity signal levels, how much of the weaker signal and its noise should be added to the stronger signal and its noise to obtain the optimum signat-to-noise ratio?" He then answers his own question by showing that a square-law is best. That is, the ratio of the two signal levels should be squared, then summed.

The method I've sketched, summing the detected signals and tying the agc buses of the two receivers together, will have a combining law that depends on the agc characteristics of the receivers. The most desirable law for Amateur purposes probably can be determined only after considerable experimentation.

The types of combining described so far are known as "post-detection combining." The combining is accomplished after detection, when the rf/i-f phase information has been removed. Additional diversity gain is possible by using "predetection combining," combining the signal before detection. This method requires that the signals be added in phase and is much more difficult to achieve.

equipment considerations — the antenna

Assuming polarization diversity, it's essential that

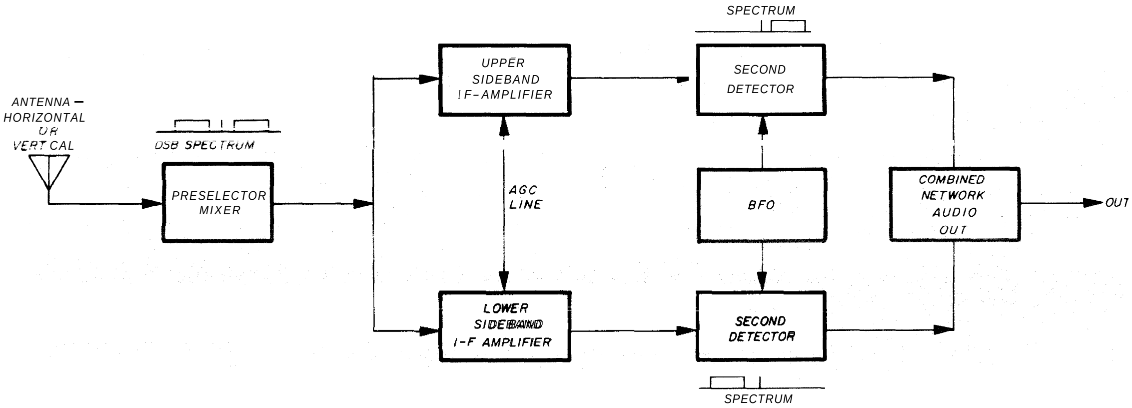

fig. 8. Simple frequency diversity receiver for independent demodulation of the upper and lower sidebands of a DSB signal.

the horizontal antenna *and its transmission line* respond only to the horizontal component of the received signal. Similarly, the vertical antenna and its transmission line should respond only to the vertical component of the received signal.

In both cases, the transmission lines are probably the biggest problems. For the horizontal antenna, the vertical down-lead is the problem area. With the vertical antenna, horizontal runs away from the antenna are potential trouble spots.

If coaxial cable is used, it should have a tight shield braid, or, better yet, be double-shielded.* A high-grade balun should certainly be used in both antennas.

frequency diversity

In describing frequency diversity, I mentioned that frequency separations as small as 400 Hz could be used to provide diversity gain. Since audio frequencies below about 300 Hz are usually filtered out in a voice transmitter, the two sidebands in an a-m or DSB signal will be at least 600 Hz apart, giving rise to the possibility of using frequency diversity.

The simplest embodiment of a frequency diversity receiving system for a-m or DSB signals is shown in **fig. 7.** Here a single antenna, receiver front-end (rf amplifier, mixer, and local oscillator) drives two i-f amplifiers. One i-f amplifier has a filter that covers the carrier and upper sideband; the other i-f amplifier covers the carrier and lower sideband. Each amplifier output is separately detected, then combined in a common audio amplifier. The agc bus of the two amplifiers may be tied together. In this way, the two sidebands are independently received and detected, then combined. Summation does not take place until after the rf phase information has been removed from both sidebands, so that multipath effects between the upper and lower sidebands will not cause fading.

This system gives a surprising amount of diversity gain, except when the carrier itself is in a fade; then the two sidebands don't have anything to beat against, so that demodulation is not possible. The receiver output sounds like a DSB signal with the BFO off.

The next obvious improvement is to provide a locally generated, fade-free, noise-free carrier to demodulate the two sidebands. This scheme is shown in ***fig.* 8.** Since the carrier is no longer needed to demodulate the sidebands, why transmit it? Put the carrier energy into the sidebands to increase talk power and transmit a DSB.

As I write this, I can imagine *ham radio* readers coming to a full stop! Didn't we fight the SSB vs DSB battle 25 years ago and decide on SSB?

The answer, of course, is, yes, we did. DSB lost for three basic reasons:

1. When the two sidebands of a DSB signal are demodulated in the same detector, the stability required of the locally generated carrier is extremely critical.

2. Multipath effects between the upper and lower sidebands cause fading.

3. Extra bandwidth is required in an already overcrowded spectrum.

In a frequency diversity receiver, items **1** and **2** don't apply because the two sidebands are demodulated in separate detectors — not in the same detector. The frequency stability required of the inserted carrier may be somewhat higher than for SSB; if it gets too far off, one sideband sounds like Smokey the Bear, the other like Squeaky the Squir-

*A point well taken. Many Amateurs tend to take coaxial transmission lines for granted. Many types of coax cable are offered for sale. Most are marginal as far as shielding is concerned. If you're interested in diversity receiving systems, it's well worth obtaining good-quality coax cable. The ARRL ***Handbook*** describes problems that can occur when using marginal-quality coaxial transmission lines. **Editor.**

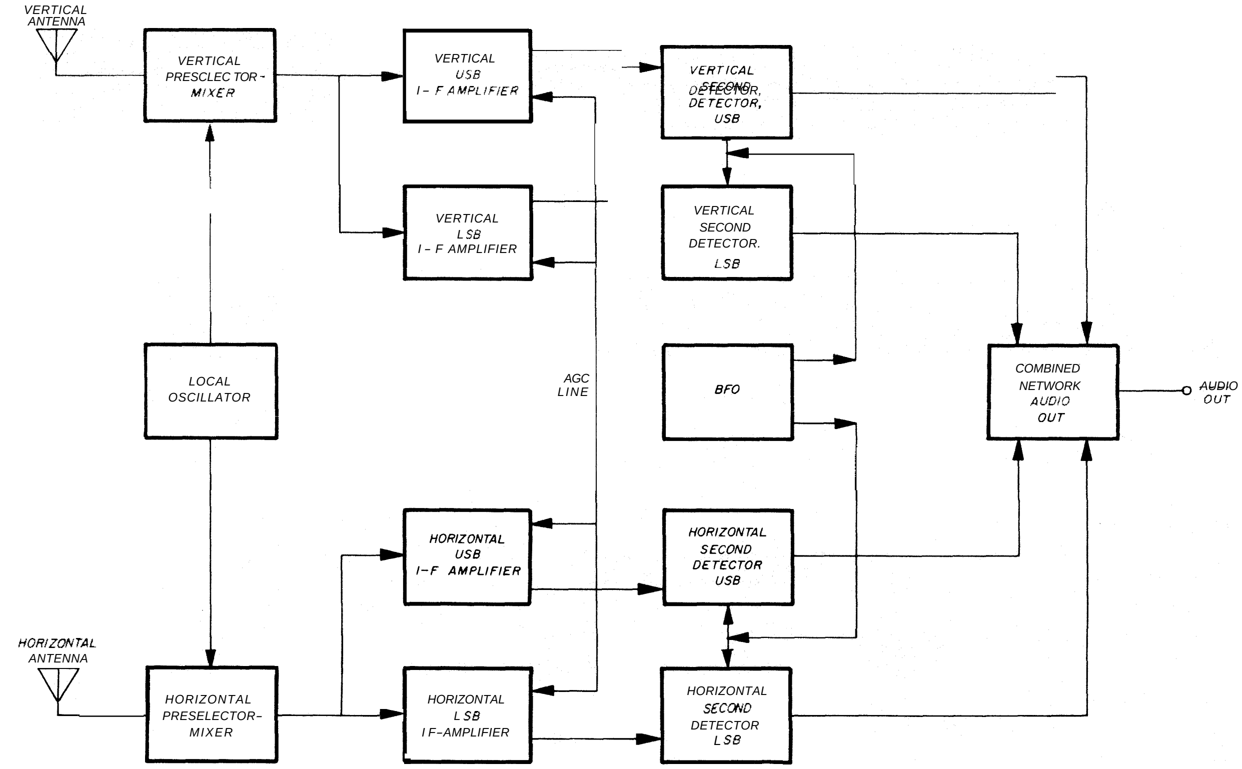

fig. 9. Simplified block diagram of a compound diversity receiver using both frequency and polarization diversity to receive an a-m or DSB signal.

rel. The stability, however, is only on the order of a few hertz instead of a few degrees.

Detecting the two sidebands separately also eliminates fading caused by the two sidebands being 180 degrees out of phase because of multipath. Furthermore, since the probability of *both* sidebands being below the MUSL simultaneously is considerably lower than that of *either* sideband being below the MUSL separately, fading should be considerably less of a problem with the sidebands independently demodulated.

The additional bandwidth will still be with us and may be considered the price paid for diversity gain.

With present-day technology, it's not impractical or expensive to build a compound diversity receiver, using both frequency and polarization diversity. A block diagram of such a receiver is shown in fig. **9.**

conclusion

I've described the advantages of the variious types of diversity reception and shown how they can be implemented with Amateur equipment. Because of space limitations I have hit only the highlights. Anyone who's going to pursue this type of work should become familiar with the references cited. Much of the original work on this technique was done 40 years ago, so I am, admittedly, reinventing the wheel. I firmly believe, however, that diversity reception will be the next step in advanced Amateur receiving techniques. I hope this article helps to start Amateurs experimenting with diversity reception and encourages manufacturers to supply equipment for this purpose.

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ham radio

measuring receiver dynamic range

How to determine receiver performance using simple test equipment and procedures

Receivers with limited dynamic range really have a tough time surviving in my neighborhood, where signals of 100 mW at the antenna are common. When the time came to shop for new equipment, my primary objective was to find a rig with good immunity to some of the problems that might affect its front-end stages. Of primary importance is information on the dynamic range[1,2] and blocking specifications for the rig in question. This information isn't usually supplied by manufacturers, so the scheme was to build some simple test equipment to measure this data and make some comparisons between different rigs available on the new and used equipment markets. An added bonus is that once the test equipment is available it may be used for other tests.

test setup

Two crystal oscillators were constructed, one for operation at 14.02 MHz and the other at 14.04 MHz. Other frequencies could have been used[3] although, in general, it's convenient to retain the 20-kHz spacing. The crystals used are ICM* units designed for OX-series oscillators. That oscillator wasn't suited to this application, so I designed a circuit that has a known power output and minimum second-harmonic content, fig. **1. I** built identical circuits on opposite sides of a piece of double-clad PC board and placed them in a tight-fitting aluminum box. I brought separate power leads through feedthrough capacitors so that the oscillators could be operated independently.

About 60 dB of isolation between the two circuits was achieved, which is adequate for the task. When two-tone signals are needed, a hybrid combiner[2] couples the oscillator outputs together with minimum interaction; a step attenuator adjusts the amplitude of the tones simultaneously (fig. **2).** This equipment plus a 9-volt battery and a few short pieces of coax and connector adaptors is all that's needed to perform the tests. Everything fits nicely into a small box, which may be easily carried to any location where tests are to be run.

Because the signal sources are high level (0 dBm), shielding of test oscillators, coax cables, and attenuators is inadequate to allow testing the sensitivity or noise floor of the receiver. Fortunately this information, although needed to calculate dynamic range, is not absolutely essential in this situation, as the input power that causes an undesired response may be

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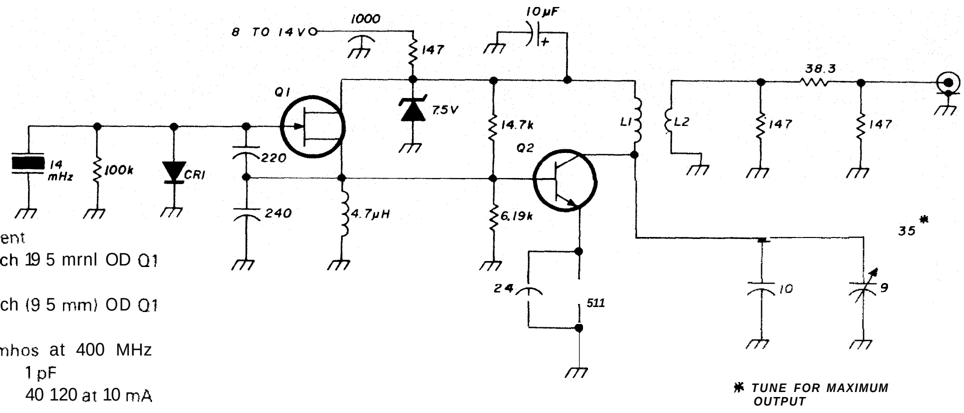

part of fig 1

CR1 Schottky diode HP 5082 2810 or equivalent
L1 11 5 turns 24 AWG (0 5 mm) on 3/8 inch 19 5 mrnl OD Q1 ferrite core
L2 1 turn no 24 AWG 10 5 mm) on 3 8 Inch (9 5 mm) OD Q1 ferrite core
Q1 2N5245 or equivalent G_{FS} 4 millimhos at 400 MHz I_{DSS} = 5 15 mA C_{ISS} 4 5 pF C_{RSS} 1 pF
Q2 2N2369 or equivalent f_T 500 MHz 3 40 120 at 10 mA C_{OB} 4 pF

fig. 1. Test oscillator schematic. The crystal is an ICM (International Crystal Manufacturing Co.) with a capacitance of 100 pF. Output is zero dBm at 14 MHz. Two oscillators are required for receiver tests: one at 14.02 and one at 14.04 MHz. The oscillators were built on opposite sides of a double-clad PC board, which was placed in an aluminum box.

compared directly. This assumes that each receiver has sufficient sensitivity to perform its task, which is normally not a problem on the high-frequency bands; on the contrary, it's common for excessive sensitivity to contribute to reduced strong-signal-handling capabilities.

procedure

Each receiver is evaluated with the agc on, normal SSB filter selected, rf gain at maximum, preselector peaked at 14.04 MHz, noise blanker and rf attenuation off, and audio set for a comfortable level. Turn on both tones, set the attenuator at zero, and tune the receiver to the third-order intermodulation-distortion product at 14.06 MHz.

No calibrated audio voltmeter was available, so my "calibrated ear" was used to determine when the undesired signal could just be heard in the receiver noise output. I've achieved good consistency with this method, although it results in a more conservative number than that obtained by using a voltmeter for measuring a 3-dB change in the audio output. However, all results obtained by this technique may be compared with the others obtained in the same manner by the same person.

Reduce the amplitude of the two tones with the attenuator until the third-order intermod at 14.06 MHz is just detectable. One convenient way to do this is to leave the 14.02-MHz oscillator running and slowly key the 14.04-MHz oscillator on and off with its battery lead while adjusting the attenuator for this just-discernible signal. Subtract the losses of the hybrid combiner and attenuator from the 0-dBm output of the test oscillators to find the input power to the receiver. This number is listed in **table 1** as the **two-tone input** power and is the receiver input power that causes a just-detectable third-order intermodulation product.

gain compression test

A second test may be performed to find the input power that causes gain compression (blocking) in the receiver. This test is usually run with one weak signal and one strong signal, but it's possible to gather some useful data by using the receiver's own internal noise as the weak signal.

An interesting thing can happen when running this test. If gain compression occurs with a strong out-of-passband signal, the noise level heard in the output will decrease. This noise originates in the first stage of the rig. Gain compression of this stage or a succeeding stage will cause a drop in the noise level. However, many times the noise output will increase when the strong out-of-band signal is present. This action is caused by reciprocal mixing with noise sidebands in the receiver local oscillators (commonly heard as a keyed hiss with a strong local CW station on a nearby frequency). A mixer really doesn't care whether it sees a strong LO and weak rf signal or a

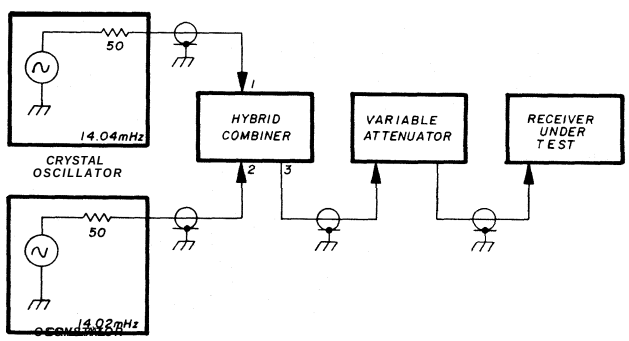

fig. 2. Test setup for making the tests described.

*International Crystal Manufacturing Co., Inc., P.O. Box 32497, Oklahoma City, Oklahoma 73132.

table 1. Test results from many popular receivers using the equipment and procedures described.

receiver	two-tone input (dBm)	gain compression input (dBm)	bandwidth (kHz) at a rejection of 60 dB	70dB	80dB	90 dB	S-meter (S9 μV, linearity)	comments
Drake TR7/DR7	– 41	– 32	3.8	5.6	6.3	6.6	20 fair	good filters, AGC pumps
Collins 75S3B	– 44	– 20	4.5	5.1	5.8	6.3	250 good	good filters
ICOM IC701	– 46	26	5.2	9.4	15.0		20 poor	HAS-65 dB hump ± 10 kHz our
Ten Tec Omni D	**–48**	– 20	4.4	6.3	10.0	—	36 good	
Ten Tec Triton IV/544	**–48**	– 30	6.0				20 poor	
Atlas 350XL	– 51	– 28		4.0		7.0	150 poor	good filters
Astro 200	– 52	– 35						
Yaesu FT901DM	– 56	– 29	3.6	7.6	17.0	—	8 poor	
Ten Tec Argonaut	– 58	– 35	4.0	6.0	14.0	18.0	10 poor	modified KVG filter
Kenwood TS820S	– 60	– 34					110 good	
Yaesu FT301S	**–64**	– 36					30 poor	
Heathkit SB303	**–64**	– 41	4.4	6.0	9.0	10.0	70 good	modified mixers
Collins KWM2	– 65	– 26	4.5	5.1	6.0	6.3	60 good	good filters, AGC pumps
Yaesu FT101E	– 65	– 36					10 good	
Yaesu FT301D	– 68	– 32					65 poor	
Kenwood TS520	– 72	– 36	4.0				70 fair	

strong rf and a weak LO signal (noise sidebands); it will generate an output in either case.

Tune the receiver to 14.04 MHz and slowly key the 14.02-MHz oscillator on and off (leave the 14.04-MHz tone off), while decreasing the step attenuator until the noise output has a just-perceptible change. An increase in noise level is an indication of reciprocal mixing with the LO noise sidebands, provided that the test oscillator output is clean. If the noise decreases, gain compression Is indicated. Note the input power to the receiver; this power is listed in **table 1** as the **gain compression input.**

Some receivers will exhibit reciprocal mixing up to 10-20 kHz from the strong signal. Then a gradual change to gain compression with a higher power input signal will occur further away from this input. In either case, the ultimate performance of the receiver will be limited if either of these phenomena occurs at too low a level. Only one strong input signal is required to cause these problems, so an input 20 dB higher than the two-tone input is probably a reasonable minimum number.

selectivity test

Another test may be run to check receiver selectivity by using one test signal and tuning the receiver to measure bandwidth. This test explores the ability of the complete receiver to reject unwanted signals, which is normally not as good as that of the filter itself because of signal leakage around the filter.[3]

Tune the receiver to 14.02 MHz and adjust the attenuator for a convenient, low S-meter reading, such as S2. About 90 dB of attenuation will be needed. Note the attenuator reading, then increase the signal by 60 dB. Tune away from the signal until it's no longer heard and the S-meter reads zero. Then tune back toward 14.02 MHz until the S-meter again reads S2 and note the receiver frequency. Now tune to the opposite side of the signal and repeat the slow approach to the signal for an S2 reading. Note this second frequency.

The bandwidth at – 60 dB is the difference between the first and second frequency readings. The procedure can be repeated for readings of – 70 dB, – 80 dB, or until the receiver runs out of signal rejection or the test oscillator runs out of power. The latter isn't a problem unless the receiver has more than 90 dB rejection.

One characteristic to watch for is a rig that may be 80 dB down ± 5 kHz away from the signal but deteriorates to perhaps 65 dB down at 10-15 kHz away and may never recover to the – 80 dB level further out. It seems clear that present-day specifications of only – 6 dB and – 60 dB are *not* adequate to determine whether a receiver has a good filter and minimizes signal leakage around it.

checking the S-meter

The next item to check is the S-meter. For example, a reading of S9 with an input of – 70 dBm (71 μV across 50 ohms) would be reasonable in view of the traditional value of 50-100 μV for S9. Linearity can be checked by increasing the input signal to 10, 20, and 30 dB over S9 and observing meter readings. Below S9, increments of 5-6 dB per S-unit may be verified.

Results to date have been dismal, Various S-meters not only read between 8 and 250 μV for S9, but the linearity was so poor that using the meter for evaluating a) gain differences between two transmitting antennas, or b) front-to-back ratios of receiving antennas is strictly a guessing game — unless the meter response has been verified. Manufacturers could do much better in this area with little additional expense and make the S-meter a useful adjunct to operating convenience.

conclusions

The various rigs tested are listed in **table 1** in order of decreasing third-order intermodulation performance. By making a few comparisons of the data, at least three conclusions may be reached:

1. Modern solid-state equipment has finally caught up with the best of 15-year-old vacuum-tube technology.

2. Manufacturers are gradually improving their products, as evidenced by newer models generally testing better than older ones.

3. Some of the newer equipment using double balanced mixers doesn't seem to extract the maximum benefits that such devices can deliver.

Other conclusions could also be reached but the point is that, without this information, only part of the material needed to evaluate the performance of any given radio is available. As this article has indicated, it's not necessary to own a completely equipped lab to gather useful information. All you need is the simple test setup described in this article.

Manufacturers should provide this data. Indeed, American manufacturers' data sheets are getting better (Atlas, Drake). Japanese data sheets either make no mention of the subject or else have enough slack built into their specifications to make meaningful comparisons impossible. So, until considerable improvement occurs in the data provided, you may want to build a couple of test oscillators before you visit your local radio store.

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ham radio

simple, low cost
440-MHz bandpass filter

Straightforward design and construction of a half-wavelength stripline resonator

Bandpass filters which use half-wavelength transmission-line resonators are attractive in many applications because of their simple construction and ease of adjustment. Such filters have been described previously in the Amateur literature.[1] Generalized design information, however, was not provided for frequencies other than those discussed in reference 1. Also, the construction techniques and materials are somewhat cumbersome. In this article I will present some simple equations which can be used to design a quarter- or half-wavelength resonator at any frequency. A simplified construction technique using copper-clad printed-circuit material is also discussed; this technique can result in considerably smaller filters. The design example is a tunable bandpass filter for receiver and low-powered transmitter applications in the 250 to 500 MHz range.

half-wave resonators

A section of transmission line one-half wavelength long and shorted at both ends will behave as a resonator because it supports the formation of a standing wave at its resonant frequency. This is illustrated in **fig. 1.** If means are provided for coupling rf energy into and out of the line section, the half-wave resonator can serve as a bandpass filter with a fairly high Q and low insertion loss. The Q of the resonator is primarily a function of the dielectric loss in the transmission line; for this reason air dielectric lines are normally used as resonators — microstripline on fiberglass epoxy substrates should be avoided. To make the resonator tunable the line length is reduced to somewhat less than one-half wavelength, and capacitive loading is applied at the center as shown in **fig. 2A.** Although a quarter-wavelength line can also be used as a resonator, the half-wavelength version is preferred because it provides better isolation between the input and output ports.

To analyze such a line, it is convenient to visualize it as two loaded quarter-wavelength lines connected in parallel as shown in **fig. 2B.** A transmission line less than one-quarter wavelength long and shorted at one end will present an inductive reactance at the open end given by

$$X_L = Z_o \tan\phi \qquad (1)$$

where Z_o is the characteristic impedance of the transmission line and ϕ is its electrical length in degrees.[2] A capacitive reactance equal in magnitude to X_L, which is placed across the open end of the transmission line will tune the line to resonance (similar to a capacitor in a parallel-resonant L-C tank circuit). The required capacitance is given by

$$C = \frac{10^6}{2\pi f_{MHz} X_L} \; pF \qquad (2)$$

where f_{MHz} is the frequency in megahertz and X_L is the inductive reactance in ohms. In the case of the loaded half-wavelength line, the total capacitance required is twice that of the loaded quarter-wave case, because the half-wave line consists of two parallel quarter-wavelength lines (see **fig. 2**).

examples

Let's design a bandpass filter centered at 440 MHz. Assume that the filter is to resonate with a center loading capacitance of 4 pF, and calculate the resonator length required to meet this condition. For the

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moment the characteristic line impedance can be chosen arbitrarily; I'll use 180 ohms in this example since that is the approximate impedance used later in the construction example (line impedance will be dis-

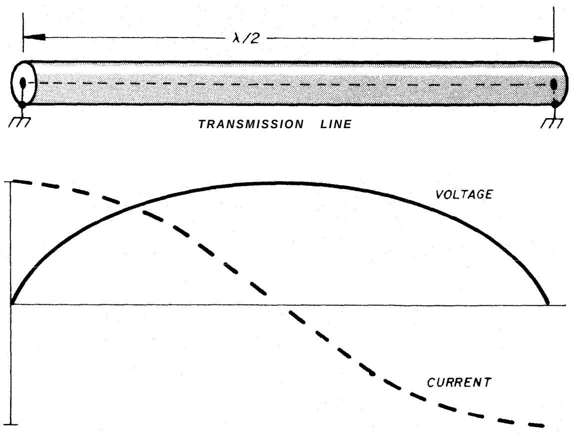

fig. 1. Voltage and current distribution along a half-wave-length section of transmission line with both ends shorted.

cussed in greater detail later). Since the entire line requires 4 pF of center loading at resonance, each half of the line should resonate with 2 pF loading capacitance. Rewriting **eq.** 2, the required inductive reactance X_L of the quarter-wavelength line is given by

$$X_L = \frac{10^6}{2\pi f_{MHz}\, C_{pF}} \tag{3}$$

$$= \frac{10^6}{(6.28)\,(440)\,(2.0)} = 180.95\ ohms$$

The electrical length of the line is

$$\phi = \frac{4\,\ell}{\lambda} \bullet (90°) \tag{4}$$

where Q is the physical length and λ is the wavelength in the line. Substituting the ϕ from **eq. 4** into **eq. 1** and solving for ℓ provides an expression for the line length in terms of wavelengths in the line.

$$\ell = \frac{\lambda}{4} \bullet \frac{arctan\,(X_L/Z_o)}{90°} \tag{5}$$

The wavelength in the line is

$$\lambda = \frac{300}{f_{MHz}\,\sqrt{E_r}}\ meters \tag{6}$$

where E_r is the dielectric constant of the line; if the dielectric is air, $E_r = 1$ (if the dielectric is air, the wavelength in the line will be the same as that in free space). Substituting this expression for λ, and assuming an air-dielectric line, provides the final expression for the length of the quarter-wave resonator.

$$Q = \frac{300}{4f_{MHz}} \left[\frac{arctan\,(X_L/Z_o)}{90°} \right]\ meters \tag{7}$$

For an $X_L = 180.95\ ohms$, a line impedance (Z_o) of 180 ohms, and a frequency of 440 MHz, the length of the air-dielectric quarter-wavelength line which will resonate with a 2.0 pF capacitor is 0.0855 meters or 8.55 cm (3.37 inches) long. A half-wavelength resonator will be twice this length (17.10 cm or 6.731 inches) and will require twice as much loading capacitance (4 pF) to tune it to resonance.

The above example assumes that the required capacitance has been chosen, then proceeds to determine the length of the resonator. Suppose, however, that the resonator length is specified and you wish to determine the capacitance required to tune it to a given frequency. Using approximately the same numbers as in the previous example, the resonator length will be 17 cm (6.7 inches); the required tuning capacitance at 440 MHz is to be calculated. This is equivalent to finding twice the capacitance required to tune a quarter-wavelength (8.5 cm or 3.4 inch) resonator to 440 MHz. Using **eqs. 6** and **4,** the electrical length of the shorted 8.5 cm (3.4 inch) air-dielectric line is 44.88 degrees. Therefore, if the line's

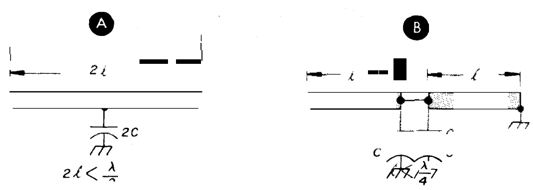

fig. 2. Capacitively loaded half-wavelength stripline resonator (A), and equivalent circuit using two quarter-wavelength lines (B). A groundplane is assumed to be present in both cases.

characteristic impedance Z_o is 180 ohms, **eq. 1** gives the inductive reactance of the line as 177.4 ohms. **Eq.** 2 provides the value of a capacitor with this reactance, 2.04 pF. Doubling this value, the capacitance required to tune the 17 cm (6.7 inch) half-wavelength resonator to 440 MHz is 4.08 pF.

You may wonder why the inductive reactance in both these examples is nearly equal to the line's characteristic impedance; it's purely coincidental. This condition occurs when the physical length of the line is close to one-eighth wavelength. The length of a so-called "quarter-wave" resonator can theoretically be anything between zero and $\lambda/4$ if the proper loading capacitance is used.

impedance calculations

One significant problem encountered in the above calculations is determining the characteristic impedance of a practical air-dielectric stripline. Normally, the stripline resonator used in a filter will be enclosed in a metal box for shielding purposes. However, enclosing the stripline in a box introduces errors into the formulas which are used to calculate line impedance because of the electric field distortion as shown in **fig.** 3. Calculating the impedance of the enclosed stripline in **fig. 3C** is rather difficult but you can get a rough idea of its magnitude by examining **figs.** 3A and 3B. It will be assumed that the width of the line and the spacing between the line and the ground plane is the same in all three cases.

The approximate characteristic impedance of a microstripline, neglecting fringing effects and leakage flux, is given by

$$Z_o \approx 377 \left(\frac{h}{w}\right) \frac{1}{\sqrt{E_r}} \qquad \textbf{(8)}$$

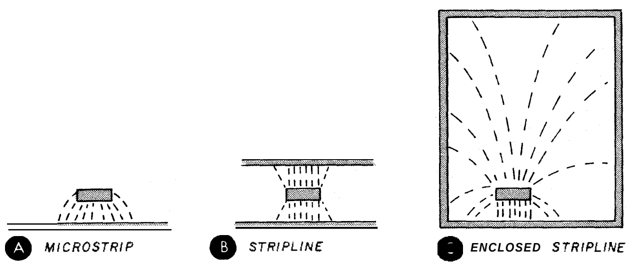

fig. 3. Electric field lines in microstrip (A), true stripline (B), and enclosed stripline (C). Enclosed stripline is used in the bandpass filter described in this article.

where w is the line width and h is its height above the ground plane.[3] Since an air-dielectric line is used, E_r can be replaced by 1. The impedance of the true stripline in **fig. 3B** is not as easily calculated, but is available in graphical form,[3,4] and is considerably lower than that of the microstripline with similar dimensions in **fig.** 3A. It can generally be assumed that the characteristic impedance of the enclosed stripline **(fig.** 3C) will be between that of the microstripline **(fig.** 3A) and the true stripline **(fig. 3B)**.

construction techniques

A filter similar to the preceding design example was built using pieces of single-clad printed-circuit material. Even though this is a uhf application, virtually any type of PC material may be used because in this case the dielectric properties of the material have no effect on circuit performance. The use of copper-clad material allows greater flexibility in the design and construction of bandpass filters than do the construction techniques described in reference 1. Fewer tools are needed, and the filters can easily be built by apartment dwellers (such as myself) without access to a machine shop.

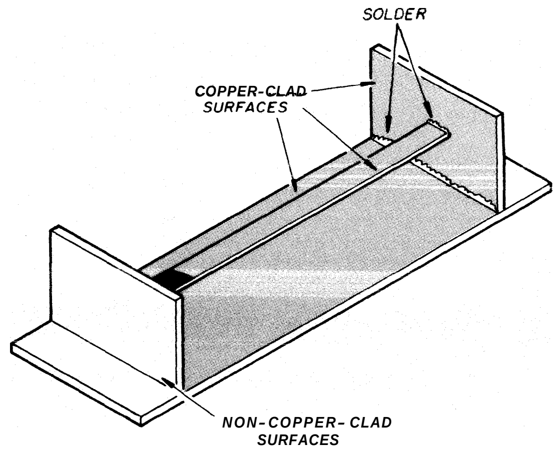

fig. 4. Construction of a half-wavelength resonator using four pieces of single-clad PC material. Dielectric of the circuit board is not important in this application.

As shown in **fig. 4,** four pieces of PC material are used — the stripline itself, the ground plane, and two end pieces to support the stripline. These can be cut out using a small pruning saw or sheet-metal shears. All pieces are soldered together at their edges. The stripline should be installed last, is nominally **18.0** cm long (7 inches) and **1.0** cm (3/8 inch) wide, and is mounted with the copper-clad side up for ease of attaching the center loading capacitor. Spacing between the stripline's upper surface and the ground plane is about **8** mm (5/16 inch). The line impedance was estimated ***a posteriori*** to be approximately 180 ohms.

The tuning capacitor was originally a **3-30** pF variable (Calectro A1-225), but half of its rotor and stator plates were carefully removed to bring its capacitance down to approximately 2 to 15 pF. The tuning capacitor is connected to the exact center of the stripline with a single ***very short*** wire.

Input and output coupling is accomplished by means of inductive coupling lines which run between

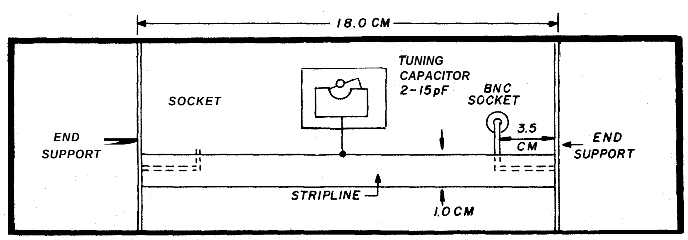

fig. 5. Rear view of the half-wavelength filter, showing the loading capacitor connections and the input/output coupling inductances.

the stripline and the ground plane in the region of the highest field intensity. As originally designed, each coupling wire starts at a BNC connector, runs under the stripline for about **3.5** cm (1 3/8 inch) and terminates in a soldered connection to the end support; this is illustrated in fig. **5.** The entire filter assembly was mounted in a **25** x 5 x **4** cm (10 x **2** x 1 **9/16** inch) Minibox for shielding purposes, with the BNC connectors and the capacitor mounting screws providing the mechanical connections between the ground plane PC board and the enclosure.

performance

The tuning range cannot be found analytically, since both the line's electrical length and the required tuning capacitance are functions of frequency. Using an iterative method, however, it is possible to determine the approximate tuning range. A frequency is arbitrarily picked (let's say **350** MHz). Eqs. **6, 4, 1,** and **2** are then solved in succession, keeping in mind that ℓ is one-half the length of the half-wavelength line. The resulting capacitance, **3.26** pF, is then doubled and this value, **6.52** pF, is observed to lie within the tuning capacitor's range, of 2 to 15 pF. By trying various frequencies, the approximate limits of the tuning range can be found, For this specific filter, the required tuning capacitance vs frequency is listed in **table 1.** Since individual construction techniques, and hence, characteristic line impedance, may vary, the tabular values should be considered only as nominal.

table 1. Center loading capacitance required to tune an 18-cm (7.1-inch) long, 180-ohm air-dielectric half-wavelength resonator to various frequencies.

frequency (MHz)	tuning capacitance (pF)
200	22.3
225	17.4
250	13.9
300	9.3
350	6.5
400	4.7
450	3.5
500	2.6
550	1.9

The filter described here had a measured insertion loss of about 1 dB and a VSWR of approximately **1.6:1** at **440** MHz. By observing the extent of mesh of the tuning capacitor plates, the capacitance required for resonance at this frequency was about 4 pF, which is in agreement with the value obtained analytically.

The VSWR was attributed mainly to mismatch at the BNC connectors between the coupling lines and the external transmission lines. The characteristic impedance of the coupling lines was estimated to be around **115** ohms, by modeling them as circular conductors between two ground planes.3 Their electrical length of **18.5** degrees at **440** MHz gives them a reactive component of +j38.5 ohms. To cancel this reactance a 10-pF silver-mica capacitor was placed in series with each input/output coupling line at the BNC socket. A schematic of the modified filter is shown in fig. **6.**

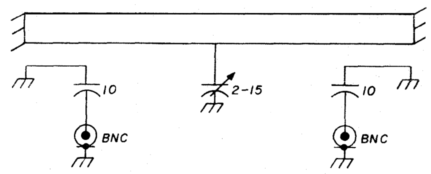

fig. 6. Schematic of the modified 440-MHz bandpass filter with improved input/output matching (see text).

Measurements on the improved filter indicated the VSWR was reduced to approximately 1.2:1 at **440** MHz, while the overall insertion loss remains around 1 dB. Unfortunately, it wasn't possible to measure the VSWR at any other frequencies, so it isn't known whether the 10-pF capacitors improve or degrade performance at frequencies far removed from 440 MHz. However, there should be very little variation over the **420** to **450** MHz range.

This filter was designed for use ahead of a wide-band **(10** MHz bandwidth) receiver operating near **440** MHz to attenuate local fm signals in the **450-470** MHz range. It should also provide significant attenuation of television signals over most of the **470-806** MHz broadcast band. Although bandwidth measurements were not performed, operating experience indicates the 3-dB bandwidth is roughly **3** or **4** MHz, or slightly less than one per cent. This bandwidth should find application among users of wideband modes, such as ATV and packet radio, who want to reduce desensitization caused by out-of-band interference. The filter should also be effective in suppressing unwanted multiplier products in low power (1-watt class) exciters.

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3. J. R. Fisk, W1HR, "Simple Formula for Microstrip Impedance," ham radio, December, **1977,** page **72.**
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ham radio

what is your real standing wave ratio?

Some quantitative answers to a question as old as ham radio

A number of magazine articles have appeared over the years discussing the relative merits of having a low standing wave ratio (SWR). Although it's been demonstrated that excellent results can be obtained with an unmatched transmission system, it's a generally accepted fact that the most straightforward method of guaranteeing acceptable performance under all conditions is by adjusting the various matching devices for minimum SWR. This is particularly true for those not completely familiar with the subtleties of transmission-line theory, as many complex effects occur in unmatched, or tuned, transmission line systems. Regardless of the reasons, most Radio Amateurs are concerned, to some degree, about their voltage standing wave ratio VSWR, or just plain SWR.

One of the reasons for the popularity of SWR as a measure of transmission-system performance is the relative ease with which it can be measured. An SWR bridge can be purchased at the nearest discount store. Even some supermarkets carry SWR bridges. Sealed in plastic packages, and intended primarily for our 11-meter friends, these inexpensive instruments provide an exceiient method for an Amateur Radio operator to evaluate the degree of "match" of his antenna system and transmission line. A typical setup for measuring SWR is shown in **fig. 1.**

As shown in the figure, the normal arrangement for measuring SWR in an Amateur station consists of connecting an SWR bridge in series with the transmission line as it leaves the transmitter or transceiver. The bridge is then used to measure the standing wave ratio *at the input to the transmission line.* If the transmission line were perfectly efficient, *i.e.,* if it had no loss, then the SWR measured at this point would equal exactly the SWR at the antenna. Unfortunately, however, all real transmission lines have some loss. This transmission line loss not only prevents all the transmitter output power from reaching the antenna but also introduces a significant error in the SWR measurement. As we will see, the standing

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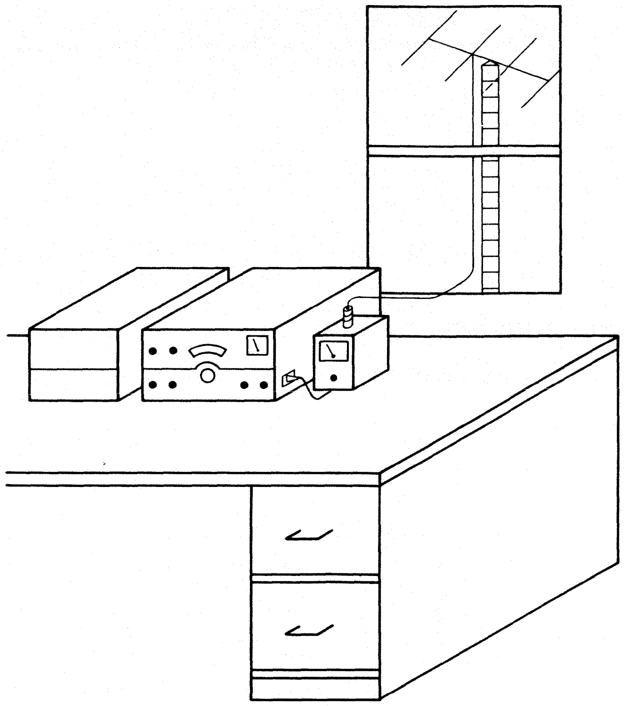


fig. 1. Typical ham setup for measuring SWR. An SWR bridge is connected in series with the transmission line as it leaves the transmitter or transceiver.

wave ratio at the antenna (fig. 2) isn't necessarily equal to the standing wave ratio at the transmitter when a lossy transmission line is used.

transmission-line losses

Fig. 3 illustrates the effect of both a lossless transmission line and one with loss on dc pulses. Both lines are terminated in a matched load:

$$Z_1 = Z_0$$

$$Z_1 = \text{load impedance}$$

$$Z_0 = \text{line characteristic impedance}$$

Note that in both cases the incident pulse is completely absorbed in the load, resulting in no reflected pulse. We may calculate the standing wave ratio at the transmitter in both cases as:

$$SWR = \frac{V_i + V_r}{V_i - V_r} \qquad \textbf{(1)}$$

where V_i = incident pulse voltage
V_r = reflected pulse voltage

Thus:

$$SWR = \frac{1+0}{1-0} - 1 \text{ or } 1{:}1$$

For the lossless transmission line the situation is unchanged at the load end. For the lossy line, however, the incident voltage is only 0.707 at the terminal end, thus:

$$SWR = \frac{0.707+0}{0.707-0} = 1 \qquad \textbf{(2)}$$

In other words the standing wave ratio for a matched line is 1:**1** regardless of *where it's measured or how much loss it has.*

unmatched loads

Now consider the case of an unmatched load. Referring to fig. **4,** we see that the standing wave ratio measured at the load end for the lossless and lossy transmission lines is respectively given by:

$$SWR = \frac{1+0.5}{1-0.5} = 3{:}1 \text{ (lossless)} \qquad \textbf{(3)}$$

$$SWR = \frac{0.7+0.35}{0.7-0.35} = 3{:}1 \text{ (lossy)}$$

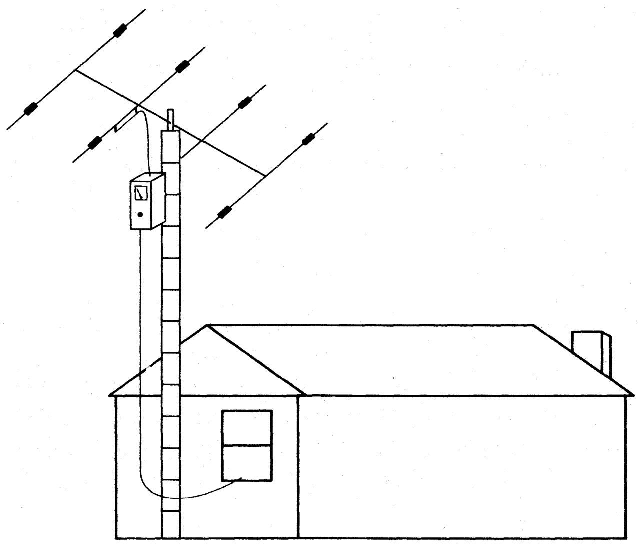


fig. 2. An SWR bridge connected at the antenna will show an entirely different set of conditions if a lossy transmission line is used.

The SWR measured at the source (transmitter) end of the line, however, is quite another situation:

$$SWR = \frac{1+0.5}{1-0.5} = 3{:}1 \text{ (lossless)} \qquad \textbf{(4)}$$

$$SWR = \frac{1+0.25}{1-0.25} = 1.67{:}1 \text{ (lossy)}$$

Thus for the lossless line the SWR is the same no matter *where* it's measured. For the lossy line, however, the SWR appears to be lower when measured

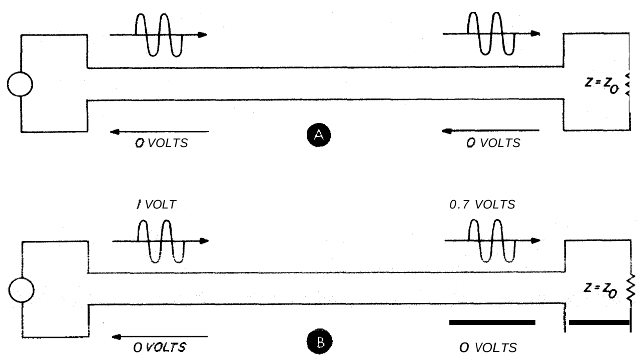


fig. 3. Effect of wave propagation on a matched transmission line. A shows a "lossless" transmission line; B shows a **lossy** line. Note that in both cases the incident puise is completely absorbed in the load, resulting in no reflected voltage.

at the transmitter end. This effect is caused by the additional attenuation suffered by the reflected wave as **it** travels down the transmission line and back, whereas the incident wave is measured *directly* at *the source.*

Fig. 5 is a plot of measured *versus* actual SWR for various amounts of transmission line loss. The error is quite significant. **Fig. 6** is a plot of typical transmission line losses versus frequency. Using these two graphs, we can estimate the actual SWR at the antenna for your installation, based on the measured SWR at the transmitter.

example

Consider a typical installation consisting of a beam antenna connected to a transmitter by 61 meters (200 feet) of RG-8/U coaxial cable. Suppose the SWR is measured as 2.5 at 28 MHz. From fig. 6, the loss of RG-8/U, when matched, at this frequency is about 1 dB per 30.5 meters (100 feet). Thus the cable loss in the example is

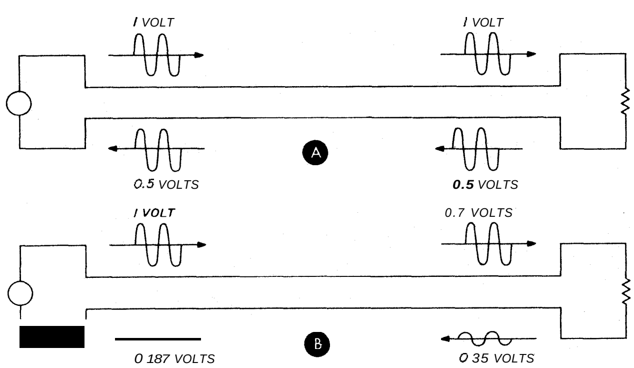


fig. 4. Wave propagation on lines terminated in unmatched load; B shows the same conditions for a lossy transmission line. For the *"lossless"* line the **SWR** is the same no matter where it's measured; for the lossy line the **SWR** appears to be lower when measured at the generatar (transmitter) end.

$$
\begin{aligned}
L &= (\text{loss per } 30.5 \text{ meters})\\
&\quad (\text{cable length}/30.5)\\
&= (1)\,(61/30.5)\\
&= 2\ dB
\end{aligned}
$$

In terms of English units:

$$
\begin{aligned}
L &= (1)\,(200/100)\\
&= 2\ dB
\end{aligned}
$$

Referring to **fig. 5,** the actual antenna SWR *is* about 5.5.

consequences

What are the consequences of the error of SWR

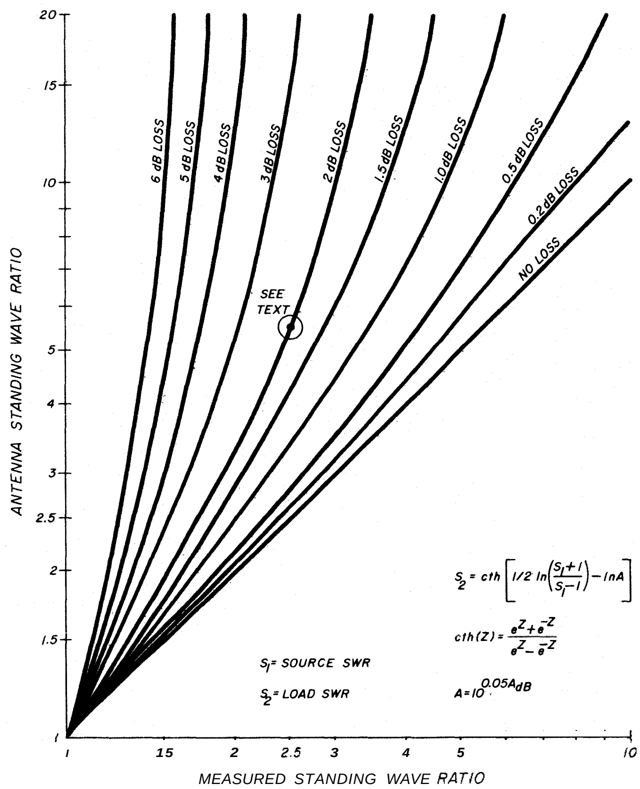


fig. 5. Antenna standing wave ratio as a function of measured standing wave ratio with transmission-line losses as a parameter.

measurement? The first consequence that comes to mind is the effect on the power-handling capability of the transmission line (see fig. **7).** In the previous example, for instance, the power rating of RG-8/U coaxial cable is about 1600 watts at 28 MHz when operated at unity SWR (fig. 7). Derating for the computed standing wave ratios, we find the maximum recommended power-handling capacity of the cable would be:

$$
P_{max} = \frac{1600\ \text{watts}}{4.2} = 381\ \text{watts} \tag{6}
$$

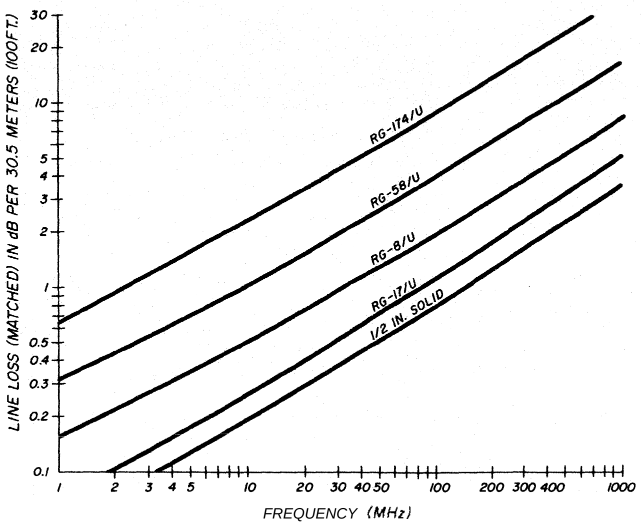


fig. 6. Transmission-line loss (matched) as a function of frequency for several types of coaxial cable.

What if we'd calculated the maximum power based on the measured SWR? The maximum power capability of the cable would have been:

$$P_{max} = \frac{1600\ watts}{2.5} = 640\ watts \qquad (7)$$

As you can see, the coax power-handling capability would have been exceeded by about 60 per cent.

testing

Another problem arises in testing antennas. Many Amateurs test their antennas by measuring the SWR at the transmitter rather than at the antenna. The

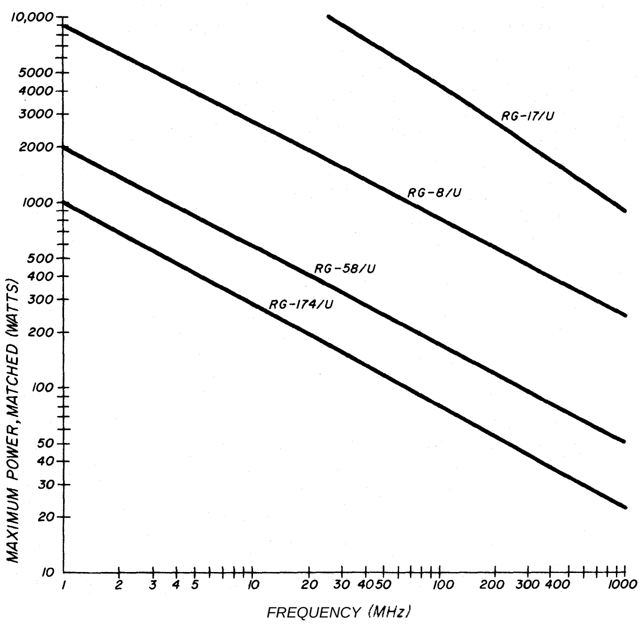


fig. 7. Power-handling capability of popular coaxial cables as a function of frequency.

result is that the antenna ***appears*** to perform better than it really does. It would be much better either to measure the SWR at the antenna, or at least make an attempt to correct for the feedline loss.

Feedline loss depends on standing wave ratio. Consider the case of a transmitter with tune and load controls adjusted for maximum voltage across the input to the transmission line **(fig. 8)**. (This implies a transmitter output impedance equal to the complex

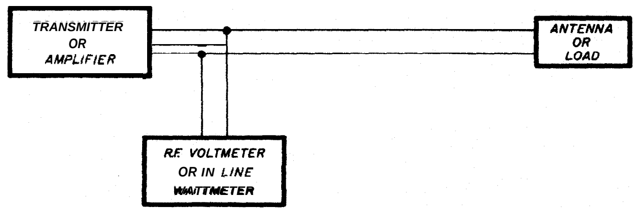


fig. 8. Adjusting the transmitter for maximum output voltage results in a conjugate match, or a condition in which the transmitter output impedance equals the complex conjugate of the impedance seen at the input to the coaxial line. This condition is assumed in figures 9 through 12.

conjugate of the effective transmission line input impedance — a condition referred to as conjugate match.)

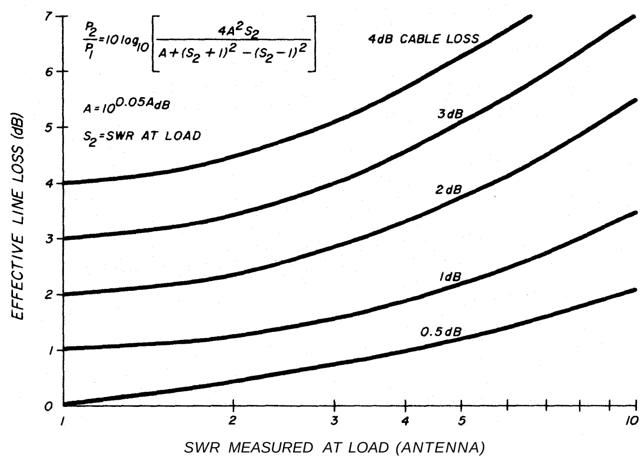


fig. 9. Effective line loss as a function of SWR measured at the load, with cable loss as a parameter.

lossless lines

For the case of the lossless line, the reflected wave is completely reflected at the source and ultimately arrives again at the antenna. Each time the wave arrives at the antenna part of it is absorbed, and part of it is reflected. The reflected portion is again reflected by the source, and so on until the entire wave is ***completely absorbed*** by the antenna.

Since the line has no loss, and since we're assum-

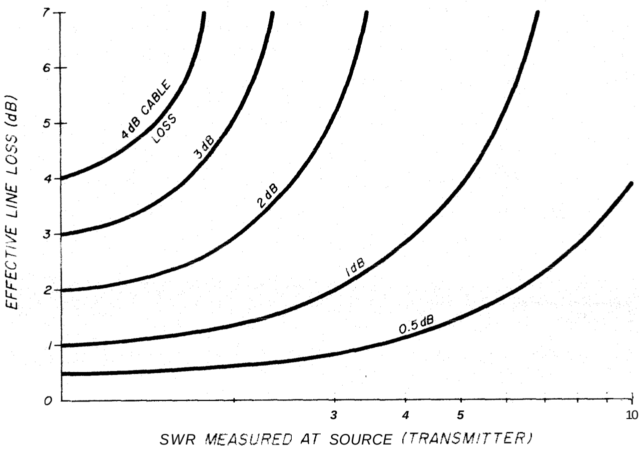


fig. 10. Effective line loss as a function of SWR measured at the source, with cable loss as a parameter.

ing complete lossless reflection at the source (transmitter), the energy is transferred to the antenna with 100 per cent efficiency regardless of whether the antenna is matched to the transmission line.

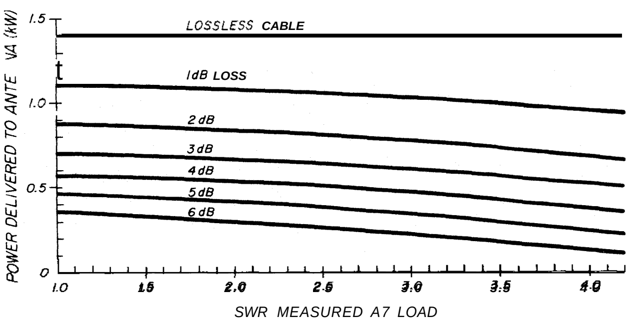


fig. 11. Power delivered to the antenna as a function of SWR measured at the load, with cable loss as a parameter.

lossy lines

Now consider the case of a lossy transmission line. For the matched load the resulting loss is simply the loss of the transmission line. The mismatched load,

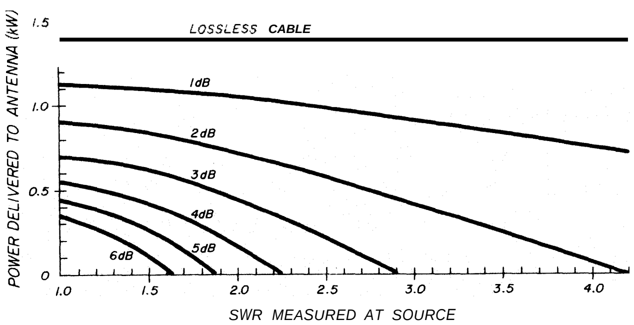


fig. 12. Power delivered to antenna as a function of SWR measured at the source, with cable loss as a parameter.

however, results in quite a different situation. Each time the wave is reflected and travels down the transmission line, it becomes smaller in amplitude by an amount equal to the transmission-line loss. Thus, even if the transmitter is perfectly tuned for "conjugate match," only part of the "re-reflected" energy

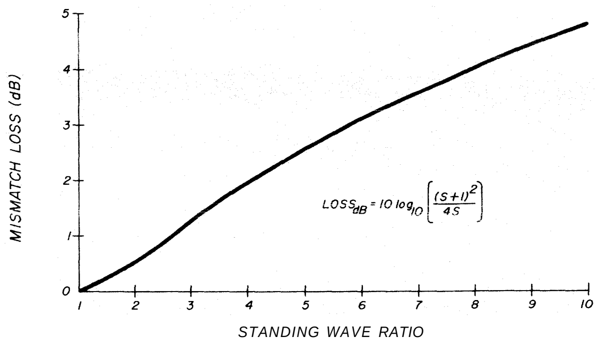


fig. 13. Mismatch loss as a function of SWR.

reaches the antenna, with the portion growing progressively smaller each time around. In other words, in addition to the portion of energy lost in the transmission line the first time, an additional amount is lost due to reflections.

Fig. 9 is a plot of effective feedline loss for various values sf standing wave ratio as measured at the antenna. **Fig. 10** is the same plot versus SWR at the source (transmitter). **Figs. 11** and **12** are plots of power delivered to the antenna by a 2-kilowatt amplifier tuned for maximum voltage across the transmission line at the source. **Fig. 11** is plotted versus load

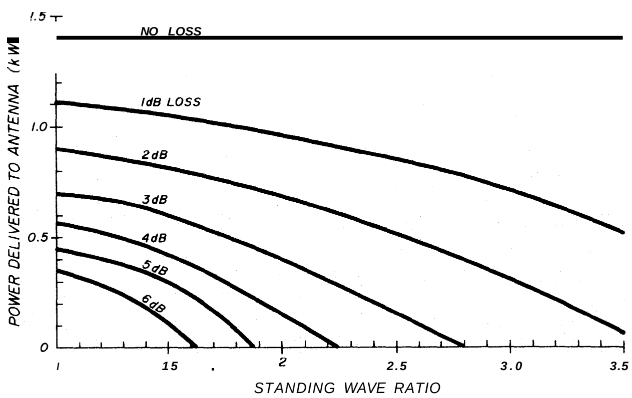


fig. 14. Power delivered to antenna, including mismatch loss for a 2-kW broadband transmitter, as a function of SWR. Output impedance is equal to cable characteristic impedance.

SWR for various line losses; **fig. 12** is plotted versus source SWR. A line loss of 2 dB, for example, would result in only 450 watts being delivered to the load if an SWR of 3:1 is measured at the amplifier output.

mismatch loss

One final comment on line loss. Many hams are now using solid-state transmitters with broadband final amplifier stages. Since there are no adjustments on this type of transmitter, it's not possible, in general, to achieve a conjugate match at the source, as discussed earlier. As a result, there is another loss to be considered when computing the power delivered to the antenna. Mismatch loss is simply the loss resulting from reflected power being absorbed by the source (transmitter) rather than re-reflected power, as discussed previously. **Fig. 13** is a plot of the additionai mismatch loss versus source SWR. **Fig. 14** is a plot of power delivered to the load, including mismatch loss for a 2-kilowatt broadband transmitter, with output impedance equal to the transmission line characteristic impedance.

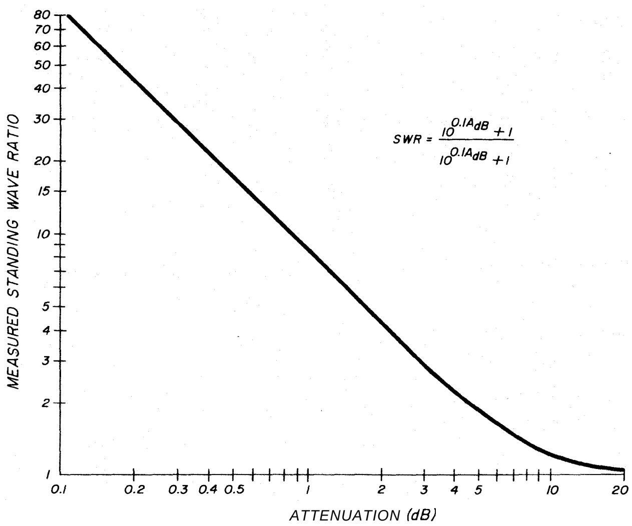


fig. 15. Measured SWR as a function of line attenuation for open or shorted lines.

measuring feedline loss

One final note. It's sometimes difficult, if not impossible, to actually measure feedline loss. An example is a repeater site at which I recently wished to measure the loss of the line from the antenna to the transmitter. One method of measuring the loss would be to carry either a power meter or a signal generator up the tower for connection to the coax at the antenna. An alternative method would be to simply short or open the transmission line at the antenna and measure the resulting standing wave ratio at the transmitter. **Fig. 15** could then be used to compute the transmission line loss. For example, a shorted SWR of 4:1 would correspond to a feedline loss of approximately 2.3 dB.

ham radio

the ham notebook

Collins 32S cooling

One of the major enemies of final amplifiers, tube or transistor, is heat. My Collins 32S-1 is no exception, and a cooling fan over the amplifier cage is attempting to lengthen the life of my 20-year-old rig.

I got tired of unplugging the fan after operating and dreaded the thought of forgetting to shut if off.

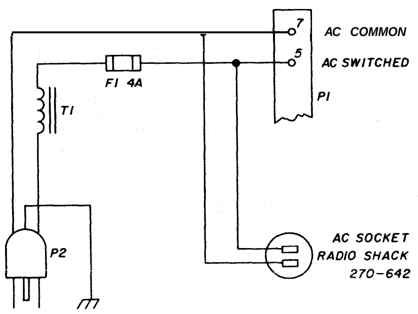


fig. 1. Partial schematic of the Collins 516-F2 supply showing a socket addition for an amplifier cooling fan.

An outboard switch was contemplated — and the thought quickly discarded. I preferred to have the fan turn on and off with the transmitter power. This is how I did it.

I mounted an ac chassis-mount receptacle (Radio Shack part 270-642) on the 516-F2 chassis in the space occupied by the stick-on, serial-number label. (I transferred the label information with an engraving pencil to the chassis.) This location almost perfectly centers the receptacle between XV1 and XV2. The socket was wired as shown in **fig. 1,** using heat-shrink tubing on the receptacle terminals.

A fan such as the Rotron ***Whisper*** may be spray painted and secured to the cabinet lid directly through the holes with M3 (4-40) hardware. A cover on this fan isn't recommended, as it will create a back pressure and hamper the cooling-system efficiency.

When the fan is plugged into the 516F-2 receptacle, the fan will turn on and off with the 32S FREQUENCY CONTROL switch to provide extra cooling.

Paul K. Pagel, N1FB

shunt-fed tower

A problem that commonly occurs when one tries to shunt feed a tower for 160, 80, and 40 meters is not having a large effective diameter for the shunt section. A small-gauge wire makes things a bit touchy and wire cages always seem to get twisted up during mounting.

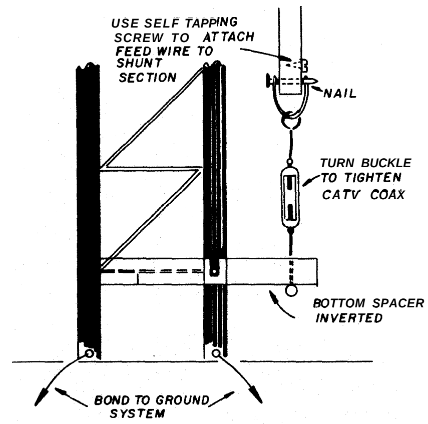


fig. 2. Bottom spacer and connection for the shunt-fed tower. Looping the spacer over the tower rung allows you to maintain tension in the gamma section, holding it rigid. Either the material for the bottom spacer or the connection between the turnbuckle and hardline must be insulated.

These problems are eliminated if the relatively large diameter CATV coax is used. This material can usually be obtained free (or, at most, for a few dollars) from the CATV company warehouse scrap yard as "reel ends."

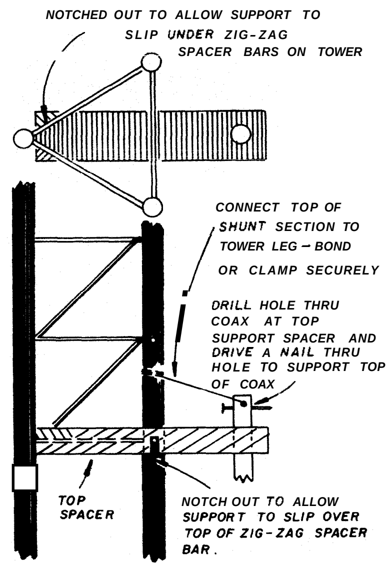


fig. 3. Diagram of the top spacer and connection to the tower.

My system requires no clamps for attaching the shunt feed element to a 36-meter (118-foot) ROHN 25 tower (see **figs. 2** and **3** for construction of spacer/clamp assembly). Some experimentation with the spacing between the tower and shunt section will be required to achieve a VSWR of 1.0:1.

Dick Bingham, N6HZ

Yagi antenna for uhf — simplified construction

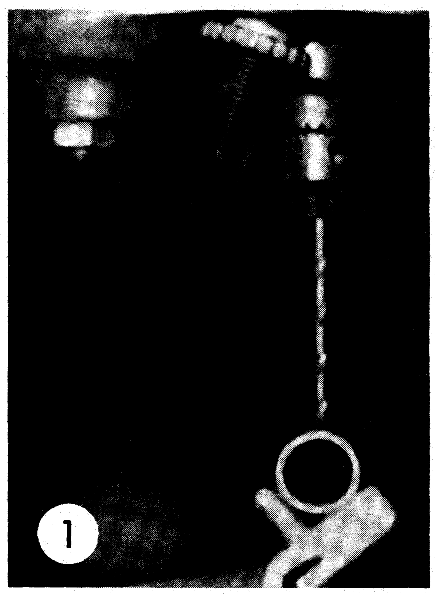
Homebrewing antennas has never been one of my strong points. Most of my beams had more of an omnidirectional characteristic than a main lobe. The problem was making all the elements point in the same direction. If your main construction tools are the same as mine — a blow torch and a sledge hammer — then the techniques I managed to acquire may be helpful in making your next beam look more like an antenna than a corkscrew.

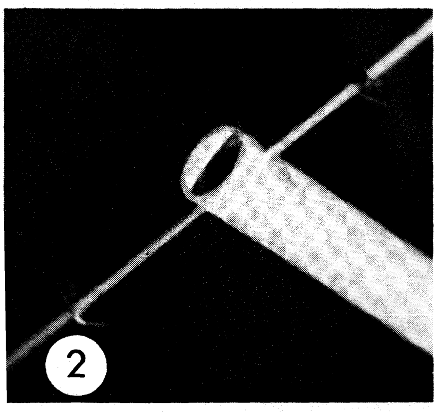
Drilling holes is probably the most critical part of construction. I used a drill stand to hold the drill in place. I made a guide from a piece of V-channel aluminum bolted to a piece of wood (see **photo 1**). The **V** block keeps the tubing from wandering during drilling. Be sure the drill is centered in both directions in the stand. Use a scrap piece of tubing to drill a test hole, then insert an antenna element to check for alignment. It took me three tries to get the drill oriented just right.

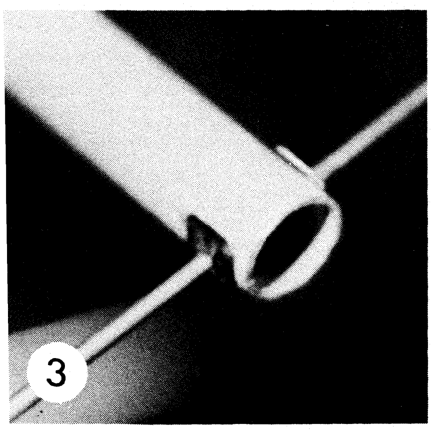
The next problem was how to hold the elements in the boom without using a lot of clamps or brackets. After many tries, I used speed nuts. They worked well and held the elements firmly in place. I pressed a speed nut onto each element end **(photo 2)**. Caution: When pushing the speed nuts up to the boom, be certain that the element is centered **(photo 3)**, because backing up the speed nuts is nearly impossible. Speed nuts are available in many sizes at most hardware stores. I applied a liberal coat of silicone rubber to protect the speed nuts from rust **(photo 4)**.
it's used to align the other holes as you look down the boom. The drill bit acts as a gunsight for alignment **(photo 5)**. If minor alignment exists, correction can be made by forming the elements.

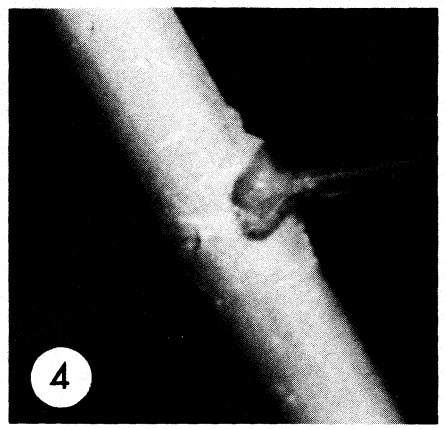
The antenna shown in **photo 6** was my first attempt to built a 432-MHz Yagi using the method described. The cost was just over $5.00, using an aluminum boom and welding rods for elements. I've also built some 2-meter Yagis. Tests have shown that these homebrew antennas are within ½ dB of their commercial counterparts. Pattern checks have shown a clean main lobe.

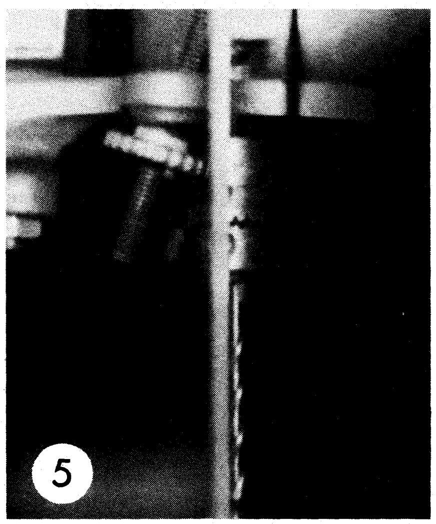
Thomas Varmecky, WA3CPH











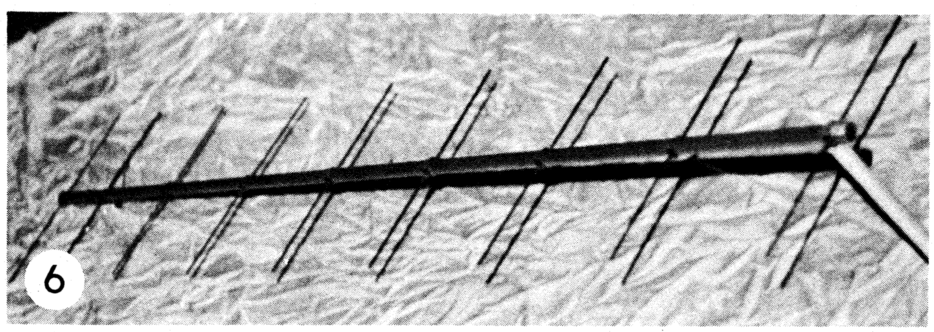


Photo 1: Drill stand for working aluminum tubing is made from a piece of V-shaped aluminum bolted to a piece of wood. Photo 2: Speed nuts hold the element to the boom. Photo 3: Detail showing element-to-boom mating using speed nuts. Photo 4: Final assembly of element to boom. A coat of silicone rubber protects the joints from rust. Photo 5: Use the drill bit as a gunsight for alignment when drilling holes for the other elements. Photo 6: Complete 432-MHz homebrew Yagi antenna. Antenna elements are welding rods; entireantenna cost just over $5.00.

comments

(Continued from page 6)

wherein L, W, and t represent length, width, and thickness.

To maintain 12 amperes/square foot, you must measure the resistance of the bath with the piece to be processed in place and apply Ohm's law or use a variable voltage supply with an ammeter in the circuit. Maintaining this current will require a power supply capable of supplying 15-20 volts at a current equal to 125-150 per cent of the calculated amount. Voltage will vary with bath temperature and alloy.

Best anodizing results are obtained by maintaining a constant current throughout the cycle. By maintaining the bath temperature between 68 and 72F (use a long glass dairy thermometer) and the current at 12 amperes/square foot, the time required to produce a given coating thickness will be 80 ampere minutes per 0.01 mil or 0.0001 inch. In other words, 6.7 minutes' time will produce 0.0001 inch of coating (80 divided by 12) if the other parameters are observed.

In substantiation of this, note that automobile trim is generally required to have a 0.3-mil coating, and most anodizers achieve this with a 20-minute treatment.

Most dyes work well on coatings of 0.3 mil and up. Note, too, that for any alloy worthy of consideration by the Amateur fraternity, the coating weight or thickness will vary no more than three per cent in either direction when coated according to these suggestions.

Proper operating practice should be observed if you expect usable results. The material must be ***clean*** as a prerequisite to anodizing. Scrubbing the piece with a good soap or detergent should suffice, provided the piece is then thoroughly rinsed. A good test for cleanness is that the rinse water falls off the surface in an unbroken fashion; that is to say, it should not form beads as does the rain on the waxed hood of a car.

Pretreatments such as buffing, wire brushing, or etching should be given some thought by the experimenter. Once the piece has been properly racked (fastened to the aluminum rod or strip for suspension in the bath), it should be carefully lowered into the electrolyte with the power off.

The power should then be applied at a low level and quickly increased to the calculated current. The bath should have some mild agitation during the whole anodizing cycle. Whatever method is used to agitate the bath must take into consideration the hazards of dealing with an ***acid*** bath. The power should be turned off before the piece is removed.

Aluminum racks are anodized along with the piece of work. Hence, before they're used again they should be sanded, wire brushed, or etched in the contact areas to ensure a good electrical contact. Alloy 2024-T3 or -T351 will work best as rack material for Amateur use. Good electrical contacts are very important to the success of any anodizing experiment.

Anodizing may be done by various methods (including ac anodizing) and for many reasons. By and large, the greater portion of such treatments represented by the H_2SO_4 processes are meant to enhance the appearance of the item treated. The use of the process by the Radio Amateur should be regarded as a means of improving his handiwork.

Any experimenter in need of further corrosion protection of an aluminum item would be well advised to take his problem to a professional anodizer. These sources are listed in the yellow pages of your local phone book.

The work cited in the reference is a three-volume set and is highly recommended to anyone interested in more information on aluminum or the processes employed to fabricate it.

Robert A. Ridout, WA9UXK
McHenry, Illinois

autotune circuit

Dear HR:

Shown in **fig. 1** is a circuit, road tested on a Heath HW-7, that can be used for "touch tuning" a vfo. S1 and S2 are momentary pushbutton switches. S1 provides down-frequency tuning and S2 allows up frequency tuning. The tuning rate is controlled by the time constants R1-C1, for down, and R2-C1, for up; the values given are for about 5 seconds per kHz. Nothing is critical, the jfet is a 4/$1.00 special, and the rf choke and capacitors were chosen by "reach." R3 provides current limiting at about 2.5 mA, and R4 is for insurance.

The idea is presented as an effort to eliminate the mechanical mishmash that is often associated with dial drives. Using this circuit with a frequency counter for readout will provide a rather neat receiver. Such niceties as variable or selectable R1C1/R2C2 time constants could be added.

Roy Propst, K4JFZ
Carrboro, North Carolina

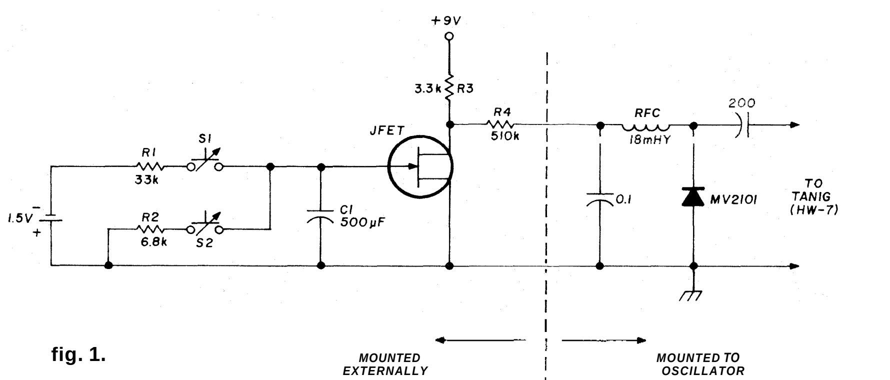


fig. 1.

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a second look

by Jim Fisk

What appear to me to be arbitrary and capricious decisions by FCC staffers in Washington have shown once again that the FCC bureaucrats are apparently interested in responding neither to public interest or public need. What I'm referring to, of course, was a recent announcement that FCC type acceptance of solid-state wideband amplifiers for the Amateur Service has been terminated ***without explanation*** by the FCC's Office of Chief Scientist. Are Radio Amateurs to be denied the use of modern solid-state technology because of the autocratic decision of an obscure bureaucrat in a ***democratic*** government? Is this not a contradiction to paragraph 97.1(c) of the FCC's own Regulations which state that one of the fundamental purposes of the Amateur Radio Service is to encourage and improve ". . . the amateur radio service through ***rules*** which provide for advancing skills in the . . . technical phases of the art?"

This recent action is just another in a long series of official FCC decisions which are contrary to the needs and desires of the Amateur Radio community — the linear amplifier ban, an unpopular and ridiculous callsign system, equipment type acceptance, the ASCII ban, recommending to the World Administrative Radio Conference (WARC) that CW be an "option" for the Amateur Radio service. This last item is a real dilly and stresses the need for closer congressional scrutiny of the Commission.

Several years ago, as part of the WARC preparations, the FCC formed the Advisory Committee for Amateur Radio (ACAR) and gave them the task of recommending, on behalf of Amateur Radio, what proposals should be made by the United States at WARC '79. ACAR carefully reviewed Article 41, which contains miscellaneous rules pertaining to the Amateur service including a Morse code proficiency requirement for operation below 144 MHz, and proposed no changes. As the WARC preparations proceeded, the FCC released Notices of Inquiry in Docket 20271 which requested public comment on various WARC draft proposals. The Commission requested comment on a proposal of "no change" to Article 41; those who responded supported that proposal. The FCC staff, however, chose to ignore both ACAR's advice and the public comments and recommended to the State Department that the United States' WARC position should include a proposal to delete the requirement for Morse code proficiency!

In Geneva the United States delegation proposed to make the code requirement below 144 MHz a "recommendation" rather than a requirement, a position that was supported by both Canada and Japan. Fortunately, some 15 administrations opposed the move as did every Radio Amateur in attendance. Brazil argued that any change to Article 41 would jeopardize existing reciprocal licensing agreements; Sweden proposed a 28-MHz cutoff; Papua New Guinea suggested 30 MHz. In the end the Papua New Guinea proposal won out and will be recommended for adoption by the conference. In effect, this lowers the frequency for a code-free Amateur Radio license to 30 MHz, a change which affects only six meters. Thus Amateur Radio lost a little — it could have been much worse — and the blame falls directly at the feet of unknown staffers within the FCC.

How could this happen? The ***publicly stated*** position of the FCC in April, 1977, regarding Article 41 reflected both the advice of ACAR and majority public opinion; its recommendations to the U.S. State Department less than two years later proposed a deletion of the Morse code requirement, in direct contradiction to the public's wishes. Apparently the change was conceived by some staff member (or members) within the Commission in direct violation of the Administrative Procedures Act, and no one in authority felt strongly enough about their public responsibility to veto it. It's no secret that the CB industry has been applying tremendous pressure for a code-free high-frequency operator's license, and this recent effort to sneak an unpopular proposal to an international forum leads one to believe they may have found a responsive element; except for Amateur Radio's friends in the international community, they would have succeeded.

Jim Fisk, W1HR
editor-in-chief

comments

linear power amplifiers

Dear HR:

I would like to comment on Bill Orr's article on linear amplifier construction in the June issue. I believe his warning against military surplus tubes is much too conservative. I'm afraid most Radio Amateurs watch costs. New high-power transmitting tubes will go a long way toward driving the cost of a homebrew linear past the cost of a commercial amplifier. I have built a linear using three 813s in parallel, and at 1500 watts PEP it performs beautifully, just as predicted in author Orr's own *Radio Handbook.* Few tubes can beat the 813 for clean response or ruggedness (carbon plate version). I have acquired a collection of six JAN 813s, mostly from other hams; all have been tested in my amplifier and all work beautifully. For a small fraction of the cost of a 3-1000Z or even two 3-500Zs, I have a good set of tubes in my amplifier and a spare set on hand.

If you're considering building your own linear amplifier, read everything you can, such as W6SAI's fine articles, then build it yourself, and *scrounge.*

William Brain, KB5EY
Houston, Texas 77040

Right. There's nothing wrong with buying a surplus 873 from a fellow ham for a few bucks and trying it out in your rig; maybe he'll even take it back if it doesn't work.

It's all a matter of judgment. How about buying a surplus 8877? You can save nearly a hundred dollars over the user's price if you buy a surplus tube. But what if your 'bargain" tube is bad? If you have bought it from a surplus dealer by mail, do you think you will get a refund? Fat chance. It all depends upon how much of a risk you want to take. You don't lose much with a surplus 813 or two.

Being in the power tube business, I am familiar with tear-stained letters from hams who have bought a JAN-branded, expensive, power tube and have been dismayed to find the tube bad and no warranty on it. But if you understand the limitations on warranty with respect to surplus tubes, and have the ability to test your tube immediately upon getting it home land stand a reasonable chance of getting your money back if the tube is no good), why not? After all, plenty of people lose a wad of money every day at the horse races. But they have the fun of watching the horses run.

Bill Orr, W6SAI
Menlo Park, California

memory keyer

Dear HR:

I have just finished building the memory keyer featured in the April issue of *ham radio;* I wish to thank Robert C. Cheek and your magazine for a beautiful and accurate article on the construction of this keyer.

I substituted 21LO2 memory chips for the 2102s and then added a 7400 gate with all inputs tied to ground. A single switch in the 5-volt supply to the 7400 will give a high or inactive output on the four gates; three of these outputs are tied to the chip-enable pins of the 21LO2 memory chips. With the addition of separate switches in 5-volt supply lines to the memory board and the keyer, and a separate supply line with a 70-ohm resistor bypassing the switch to memory board, you can hold the memory in a power-down mode with a drain of less than 40 mA (compared with over 300 mA to the memory board alone at 5 volts).

I found the power supply ran a little warm when the drain was high and the keyer was left on overnight to re tain the memory. The 70-ohm resistc reduces the supply voltage to 1. volts to the memory board; the chip enable pins on the 21LO2s are floa ing at about 4.5 volts from gates (the 7400 chip. With the 5-volt supp to the 7400 switched off, the ga outputs are inactive and the 21LC functions as before the modificatio Switching to memory power-dow mode must take place before the volt supply to the memory board opened or else memory is lost. The is a zero rime factor from power on power down.

William Hans
Glenwood, Illinc

split-band speech processor

Dear HR:

Congratulations to Wes Stewa N7WS, for his fine article, "Sp Band Speech Processor," in Septe ber, 1979. Wes mentions that the (cuit is sensitive to rf and that t proper use of ferrite beads, bypa capacitors, and rf shielding is imp tant; he is correct. To this end would like to suggest that rf byp capacitors be added in parallel w the 1N914 clipper diodes to preven mixing. Values of several hund picofarads should be sufficient.

Since symmetry is extremely portant in the prevention of sec order harmonics, I would also rec mend replacing the 1N914 clipper odes with diode pairs such as Mo ola's MSD6150. Using two dio manufactured on the same subst provide close matching of V_f other electrical characteristics. I also an ideal means of keeping t diodes at the same temperature, suring the best clipping symm possible. The additions have prc themselves in the processor I using.

James D. Allen, WA2S
Rochester, New Y

(Continued on page 12)

comments

(Continued from page 6)

lightning protection

Dear HR:

While I was pleased to see two letters in the July issue complimenting my article on lightning protection, I must take exception to W6RTK's suggestion that the ground wire on a wooden pole be broken into short lengths, with small spark gaps between segments.

The main ground conductor on a wooden pole is one of the most important items in the protection system. The establishment of a low impedance path from the air terminal on top of the pole to ground is necessary to send the greatest possible percentage of the total lightning stroke current directly to ground. Although lightning will certainly jump across the small gaps recommended, the presence of these gaps will have a negative effect on the performance of the overall protection system. I don't know how to quantify the amount of degradation, but I don't think it's wise to take a chance. Also, even if lightning doesn't strike the pole, the breaking up of this ground lead may allow the entire antenna system to acquire a large static charge, possibly sufficient to cause minor equipment damage.

Mr. Caldwell is concerned that this ground wire may have some undesirable effects on the performance of the antenna system; when considered from the standpoint that the ground wire only makes the wooden pole look electrically equivalent to a metal tower, I think it is safe to say that this effect must be minimal.

John E. Becker, KSMM
Prospect Heights, Illinois

quartz crystals

Dear HR:

It has been brought to my attention that a statement in my article on quartz in the February issue was misleading if not incorrect. While it is true that crystals have high inductance and low motional capacitance, this is not the reason for high Q; Q is a ratio of the charge stored to the charge dissipated. In crystals the charge is primarily stored by the inductance, so Q is determined by the value of the motional inductance divided by the motional resistance:

$$Q = \frac{2\pi f K1}{R1}$$

Most of us associate high inductance with a large piece of iron wrapped in copper wire — an arrangement which is completely ineffective at rf. With quartz, you must rethink the problem.

Don Nelson, WB2EG
Voorhees, New Jerse

10 meters for satellite communications

Dear HR:

I write in response to W1GMF who suggests (July *ham radio)* thi we discontinue rhe use of 29.360 29.502 MHz for satellite cornmunici tions. To put this into proper perspe tive, the frequency spectrum reserve for this purpose is no wider than tw 25-kHz wide repeaters, adding bo input and output bandwidths, y serves fifty times as many stations on an intercontinental basis.

Apart from the aspect of commun cations, the use of 10 meters h been the basis of valuable resear into sub-horizon communications, and aurora detection and forecastin and low-level signal techniques; it also of great value in using Amate Radio for teaching practical physic geometry, trigonometry, astronom and mathematics through the use a simple antenna and receiver.

The present maximum in the sc cycle will soon begin to decay, lei ing only the satellite devotees effectively occupy the high end of meters. This will help prevent int sion and takeover of the top part the band, safeguarding it by regu valuable usage.

Pat Gowen, G3I
Norwich, Engla

More Details? CHECK — OFF Page 126

CMOS 2-meter synthesizer

Construction details of a 2-meter synthesizer featuring choice of output frequency and CMOS design

Soon after joining the 2-meter fm crowd about three years ago, I learned how limiting "rockbound" mobile operation can be. At the same time, I had been wanting to learn more about frequency synthesizers, and designing and building one looked like the perfect answer to both needs. The result of my labor tunes from 146.000 to 147.995 MHz in 5-kHz steps and provides a variety of output frequencies to permit use with quite a number of rigs. This article will give full details on how the design was thought out, as well as how to build a copy. If you are interested in the subject of synthesizers, want to design one of your own but wonder where to start, or have soldering iron in hand ready to begin building, this article is for you.

design requirements

In addition to the above description, I expected the completed design to meet the following requirements:

1. Thumbwheel switch selection of receive and transmit frequencies

2. High output purity, at least 60-dB spur rejection

3. Self-contained; simple construction and cir cuitry

4. Minimum of test equipment needed to align an test

5. Minimum modification of 2-meter rig

6. Capable of mobile operation

operating frequency

The first choice was an operating frequency for th synthesizer. By looking at schematics and informa tion on a number of common rigs, I learned that mos use a receive crystal near 45 MHz. Transmit frequer cies are less consistent and include *f/6*, *f/12*, *f/1* and *f/24*. The circuit simplicity and purity goals rule out the use of multipliers; therefore, I picked 45 to 4 MHz, or one-third the channel frequency.

synthesizer concepts

The next step was finding the most suitab method of synthesis. A literature search showed th the most popular type of synthesizer today is an elal oration of the phase-locked loop (PLL). **Fig. 1** show the block diagram of such a system. In this, the VC (voltage-controlled oscillator) is made to run at times the reference frequency, f_R, which is normal fixed. Because of loop feedback, changing the divi ratio, N, also changes the VCO frequency to mainta the frequency relationship shown.

Because of its apparent simplicity, this kind of sy thesizer seemed like an ideal approach for my desig but, after studying the logic required, some serio complications were obvious. The need for an i-f sh to go from transmit to receive made the design a r mess.

By Tom Cornell, K9LHA, RR2, Box 531 Greentown, Indiana 46939

Several synthesizer schematics showed a different approach that looked like it had real promise; **fig. 2** shows the basic block diagram. The design freedom introduced by choice of crystal frequency allows the same variable divider ratio to be used in both transmit and receive. This results in great simplification of the overall synthesizer system.

practical design

This section will cover the more important design considerations.

VCO. The simplest form of VCO is a varactor-tuned oscillator, and a common circuit is shown in **fig.** 3. While a VCO can be developed by trial and error, it is much easier to calculate tank circuit values using the equations shown in the **Appendix** of this article.

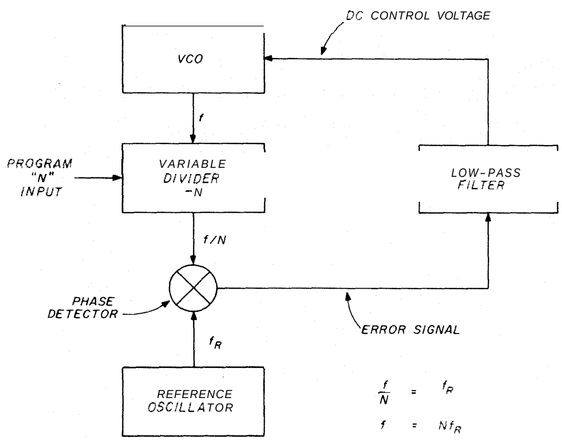


fig. 1. Block diagram of a simple synthesizer using a phase-locked loop.

Assuming a receiver i-f of 10.7 MHz, the minimum and maximum VCO frequencies are:

$$f_{max} = 147.995/3 = 49.33166\,MHz$$
$$f_{min} = (146.000 - 10.7)/3 = 45.1\,MHz$$

After picking some varactors (Motorola MV-2209) and suitable end point voltages, I was able to begin using the equations.

$$C_{max} = 50\,pF\,(1V) \qquad C_{min} = 28\,pF\,(6V)$$

Letting $C1 = 330\ pF$ and $C2 = 33\ pF$ and assuming 3 pF of transistor and stray capacitance, the total fixed capacitance, T, was 33 pF. The equations then gave $C_p = 88.2\ pF$ and $L = 0.19\ \mu H$. The circuit was built using these values and worked just about exactly as intended.

Oscillator/mixer. Design of the oscillator/mixer has quite an impact on the variable divider, and, after much study I decided on:

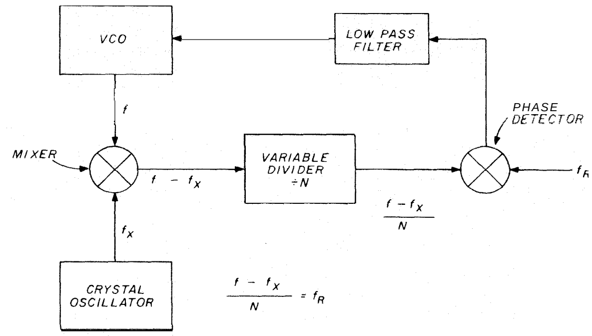


fig. 2. Diagram of a mixing-type synthesizer where the same divide ratio can be used for either transmit or receive, with the change in frequency accomplished by shifting the crystal oscillator frequency.

$$N\ at\ 146.000\,MHz = 400$$
$$f_{xtal}\,(TX) = 48\,MHz$$
$$N\ at\ 147.995\,MHz = 799$$
$$f_{xtal}\,(RX) = 48 - i\text{-}f/3$$

These numbers are a good illustration of synthesizer operation, and it might help your understanding if you plug them into the equation shown in **fig. 2.**

After selecting these points, design of the oscillator and mixer was relatively uncomplicated. A dual-gate MOSFET with an untuned output circuit was selected as the mixer, and two separate oscillators were used for receive and transmit.

Variable divider. The variable divider design has a lot to do with the complexity of a synthesizer circuit, and an intelligent choice is very important. Some of the divider requirements have already been covered, and the remaining important characteristic is speed. For this kind of synthesizer, the highest divider speed is:

$$f_{in}(max) = N_{max} \times f_R = 1331.66\,kHz$$

After studying the above requirements and the data sheets of a number of prospective devices, I chose the RCA CD4059 as the most suitable. This IC is a CMOS, 5-stage, BCD-programmable counter which has exactly the capability needed in this design. Additional factors favoring this choice were the inherent properties of CMOS. This logic family offers greater circuit density than TTL. It also consumes far less power, which incidentally means that there will be much less high-frequency energy produced to cause interference in other parts of the synthesizer.

Phase detector. A second CMOS IC, the CD4046, was selected as the phase detector. This is a special purpose device designed specifically for such an

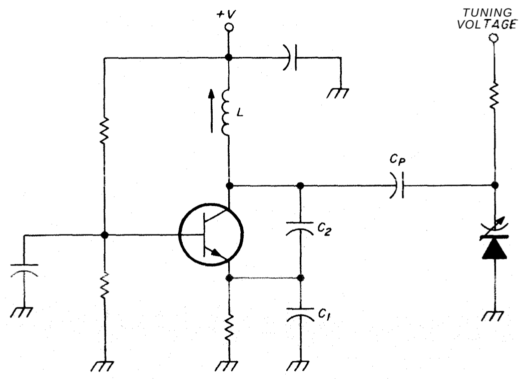


fig. 3. Schematic diagram of the basic VCO oscillator. The computations for the component values are shown in the appendix.

application. In addition to the phase detector, the CD4046 has a lock detection circuit that will be described later.

Lowpass filter. To keep the phase detector switching products from frequency modulating the VCO, a lowpass filter is inserted in the tuning line to the VCO. Because of its inherent phase shift, this filter is also to a large degree responsible for determining the PLL stability. And, while there are formulas for calculating filter values, I felt that the complex relationships involved would dictate some "cut and try" anyway, so that was the design approach I used. A circuit from another synthesizer was used as the starting point, and experimentation helped to determine the final values.

Fig. 4 shows the basic circuit. R1 and R2 together with C1 establish the main cutoff frequency, which must be somewhat less than the system reference frequency. C2 and C3 must be several times smaller than C1 to avoid instability and are included to add to the filtering action. R3 dampens the filter to control overall synthesizer system stability. My design method was to listen to VCO harmonics on an fm receiver and to make a sudden change in synthesizer frequency. R3 was then adjusted until the system demonstrated stable transient behavior.

Reference-frequency circuit. The reference frequency of a synthesizer is normally equal to the channel spacing. For this design, the spacing is 5 kHz and the reference frequency is 5/3 kHz, since the VCO operates at one-third the output frequency. For reasons of stability and accuracy, crystal control is usually considered a must.

After looking at several alternatives, I chose a crystal frequency of 2.56 MHz and a CD4060 CMOS oscillator/divider IC to generate the 5-kHz signal. A CD4027 dual J-K flip-flop was then used to divide by three to get 5/3 kHz.

System tests. When all of the preceding synthesizer circuits were hooked together, Murphy put in his first appearance. Switching from receive to transmit invariably causes the loop to drop out of lock, and output signal purity was awful. The first problem resulted from something I overlooked; transmit1 receive switching caused the input frequency to the variable divider to jump by i-f/3, which could exceed the reliable counting speed of that IC. Adding a frequency-shifting circuit to the VCO to retune the tank for the receive and transmit ranges solved the problem.

A buffer amplifier was placed between the VCO and mixer because it was found that the crystal oscillator was the cause of spurs in the VCO output: and the oscillator signal was getting to the VCO through the mixer. Purity now measured better than 60 dB, so I figured the design was adequate. After finishing the rest of the circuits, doing printed circuit artwork, building the synthesizer and hooking it up to my rig, I learned that Murphy doesn't give up very easily! The oscillator/tripler of my rig's receiver degraded purity to only 55 dB. Since the crystal oscillator was still the source of unwanted signals, lowering the output of the crystal oscillator was the logical way to reduce

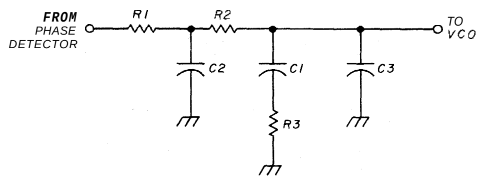


fig. 4. Schematic diagram of the lowpass filter that is inserted between the phase detector and the VCO.

the spurs. Unfortunately, this change resulted in insufficient drive to the variable divider, and new circuit boards were required to add the extra amplifier between the mixer and divider needed to bring the level back up. But the effort was worth it. Purity at the synthesizer output improved to about 75 dB, and at the receiver mixer 65 dB.

Output divider. To obtain the various output frequencies needed from this synthesizer, I made the logical choice of a divider stage driven by the VCO. Of the frequencies listed earlier, only f/18 presented any problems. The others could easily be derived by dividing the VCO frequency by 2, 4, or 8. Not providing an f/18 output did seem to be a compromise of the original requirements, but the improvement in circuit simplicity looked like a desirable trade-off, especially since only one rig that I knew of (Regency HR-2B) used f/18. Builders are still encouraged to consider this synthesizer design even though they may have to substitute for a small portion of the circuitry in order to get the precise frequencies.

operation

This section will briefly describe the function of the major circuit elements shown in **fig. 5.** The VCO is composed of 0 3 and surrounding components. CR1 is the tuning varactor. Three buffer amplifiers and U6, the output divider, follow up the VCO. Buffer 0 4 acts as a squaring amplifier to convert the sinewave VCO output to a squarewave suitable to drive NAND gate buffers U5D and U5B. U5D amplifies the f13 VCO signal, and U5B isolates this output from U6 and its back-fed divider products. Any of the four divider outputs from U6, as well as f/3 from U5D, can be jumper-selected as the synthesizer receive and transmit output frequencies. NAND gates U5A and U5C actually supply these outputs to the transceiver.

To the left of the VCO are Q2 (the MOSFET isolation buffer) and Q1 (the VCO frequency-shifting transistor). During receive, Q1 is turned on and places C14 across the VCO tank. A TTL logic-level signal at the Q1 collector also serves to turn off the transmit output during receive.

Below Q1 are the two crystal oscillators, which together with mixer 07 were added to simplify the variable divider design. Either Q5 or Q6 is turned on by application of supply voltage. LC tanks are in series with both crystals to allow slight adjustment of actual oscillator frequency.

The mixer output signal is amplified by Q8 and 0 9 to an adequate level to drive U1, the variable divider. As shown, U1 looks deceptively simple; actually, it is very busy inside. The divide ratio is loaded from the switch inputs, U I counts down N pulses to zero, produces a single output pulse, and then reloads the divide ratio to begin the cycle again.

All the frequency selector switches are shown in two groups below U1. Diode OR-gates between the transmit and receive switches isolate the two groups of switches and allow selection of transmit or receive operation by mere application of supply voltage. Toggle switches are used for both MHz and 5-kHz ranges, since they are all that is necessary and are much cheaper than thumbwheels.

To the right of U1 is the phase detector, U2. This IC provides a tuning voltage for the VCO at pin 13 that is filtered by the lowpass filter composed of R23, R24, R25, C27, C28, and C29. At pin 1 of U2 is the lock-detector output, which is normally high when the PLL is locked and goes low in a series of pulses when out of lock.

010 and Q11 amplify and stretch the lock detector pulses of U1 to produce a continuous logic-level signal that both lights the UNLOCKED indicator and shuts off the synthesizer transmit output. C26 serves to slightly delay the turn-on of 010 so that slight disturbances of the loop don't shut down the transmitter.

ICs U3 and U4 generate the synthesizer reference frequency. U4 contains a crystal oscillator running at 2.56 MHz and a divide-by-512 circuit (in this application) to produce 5 kHz. U3 then divides this frequency by three to produce 513 kHz. C23 allows for exact adjustment of oscillator frequency.

In the power supply, three-terminal regulators U7 and U8 provide 5 and 8 volts respectively. Control gates 012 and Q13 are driven by the transceiver push-to-talk line and select either receive or transmit operation of the synthesizer by providing logic supply voltages. The input LC filter, L8 and C49, serves to protect the synthesizer from transients and noises from the car's electrical system.

construction

The two synthesizer circuit boards may be assembled in any order, with the exception of the CMOS IC's which should be saved to the last to avoid damage from static electricity. (See **figs. 6** and **7,** respectively, for the circuit board pattern and parts placement diagram.)" Clip a ground wire from the soldering iron to the ground copper of the board when soldering these ICs.

Fig. 8 shows construction of the VCO coil.† Since this coil form was chosen for reasons of mechanical rigidity, you will need to cover the completed winding with Q-dope or airplane cement to ensure coil stability. Tighten the shield can to the base by making small impressions on at least two sides of the can into the plastic base with a center punch.

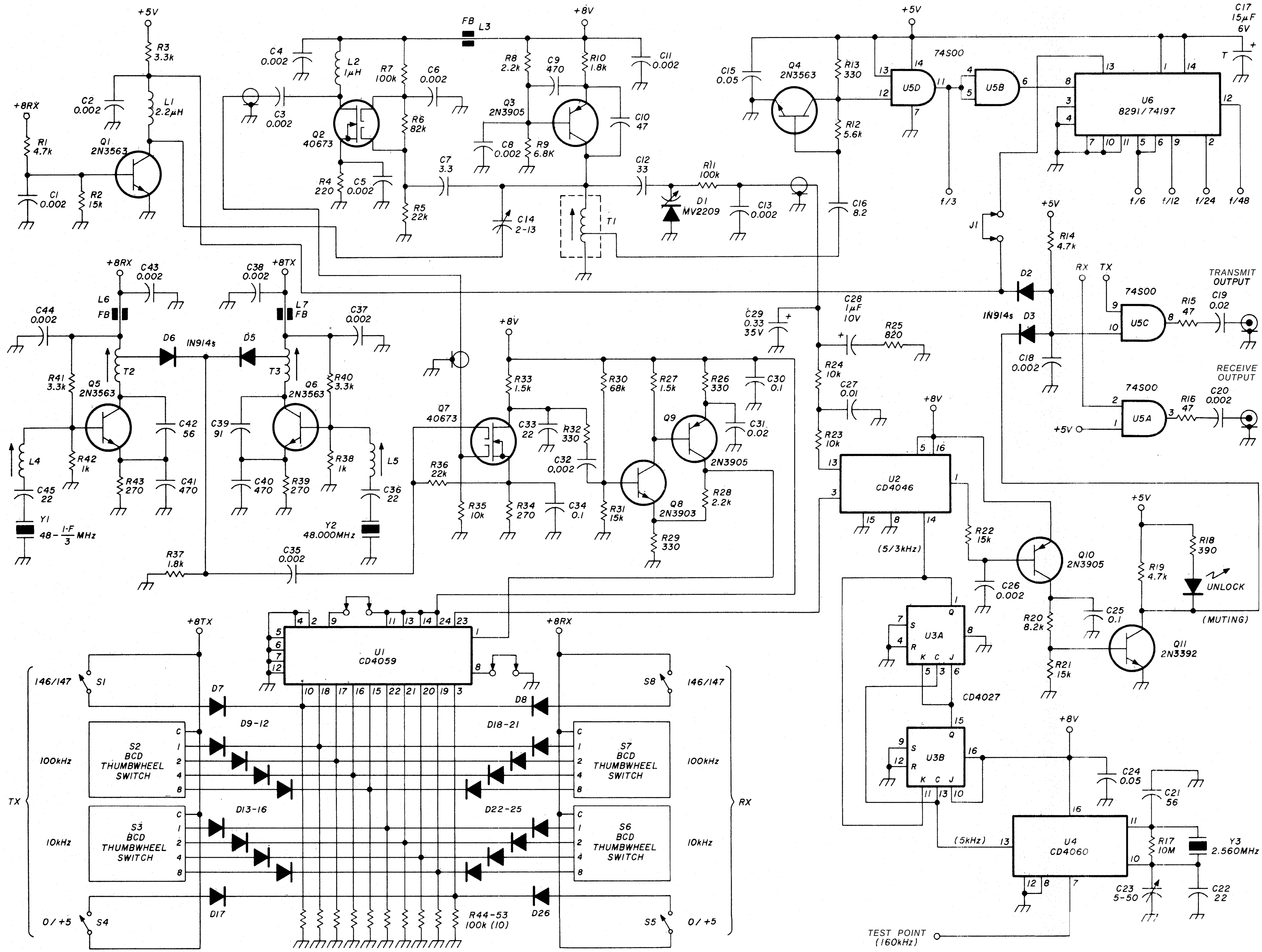
The plastic-molded coils used in the crystal oscillators may prove hard to find, and you can probably substitute most any good-quality coil forms of suitable size. Information on the number of turns is shown in the parts list.

L8, the supply filter choke can be made, or a suitable commercial part used. To make the choke, cut the heads off some small-diameter nails and tape them together to form a core roughly 30-mm (1 1/4-inch) long by 5-mm (3116-inch) in diameter. Wind a coil of about one-hundred turns of no. 22-26 AWG (0.6-0.4 mm) wire over the core. Finish by covering with electrical tape. Form the leads to fit the circuit board and mount to the board with ordinary string or wire wrapped over the body of the coil.

Install jumpers to select the correct output frequencies for your rig. For receive, connect a small piece of insulated wire from pin 2 of U5 to one of the divider outputs. For transmit, the jumper goes from

*The circuit boards and many components to build the synthesizer are available from Radiokit, Box 429, Hollis, New Hampshire 03049.

†The VCO coil form is a standard 10-mm i-f transformer form. If you are unable to find such a part, it may be purchased from the author for $1.00.


+5V
R3 3.3k
C2 0.002
L1 2.2μH
+8RX
R1 4.7k
Q1 2N3563
C1 0.002
R2 15k
C4 0.002
L2 1μH
C3 0.002
Q2 40673
R7 100k
C6 0.002
R6 82k
R4 220
C5 0.002
R5 22k
FB L3
+8V
R8 2.2k
C9 470
R10 1.8k
C11 0.002
Q3 2N3905
C8 0.002
R9 6.8K
C10 47
C7 3.3
C14 2-13
T1
C12 33
R11 100k
D1 MV2209
C13 0.002
C16 8.2
+5V
C15 0.05
Q4 2N3563
R13 330
R12 5.6k
74S00
U5D
U5B
f/3
U6 8291/74197
C17 15μF 6V
f/6
f/12
f/24
f/48
J1
+5V
R14 4.7k
D2
IN914s
D3
C18 0.002
RX
TX
74S00
U5C
R15 47
C19 0.02
TRANSMIT OUTPUT
74S00
U5A
R16 47
C20 0.002
RECEIVE OUTPUT
+5V
+8RX
C43 0.002
L6 FB
C44 0.002
R41 3.3k
T2
Q5 2N3563
D6
IN914s
D5
C38 0.002
+8TX
L7 FB
C37 0.002
T3
R40 3.3k
Q6 2N3563
C39 91
C42 56
R42 1k
L4
C45 22
R43 270
C41 470
C40 470
R39 270
R38 1k
L5
C36 22
Y1 48 - 1-F/3 MHz
Y2 48.000MHz
R37 1.8k
C35 0.002
+8V
Q7 40673
R33 1.5k
C33 22
R32 330
R36 22k
R35 10k
R34 270
C34 0.1
R30 68k
R27 1.5k
R26 330
C30 0.1
C32 0.002
Q9
C31 0.02
2N3905
R31 15k
Q8 2N3903
R28 2.2k
R29 330
C29 0.33 35V
C28 1μF 10V
R25 820
R24 10k
C27 0.01
R23 10k
+8V
U2 CD4046
(5/3kHz)
R22 15k
C26 0.002
Q10 2N3905
+5V
R18 390
R19 4.7k
UNLOCK
C25 0.1
(MUTING)
R20 8.2k
R21 15k
Q11 2N3392
U3A
CD4027
U3B
+8V
C24 0.05
C21 56
(5kHz)
U4 CD4060
R17 10M
Y3 2.560MHz
C23 5-50
C22 22
TEST POINT (160kHz)
U1 CD4059
+8TX
+8RX
146/147
S1
D7
D8
S8
146/147
TX
100kHz
S2 BCD THUMBWHEEL SWITCH
D9-12
D18-21
S7 BCD THUMBWHEEL SWITCH
100kHz
RX
10kHz
S3 BCD THUMBWHEEL SWITCH
D13-16
D22-25
S6 BCD THUMBWHEEL SWITCH
10kHz
0/+5
S4
D17
D26
S5
0/+5
R44-53 100k (10)

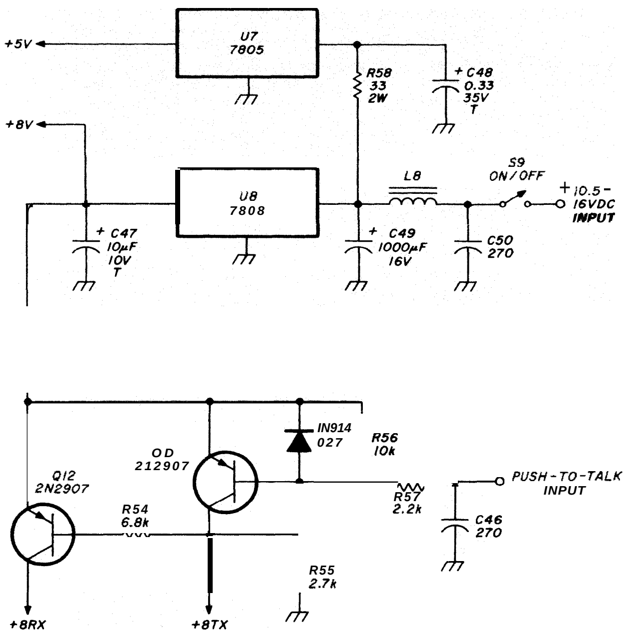


fig. 5. Schematic diagram of the CMOS **synthesizer (opposite). The individual portions of this schematic are discussed in the text. All small-valued capacitors are** NPO **ceramics. Power supply is seen above; parts list is below.**

L4 11 112 turns 6.5 mm (114 inch) diameter, slug-tuned, close-spaced molded plastic form

L5 10 112 turns, same as L4

L8 see text

T1 5 112 turns, tap at 2 turns, no. 32 AWG (see **fig. 8**).

T2 6 112 turns, tap at 1 314 turns, spaced 1 wire diameter, no. 26 AWG (0.4-mm) wire, plastic-molded 6.5 mm (114-inch) diameter form, J-iron core

T3 same as T2 except aluminum core

Y1 f = (48 – i-f/3) MHz, see text (44.4333 MHz, i-f = 10.7), same as Y2 except frequency

Y2 48.000 MHz, series mode, third overtone, 0.0025 per cent tolerance, HC-18/U case

Y3 2.5600 MHz, parallel mode, fundamental, 32-pF load, HC-6/U case with wire leads, 0.005 per cent tolerance

pin 9 of U5 to a divider output. Note that the receive frequencies actually contain an i-f offset and are really (f – i-f)/3, (f – i-f)/6, etc. If the f/3 receive option is used, install jumper J1 as shown in **fig. 7.** This will turn off U6 during receive and eliminate some low-level subharmonic spurs U6 produces. If f/3 is not used for receive, install jumper J2 instead to allow U6 to operate in both transmit and receive.

Temporarily install interconnecting wires between the two boards to allow circuit alignment. Connect the following: 5 volts, 8 volts, 8 volts RX, VCO tuning voltage (coax), and the VCO buffer output (coax). The synthesizer will operate on 146.000 MHz in this condition.

alignment

Alignment of the synthesizer requires the following equipment: a dc voltmeter (VTVM or high-input impedance), and a-m/fm radio (a portable set is fine), the diode detector probe shown in **fig. 9**, and a regulated power supply (preferably current limited). Other useful equipment includes a frequency counter (50 MHz, high-input impedance), a grid-dip meter, and a general-coverage receiver.

Connect the synthesizer to a 12-volt supply; it should draw approximately 125 mA. Next, check the 5- and 8-volt supplies, which should be within 5 per cent of the correct value. Test the 8V-RX and 8V-TX lines. With the push-to-talk line open, 8V-RX should read 8 volts and 8V-TX about a volt. Grounding the push-to-talk input should bring the 8V-TX up to 8 volts and drop 8V-RX to zero.

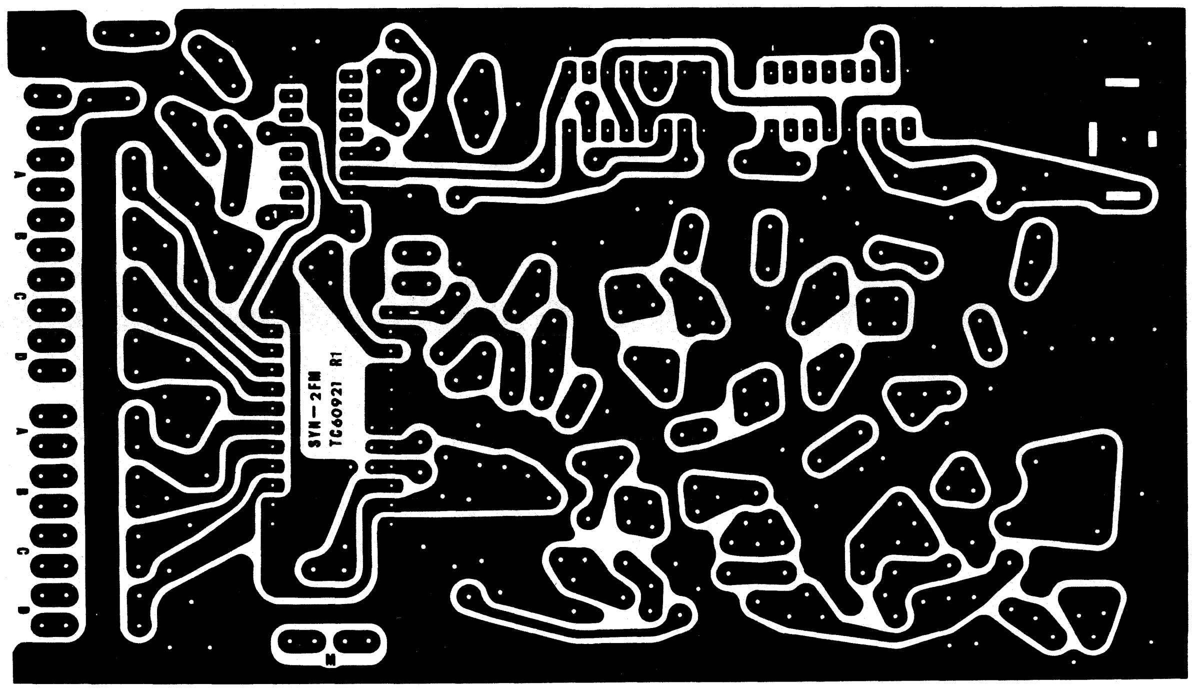
Next, some kind of check on the 2.56-MHz oscillator should be made. There are several possibilities, including connecting the diode probe to pin 7 of U4 (dc output voltage should be around 6.5 volts); measuring the same point with the counter (it should be 160.000 kHz, adjust with C23); and listening with the a-m/fm receiver (antenna near U4) at 640 or 1280 kHz, or listening with the communications receiver at 2.560 MHz. Alignment can be by the counter or by comparing one of the U4 divider products (1.28 MHz, 640 kHz, etc.) with a known frequency such as an a-m radio station.

The easiest method for testing the receive crystal oscillator is to hold a grid-dip meter near T2 as the slug is adjusted. An fm receiver tuned to the second harmonic of the oscillator can also be used, as well as the diode detector probe connected across R37. Adjust the slug of T2 for maximum output and then turn it toward the top of the coil until the dc voltmeter connected to the diode probe reads 0.25 volts. Ground the push-to-talk line, and then make the same adjustment and check on the transmit crystal oscillator and T3. If you are able to use an aluminum slug in T3, remember that, compared with an iron slug, it works backwards.

If the oscillators refuse to run, temporarily bypass the base to ground with a 0.001-to-0.002 µF capacitor. You can then find out what the free-running frequency of the oscillator is and make corrections in T2 or T3 or the value of C39 or C42. If you can adjust the oscillator in this condition to the crystal frequency, then crystal control should work, too.

If you have a high-impedance counter available, connect it across R37. Adjust the slugs of L4 and L5 to fine-tune the frequency of each oscillator. Without a counter, you may be able later to arrange some type of on-the-air check to adjust frequency.

Measure the VCO tuning voltage with your dc volt-



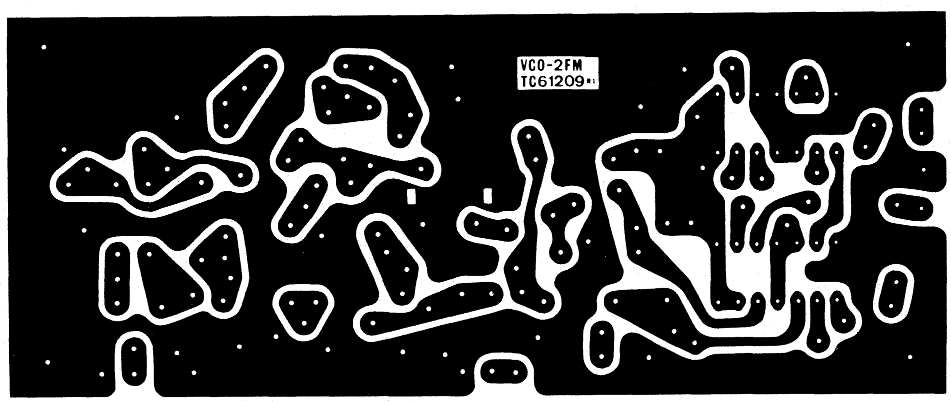


fig. 6. Foil pattern for the synthesizer board (above)and the VCO **board (below).**

meter and ground the push-to-talk input. If the voltage is 8 volts, turn the slug of T1 toward the top of the can. If the voltage is, instead, zero, turn the slug into the coil. Adjust for a final reading of 1.0 volt. At this point, you should be able to hear the VCO harmonic at about 97.3 MHz on the fm receiver.

If the tuning voltage cannot be adjusted, the problem may be one of several things. A high tuning voltage is caused by a high VCO coil inductance, and a low voltage by low inductance. If slug adjustment is insufficient, the coil turns may need to be changed; the turns can be spread apart or squeezed together. As an alternative, C10 can be changed slightly to get the right tuning range. A dead VCO, mixer, crystal oscillator, or mixer output amplifier will also cause the tuning voltage to go to 8 volts.

Once the VCO works correctly in transmit, remove the ground from the push-to-talk line. Adjust C14 for a tuning voltage of 1.0 volt and verify operation by listening to the VCO's second harmonic on the fm radio at 90.2 MHz.

final construction

The synthesizer boards may now be boxed to your preference. My experience has indicated two possible critical areas: the VCO board will very likely require a complete shield to keep transmitter rf away from the VCO. For the same reason, wires between the two circuit boards should be kept away from the power/control wires entering the box from the transceiver. Once the unit is assembled, I usually recommend a touch-up alignment of the VCO and crystal oscillators to compensate for any stray capacitance added by the case. When the frequency selector switches have been connected, you will have your first opportunity to check full operation of the syn-

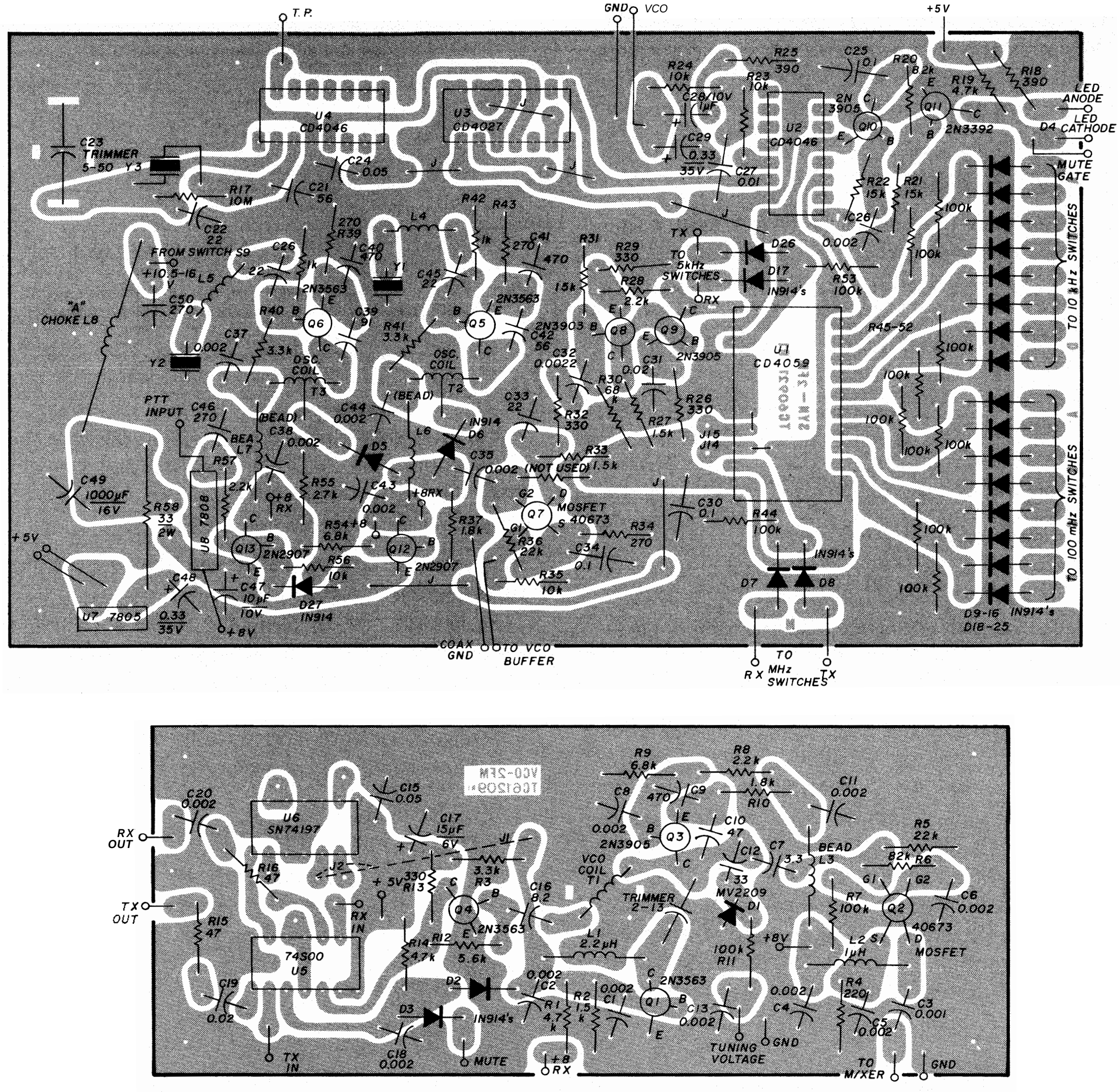


fig. 7. Component placement diagrams for the synthesizer board (above) and the VCO board (below).

thesizer by listening to it on the fm radio or your 2-meter rig.

connecting your rig

Fig. **10** shows the circuits I used to couple the synthesizer to my rig's receive and transmit oscillators. Install these right at the crystal sockets of your rig, drill holes, and mount two coax connectors on the rear of your rig. Connect up the entire system using coax cable. Make sure the inductor tunes to the transmit crystal frequency with the capacitors of your oscillator circuit, and adjust the resistor (470 ohms in **fig. 10)** to keep the transmit oscillator from running on its own (you will be mighty unpopular on 2 meters if it does).

Next, connect the power and push-to-talk lines from your rig to the synthesizer. Shielded cable is strongly recommended for this purpose.

The synthesizer and transmitter should now be thoroughly tested in the transmit mode, first on a dummy load and then on an antenna. Any rf that gets into the VCO can cause instability and flickering or illumination of the LED indicator. The dummy load check will determine if your rig is feeding rf back from its oscillators into the synthesizer. This condition may be corrected by insertion of a lowpass filter, having a cutoff frequency just above the synthesizer's output frequency, in one or both of the synthesizer output lines. The antenna test is somewhat more complicated in that certain antenna types, especially the gutter-mount and magnet-base variety can cause appreciable ground currents to flow on the coax. The solution in these cases is wire rerouting away from the synthesizer and possibly within the synthesizer, and use of additional shielded wires plus the VCO shield.

additional possibilities

The simple BCD programming of this synthesizer makes it easily adaptable to some interesting fre-

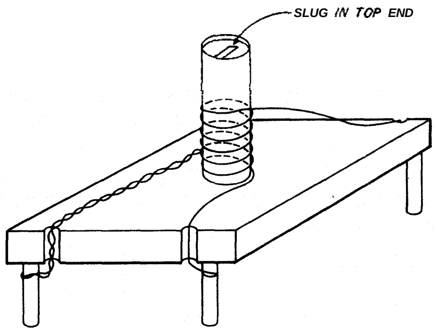


fig. 8. Detailed diagram of the VCO coil. The form is a standard 10-mm shielded slug-tuned form.

quency control methods. Replacing the switches with an up-down counter will allow scanning as well as LED frequency readout. A memory can be used to store favorite channel frequencies, or a microprocessor can be added for all kinds of control functions including scanning all channels, a group of channels, or those you preset. Your imagination is the limit.

When this synthesizer was developed, I intended to build at least one unit to cover 150 to 159.995 MHz to tune some of the vhf mobile channels, That is the

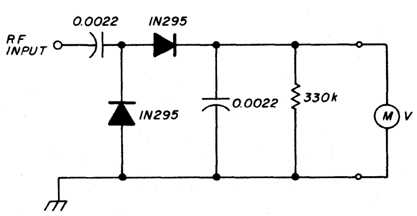


fig. 9. Schematic diagram of an rf diode probe suitable for tuning the synthesizer.

reason for the two jumpers beneath U1. Although I've not had time yet to try this, expanded coverage might appeal to you. This design will, therefore, permit operation over the expanded 2-meter band as proposed by the FCC.

final comments

I hope that this article has proven valuable to you. Should you have questions about the design or prob-

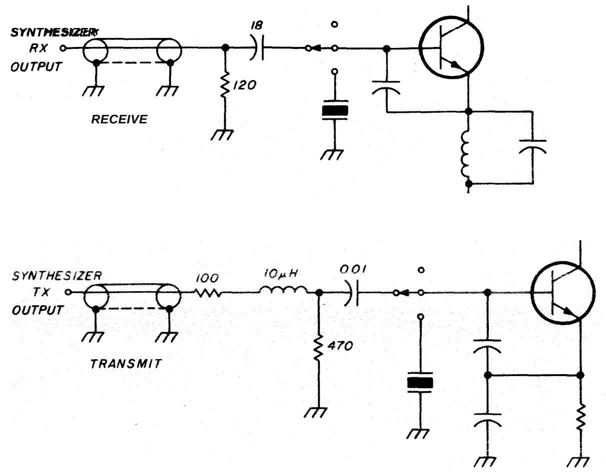
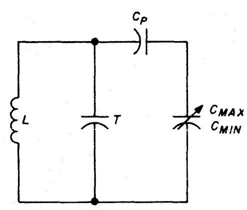


fig. 10. Diagrams of the interface circuits between the synthesizer and the rig's receive and transmit crystal oscillators.

lems in construction, please feel free to write me, but do enclose an SASE.

I'd like to offer my thanks to those who helped me in this project: Dib, K9HLG, and Tom, W9IJ, for their encouragement, counsel, and interest in the project; to Russ, K9AYD, for his valuable ideas on logic and synthesizer design; and to Bill, WA9GUY, for his mechanical help and engraving of the front panel.

appendix



$$\alpha = \left(\frac{f_{max}}{f_{min}}\right)^2$$

$$C_p = \frac{-b + \sqrt{b^2 - 4c}}{2}$$

$$b = \frac{(1-\alpha)\,[T(C_{max} + C_{min}) + C_{max}\,C_{min}]}{(1-a)\,T + C_{max} - \alpha C_{min}}$$

$$C = \frac{(1-\alpha)\,T\,C_{max}\,C_{min}}{(1-\alpha)T + C_{max} - \alpha C_{min}}$$

$$C_T = T + \frac{C_p\,C_{max}}{C_p + C_{max}}$$

C_T = total tank C at f_{min}

$$L = \frac{1}{(2\pi f_{min})^2\,C_T}$$

$$= \frac{25330.34\,\mu H}{(f_{min})^2 C_T}$$

f in MHz
C_T in pF

ham radio

environmental aspects of antenna radiation

How to calculate approximate near-field radiation levels to meet existing environmental standards

At present there's a great **deal** of interest and controversy regarding non-ionizing electromagnetic radiation and its effect on the environment. This issue, of course, affects Amateur Radio. Suggestions have been made that all nonionizing electromagnetic radiation be eliminated from residential areas, or that such radiation be limited to levels that would make Amateur Radio operation impossible.

Some groups, in a wave of hysteria, are attempting to make *all radiation* illegal. As in most situations of this type, when one looks at the facts, the picture becomes clearer.

In a report by the U.S. General Accounting Office dated March 29, 1978,[1] it states that 10 mW/cm^2 is the maximum level to which a human should be exposed for 6 minutes per hour, and that 1 mW/cm^2 is the maximum continuous exposure. In other words, to be completely safe, one should stay at levels of 1 mW/cm^2 or less. These levels are recommended by the American Natio'nal Standards Institute (ANSI) for frequencies between 10 MHz through the microwave region.

analysis

I have calculated the approximate separation distances between the radiation source (Amateur antennas) and humans to meet the recommended levels in reference 1. These data are shown in figs. **1A** through **1D** for four Amateur antennas: half wave, quarter wave, eighth wave, and sixteenth wave. Parametric curves show the input power to the antenna at two field-strength levels, 1 mW/cm^2 and 10 mW/cm^2.

Looking at fig. **1A,** one can see that if an Amateur operates on 7 MHz using a half-wave antenna with 100 watts input, the antenna must be at least 4.6 meters (15 feet) from any human to keep the field strength at 1 mW/cm^2 or less. At a power of 1 kW input to the antenna, the field at 4.6 meters (15 feet) increases to 10 mW/cm^2. Thus it's necessary to move the antenna a distance of 7.3 meters (24 feet) from any human to reduce the field to 1 mW/cm^2.

If the antenna has 10 dB gain in one direction, the equivalent antenna input would be 10,000 watts instead of 1000 watts. The field at 7.3 meters (24 feet) would increase to 10 mW/cm^2 in the direction of the antenna gain.

The apparent free-space field strength near any antenna can be approximated by:

$$p_{field} = \frac{755 L^2 P_{ant} K}{Z_{ant} \lambda^4} \text{ watts/meter}^2 \qquad (1)$$

where
L = length of antenna (meters)
P_{ant} = power input to antenna (watts at Z_{ant})
Z_{ant} = input impedance of antenna (ohms)
λ = wavelength in meters ($300/f_{MHz}$)
$K = \frac{3(\alpha r)^2 + 5}{(\alpha r)^6}$ (values in table 1)
$\alpha = 2\pi/\lambda$
r = distance from the antenna (meters)
f = frequency

After the apparent free-space field strength has been calculated in watts/meter^2 it can be converted to mW/cm^2 by multiplying the calculated value by 0.1. In other words, 100 W/m^2 is the same as 10 mW/cm^2. If the field strength in volts/meter is desired, the following expression can be used:

$$E_{field} = \sqrt{120\pi (P_{field}\ W/m^2)} \text{ volts/meter} \qquad (2)$$

By John Abbott, K6YB, P.O. Box 66, Newhall, California 91322

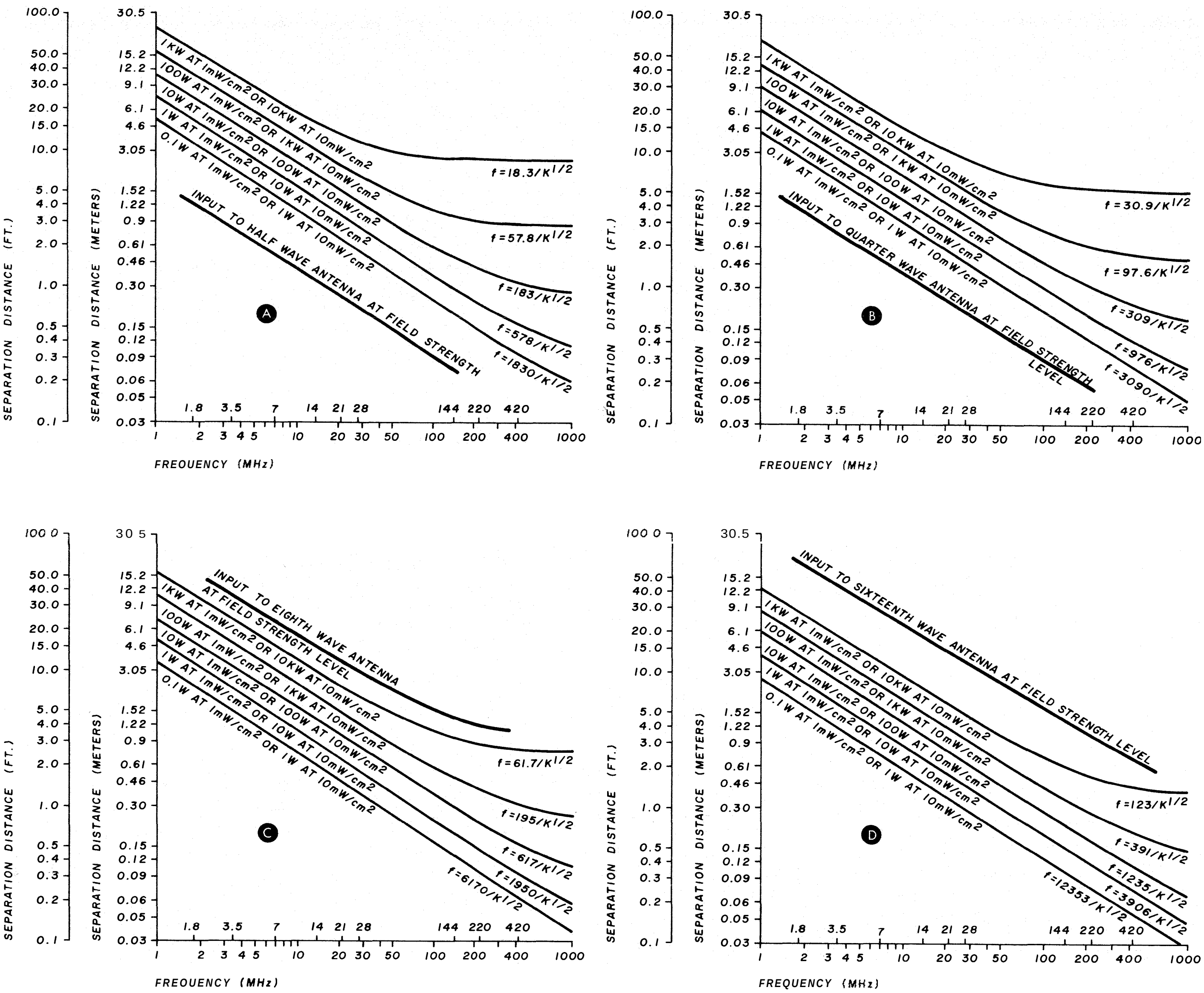


fig. 1. Separation distance as a function of frequency for recommended input power to four Amateur antennas: (A) one-half wavelength; (B) one-quarter wavelength; (C) one-eighth wavelength; and (D) one-sixteenth wavelength.

where 120π is the impedance of free space. Using this expression, 10 mW/cm^2 is the same as 194 volts/meter field intensity.

Kvalues (**eq. 1**) are shown in **table 1** as a function of r/λ, the ratio of the distance from the antenna in meters to the wavelength in meters. Using this table and **eq. 1** for P_{field}, it's possible to calculate the fields at various distances from an antenna to obtain the apparent free-space field intensity. Otherwise, use **fig. 1** to make sure that your antenna is always at a separation distance with less than 1 mW/cm^2 field intensity. (A mathematical derivation of **eq. 1** is available from ***ham radio*** upon receipt of a self-addressed stamped envelope).

practical considerations

There should be no problem for most Amateurs in installing an antenna away from houses and areas occupied by humans, except for the 160- and 80-meter bands. In these cases it may be necessary to limit power input to the antenna if necessary distances can't be maintained. The real difficulty lies in the operation of handheld portables above 25 MHz. If adequate separation from the body is maintained, it will be difficult to talk into a handheld unit. You'll have to decide if the risk is worth the exposure.

Mobile operation above 25 MHz should be no problem if simple precautions are followed. **Tables 2** table 1. Proximity coefficient K as a function of the ratio of distance from antenna, r, to wavelength, λ.

ratio of distance from antenna, r, to wavelength, λ r/λ	proximity coefficient, K
0.01	81,000,000.000
0.015	7,170,000.000
0.02	1,282,000.000
0.04	20,620.000
0.06	1,890.000
0.08	364.700
0.10	103.000
0.15	12.300
0.20	3.110
0.25	1.186
0.30	0.640
0.35	0.388
0.40	0.252
0.45	0.187
0.50	0.135
0.60	0.089
0.70	0.0596
0.80	0.0452
0.90	0.0346
1.00	0.0272
1.20	0.01872
1.40	0.01368
1.60	0.0104
1.80	0.0080
2.00	0.00654
2.50	0.00416
5.00	0.001028

and 3 show a summary of approximate operating distances that should prevent overexposure for most Amateur installations.

I'd like to emphasize that, in this article, I make no attempt to account for the shielding effects of buildings or the susceptibility of humans to radiation at any given frequency. The field levels are simply calculated at each frequency shown. It may well be that 1 mW/cm^2 is more of a hazard at 420 MHz than at 1.8 MHz. Such matters will have to be explored by medical research.

The data presented here will allow Amateurs to estimate field-strength levels from the antennas described in a manner that will meet present recommended criteria. Furthermore, the data will provide ammunition with which to fight pressure groups who are trying to abolish Amateur Radio!

addendum

The effects of radiation from electronic equipment on the environment has become a hot issue of late. The FCC has issued a Notice of Inquiry (NOI), General Docket 79-144 (June 15, 1979) which states in paragraph 33:

"It may be desirable for the Commission to consider the need for applying to the subjects of its jurisdiction one of the existing safety criteria, such as the 10 milliwatt per square centimeter (10 mW/cm^2) short-term exposure limit used by ANSI and OSHA . . ."

Furthermore, ANSI is considering reducing this level to 1 mW/cm^2. What does all this mean to Amateur Radio? The answer is presented in the article above.

If you are concerned you'll want to file comments to the FCC/NOI mentioned above before the December 15, 1979 deadline.

hr report has been publishing material on this subject since early March, 1979. The following excerpts from ***hr report**** are for those wishing more background information:

> ***PROHIBITION OF RADIO TRANSMISSIONS*** in residential areas is being considered by the Oregon State Senate. Senate Bill 423, sponsored by Senator Ted Hallock of Portland, proposes sharply restricting all electromagnetic emissions in residential areas.
> ***In Testimony Favoring*** the bill Merrie Buel, government affairs coordinator for the Oregon Environmental Council, said that medical studies "have found that persons living next to electromagnetic sources often experience serious health effects, including rashes, headaches, dizziness and tingling sensations."
> ***Power Transformers*** and transmission lines as well as radio and TV transmitters would be curtailed under the bill's provisions, though Senator Hallock and members of the Senate Committee on Environment and Energy have been discussing removing transmission lines from its coverage.
> ***As Written*** the bill would become effective January 1, 1983, after which violations of the standards established for it would be a misdeameanor punishable by a $250 fine. However, Ms. Buel termed the $250 fine "merely a slap on the hand," stating that her group felt that "endangering

table 2. Approximate operating distances between an antenna and humans for 1 mW/cm^2 or less exposure.

frequency (MHz)	antenna length and minimum separation meters (ft.) with 100 W antenna input for 0.1 mW/cm^2 field or 1000 W antenna input for 1 mW/cm^2 field: half wave	quarter wave	eighth wave	sixteenth wave
1.8	16.8 (55)	13.7 (45)	11.0 (36)	8.5 (28)
3.5	13.0 (36)	9.1 (30)	7.0 (23)	5.5 (18)
7.0	7.3 (24)	5.8 (19)	4.6 (15)	3 7 (12)
14.0	4.9 (16)	4.0 (13)	3.0 (10)	2.4 (8)
21.0	4.0 (13)	3.0 (10)	2.4 (8)	1.8 (6)
28.0	3.7 (12)	2.7 (9)	2.1 (7)	1.5 (5)

**hr report* is published by Communications Technology, Inc., Greenville, New Hampshire 03048.

table 3. Approximate operating distances between **portable/mobile** antenna and humans.

frequency (MHz)	half wave 10W	half wave 1W	quarter wave 10W	quarter wave 1W	eighth wave 10W	eighth wave 1W	sixteenth wave 10W	sixteenth wave 1W
	Antenna length and minimum separation, cm (in.) with 10 W or 1 W antenna input for 10 mW/cm²* (divide power by 10 for 1 mW/cm² field)							
50	55.9 (22)	40.6 (16)	48.3 (19)	33.0 (13)	35.6 (14)	25.4 (10)	27.9 (11)	20.3 (8)
144	27.9 (11)	17.8 (7)	25.4 (10)	15.2 (6)	17.8 (7)	12.7 (5)	15.2 (6)	10.2 (4)
220	22.9 (9)	15.2 (6)	17.8 (7)	12.7 (5)	15.2 (6)	10.2 (4)	10.2 (4)	7.6 (3)
420	15.2 (6)	10.2 (4)	12.7 (5)	7.6 (3)	10.2 (4)	7.6 (3)	7.6 (3)	5.1 (2)

"Maximum exposure at 10 mW/cm² should be limited to 6 minutesihr.

people's health should be considered a much more serious offense." Furthermore, she said, the OEC wants the bill to become law much sooner since, "we suggest that the sooner electromagnetic radiation is under control, the safer the public health." (*HRR* 245, March 16, 1979).

EFFECTS OF CB ANTENNA RADIATION on the bodies of nearby people is being investigated by the Department of Heath, Education and Welfare. The first study, published in a 24-page booklet titled "Measurement of Electromagnetic Fields in Close Proximity of CB Antennas," discusses bumper, trunk lid and rooftop-mounted mobile antennas as well as those on hand-held units. Near field radiation distribution of each type is presented graphically, in an attempt to determine what hazard, if any, radiation presents.

The Study Concludes: "The health implications (of CB antenna radiation) are not clear at this time. The Bureau of Radiological Health is continuing to investigate this matter." HEW is obviously quite concerned with the effects of RF on the population, and with Amateurs running 200 times the power of CBers on frequencies from 1.8 MHz through millimeter wavelengths, our operations are sure to come under careful scrutiny as well – if they haven't already. (*HRR* 247, March 30, 1979).

AMATEUR RADIO WAS ATTACKED as "one of the main non-ionizing radiation hazards in the United States" at an April 9-10 meeting of the Subcommittee on Public Health Aspects of Energy, in New York. The group is an arm of the New York Academy of Medicine's Committee on Public Health, reports K6YB, who has an article on the effects of Amateur RF radiation on family and neighbors coming out in ham radio magazine later this year. (*HRR* 253, May 18, 1979).

RF RADIATION HAZARDS are the subject of a new FCC Notice of Inquiry, General Docket 79-144, agreed to by the commissioners earlier this month. Although the Commission noted that promulgation of RF radiation health and safety standards is the responsibility of health and safety agencies, it also recognized that it would have to consider radiation exposure standards adopted by other Federal agencies in its licensing activities.

Full Text Of This Potentially very important NOI, which is reported to contain a number of questions on specific areas of concern, hasn't yet been released. With the environment currently a hot public issue, this NOI could easily become a crucial one for Amateur Radio as well as most other radio services.

Comment Date for Docket 79-144 is December 15, with Reply Comments due March 15 of next year. (*HRR* 258, June 22, 1979).

ANOTHER FCC PROPOSAL that could affect Amateur Radio is in General Docket 79-163, which proposes changes in the Commission's environmental impact rules. At present those rules offer some leeway with respect to prospective stations that could have a "Major Impact" on the environment when the actual impact would seem less significant. Under the proposed tighter restrictions, it appears formal impact statements would be required of many more applicants, probably including a number of Amateurs, and the Commission would then have to prepare and distribute a written environmental assessment for each such case.

In A Dissenting Statement to the proposed change, Commissioner Washburn makes the point that environmental impact is not the Commission's business and to make it so would add to their already heavy workload and thus increase licensing delays.

Comments On Docket 79-163 are due by August 1. (*HRR* 260, July 6, 1979).

HIGH LEVELS OF RF RADIATION have been detected by the FCC in its test of some popular personal computers. Tests of computers manufactured by Atari, Apple, Commodore, Heath, Southwest Technical, and Radio Shack have reportedly shown that, in most cases rf radiation levels far exceed allowable Class 1 TV limits.

With The Popularity of home computers sharply on the rise, the FCC plans to use the data it's collected to set up new rules governing all computers that could be used in the home. It will probably be several months before the FCC decides what action to take and files a notice of proposed rule-making. (*HRR* 261, page 2, July 13, 1979).

references

1. *Efforts by the Environmental Protection Agency to Protect the Public from Environmental Non-Ionizing Radiation Exposure, U.S. General Accounting Office, Report CED 78-79. March 29, 1978.*
2. *Richard A. Tell, "Broadcast Radiation: How Safe is Safe?", Spectrum, IEEE, August, 1972, pages 43-51.*
3. *Reference Data for Radio Engineers, Chapter 25, "Antennas," Howard W. Sarns & Co.,* Inc., *Fifth Edition,* pages 25-1 *through* 25-3.

ham radio

the Hellschreiber
a rediscovery

European Amateurs are using a teleprinting system made from World-War II surplus — will it replace RTTY?

The Hellschreiber is a teleprinting machine based on a principle entirely different from that of the RTTY teleprinter. The Hell sytem (named after its inventor, Dr. Rudolf Hell) could have been invented with the requirements of the Radio Amateur in mind, but strangely enough the Hell system has never been fully accepted by the Amateur fraternity. The reason may be that an enormous number of used RTTY machines flooded the market at low prices after World War II.

Hell and RTTY existed simultaneously for a long time for both military and commercial use. However, Hellschreibers have now disappeared, mainly as a result of the introduction of protected RTTY systems with automatic-request and error-correcting circuits. Most hams have probably never heard of the Hell system as a means of communications.

the Hellschreiber

What is the Hellschreiber? In contrast to the RTTY machine, in which received pulses determine the character to be printed, the Hellschreiber uses the transmitted pulses to ***directly*** write images of characters on paper tape. Thus, Hell writing could be considered a simple form of facsimile, covering seven image lines per character, with seven elements per line.

Not only has this system of printing character images some very important advantages to offer, but the simple way in which the Hell teleprinter works is extraordinarily elegant. The thread of a fast-turning worm shaft wipes, with high speed, transversely across a slowly moving paper tape. This worm thread is wet with printing ink. Every time the paper is

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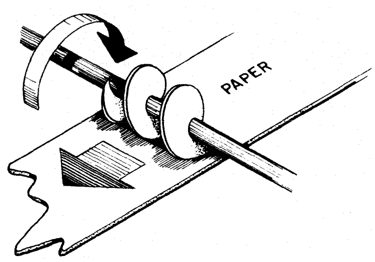
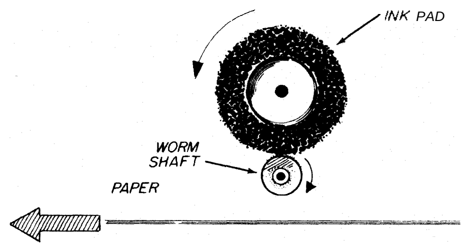
How The Hellschreiber Works

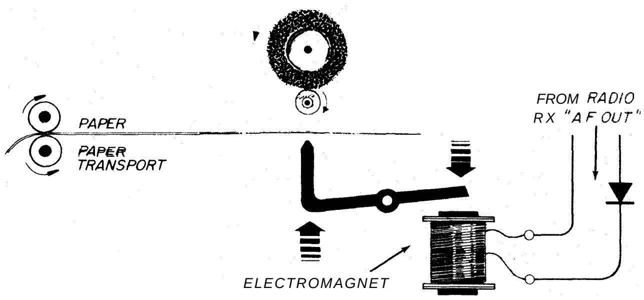
A. Imagine a fast turning worm shaft above a relatively slow-moving paper tape:



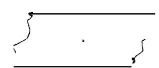
B. The thread on this worm shaft is kept wet with printing ink:



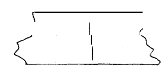
C. Under the paper is a mechanism that taps the paper against the worm shaft by means of an electromagnet:



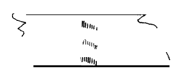
D. What is printed on the paper depends upon the rhythm and the length of time the electromagnet is actuated. For example, if the paper is just tapped, one gets:



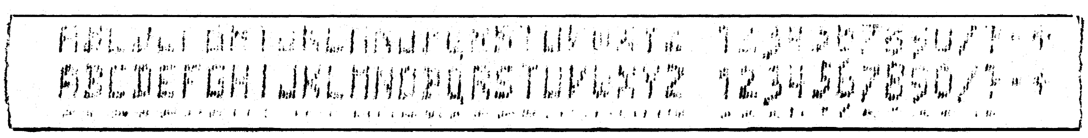
What you see are the little dots where the paper touched the fast-turning worm shaft. If the thread sweeps fast over the paper, and if the electromagnet pushes a bit longer, a little line is printed:



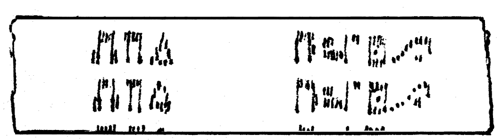
If the tape is tapped in rhythm with the revolutions of the worm shaft, a sequence of little dots is printed:



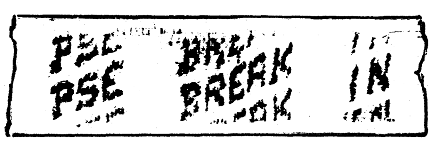
E. Thus, all sorts of simple images can be written; for instance, all the characters of the alphabet:

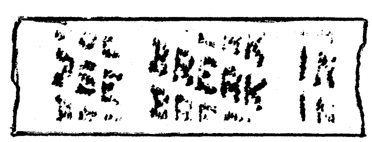


F. Or, if necessary, the characters of anybody else's alphabet, such as Greek, Arabic, or Chinese:



G. What happens if the worm-shaft speed is not quite correct? Nothing serious; the lines of the Hell text threaten to run off the paper tape:





This provides a simple method for determining the correct speed. If, for example, the lines show a tendency to drop, the motor speed must be increased until the lines run straight again. But whatever happens, the text remains legible.

Hell writing. This enlargement shows, how each character takes the space of seven image lines. As a result of the relativeiy siow-moving tape, the characters hang slightly over.

tapped against the turning wcrm shaft, little lines are formed across the paper tape. Several of these lines together form a character.

The Hellschreiber of the World War II Wehrmacht type we're using runs somewhat slower than the RTTY machine: 2½ characters per second. Nevertheless, a respectable 25 words per minute is achieved. This CW terminology is not misplaced, as Hell and CW have much in common. In fact, given a certain bandwidth, the reliability of Hell communications approaches that of CW.

QRM proof?

During World War II the Hellschreiber proved its reliability. Users recognized that a Hellschreiber could be the only link between an isolated military unit and its headquarters. When all other means of communications failed, often the Hellschreiber managed to get the message through, even when only barbed wire and an earth connection were available as a signal path.

Amateur applications

Our Hell QSOs occur on 80 meters (over here, the official RTTY segment is between 3575-3625 kHz). It's difficult to think of a better part of the radio spectrum for putting the Hell system to the test because of the high QRM level in this portion of the band.

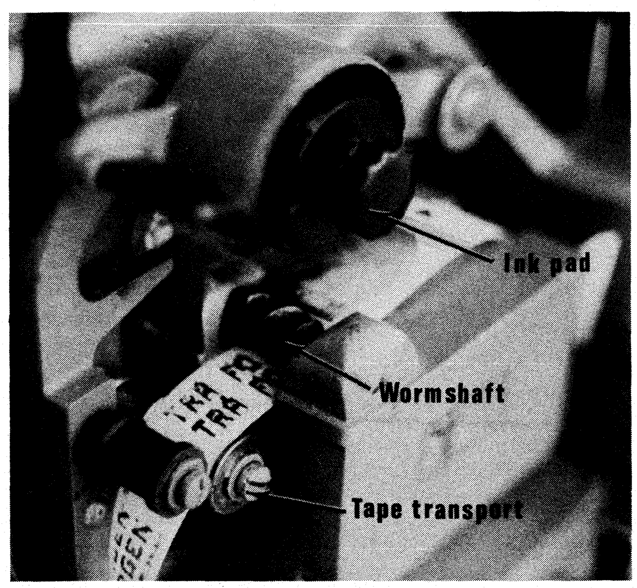
In this context I'd like to mention an interesting side effect. Our modest ***prrt, prrt, prrt*** Hell signals apparently tend to provoke fury among some hams, who seem to be convinced that the unusual sounds are caused by commercial stations. This turns our little Hell channel into the center of zero-beating and QRZ-blaring stations. This intentional interference does, however, provide us with an invaluable opportunity to test the communications system under highly adverse conditions and is, therefore, to some extent, not unwelcome.

Of course, the interfering transmitter determined to cause serious trouble by tuning carefully zero-beat with our Hell signals may eventually manage to temporarily destroy our communications, provided, of course, that the signal is stronger than ours. By maneuvering with tuning, bandwidth, and threshold level it's possible to get through. We might lose contact for a moment; however, contact is restored through the foggy QRM clouds on our printouts, and we pick up the text as soon as the characters become distinguishable again. This sort of working on the threshold is possible with Hell: The text, even under the worst conditions, is never subject to errors of a substitution-oi-characters rype. The character may, however, be difficult to read because of mutilation.

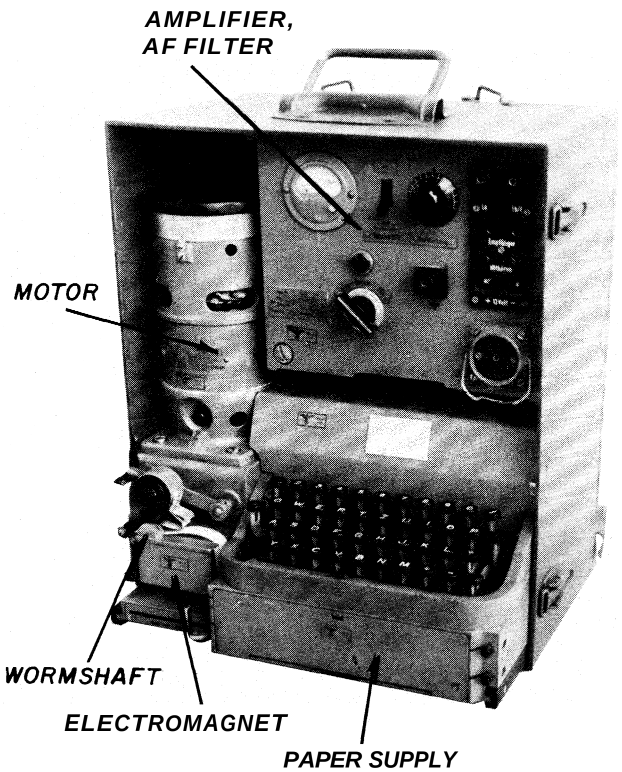
Hell *versus* RTTY

Under certain circumstances the communications reliability of Hell can be even better than that of CW. The received Hell signal is printed in its original form. At the moment of reception no decision has to be made such as, "Did I hear correctly?" Thus wrong decisions are avoided. The Hell printer enables the reader to decide later on, at his ease, what was actually sent by the distant station.

Some examples are shown of radio Hell-communications in which the printer obviously has great trouble in keeping the text intelligible because of a high noise level or heavy QRM. The examples contain considerably more information than can be deciphered on first sight. If you really take the trouble to read the text, you immediately realize to what the Hellschreiber owes its superior qualities: it calls in the services of a computer, ***ie.,*** our human ability to recognize pictures in a chaos of little specks and lines.



Printing mechanism of the Hellschreiber. The ink pad (felt roll) has been lifted to show the worm shaft. The paper tape is slowly moved by the transport capstan. The electromagnet (not visible) taps the paper tape from below against the fast turning worm shaft.



The Feldfernschreiber. Hellschreiber of the German Wehrmacht (1938) as used on a large scale during World War II. It is with this type of machine that such remarkable results were obtained on the Amateur bands.

The Hell system is less sensitive to interference than RTTY because the Hellschreiber prints the interfering clutter as well as the desired text. This may sound paradoxical, but it becomes understandable if you realize that a teletype printer must translate its received signal into a character before it can decide which key must be pressed. It cannot count upon the services of a "computer." Thus, with RTTY, a single interfering rf spike may result in a wrong decision, turning out a character that has no resemblance whatever to the actual character transmitted. The unprotected teletype character can't warn the reader that it is in error; it can't even indicate that a certain amount of doubt existed during the moment of its selection!

The Hellschreiber, on the other hand, requires no such decisions. The machine just prints, complete with all the received interference. But (and this is the important distinction) although the interference may give the image of the characters an untidy appearance, the Hellschreiber is not capable of changing it. In other words, the Hellschreiber simply leaves to the boss the problem of sorting out the text from the rubbish and doesn't try to disguise the difficult reception conditions.

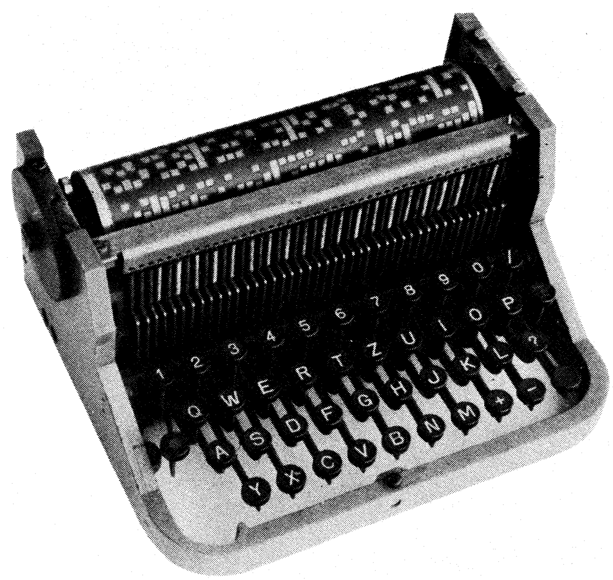
This is the explanation for the rather amazing fact that you may read Hell text from signals that are only barely audible through an overwhelming amount of QRM; indeed, that it's even possible to decipher Hell signals received *under* the noise level. No wonder we're highly enthusiastic about this fantastic system.

experience with Hell

For three years, almost every week, our little international Hell group (five Dutch, one German, one French, one British) make our regular Hell QSO of an hour or so, using one of the most crowded portions of the 80- and 40-meter bands. Our Hellschreibers are ex-Wehrmacht printers, some of them 40 years old and in fact valuable museum pieces.

As with CW and RTTY, the modest bandwidth requirements of Hell are a great advantage. They are determined by the shortest pulses contained in the signal, being 8.16 ms. This produces a speed of 122.5 baud, requiring a minimum bandwidth of 61 Hz. Even in an overcrowded band it's possible, with a sharp CW filter, to remove most of the QRM or, in case of telephony interference, to keep the bulk of the speech sidebands out of the picture.

Watching a Hellschreiber printer in operation, you can't help being impressed by its imperturbability: While the radio receiver produces the most frightening sort of QRM noises, the machine swallows it all. Quietly, apparently hardly disturbed by it all, it goes on spelling out its characters. Often the QRM is so bad that you need a Hellschreiber to establish that there's still a Hell signal in the air.



Transmitter section of the original Hellschreiber. The coded drum turns one revolution per character. Every time a key is pressed, one turn of the drum produces a series of pulses by the contact with one series of lamellas.

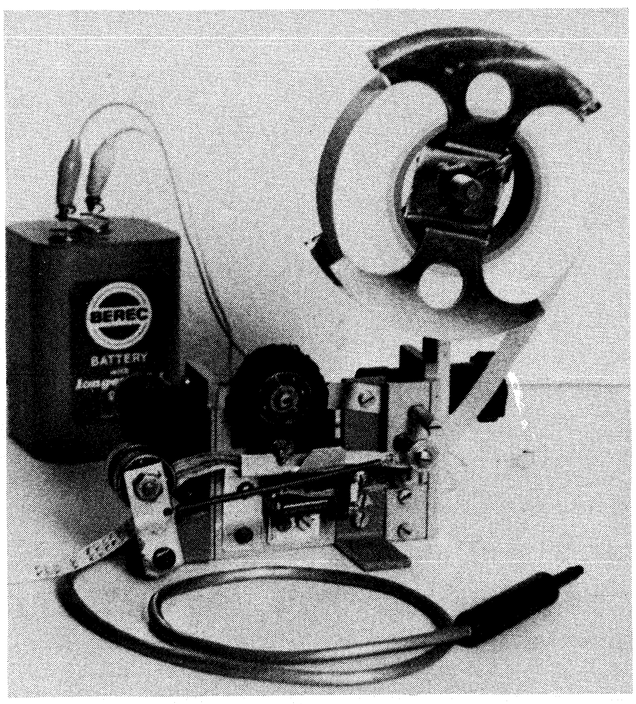
Between transmitter and receiver a certain amount of synchronization is needed, which requires a means of regulating receiver-motor speed. Not that this synchronization is very critical; contrary to what you might expect from a synchronous image-line system, the good old Hell machine is not so easily disturbed by the wrong motor speed. The only thing that might happen is that the written text might drop over the edge of the paper. The text remains legible, however, and, while continuing to read the text, you correct the motor speed by hand until the text prints correctly along the plane of the paper strip. It is this reliable, almost undisturbable, character of the Hell-schreiber that makes it such a fine instrument for Amateur Radio communications.

The CW-like disposition of Hell signals permits break-in. Spaces don't produce signals (the tape just runs without printing), so it's possible to cut in between words of the distant station's text. You can even keep watch on the QRM situation between transmitted words.

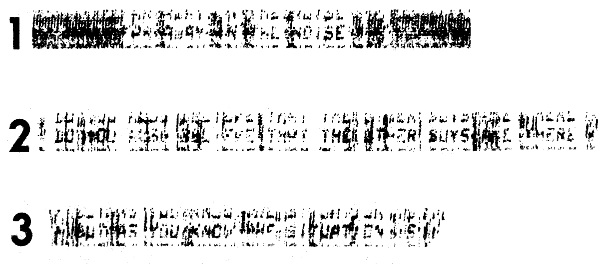
Hell is economical with transmitted energy. With considerable fewer marks than spaces in its signals, and without start and stop pulses, the average output is about 25 per cent of the maximum output. This low duty cycle permits increased transmitting power.

quo vadis?

It's possible to make a Hellschreiber yourself — something that can't be said for any ordinary teleprinter. The actual printer consists of only a simple mechanism. This is another advantage of the Hell-schreiber. The receiving part is easy to build and may be a good starting point. After gaining some experience with receiving Hell QSOs, you can decide whether it is worthwhile building a Hell transmitter.



Home made Hellschreiber.



Reception of Hell signals under extreme conditions.

1. Very weak signal, drowning in the noise. On first sight it's unusable; however, our ability to recognize pictures in a chaos of little specks permits us to read the text into the noise.

2. Interference by a strong SSB telephony signal on the same channel. (Text: "Do you also believe that the other boys are there".)

3. Hell signal exactly zero-beat with equally strong 14-wpm CW signal. (Text: "but as you know the situation is".)

We have already built some mechanical Hell printers. Of course, electronics have advanced considerably since 1938, and the dimensions of our modern Hellschreiber can no longer be compared with those of that bulky German design. We now have small electric motors with solid-state speed regulation and we can use refinements such as coils with ferrite cores to pick Hell signals out of overwhelming QRM. Accurately defined Schmitt-triggers are available for separating signals of different levels.

You could even go as far as PAØWV, who has developed a microprocessor displaying received signals as a slowly moving line of characters, complete with interfering pulses (thus fully maintaining all qualities of the Hell system) on an oscilloscope screen.

The transmitter part, "pulse machine," is somewhat more complex to build. In the original Hell-schreiber the transmit pulses were produced by a coded drum requiring some mechanical refinements. But a solid-state solution exists here. It was PAØWV again who built the first clock-plus-matrix system that can be hidden under a small keyboard, producing all characters in complete silence.

A converter isn't required for receiving Hell signals. The Hellschreiber can be plugged directly into the headphone jack of any radio receiver (or any telephone line, for that matter). The transmitter output plugs into the KEY jack of any CW transmitter, that's all.

ham radio

log-periodic antenna design

The LP is a useful antenna for Amateur applications — it provides constant gain and a low VSWR over a wide frequency range

The log-periodic (LP) array is a moderate-gain antenna useful for many Amateur applications. It has the desirable characteristics of constant gain and low VSWR over a wide frequency range. It's very forgiving of construction and design tolerances. Accordingly it doesn't require fancy test equipment or interminable pruning to achieve satisfactory operation. Minor errors may result in somewhat reduced gain but won't markedly affect the basic radiation pattern or front-to-back ratio. Once a design has been completed, it's seldom necessary to make adjustments after the antenna has been erected.

This article deals with the design of LP antennas using simple formulas that can be worked on any 4-function calculator. Also given is a simplified approach using only tables and elementary arithmetic, which allows you to design single or multiband LPs to fit into an available space. Examples shown are for wire antennas. For vhf arrays using tubing, appropriate changes should be made to obtain the effective element length. Robert Carrell presented an excellent paper, "The Design of Log-Periodic Dipole Antennas," which is in the IEEE International Convention Record for 1961. Data for the article here was, in a large part, derived from that paper.

The design gains shown here are approximate and may seem low to many readers. They are given as the free-space pattern gain with reference to a dipole (dBd). Many antenna designs are quoted as dB above isotropic (dBi) and sometimes include ground reflection gain over a perfect reflecting surface. Such approaches are misleading and can yield numbers anywhere from 2.2 to 8 dB higher over a dipole.

Amateur applications

The LP is particularly useful in split-band operation such as working DX on 40 or 75 meters, where the frequency separation of U.S. and foreign bands is frequently greater than the bandwidth of many other types of antennas, such as Yagis.

In many areas of the world, material such as telescoping aluminum tubing is difficult and expensive to obtain. The LP, either in a fixed configuration or rotatable in a design using forward V-shaped horizontal-wire elements supported by a shaped stress-line diamond, may be built from simple available materials: wire, bamboo, and nylon line. YV5DLT, Ansel Eckels, has built several of the latter configuration that have worked very well.

In all probability the number of Amateur bands will increase in the not too distant future; new bands have been proposed at 10.1, 18.1, and 25.25 MHz. An LP can be designed to cover 10-30 MHz, with performance and size making it comparable to many current triband beams. At least two Amateur manufacturers (KLM and Telrex) have a practical-size LP rotatable array that comes close to meeting this requirement. Alternatively, a multiband Yagi covering six bands would be quite a mechanical challenge. Unless extreme care is taken in trap design, it would be quite lossy.

For those who operate in the vhf bands, a single LP can yield good performance from 50 MHz through 432 MHz. After many years and many other design approaches, the LP has become the standard configuration for most TV antennas.

Don Bostrom, N6IC, in the DXpedition to Wallis Island in 1974, used a homemade LP with good suc-

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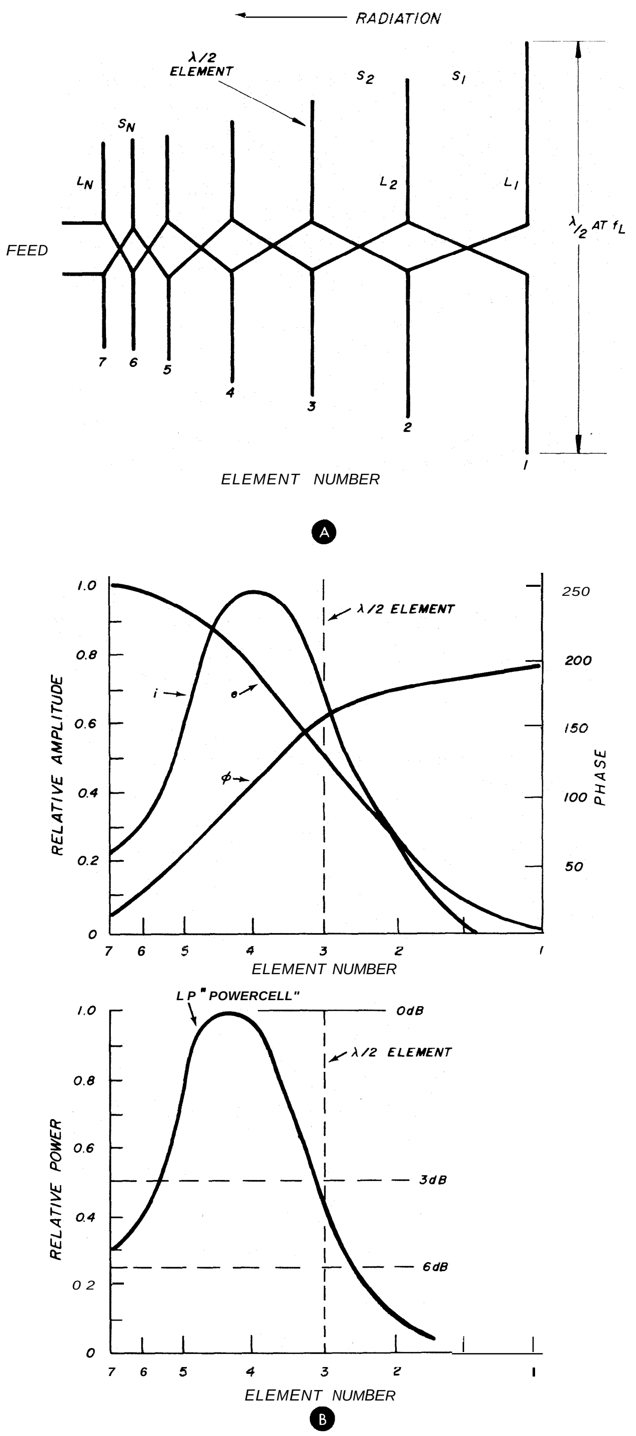

fig. 1. A 7-element LP antenna (A) with equations defining taper factor, τ, in terms of element length. L, and element-spacing relationships, S_1, S_2 . . . S_N. Characteristics are shown in (B). Upper curve: voltage, current. and phase at an arbitrary frequency higher than lowest desired operating frequency, f_L. Lower curve: approximate power distribution, illustrating the "LP cell," which occurs in the region where the elements are less than λ/2 long.

cess. It was a vertically polarized 5-element wire array for 10, 15, and 20 meters.

Compared with the Yagi, the LP will usually have somewhat lower gain. However, it does have a significant advantage in bandwidth and maintains its front-to-back characteristics over the entire design frequency. A properly designed LP working through a balun will allow a solid-state transmitter to operate efficiently over a wide frequency range without an antenna tuner. For example, a 7-element wire beam covering the low end of 80 meters to the high end of 40 meters, and with a good match, can be erected in a space roughly 43 meters square (140 feet square) and will have a gain of about 6 dB.

description

The basic LP consists of a number of dipoles arranged in a plane **(fig. 1A)**. The element lengths, L, and the relative spacing, *a,* are arranged in a geometric progression with a taper factor, τ. Each element is connected to the feeder in an alternating manner. The feedline is transposed between each set of elements as the easiest method with wire elements and as an intra-element feedline.

The array operates as a backward-wave antenna; radiation is in the direction of the feed. Propagation velocity is about 0.35. The free-space pattern in the plane at right angles to the elements is similar to a cardioid: egg-shaped in the radiation direction in the plane of the elements. The LP operates over a frequency band defined by the longest element, about λ/2 long at the lowest frequency, and the shortest element, about λ/4 at the highest frequency. The gain is constant over this frequency interval; therefore the beam width in E and H planes is constant in free space. Over real ground, the elevation beam maxima will change in angular position with frequency because the effective height in wavelengths will change.

Many configurations have been discussed in *ham radio* and other literature. These include the inverted T monopole, which is a vertically polarized ground-plane array; a configuration using inverted V elements; and truncated LPs, which use fewer than the optimum number of elements.

characteristics

Power to the elements is maximum in the region where the elements are somewhat less than λ/2 long. This is called the "LP cell" **(fig. 1B)**. The cell at a particular frequency encompasses the longest element, λ/2, and a few shorter elements. The shortest

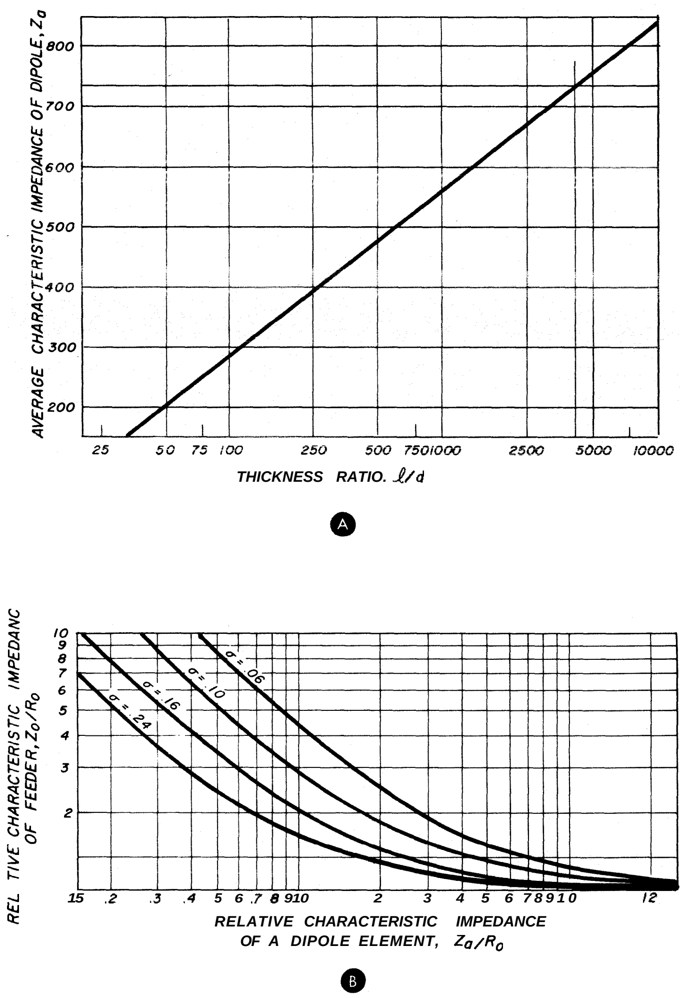

fig. 2. Data for the LP intra-array feedline design. The "natural" dipole impedance is selected from curve (A), which shows the average characteristic impedance of a dipole, $Z_a = 120\ (\ln \ell/d - 2.25)$. (B) shows relative feeder impedance as a function of dipole impedance with spacing factor, σ, as a parameter.

element is approximately $\lambda/4$ long.

The array will operate as an antenna at frequencies lower than that defined by the longest element of $\lambda/2$; however, the front-to-back ratio will degrade rapidly, and the gain will be impaired.

For large taper factors (τ near unity) many elements are within the cell and the array has high gain. For small taper factors only a very few elements will be within a cell, and the gain will be much lower. Table **1** shows how the gain for an optimum-gain design varies as a function of the frequency range and number of elements.

feed system

The LP at low frequencies is usually fed by coax through a balun. The LP feedpoint impedance, unlike that of other antennas, is a function of the natural dipole impedance and the spacing factor, σ. A convenient feedline impedance can be used with a balun at the antenna to obtain a low VSWR (fig. **2).** Air dielectric should be used to prevent excessive phase shift within the array feed. Any other dielectric increases the spacing factor, σ, in a complex manner. From tests run by W4AEO, the driving point impedance for low-frequency arrays at $\lambda/4$ high is on the order of 225 ohms, with an intra-array feedline characteristic impedance of 450 ohms. Typically the intra-array feedline is formed from 14 AWG (1.6-mm) wire spaced for a feedline impedance of about 450 ohms. This, with a 4:1 balun, results in a close match to 52-ohm coax.

array gain considerations

For each value of taper factor, τ, there is a corresponding value of spacing factor, σ, which yields maximum gain (fig. 3). Smaller-than-optimum spacing factors may be used with consequent loss of gain but without pattern degradation. Larger-than-optimum values of spacing factor cause undesirable lobes. The optimum spacing factor, σ, is approximately 0.19 times the taper factor, τ.

Some commercial rotatable or space-saving arrays have spacing factors as low as 0.03. For example, one may build an array covering 7-30 MHz on a 10-meter (33-foot) boom (sixteen elements, τ = ***0.895; σ = 0.03).*** It would have about 5.5 dBd gain over this band. With the optimum spacing factor of 0.17, the gain would increase to 7.3 dBd, but the boom length would become 59.5 meters (195 feet). Thus by compromising gain for bandwidth you can build an effective, very broadband array with a practical boom length. This may be more desirable than stacking severai potentially interacting Yagi arrays.

mathematical design

Here's a design approach for LP antennas using simple mathematics that can be worked on a 4-function calculator. It is presented to show how the basic design is evolved.

Definition of terms. All of the LP design terms needed for calculator implementation are as follows:

f_H = highest desired operating frequency

f_L = lowest desired operating frequency

$B = \dfrac{f_H}{f_L}$ = desired frequency ratio

B_s = structure bandwidth; ratio of length of longest-to-shortest element

B_{ar} = array bandwidth of active region (amount by which B_s is reduced to obtain usable bandwidth; $B = \left(\dfrac{B_s}{B_{ar}}\right)$

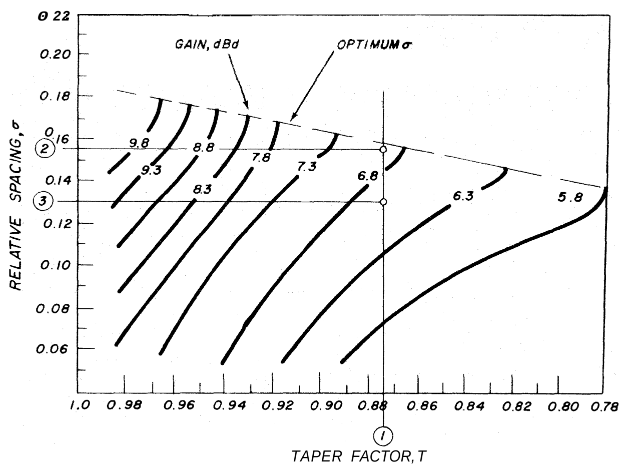

fig. 3. Relative spacing, σ, as a function of taper factor, τ, with contours of constant gain, dBd, as a parameter. Circled values are used in a simplified LP design approach as described in the text.

τ = taper factor

σ = spacing factor

λ_0 = free-space wavelength

$-\ \frac{300}{f_{MHz}}$ in meters; $\frac{984}{f_{MHz}}$ in feet

λ_L = free-space wavelength at lowest frequency

λ_1 = antenna wavelength using wire

$= \frac{286}{f_{MHz}}$ in meters; $\frac{936}{f_{MHz}}$ in feet

(λ_1 is **5** per cent less than the free-space number because of end effects.)

L = element length

= array length

S_1 = spacing of first two elements

dBd = gain above dipole, approximately 2.2 dB below isotropic gain (dBi)

design steps for log-periodic antenna

1. Select desired band ratio, $B = \frac{f_H}{f_L}$

2. Select an initial set of values for τ, the taper; and **a**, the element spacing factor from **fig.** 3.
Spacing of first element $= \sigma \times \lambda_L$
Other elements spaced at τ **X** previous S

3. Calculate array bandwidth, B_{ar}:
$B_{ar} = 1.1 + [30.8\,(1-\tau)\,\sigma]$

4. Calculate structure bandwidth, B_s:

$B_s = B \times B_{ar}$

5. Calculate array length:

$$\ell = \left(\frac{\left(1 - \frac{1}{B_s}\right)\left(\frac{4\sigma}{1-\tau}\right)}{4} \right)(\lambda_L)$$

6. Calculate number of elements, N:

$N = 1 + \frac{\log B_s}{\log \frac{1}{\tau}}$ and round to next largest number,

7. Calculate new length

$$\ell' = \frac{\text{rounded } N}{N} \times \ell$$

simplified design approach

Design of an LP array using the basic equations above is arduous, since the designer has little feel for the size of the array until he's made one or more iterations. Furthermore, most first tries won't equate to an integral number of elements. Accordingly, a computer program was written to yield initial data on key array characteristics and in terms of antennas with an integral number of elements, **table 1.** This table allows an antenna design based on desired parameter such as number of elements, length, and gain. The following example of antenna design uses the curves of **fig.** 3 and **table 1.**

A three-band **LP** antenna isdesired covering **14-29** MHz. Length is about 15 meters (50 feet). Proceed as follows:

1. $B = \frac{f_H}{f_L} = \frac{29}{14} = 2.1$ (desired frequency ratio)

2. Desired length is 15 meters **(50** feet)

$\lambda_L = \frac{300}{14} = 21.4$ *meters* **(70** *feet)*

$Length = \frac{15}{21.4} - 0.71\lambda$

3. From **table 1** choose ($B = 2$), and $\ell/\lambda = 0.87$, or $\ell = 0.87\lambda$. The desired length is slightly shorter to stay on the correct, or lower, side of the optimum spacing factor, a.

$N = 10$, $\ell/\lambda = 0.87$, $\tau = 0.875$,
$\sigma/\lambda = 0.155$, $gain = 7\,dB$

4. From **fig.** 3 proceed as follows:

a. Modify **a** of **0.155** by the ratio

$$\frac{0.71\ (desired)}{0.87\ (\textbf{table 1})} = 0.82 \times 0.155 = 0.13 = \sigma'$$

b. Draw a vertical line through $\tau = 0.875$ (see example in **fig. 3**).

c. Draw a horizontal line through a $= 0.155$.

d. Optimum gain occurs a the intersection i7.0 dB).

e. Modify by drawing a horizontal line through $\sigma' = 0.13$ to intersect vertical through $\tau = 0.875$.

f. A new gain occurs, ≈ 6.7 dB (down only 0.3 dB from optimum).

element parameters

$f_L =$ MHz, $f_H = 29$ MHz,

$\tau = 0.875$, $\sigma' = 0.13$

for a wire antenna, $\lambda/2$

$$L = \frac{143}{f_{MHz}} \equiv 10\ 2 \text{ meters } (33\ 4 \text{ feet}) \text{ at } f_L$$

$$S_1 = \sigma' \cdot \lambda_L = 0.13 \times 21.4$$

$$= 2.8 \text{ meters } (9.2 \text{ feet})$$

table **1.** Number of elements and array length with optimum taper and spacing.

N	, /λ	τ	σ/λ	B_s	gain over dipole dBd
1 one-band operation ($B = 1$)*					
4	0.34	0.79	0.142	2.02	5.9
5	0.52	0.88	0.155	1.67	7.0
6	0.73	0.92	0.170	1.52	8.0
7	0.91	0.927	0.175	1.39	8.9
8	1.12	0.963	0.178	1.30	9.7
9	1.33	0.972	0.180	1.26	10.2
10	1.48	0.978	0.181	1.22	10.6
2 $B = 1.5$					
6	0.49	0.810	0.142	2.9	6.1
8	0.77	0.875	0.158	2.56	7.0
12	1.35	0.93	0.172	2.21	8.2
15	1.79	0.95	0.175	2.05	9.0
3 $B = 2$					
7	0.52	0.795	0.142	4.0	5.9
10	0.87	0.875	0.155	3.39	7.0
14	1.37	0.917	0.170	3.07	8.0
19	1.97	0.943	0.175	2.8	8.9
4 $B = 3$					
9	0.59	0.8	0.142	5.93	5.9
15	1.18	0.895	0.155	4.80	7.0
22	1.90	0.932	0.170	4.38	8.0
5 $B = 4$					
11	0.65	0.815	0.142	7.64	5.9
17	1.32	0.89	0.160	6.57	7.0
24	2.03	0.925	0.170	5.97	8.3

*B desired band ratio (harmonic number) $= \frac{f_H}{f_L}$ For an array covering 14, 21, 28 MHz, $B = \frac{f_H}{f_L} = \frac{28}{14} = 2$

$S_2 = 2.8 \times \tau$, etc.

$L_1 = 10.2$ meters (33.4feet)

$L_2 = 10.2 \times \tau$, etc.

N	L, in meters (ft)	S, in meters (ft)
1	10.2 (33.41	2.8 (9.2)
2	8.9 (29.3)	2.4 (8.0)
3	7.8 (25.6)	2.1 (7.0)
4	6.8 (22.4)	1.9 (6.1)
5	6.0 (19.6)	1.6 (5.3)
6	5.2 (17.2)	1.4 (4.7)
7	4.6 (15.0)	1.3 (4.1)
8	4.0 (13.1)	1.1 (3.6)
9	3.5 (11.5)	0 9 (3 2)
10	3.0 (10.0)	
	array length	15.5 meters (51 ftl

conclusion

We have presented design details for a log-periodic antenna using simple mathematical formulas. We have also given a simplified approach to LP antenna design using tables and elementary arithmetic. The LP antenna certainly has a place in Amateur Radio. It has advantages of bandwidth which Yagi and quad antennas don't have. A well-designed LP will provide acceptable forward gain and front-to-back ratio over a wide band of frequencies. The LP can be designed with wire elements for lower frequencies — another advantage when aluminum tubing is hard to obtain. An LP antenna can be designed to cover 10-30 MHz with performance and size comparable to that of many current triband beams.

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appendix

intra-array feedline design

The characteristic impedance of the feedline is a function of the natural dipole impedance, Z and the spacing factor, σ. The natural dipole impedance is related to the length-to-diameter ratio of the elements. The optimum intra-array feedline impedance is determined as follows:

a. Select the desired driving impedance, R_0

b. Determine the mean spacing factor $\sigma'' = \frac{\sigma}{\sqrt{\tau}}$

c. Determine the natural dipole impedance:

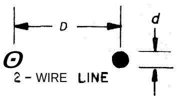
$Z_a = 120\ (\ln \frac{\ell}{d} - 2.25)$. or from fig. **A1**, where $\frac{\ell}{d}$ is length-to- diameter ratio

d. Calculate $\frac{Z_a}{R_0}$

e. Determine $\frac{Z_0}{R_0}$, the intra-array feedline modifier, M.

$$\frac{Z_0}{R_0} = \frac{1}{8\sigma'' = \frac{Z_a}{R_0}} + \sqrt{\frac{1}{(8\sigma''\frac{Z_a}{R_0})^2} + 1;}$$

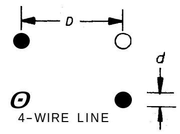
or from fig. **A1**,



$Z_0 = R_0 \times M$

$Z_0 = 276 \log_{10} \frac{2D}{d}$ for 2-wire line or

$138 \log_{10} \frac{\sqrt{2D}}{d}$ for 4-wire line,



where D = center-to-center spacing, and d = wire diameter.

If a non-optimum intra-array feedline impedance is used, the antenna VSWR will vary considerably with frequency, in a somewhat periodic manner.

design example

Array parameters:

$\tau = 0.875 \quad \sigma = 0\ 13$

$R_0 = 200\ ohms$ (with 4:1 balun from 52 ohms)

Mean spacing factor

$$\sigma'' = \frac{\sigma}{\sqrt{\tau}} = \frac{0.13}{\sqrt{0.875}} = 0.139$$

Natural dipole impedance using no. 14 AWG (1.6-mm) wire:

$\frac{\ell}{d} - \frac{12 x 21}{0.064} = 3938;$ since Z_a varies as $\log_n$

the mean $\frac{\ell}{d}$ is usually sufficiently accurate

From fig. **A1**, $Z_a = 72\ ohms$, and

$$\frac{Z_a}{R_0} = \frac{720}{200} = 3.6$$

Feedline impedance from fig. **A1**:

$$\frac{Z_a}{R_0} = 3.6\ \sigma'' = 0.139\ M = 1.3$$

$$or\ \frac{Z_0}{R_0} = \frac{1}{8\sigma''\frac{Z_a}{R_0}} + \sqrt{\frac{1}{(8\sigma''\frac{Z_a}{R_0}} + 1}$$

$$= \frac{1}{8 \times 0.139 \times 3.6} + \sqrt{\frac{1}{(8 \times 0.139 \times 3.6)^2} + 1} = 1.28$$

$$= 0.245 + 1\ 93 = 1.28$$

$$Z_0 = M R_0 = 1.28 \times 200 = 255\ ohms$$

ham radio

compact and clean
L-band local oscillators

A clean, L-band local-oscillator system featuring spurious rejection greater than 30 dB with simple test equipment

The recent popularization of microstripline construction for Amateur uhf equipment[1] has made it relatively simple for countless experimenters to build state-of-the-art multipliers, amplifiers, mixers, filters, and the like directly from magazine construction articles. A major exception, unfortunately, has been in the area of microwave local oscillator chains. Most will agree that the LO is the weak link in just about every microwave transmitter, receiver, or converter. Local oscillators generally require extensive tweaking on costly spectrum analyzers, even then often falling short of the required calibration tolerance, stability, and spectral purity. And, since a spurious or drifting LO can negate all the benefits of the very finest low-noise front end or ultra-linear power amplifier, it is evident that the LO requires a great deal of attention. I've been making an effort in recent months to take some of the mystique out of local oscillator design and construction. In this article I shall present the results of that effort — a high-stability, crystal-controlled LO for 1.1 through 1.6 GHz that is only 6.4 x 10 cm (2.5 x 4 inches) and employs microstripline construction (with absolutely no coils to wind). It can be built for about $70 in parts, and can be completely aligned with only a VOM and a diode detector.

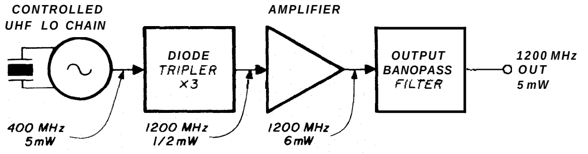


fig. 1. Block diagram of a module LO system for L band. Each block represents a physically separate stage, individually boxed and interconnected by short lengths of coaxial cable.

The basic approach that numerous experimenters (myself included) have used for L-band local-oscillator chains over the past several years is blocked out in fig. 1. I implemented this system modularly, with each of the blocks representing a separate, shielded box, the various modules being interconnected via coaxial cable. The basic 400-MHz oscillator is a version of my recently published uhf LO chain,[2] driving a diode tripler built on a microstrip bandpass filter board.[3] In order to make up the power lost in the passive multiplier, a buffer amplifier is employed.[4] The final microstrip filter keeps all spurious frequency components down better than 40 dB with respect to the desired output.

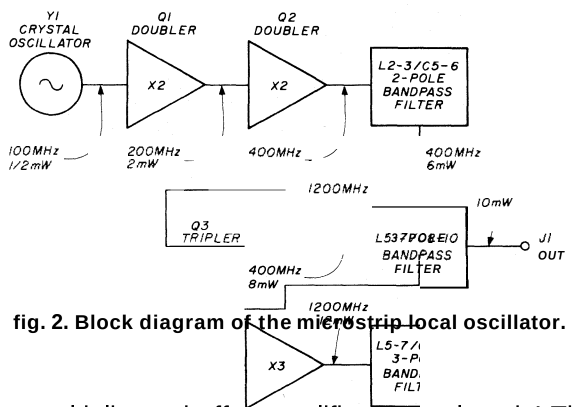


fig. 2. Block diagram of the microstrip local oscillator.

The modular LO chain has some major drawbacks. It is physically large, since each stage must occupy its own separate enclosure to achieve acceptable spurious rejection. The cost of jumper cables, coax connectors, and die-cast aluminum boxes can be

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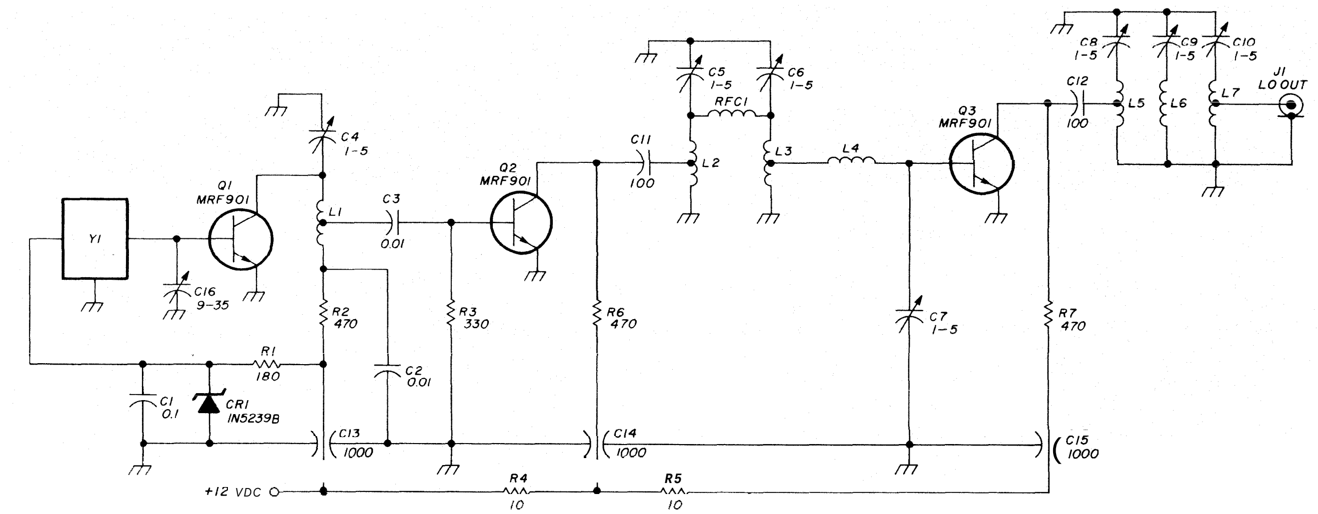

designation	description
C01-03	miniature ceramic disc capacitor, 0.01 μF
C04-10	ceramic piston trimmer capacitor, Triko 202-08M, 1-5 pF
CII-12	chip capacitor, ATC type 100B, 100 pF
C13-15	ceramic feedthrough capacitor, Erie 2404-000-X5UO-102P, 1000 pF
C16	ceramic trimmer capacitor, 9-35 pF
D01	zener, 9.1 V ± 5%, 400 mW. 1N52393
J01	SMA receptacle, E. F. Johnson 142-0298-001
LO1-07	microstrip inductor (PC artwork LO-1200-601)
001-03	uhf silicon bipolar transistor, Motorola MRF-901
R01	carbon composition resistor, 180 ohm 10% 1/4 watt
R02, 06, 07	carbon composition resistor, 470 ohm 10% 1/4 watt
R03	carbon composition resistor, 330 ohm 10% 1/4 watt
R04-05	carbon composition resistor, 10 ohm 10% 1/4 watt
RFC01	0.33 μH molded choke, Nytronics mini-ductor
YO1	crystal oscillator assembly, International Crystal OE-5, 90 to 130 MHz (output frequency ÷ 12)

fig. 3. Schematic diagram of the L-band local oscillator. **C1**, C2, and C3 are miniature ceramic disc capacitors. C4 through **C10** are Triko **202-08M**, 1-5 **pF** ceramic piston trimmers. ATC type **100B** chip capacitors are used for **C11** and C12. All resistors are ¼ watt, 10 per cent tolerance. The rf choke is a Nytronics Mini-ductor. The crystal oscillator assembly should be in the frequency range of 90 to 130 MHz, depending upon the desired output frequency (LO output frequency divided by 12).

prohibitive. Plus, the use of a passive multiplier followed by a buffer amplifier is a crude and inefficient way to generate the required 5 to 10 mW of LO output.

I toyed with the idea of integrating the LO chain onto a single board, but became convinced that first it would be necessary to develop a reliable active multiplier circuit to take the place of the diode tripler and buffer amplifier modules. For a time I considered the push-pull tripler approach which Wade had used in his 1296-MHz LO,[5] but in studying the spectrum analyzer photos from his article, I noticed a few potential difficulties. Wade's output filter had the advantage of requiring no tuning whatever, but it afforded only about 20 dB of spurious rejection. His active multiplier, though far easier to tune than my diode triplers, appeared to offer about the same degree of conversion loss. Since I was seeking multiplier gain of not less than unity, I decided to try a single-ended active tripler followed by a tunable, 3-pole microstrip filter to keep the spurs down.

What finally produced acceptable results was an active parametric multiplier, a circuit technique I had employed in an earlier 1296-MHz converter.[6] The secret is to place a series tank circuit, which resonates at the desired output frequency, in the base lead of a standard class-C common-emitter multiplier. This throws the base into a negative-resistance region at the stage's *output* frequency, enhancing the gain of the desired frequency multiple relative to that of the other multiples. The result is an improvement in the multiplier's spurious rejection without resorting to harmonic-cancelling circuits like push-pull and push-push.

With the active multiplier and an output filter tacked onto the end of one of my 400-MHz LO chains, I found I was getting as much output, with as clean a signal as my modular LO had yielded. Plus, it took up only half the space, and without the various coaxial jumpers between stages. Fig. 2 shows the block diagram of the new LO, and the complete schematic is shown in fig. 3.

circuit description

Refer to fig. 3 and the accompanying parts list. The stages to the left of Q3 are essentially the same

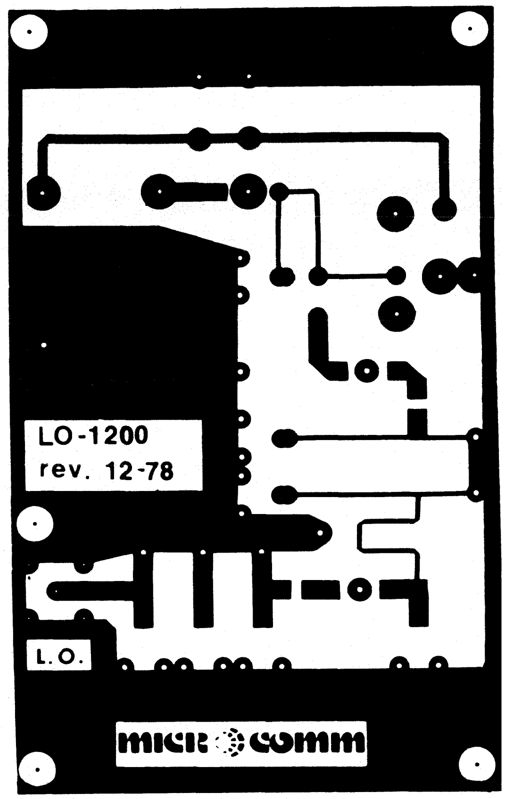

fig. 4. Full-size artwork for the printed circuit board. The reverse side remains unetched, acting as a groundplane.

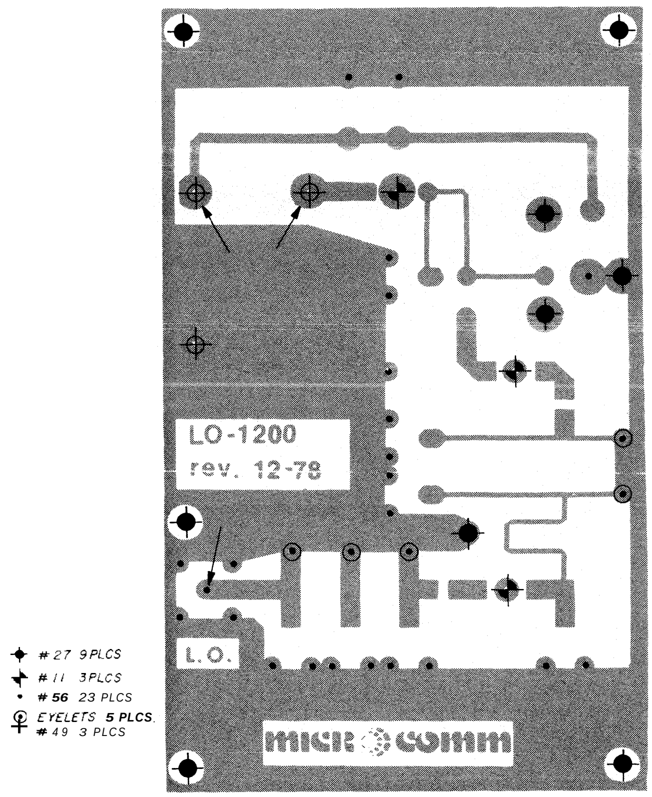

fig. 5. Drilling diagram for the etched side of the circuit board. In addition to the etched side, the three locations marked with arrows are also countersunk on the groundplane side.

as those used in my uhf LO, except for a few component substitutions. Oscillator module Y1 is a high-stability, crystal overtone oscillator producing −3 dBm (1/2 mW) of output near 100 MHz. This assembly requires a regulated +9 volt supply, which is furnished by zener diode CR1. The output port of Y1 exhibits dc continuity to ground, this continuity being essential in providing a bias return for the following stage.

At the input to Q1, common-emitter class-C double stage, C16 is used to resonate Yl's output inductive link, creating a double-tuned, interstage transformer between the oscillator stage and the first multiplier. The output circuit for Q1 consists of microstripline inductor L1 resonated by C4 at 200 MHz.

Collector current for Q1 is limited to 10 mA at R2. This resistor, along with C2 and C13, provide power supply decoupling for the first multiplier stage.

Q2 serves as a second class-C common-emitter doubler. Its input is fed via C3, which is tapped down on L1 for impedance matching. R3 provides base bias return. The collector is shunt-fed by R6, with a collector current of approximately 15 mA. The output circuit for Q2 consists of two microstripline inductors (L2, L3) resonated at 400 MHz by two piston trimmers (C5 and C6). Inductive coupling between filtering poles, provided by RFC1, suppresses higher-order harmonics at the output of the second doubler. The conversion gain of each doubler — exclusive of any filter losses — is on the order of +6 dB.

As mentioned previously, tripler Q3 operates as a parametric multiplier. The input is applied via a low-pass filter consisting of microstripline inductor L4 and piston trimmer C7. The series inductance of C7 is such that it self-resonates at the desired output frequency, maximizing gain at that particular frequency by driving the base impedance of 03 negative. For this reason, use only the specified capacitor at C7. Shunt collector feed for the active tripler is via R7, with dc decoupling provided by R5 and C15. Collector current for Q3 is on the order of 15 mA, and the stage operates at approximately 3-dB gain.

The output of Q3 is capacitively coupled via C12 into a 3-pole output filter consisting of microstripline

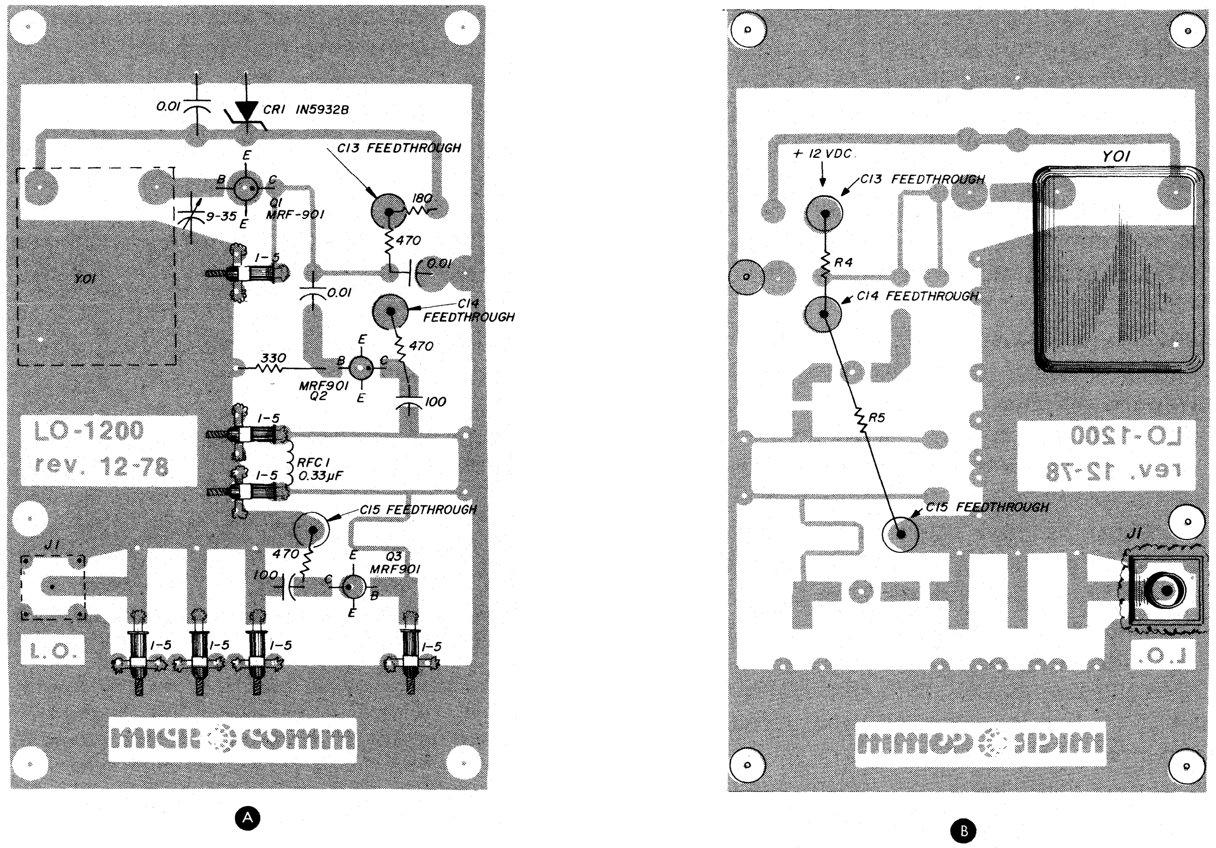

fig. 6. Component placement diagrams for the microstripline and groundplane sides

inductors L5, L6, and L7, resonated at 1.2 GHz by piston trimmers. Coupling between filter poles is a result of the proximity of the piston trimmer stators; hence, spacing between filter poles is critical.

The final + 10 dBm (10-mW) LO output is available on connector J1. This level is suitable for driving most transmit and receive diode balanced mixers.

construction

Assembling the LO is relatively simple, since all circuitry is mounted on a single printed circuit board and the bulk of the critical components are implemented as etched microstriplines. The circuit should be etched on fiberglass-epoxy circuit laminate 1.5 mm (0.063 inch) thick, clad on both sides with 1 ounce per square foot copper. One side of the board is etched in accordance with the artwork supplied in **fig. 4,** the other side remaining fully clad and serving as a groundplane. It is essential that the pattern of the printed circuit artwork be followed exactly (photo etching is recommended), since the dimensions of the microstriplines are critical and the placement of the circuits on the board determines the degree of spectral purity achieved. In fact, the layout of the board was changed several times during the development phases in order to optimize performance and ease of tuning.

After the board is etched, it should be drilled as in **fig. 5.** Be sure to remove a small portion of groundplane metallization from around the holes that will accommodate the center pin J1 and the output and power pins of oscillator Y1. Neglecting this crucial step will result in these pins being shorted to ground, which will obviously have a detrimental effect upon circuit performance! Note that five of the microstripline inductors (L2, 3, 5, 6, and 7) must be grounded through the board. This is best accomplished, as outlined in reference 3, with eyelets 0.5-mm in diameter set with a punch and soldered on both sides of the board.

When mounting components on the printed circuit board, you will find it helpful to refer to the photographs, the schematic diagram in **fig.** 3, the layout drawings shown in **fig. 6,** and perhaps to reference

2. I personally find it easiest to install J1 first, soldering the five pins to their respective pads on the microstripline side of the printed circuit board and then running a smooth bead of solder around the body of the connector, securing it electrically and mechanically to the groundplane side. Next, I install feedthrough capacitors C13, 14, and 15. Here I prepare a small solder preform (1 turn of multi-core solder wrapped around the body of the capacitor just under the flange!, position the capacitor in its mounting hole, and apply heat to the flange from above. The solder preform will flow, filling the space between the flange and the groundplane. This technique prevents excess solder from accumulating on the groundplane side of the printed circuit board.

Installing the resistors and capacitors according to the layout diagram is relatively easy. Do not install power decoupling resistors R4 and R5 at this time; they will be added during the tune-up sequence. When installing the three transistors, note that the raised dot on the plastic package indicates the collector lead. The base lead emerges from the opposite side of the transistor package, with the two emitter leads appearing at right angles to the collector and base. Bend the two emitter leads of each transistor down sharply before installing the transistors in their hoies. That way the emitter leads can protrude through to the groundplane side, where they will be bent over and soldered directly to ground.

When installing oscillator stage Y1, care should be taken to prevent traces on the oscillator's printed circuit board from shorting to the groundplane of the LO main circuit board. I recommend installing a thin insulating washer between the oscillator can and the groundplane.

The only additional advice I might offer in microstripline projects is that it is not unusual for component leads to be laid on and soldered directly to printed circuit board traces or pads, rather than running the component leads through holes in the board. For this reason, it is generally helpful to preform and pretrim the component leads prior to installation.

tune up and test

I cannot overemphasize the importance of employing a systematic, orderly approach in tuning up local-oscillator chains. Tuning for maximum smoke (a favorite Amateur pastime) is a surefire way to make one or more of the multiplier stages oscillate (see **fig. 7**). Further, since the LO chain was designed to provide maximum user flexibility, it may be built for a fairly wide range of frequencies and applications. As a consequence, each of the resonant circuits has a relatively wide tuning range, and it is entirely possible to tune up any one of the multiplier stages on the wrong multiple, if maximum apparent output power is the only criterion.

In fact, whether a microwave spectrum analyzer is available or not, it's a good idea to pre-align the various piston trimmers to the appropriate part of the spectrum. This is easier than it may sound. If the intended output frequency is below about 1.3 GHz, adjust all seven of the piston trimmer capacitors to maximum capacitance (screws all the way *in*). If the

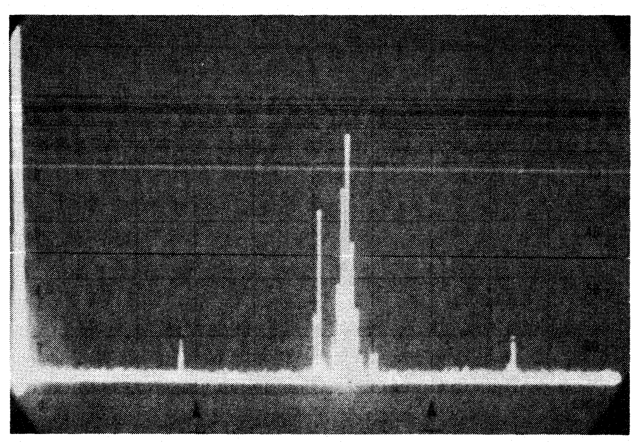

fig. 7. Aligning microwave LO chains for maximum output power often results in a spurious output spectrum. In this case, the LO was tuned for maximum output. The numerous spurs near the desired output are not harmonically related to the crystal frequency. This suggests that they are the result of one of the active stages going into self-oscillation. The horizontal scale is 0 to 2 GHz, with a vertical calibration of 10 dB/cm.

intended output frequency is above about 1.4 GHz, adjust all piston trimmers for minimum capacitance (screws almost all the way out). And, if the oscillator is intended to operate between about 1.3 and 1.4 GHz, pre-adjust all piston trimmers at about mid-range. Now, as you proceed with the alignment procedure, it should not be necessary to adjust any of these capacitors by more than a couple of turns. Keep this in mind, because if you find yourself adjusting the trimmers more than just a little, you're probably enhancing the wrong frequency component.

The approach I recommend for tuning this L-band LO requires no test equipment other than a VOM and a diode detector (or some other means of monitoring relative rf power). It is based upon the principle that, as a class-C multiplier stage is tuned, the signal level applied to the next stage (hence the next stage's collector current) will vary. By knowing what kind of variations to expect and by monitoring stage current closely, it is possible to tune the LO chain to produce an output spectrum such as that shown in **fig. 8.** But it is necessary to monitor the various stage currents separately, to be sure you're observing proper multiplier action and not oscillation.

With the piston trimmers pre-adjusted as outlined, the next task in aligning the L-band LO is to get the oscillator stage oscillating. On the side of the crystal oscillator can is a small access hole, behind which is found the ceramic trimmer capacitor which resonates the oscillator's collector tank circuit. This trimmer is pre-adjusted at the factory to ensure that the oscillator will start each time power is applied; it should not be adjusted at this time. Rather, it should be possible to optimize drive to the first multiplier stage merely by adjusting C16, which, you'll recall, resonates the oscillator stage's output coupling link.

Apply a well-regulated voltage between +12 and +13 volts to feedthrough capacitor C13. This powers both the oscillator stage and the first doubler. Monitor the current drawn by these two stages as C16 is adjusted through its range. Since the sum of the current drawn by the oscillator stage and its zener regulator will remain constant at between 16 and 22 mA (depending upon the power supply potential), any variation in current as C16 is adjusted represents the collector current of Q1. There is a point in C16's tuning range where the current at C13 will rise smoothly to about 10 mA above its minimum value (that is, 26 to 32 mA, total), and this is the point to adjust C16. Now, momentarily remove V_{cc} from C13. If the current returns to the previous value, all is well. If on the other hand Q1 appears not to be drawing any current (that is, total current at C13 decreases to between 16 and 22 mA), then the oscillator stage is not starting smoothly and it will be necessary to readjust Y1's trimmer. Do so ***carefully;*** it should be necessary to rotate the trimmer only about ten degrees one way or the other, and the current should rise again, indicating oscillation. Now, repeak C16 for the proper rise in current, and again remove and reapply power. The adjustments of C16 and the oscillator's trimmer capacitor are somewhat interactive, so repeat the above procedure until the oscillator starts reliably each time power is applied.

Once the adjustment of C16 and the oscillator trimmer is completed, do not under any circumstances change their settings while aligning the balance of the local-oscillator chain. I usually paint a dot of nail polish on C16 to lock it down and tape over the access hole in the side of Y1, lest I be tempted to backtrack and screw things up completely! Remember, the objective is to perform a reasonably clean LO alignment without the use of any costly test equipment, so don't jump sequence.

The easiest way to resonate the collector tank of the first multiplier stage is to monitor the current drawn by second doubler, Q2. Apply operating potential to feedthrough capacitors C13 and C14, and this time monitor the current drawn at C14. This current should peak smoothly at 10 to 12 mA while adjusting C4 no more than two or three turns from its preset position.

Adjusting the interstage circuitry between the second doubler and the parametric tripler is perhaps the trickiest part of aligning this LO because there are three separate trimmer capacitors and the adjust-

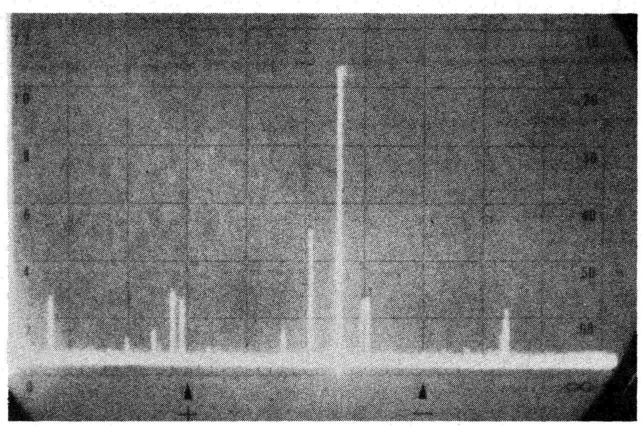

fig. 8. This photograph shows the output spectrum of the L-band LO system after being aligned as described in the article. Spurious rejection of close to 30 dB was achieved with the simple test equipment called for in the article.

ments are all interactive. Note that at this point the only clue you have to proper alignment is the collector current drawn by Q3, so watch it closely. Apply operating potential to all three feedthrough capacitors (C13, 14, and 15), this time monitoring current at C15. First, adjust C7, slightly, just to the point that a few milliamperes of current flow through C15. Now, carefully adjust both C5 and C6 to maximize this current. As before, a peak should occur before the trimmers have been adjusted very far frorn their preset positions. Once a peak has been found with C5 and C6 both set at approximately the same point, re-adjust C7 slightly. At this point, C4 (the collector tank of the first multiplier) may be adjusted ever so slightly to maximize current at C15. Now, back to C5 and C6 again, then C7 if necessary, and so on until the current at C15 settles in at about 15 mA. Note that when you're done, both C5 and C6 should have their tuning screws protruding by about the same amount.

All that remains is to align the output bandpass filter. An rf-diode detector can be connected to the output connector, the dc from the diode assembly being fed to a sensitive microammeter as an indication of relative rf output. Any other method of measuring relative rf power (bolometer bridge, calorimeter, or similar) may also be employed. The object is

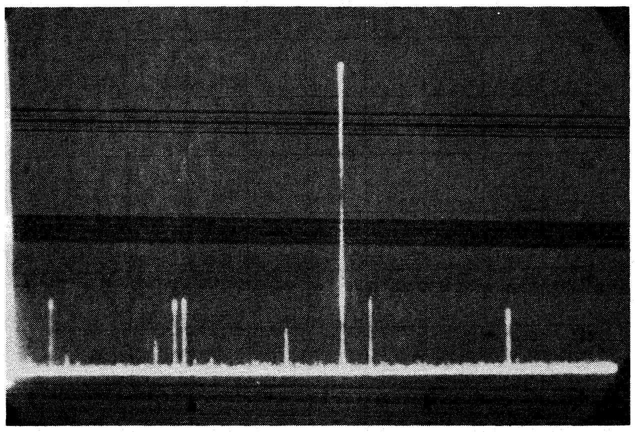

fig. 9. Example of the output spectrum after aligning the local-oscillator chain with a microwave spectrum analyzer. In this case, all spurs are down greater than 40 dB.

to adjust C8, C9, and C10 simultaneously for maximum indicated output (the three trimmer capacitors should all track relatively closely). Keep monitoring the current at C15. If it jumps abruptly, the tripler stage is self-oscillating. It can be tamed down by slightly adjusting C7 until current at C15 returns to its proper value.

With C8, C9, and C10 all peaked at approximately the same setting (not too far, you hope, from the preset position) and output power maximized, one last adjustment to C7 is in order. Adjust this capacitor for the maximum output level obtainable *without causing an increase in the current at C15.* At this point, you may be tempted to go back and repeak all the other trimmer capacitors in the circuit; resist that temptation. You can only disrupt what would in all likelihood be a very clean output spectrum, such as that shown in **fig. 8.**

Of course, if you are fortunate enough to have access to a microwave spectrum analyzer, adjusting the trimmer capacitors ever so slightly can indeed clean up the output spectrum still further (see **fig. 9**). But this should be attempted only after the current-sensitive tuning method has been completed and decoupling resistors **R4** and R5 installed.

One final thought. Those super-purists fortunate enough to possess a complete laboratory of microwave test equipment will doubtless notice that any tuning adjustment can potentially affect output power, spectrum, *and* frequency. Thus, you may wish to simultaneously monitor all three parameters during alignment. **Fig. 10** is the lab setup I use in aligning these LO chains. The key to the success of this method is the resistive three-way power divider, which applies equal samples of the LO's output signal to the counter, spectrum analyzers, and power meter. The divider is built simply from four 27-ohm, 1/8-watt, carbon-composition resistors, arranged symmetrically in a small shielded box which supports four coaxial connectors.

parts procurement

Readers of my construction articles frequently write asking if I can supply a complete kit of parts for a given project. Unfortunately, I have neither the time nor the inclination to get into that business. But I am not heedless of the plight of the home constructor, and as much as possible like to help identify (or sometimes create) sources for some of the less-common components.

For example, I have in the past endeavored to make etched, drilled, and plated circuit boards available at cost, for the benefit of those experimenters who prefer not to fabricate their own. This project is no exception. I will supply the boards for $10 each, postpaid anywhere in the U.S. or Canada ($11 elsewhere).

In a previous article, I mentioned a source of supply for the Triko trimmer capacitors I employ in this and other modules. Unfortunately, I later discovered that the importer had a $50 minimum order requirement. Thus, I have recently obtained a quantity shipment of the piston trimmer capacitors used in the LO chain, and will gladly supply them to Amateurs in sets of seven pieces (the quantity needed for each LO chain) at $10.50 per set postpaid in the U.S. or Canada, $11.50 elsewhere.

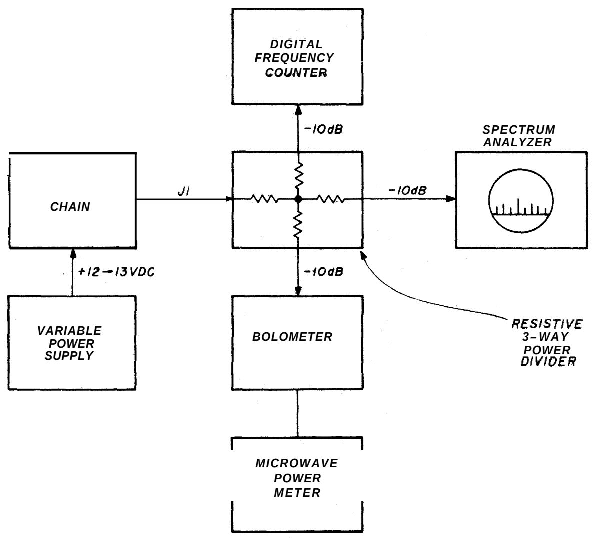

fig. 10. Test configuration for monitoring power, frequency, and output spectrum simultaneously during alignment of the LO chain.

The crystal oscillator assembly designated Y1 is available only from International Crystal Manufacturing Company. I recently spoke to Mr. Royden Free-land, himself a ham and microwave experimenter, and he assures me that this module will be sold to individual experimenters in single quantities. Be sure to allow six to eight weeks for delivery, as the units employ custom-ground crystals.

The MRF-901 transistors used for 01, Q2, and Q3 are available from Motorola Semiconductor Company. When I first used this particular transistor in a 1296-MHz preamp a few years ago, the price was $9 each. Quantity production and improved yield brought the price down to $4.30 in 1977, and to an unbelievable $1.45 today. At that price, I'd recommend against trying to substitute any other transistor.

The rest of the components used in the LO are, for the most part, garden-variety. Though not necessarily available at your corner Radio Shack, they should nonetheless be obtainable by most experimenters after a bit of scrounging.

Of course, there are always those who need a microwave LO but prefer not to do the scrounging, building, tuning, and testing themselves. To such individuals I am able to offer a completely built, tuned, and tested Model LO-1200 Oscillator Module, packaged in an enameled die-cast aluminum box, operating at your specified frequency between 1150 and 1555 MHz, for $160 postpaid. Foreign orders please add $5 additional postage. This offer is extended to licensed Radio Amateurs only (state your call when ordering), and is restricted to units used for personal, noncommercial applications only. All orders for printed circuit boards, capacitors, or complete LOs must be prepaid in U.S. dollars, and all inquiries must be accompanied by a stamped, self-addressed envelope.

Frankly, I hope nobody takes me up on the above offer. I'd rather design gear than build it for others, and, besides, you're missing out on quite a feeling of accomplishment if you buy your gear ready-made. After all, it is the homeconstructor to whom this article is dedicated.

Happy building!

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ham radio

More Details? CHECK – OFF Page 126

vhf preamplifiers

A survey of today's devices and techniques for building a truly low-noise preamp for vhf reception

The vhf converters and preamplifiers of ten and fifteen years ago are inadequate by today's standards. This article presents the results of my experience with many low-noise preamps using modern devices and techniques, all of which have been built and tested as a part of a program to update my vhf station. The problems of wide dynamic range and intermodulation distortion are not addressed in this article.

analysis

Consider a typical vhf receiver (fig. 1). The receiver consists of X stages, where X is some number. Each stage has a noise figure, NF_X, and therefore a stage noise factor, F_X. Each stage has a gain, G_X, which may be an actual gain or it may be a loss as in the feedline, attenuator/filters, or even in the mixer. Noise figure, NF, is expressed in dB; noise factor, F, is nondimensional. NF and F are related by

$$NF = 10 \log F$$
$$F = \log^{-1}(NF/10) \qquad (1)$$

The most common error in analyzing system noise performance is to lump noise figures and noise factors — a simple but bold dB symbol after each noise figure helps to differentiate the noise figure numbers from noise factors.

It's well known[1-4] that the system noise factor, $F_{S,0}$, as presented by the receiver at the antenna, is influenced to some extent by the noise factors and gains of the following stages:

$$F_{S,0} = F_0 + \frac{(F_1-1)}{G_0} + \frac{(F_A-1)}{(G_0G_1)} + \qquad (2)$$
$$\frac{(F_2-1)}{(G_0G_1G_A)} + \frac{(F_B-1)}{(G_0G_1G_AG_2)} + \frac{(F_C-1)}{(G_0G_1G_AG_2G_B)} +$$
$$\frac{(F_3-1)}{(G_0G_1G_AG_2G_BG_C)} + \frac{(F_4-1)}{(G_0G_1G_AG_2G_BG_CG_3)}$$

The "gain" of passive elements, such as the feedline and interstage networks (and often the mixer), is actually a loss. That is, a negative gain occurs, which is expressed in dB. When converted into a numerical ratio for use in eq. 2 this "gain" will be less than unity.

Thus a feedline with a 2.2 dB loss has a ratio of 1.66, which means that about 39.7 per cent of the input power is lost in the feedline. The gain, in this case, is -2.2 **dB**, or $1/1.66 = 0.6$. Noise figure in this portion of the system is essentially equal to the loss, $NF_0 = G_0$, and noise factor $F_0 = \log^{-1}(-G_0/10)$. Therefore $F_0 = 1.66$ and $G_0 = 0.6$.

Similarly, a double-balanced mixer has a negative gain. A double-balanced mixer is a passive circuit to which rf signals and LO inputs are supplied. It uses nonlinear elements (diodes) to derive the i-f signal. However, the double-balanced mixer has a noise figure about 1-3 dB greater than the conversion loss. Thus a typical balanced mixer with a conversion loss of, say, 6 dB (numerical $loss = \log^{-1}(6/10) = 3.98$) has a numerical conversion gain $G_C = 1/3.98 = 0.25$ but may have a noise figure of 8 dB, or a noise factor of 6.3. The noise figure is greater than mixer conversion loss.

system noise factor: some simplifications

Several simplifications to eq. 2 can be made. Each interstage network can be considered as a portion of the rf amplifier output circuit preceding the interstage circuit. The noise factor, F_4, of the subsystem following the first stage can be neglected if the first i-f has a gain, G_3, of at least 100 120 dB).

The effects of the feedline (G_0 and F_0) can be removed if the first rf amplifier is placed directly at the antenna, and if the first rf amplifier has a gain, G_1, at least 10 dB greater than the losses in the fol-

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lowing interstage network and feedline (G_A and G'_0, fig. **2A**).

Thus the receiving system (fig. **2B**), for noise considerations, can be reduced to:

a) A low-noise preamp at the antenna, which may contain input and output filters as well as the feedline to the second rf amplifier, and which may have an overall stage noise factor, F'_1 and stage gain, G'_1.

b) The second rf amplifier with noise factor and gain of F_2 and G_2.

c) The frequency conversion stage Far image and noise rejection it includes a bandpass filter as part of an interstage network. It has a noise factor, F'_C, and a stage gain (or loss) of G'_C.

d) I-f amplifier, which has an input noise factor, F'_3.

The system noise factor, F'_{S0}, is now

$$F'_{S0} = F'_1 + \frac{(F_2 - 1)}{G'_1} + \frac{(F'_C - 1)}{G'_1 G_2} + \frac{(F'_3 - 1)}{G'_1 G_2 G'_C} \qquad (3)$$

noise factor and noise figure

The system noise factor establishes the overall receiving-system signal sensitivity; a low noise factor (and, therefore, a low noise figure) is desirable above 100 MHz. Occasions arise when low noise figures are also desirable below 100 MHz, as in an extremely quiet location for 50-MHz operation. Accordingly, an evaluation of low-noise amplifier stages for 30, 50, 144,220, and 432 MHz is useful.

Many receiving systems may use more than one preamplifier. In the systems of **figs. 1** and **2,** first and second preamplifiers are used between antenna and mixer. Regardless of the number of preamplifier stages between antenna and mixer, the general rule is to first adjust the first stage (at the antenna) for minimum noise figure, and then adjust the remaining stages, between the first preamplifier stage and the mixer stage, for maximum power gain. However, as shown by the system noise-factor equations, even the remaining rf amplifier stages should have reasonably tow noise figures.

As mentioned, the rule of thumb is that each stage should have at least 10 dB more gain than the noise figure (in dB) of the subsequent portion of the system. However, the noise figure of the subsequent portion is predominantly established by the noise figure in its first stage. Therefore, this stage should have a reasonable noise figure even if adjusted for maximum gain. Thus, if the noise factor presented by the conversion stage is $F'_{S2} = 10$ (or $NF'_{S2} = 10$ dB), the second amplifier stage should have a target gain, G_2, of $(10\ dB + NF'_{S2}) = 20\ dB$ and a relatively low noise factor, F_2.

If, after the second rf amplifier stage is built and tested, the noise factor, F'_{S1}, of the second rf amplifier/frequency-conversion/i-f system is measured, for example, as $F'_{S1} = 2$ $(NF'_{S1} = 3\ dB)$ the requirements for the first preamplifier stage are that the gain, G'_1, be equal to $10\ dB + NF'_{S1}\ (= 13\ dB)$, with the smallest noise factor, F'_1, (and noise figure) achievable at that gain for minimizing the total system noise factor, F'_{S0}. The actual value of the system noise factor/noise figure will depend on the device selected far the preamplifier stage.

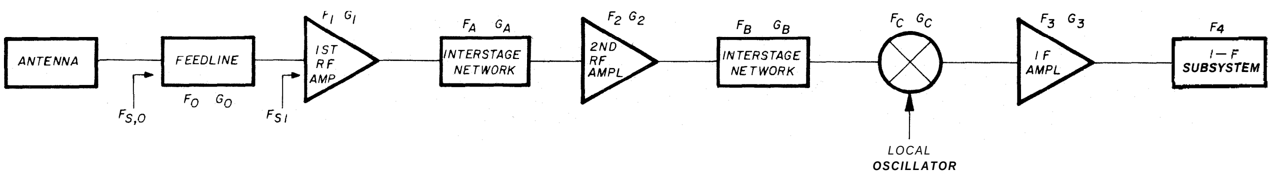


fig. 1. Typical vhf receiver used in the analysis.

background information on devices

What devices are available for low-noise vhf preamplifier use, and what performance can be achieved? To give a proper perspective, consider first a short history of low-noise vhf preamplification.

About twenty years ago, in 1958, the predominantly used vhf bands were 50 and 144 MHz. Little was done, except by a few adventurous Amateurs, with the higher frequency bands. The "low" noise figures then achievable were about 4 dB on 50 MHz, typically with cascade-connected triode vacuum tubes such as the twin triodes of the 6BZ7 variety. About 5 dB was obtained with single triode vacuum tubes such as the 6AM4 on 144 MHz.

Serious experimenters tried to obtain type 58421 417A triodes, which could be used to achieve 3-dB noise figures up to about 250 MHz and about 5-dB noise figures at 432 MHz. A better low-noise tube for 432 MHz was the gold-plated 416B, which was occasionally available as pullouts from microwave relay transmitters. The problems of using pullouts, particularly at the relatively high current levels required for low noise figure operation and the high cost and low availability of special-purpose tubes, deterred widespread Amateur use. Many 432-MHz and most 1296-MHz mixers were of the diode-mixer type, with no preamplification used in many 432-MHz and almost all 1296-MHz Amateur receivers.

Between about 1958 and 1963, vacuum-tube technology progressed to introduce the Nuvistor,® a miniature ceramic-metal tube best typified by the 6CW4, which had the high transconductance and low interelectrode capacitance necessary for high-gain, low-noise preamplification. Noise figures of 3 dB at 50, 144, and 220 MHz were possible, by 1963, in a host of commercially available vhf converters, transverters, and at least one complete receiver (the Clegg Interceptor, for 6 and 2 meters). Versions of the Nuvistor,® would even yield 3-5 dB noise figures at 432 MHz, and a few commercial converters and preamplifiers were offered.

With the advent of relatively inexpensive, relatively low-noise preamps, interest in 220 MHz increased, although not as rapidly as in 432-MHz operation. Interest in 432 MHz was greater because the output of a 2-meter transmitter could be tripled for operation on 432 MHz, whereas a completely new transmitter had to be built for 220 MHz.

Parametric amplifiers were noted for very low noise operation but weren't very popular because of the need for a pump oscillator above 1 GHz and for special components.

Solid-state devices. Early silicon transistors did not, in 1963, appear to have any advantage over the new vacuum tubes. Consequently they saw limited use. However, in another five years, by 1968, the pressure of the commercial home-entertainment market forced development of relatively inexpensive and relatively low-noise silicon transistors. Several transistor types became available at sufficiently low cost for Amateur use. The 2N3819 family was often used for i-f amplifiers. The 2N3823 was specified for rf amplification to more than 150 MHz. Devices such as the Motorola MPF-102 family, the 2N2857, and the 2N4416 appeared. Germanium transistors, notably the 2N1742 and the Philco T2028, were available with even better noise figures than most silicon devices. Transistorized preamplifiers and converters were built with noise figures of 2 dB on 10 and 6 meters, 3 dB on 2 meters and 220 MHz, and 4-5 dB on 432 MHz. The use of vacuum tubes in Amateur vhf receiving gear diminished and all but ceased.

By 1973 a host of solid-state devices were available for vhf receiver use. Silicon devices, with better performance vs temperature characteristics than germanium devices, had, in this period, also achieved better noise figures than their germanium counterparts.

Best remembered of that group were the devices available from KMC Transistors and from Fairchild Microwave Transistors. The KMC devices (now available under different device numbers from Microwave Associates) were capable of less than 2 dB noise figure at 144 and 220 MHz. With selected devices a 2-dB noise figure could be achieved at 432 MHz. If the preamplifier configuration and matching network were chosen and built with care, some of the FMT devices, such as the FMT 4575, reference 5, were capable of even better performance, but cost was generally prohibitive. The availability of the best FMT devices apparently ceased, as later batches of these devices were rumored to have poor noise figures, compared with the earlier batches, and a "lost recipe" was widely rumored as the cause. Solid-state low-noise vhf preamplifiers, were, however, firmly entrenched by this time.

Devices available today. As this is written, some twenty years after the 6BZ7 and 417A era, noise figures of 1 dB are easily obtained on the vhf Amateur bands. I built and tested forty-four preamplifiers to ascertain the performance that might be expected with a wide range of devices. Not all devices available to Amateurs were tested. Financial considerations limited this program to building units only with devices donated by various sources (including many engineers who wished to have a device evaluated and who arranged for these transistors to be made available to me). Results are shown in **table 1** by band and in order of increasing noise figure. **Table 1** also lists

a) The minimum noise figure, M:

$$M = 10 \log \left(1 + \frac{F-1}{1-\frac{1}{G}}\right) \tag{4}$$

which is indicative of the minimum noise figure obtained with a cascade connection of several such stages.

b) The return loss, R, through the preamplifier from output to input, which is a measure of preamplifier stability and should be at least 8 dB greater than the forward gain, G, (reference 5).

c) A performance ratio factor, P, indicative of the noise figure vs preamplifier cost, which is changing between units of the same noise figure because of device price.

All preamplifiers in this article require a low-noise active device and a pair of matching networks. The input matching network matches the device input impedance to the antenna impedance (50 ohms). Device input impedance may be different, for mini-

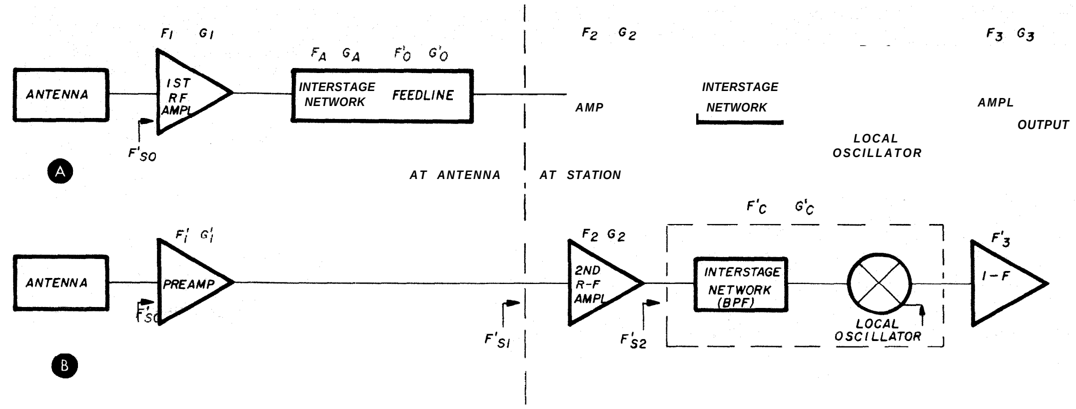


fig. 2. System block diagrams to illustrate how simplifications can be made to the system noise factor, $F_{S,O}$ (eq. 2). The system is divided into two parts: antenna and station. Sketches A and B are used in the text to explain the derivation of the simplified system noise factor, $F'_{S,O}$ (eq. 3).

mum noise-figure operation, from the data sheet values typically given for maximum gain operation. The output matching network matches the device output impedance to 50 ohms.

device choice

Choice of a suitable device should, in general, be governed by the manufacturer's data sheet. It not only lists expected noise figure and associated gain (at that noise figure) at some optimum bias level (generally stated in terms of V_{ce} and I_c for a bipolar transistor, or the equivalent V_{ds} and I_d for an FET device), but also gives

a) Rf parameters for determining stability and matching-network design.

b) A frequency range in which best operation of a device may be expected.

These data are advisory – they are typical values, which may be determined by testing a group of devices. They may not always be obtained for all devices of that type but are good starting points.

stability

If the selected device will operate in the desired frequency range, its stability is always the most important consideration. Nothing is achieved if a low-noise preamplifier is unstable with the input or output impedances found in the system, Instability may result in oscillation and may produce birdies, blocking, or other undesirable results.

Tests for stability are well covered elsewhere. However, two concepts should be kept in mind: First, the device should be stable over a wide frequency range, including the frequency of interest, and second, even potentially unstable devices can be used if care is taken in matching-network design. In this respect, the use of a collector resistance of relatively low value is often helpful and may be found in many of the low-noise preamplifiers discussed in later portions of this article.

bias voltage

The bias voltage and current figures on a device data sheet should always be considered as *nominal* values and should always be varied to achieve best noise figure, assuming that a noise figure test setup is available.

I've tuned several identical preamplifiers containing devices taken from the same shipment. I found

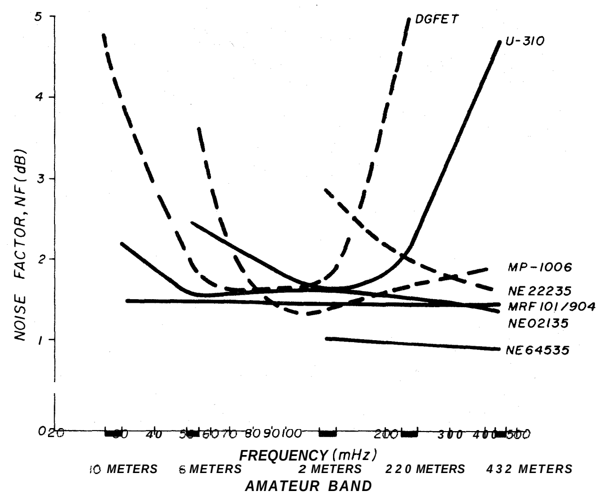
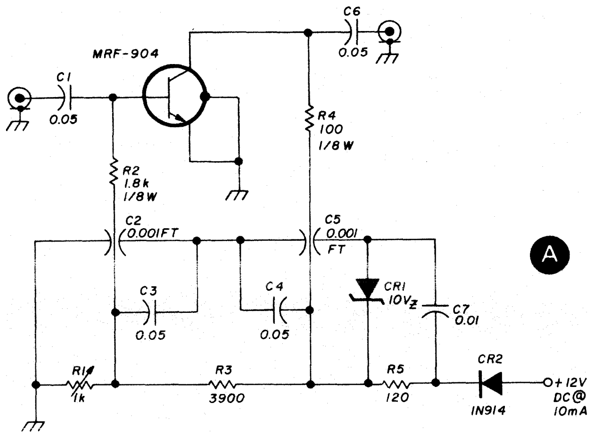
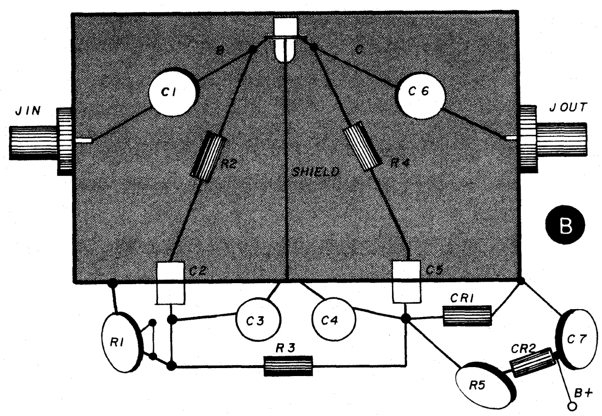


fig. 3. Noise figure as a function of frequency with several devices as a parameter. Note the apparent "useful frequency band" effect, particularly with the dgfet and the U310. Note also the lower frequency limitations shown by devices such as NE22235 and MP-1006.





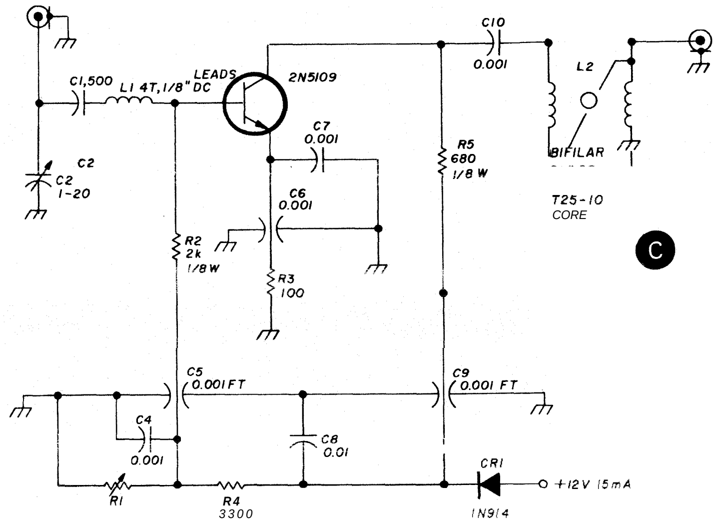


fig. 4. Preamplifiers for 30 and 50 MHz. (A) shows a circuit using the MRF-904, which has very wide bandwidth. (B) is a pictorial layout of the MRF-904 unit. In (C) a 2N5109 is used. which has a 1.7-dB noise factor. Many other devices are usable at this frequency (see text).

that, while almost identical noise figures and gains were achieved, each device had to be biased differently. For example I built and tested four units using NE64535s at 432 MHz; each unit was initially adjusted for data-sheet bias values of 8 volts V_{ce} and 7 mA I_c. All units were then adjusted for minimum noise figure in a test setup using a precision automatic noise figure indicator and noise source:

unit	ultimate NF (dB) (Y method, +0.3dB; −0.1dB)	V_{ce}	I_c
1	0.86	8.6	7.5
2	0.84	7.9	6.4
3	0.91	8.1	10.0
4	0.90	4.9	5.0

S-parameters

Most device data sheets give rf characteristics in terms of S-parameters for gain amplification. While S-parameter design may be new to many Amateurs, several very good articles have appeared.[9-13] The design procedures are relatively easy for devices that are unconditionally stable; *i.e.*, having a stability factor, K, greater than unity at the frequency of interest. However, if the device is only conditionally stable (K less than 1) and the input and output S-parameters (S_{11} and S_{22}) are less than 1, a low-inductance resistor (between 50-200 ohms) may often be used as a collector load to provide a stable preamplifier configuration.[14]

Many data sheets don't give noise-related information. Most particularly, the optimum source reflection coefficient, Γ_{os}, is often missing. This is the parameter that should govern the entire design of a low-noise preamplifier.[15-16] Given Γ_{os}, the input matching network is designed to transform the noise matched input impedance, derived from Γ_{os}, to the antenna impedance (generally 50 ohms). The output matching circuit may be designed once the device output impedance, for the device input connected to the optimum noise-match impedance, is known. Thus, both S and optimum-noise parameters should be known.

device frequency range

The range of device operation should be carefully determined from the data sheet. Many devices appear to be designed for a specific application and, in meeting those application requirements, have parameters that limit the useful frequency range. As an example, the dual-gate FET (DGFET) of **table 1** is an experimental device designed for a noise figure of less than 2 dB over the frequency range of the commercial fm broadcast band. A low noise figure is obtained near this band at 50 and 144 MHz, while the noise figure at 30 and 220 MHz is very high. These latter frequencies were of no interest to the device designer, so the device characteristics are relatively uncontrolled at these frequencies. Note also the noise figure (fig. 3) for the NE22235 and NE64535 devices, which are intended for use at 1-4 GHz and 0.5-4 GHz respectively. The noise figure curves show that these two devices were designed for optimum noise figure at frequencies above about 500 MHz. On the other hand, from the noise figure curves of fig. 3, devices such as the MRF901/904 were apparently designed for broadband application use from dc through the vhf range.

Device data sheets frequently give a circuit in which the device is tested, and which often makes a good starting point for design of an Amateur circuit. As the frequency increases, care must be taken in circuit layout. Short lead lengths and high-quality components (of types intended for vhf, uhf, or microwave use) are required.

band-by-band discussion

30-MHz. This band is primarily of interest for either OSCAR downlink or as an intermediate frequency for microwave equipment, such as the Microwave Associates Gunnplexers.® In i-f preamplifier application, the preamplifier design following the microwave mixer (generally a diode having 6-10 dB of conversion loss and noise figure) will be an important factor in establishing the overall receiver noise figure.

The MA42001-509, as used in W1HR's two-stage i-f amplifier,[17] certainly has the lowest noise figure of units thus far built and tested. However the cost (about $16.50 at the time of writing) of the two devices required for each such preamplifier must be balanced against the small loss in system sensitivity if you use a single-stage i-f preamplifier with a less-expensive 2N5109 or MRF901 device. Use of older devices, such as the 2N4416 and MPF102, is not advised; preamplifiers using these devices have been found to be only conditionally stable, even with heavily loaded output circuits.

The MRF-904 is of particular interest. Apparently it was designed from minimum noise figure when inserted into a 50-ohm system. As shown in **fig. 4A,** matching networks aren't required on either input or output of this transistor in a 50-ohm system. Coupling capacitors C1 and C6 provide dc isolation between input and output feedlines and the device bias circuit. This preamplifier has a very wide bandwidth, as no input or output filtering is used. Output filtering is usually provided by the receiver or converter. The need for, and degree of, input filtering is established by a particular use and location.

For the OSCAR down-link application almost any of the devices yielding less than a 2-dB noise figure should be acceptable, especially in front of a 10-meter receiver having a noise figure of 6-10 dB. I believe the antenna pattern and pointing accuracy are of greater concern than noise figure in this application.

***50* MHz.** The background noise at 50 MHz is high enough so that a system noise figure of less than about 2 dB is unnecessary. Most of the devices tested achieved this noise figure for a single-stage preamplifier. The noise contribution of succeeding stages (existing converter and the like) will dictate the preamplifier gain and also indicated that a first-*stage* noise figure somewhat lower than 2 dB is required if the 2-dB *system* noise figure is to be achieved.

For example, I presently use a 50-MHz converter with a single 6CW4 rf amplifier preceding a mixer (a remnant of earlier 1960s equipment not yet replaced).

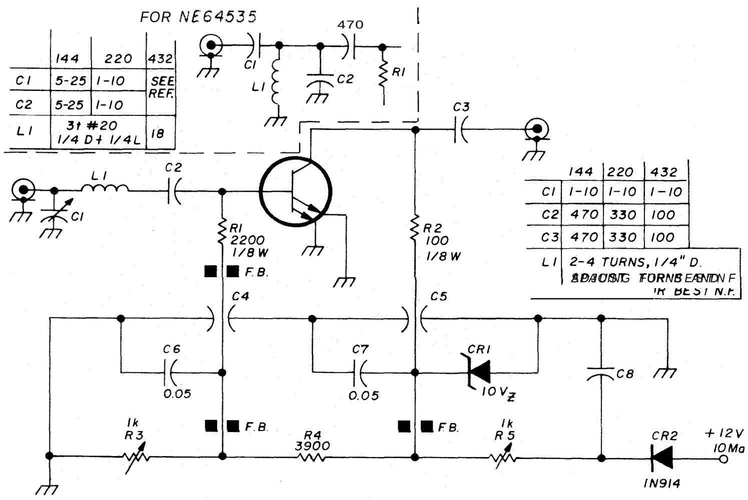


fig. 5. Preamp for 144 and 220 MHz using the MRF-901, NE02135 or NE22235. Design is similar to that by WB5LUA (reference 18).

The converters have input noise figures between 3.5 and 4.5 dB. If the highest expected converter noise figure (4.5 dB) is taken for worst-case analysis, and if a single-stage rf preamplifier with at least 14.5 dB (equal to 10 dB plus the maximum noise figure of the following stage, 4.5 dB) is to be used, then, by the noise-figure equation, $F_S = F_1 + (F_2 - 1)/G_1$, where the value of $F_2 = \log^{-1}(4.5/10)$, with $G_1 = \log^{-1}(14.5/10)$ and $F_S = \log^{-1}(2.0/10)$. Thus $F_S = 1.585$; $F_2 = 2.818$, and $G_1 = 28.184$. These values are substituted into the equation and give

$$1.585 = F_1 + (2.818 - 1)/28.184 = F_1 + 0.065$$

Therefore, $F_1 = 1.585 - 0.065 = 1.52$, and the preamplifier noise figure, NF_1, should be no greater than $10 \log(1.520) = 1.82\,dB$.

Devices tested at 50 MHz that are usable for this particular applicatiion include MA42001 (0.95 dB NF, K = 18.92); NE02135 (1.55 dB NF, K = 19.08); MRF-901 (1.67 dB NF, K = 16.87); KD6003 (1.7 dB NF, K = 22.88); and 2N5109 (1.7 dB NF, K = 16.62). An example of a preamp using the 2N5109 is shown in fig. 4B.

Other factors, dependent on your needs, may now be considered, such as circuit simplicity, resistance to overload, and preamplifier bandpass characteristics.

Note that many of the low-noise preamplifier circuits are broadband; a high-Q input circuit will add undesired noise before the desired signals can be amplified by the device. Of course, if use is intended in an environment near other rf sources, then input filtering may be mandatory to prevent receiver overload. The amount of input bandpass filtering is determined on a case-by-case basis, although use of low-intermodulation-producing, and therefore overload-tolerant devices, such as the 2N5109 provides some leeway in achieving the desired result. I prefer to place a separate low-insertion-loss filter such as a helical resonator or interdigital filter, having steep

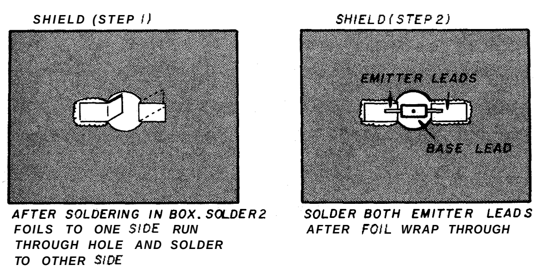


fig. 6. Details of preamp shield construction.

skirts and high out-of-band loss, before the first preamplifier. Such filters have a varying impedance outside the passband of interest, so this reactance may cause a potentially unstable preamplifier to oscillate.

If a 50-MHz i-f system is used with higher frequency (microwave) equipment, the i-f amplifier noise figure may, depending on its total noise figure contribution, require a very low noise configuration using a device such as the MA42001. However, use of such a low-noise-figure preamplifier would usually be expected only when no additional rf amplification is inserted between the antenna and the higher-frequency mixer.

Devices are available for low-noise preamplifiers for the bands up to at least 3300 MHz. Image-rejection considerations at even higher bands, above 5650 MHz, dictate the use of i-fs higher than 50 MHz, so little is gained by using low-noise 50-MHz i-f amplifiers following a microwave mixer. Exceptions always occur — one is a 50-MHz i-f with the mixer portion of a 10-GHz Gunnplexer® transceiver in which a low-noise-figure device, such as the MA42001, would be advisable. Additional information for 50-MHz preamplifiers is shown in figs. 4 through 7.

144 and 220 MHz. The 2-meter band has a sufficiently low background noise to allow very low-noise-figure preamplifiers to be used to advantage. The frequencies of the 2- and 1¼ meter bands are relatively close (less than one octave apart). Similar device and circuit design and selection criteria are valid. Furthermore, the 2-meter band is a common i-f for 1296 and 2304 MHz converters, wherein low-noise-figure i-f amplifiers may be used to advantage. The same 2-meter preamplifier may be used between a 2-meter receiver and either a 2-meter antenna or the i-f output of a 1296- or 2304-MHz mixer.

For truly low noise figures, the NE64535, priced at about $17, provides a 1-dB noise figure in a relatively simple circuit. This device is potentially unstable at all Amateur bands of interest; therefore, the collector circuit (fig. 5) is heavily loaded with a 100-ohm, 1/8-watt resistor.

The design is similar to that of WB5LUA (reference 18), but with added shielding between the device base and collector terminals. This shielding must be properly designed, because with improper shielding, the preamplifier exhibits a strong tendency towards oscillation, as noted by WB5LUA in his article. While the 432-MHz preamplifier of that article did not require shielding, at the lower frequencies of 144 and 220 MHz, the greater forward gain of these devices makes full shielding mandatory.

The preamplifiers are built in a box (fig. 7) formed of double-sided PC board with a double-sided PC-board shield (fig. 6), into which a small hole is drilled. The hole diameter is slightly larger than the device package dimension between the opposed emitter leads. The edges of both copper-clad sides of the shield are soldered to the bottom and two side walls of the basic enclosure. Most important, at least two small strips of copper foil, as found in most hobby shops, are passed through the hole and soldered on each side of the shield.

The opposed pair of emitter leads are soldered to the copper foil straps on the *input* side of the shield after the collector lead is passed through the hole in the field (see figs. 6 and 7). Thus, the base and emitter leads are on the input-circuit side of the shield, and the collector lead extends through the shield to the output-circuit side of the box. Failure to use the copper foil straps under the emitter leads, or soldering the emitter leads to the output-circuit side of the shield, will invariably cause oscillation, even with a resistive collector load.

Note particularly the use of ferrite beads, low-inductance coaxial feed-through capacitors, and the 0.05 μF low-frequency bypass capacitors in parallel with the feed-throughs, to prevent oscillations at fretable 1. Test results obtained by the author on solid-state devices used in forty-four preamplifiers covering 30-432 MHz.

frequency (MHz)	device	mfg.	cost	NF (dB)	gain (dB)	R(dB)	M(dB)	P	BW (MHz)
30	MA42001	MA	11.50	1.05	18.0	−33	1.06	20.67	1
	2N5109	—	1.55	1.44	14.4	−39	1.49	14.23	BB
	MRF-901	M	2.10	1.47	26.2	−34	1.47	14.85	BB
	MPF-102	M	0.35	1.60	16.0	−16	1.63	13.61	1
	2N4416	—	0.50	1.62	13.1	−22	1.69	14.67	1
	MRF-904	M	1.25	1.65	23.0	−34	1.66	15.36	BB
	NE02135	NEC	4.00	2.15	16.7	−47	2.19	26.28	BB
	DGFET	—	1.00[1]	4.80	12.0	−58	4.99	44.91	1
50	MA42001	MA	11.50	0.95	15.3	−34	0.97	18.92	2
	2N4416	—	0.50	1.37	13.0	−15	1.43	12.16	2
	NE02135	NEC	4.00	1.55	15.0	−27	1.59	19.08	BB
	MRF-901	M	2.10	1.67	25.0	−32	1.67	16.87	BB
	KD6003	MA	5.00[2]	1.70	13.7	−28	1.76	22.88	T144
	2N5109	—	1.55	1.72	18.0	−29	1.74	16.62	T50-BB
	DGFET	—	1.00[1]	2.00	13.5	−46	2.07	18.63	2
	MPF-102	M	0.35	2.00	13.2	−27	2.08	17.37	2
	MRF-904	M	1.25	2.00	28.0	−34	2.00	18.50	BB
	U-310	S	4.00	2.50	10.3	−26	2.69	32.28	2
	MP-1006	AND	11.00	3.70	27.0	−29	3.70	70.30	BB
144	NE64535	NEC	17.00	1.00	22.0	−36	1.00	25.00	BB
	MP-1006	AND	11.00	1.37	16.6	−28	1.40	26.60	BB
	MRF-901	M	2.10	1.40	23.0	−28	1.41	14.24	BB
	MRF-904	M	1.25	1.41	17.0	−32	1.43	13.23	BB
	U-310	S	4.00	1.60	12.0	−27	1.69	20.28	2
	KD-6003	MA	5.00[2]	1.67	21.5	−25	1.68	21.84	BB
	DGFET	—	1.00[1]	1.75	17.0	−28	1.78	16.03	3
	NE02135	NEC	4.00	1.81	23.5	−31	1.82	21.84	BB
	2N4416	—	0.50	1.90	17.5	−20	1.93	16.40	2
	2N5109	—	1.55	2.45	12.5	−23	2.56	24.45	T50
	NE22235	NEC	17.00	2.80	17.0	−38	2.84	71.00	BB
	MPF-102	M	0.35	3.65	11.0	−24	3.86	32.23	2
	J-308	S	1.25	6.20	13.0	−18	6.37	58.93	(several measured)
220	NE-64535	NEC	17.00	0.96	19.0	−34	0.97	24.27	BB
	MRF-904	M	1.25	1.35	14.5	−28	1.39	12.86	BB
	MRF-901	M	2.10	1.40	18.1	−24	1.42	14.34	BB
	MP-1006	AND	11.00	1.66	15.1	−28	1.71	33.35	8
	NE-02135	NEC	4.00	1.87	20.8	−30	1.88	22.56	BB
	KD-6003	MA	5.00[2]	1.95	16.5	−23	1.99	26.87	T144
	2N4416	—	0.50	2.00	10.0	−20	2.17	18.48	4
	U-310	S	4.00	2.00	9.1	−28	2.22	26.70	4
	NE22235	NEC	17.00	2.10	29.3	−26	2.10	48.71	BB
	2N5109	—	1.55	4.60	7.9	−21	5.10	51.23	T50
	DGFET	—	1.00[1]	5.00	13.5	−22	5.14	46.26	4
432	NE24483	NEC	120.00	0.74	15.3	−25	0.76	114.00	GaAs fet
	NE64535	NEC	17.00	0.86	16.0	−24	0.88	22.00	BB
	MSC-H001	MSC	40.00[3]	(1.00)	(20.0)	—	(1.00)	—	GaAs fet
	NE02135	NEC	4.00	1.27	11.2	−27	1.36	16.37	BB
	MRF-904	M	1.25	1.38	11.0	−25	1.48	13.69	BB
	MRF-901	M	2.10	1.40	16.1	−22	1.43	14.44	BB
	NE22235	NEC	17.00	1.60	11.0	−26	1.71	48.25	BB
	MP-1006	AND	11.00	1.90	17.7	−28	1.93	37.64	55
	MA42161	MA	>15.00[4]	(1.90)	(16.0)	—	(1.94)	—	—
	MP-1004	AND	14.00	2.00	12.2	−34	2.11	47.48	BB
	MP-1001	AND	6.00	2.40	16.5	−23	2.46	34.44	BB
	KD-6003	MA	5.00[2]	4.20	7.7	−23	4.73	62.79	T144
	U-310	S	4.00	4.70	6.0	−30	5.57	66.84	10

notes

(1) Experimental device — not generally available.
(2) No longer available; MA42003 is a substitute. KD6003 circuit (reference 6) tuned to 145 MHz for all measurements.
(3) Not built. See reference 7 for data.
(4) Not built. See reference 8 for data.

legend

BB Broadband. Input noise matched at measured frequency.
M Minimum noise figure (dB). (See text.)
P Performance ratio factor. (See text.)
R Return loss through preamp (dB). (See text.)
Txxx Tuned at frequency indicated for all measurements.

manufacturers

AND AND Transistors, 770 Airport Blvd., Burlingame, California 94010.
M Motorola Semiconductors, 4221 East Raymond, Phoenix, Arizona 85040.
MA Microwave Associates (G.R. Whitehouse & Co., 17 Newbury Drive, Amherst, New Hampshire 03031).
MSC Microwave Semiconductor Corporation, (HAM-TRANS, P.O. Box 383, South Bound Brook, New Jersey 08880).
NEC Nippon Electric Company (California Electronic Labs, 1 Edwards Court, Burlingame, California 94010).
S Siliconix, 2201 Laurelwood Road, Santa Clara, California 95050.

quencies removed from the frequency to which the preamplifier is tuned. Also note the biasing arrangement. A zener between transistor collector and base is not recommended because zeners are themselves often used as noise sources. Thus, even if bypassed, they will inject noise.

As previously stated, arrangement is made for adjusting both the voltage and current at the device collector; noise figure is affected by these parameters. The zener limits the maximum V_{ce} to avoid device burnout; the actual V_{ce} value, adjusted for minimum noise figure, is less than the zener voltage.

The preamplifier may be initially tuned for maximum gain but should be finally tuned using a noise generator. A vacuum-tube noise generator, such as one using a 5722 tube, is discouraged: voltage spikes

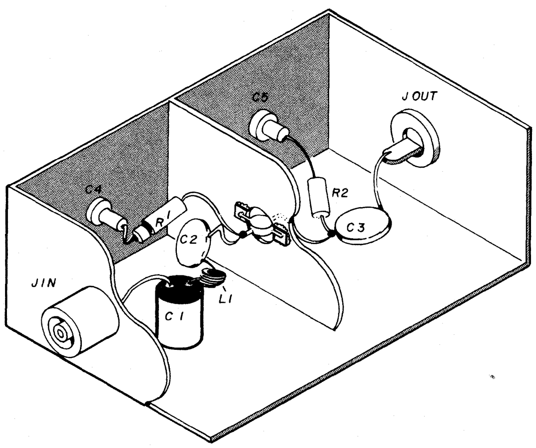


fig. 7. Isometric drawing showing internal construction

of sufficient amplitude to destroy the transistor may be present.

A semiconductor noise generator is preferred, if available. If a precision noise-figure test set isn't available, you'll find that the units of this preamplifier are close to minimum noise figure when adjusted for maximum gain, at the manufacturer's suggested optimum bias level of 8 volts and 7 mA.

Slightly higher noise figures can be achieved with a variety of lower-cost devices, as listed in **table 1.** Of particular interest is the 2-meter preamplifier using the U310, which is unconditionally stable in a grounded-gate configuration.[19] It has tuned input and output circuits, achieving a relatively narrow bandpass characteristic. The narrow bandpass characteristic prevents preamplifier intermodulation and blockage problems caused by strong signals in the aircraft, business, and other adjacent bands.

432 MHz. This band may, to the purist, be considered above the vhf frequencies. However, it is the highest-frequency Amateur band at which point-to-point wiring techniques have been found to be generally usable. So it's included as the highest frequency band at which vhf-type preamplifiers may be easily built.

The listings of **table 1** illustrate that high-frequency versions of the 144- and 220-MHz preamplifiers previously discussed do. indeed, give performance unheard of twenty years ago. The best performance is, however, obtained by using microwave gallium arsenide field effect transistors (GaAs fet) devices such as the NE24483[20] or MSC-H001.[7] Because of the inordinately high cost and great susceptibility to damage of these devices, they are used mainly by moonbounce operators. The NE64535 or one of the MRF901 or NE02135 devices in a 432-MHz preamplifier if followed by a converter with a 3-dB input noise figure, will fulfill the needs of most operators of this band. The preamplifier must be installed at the antenna to realize this increase in sensitivity.

Values are shown for 432-MHz operation in the figures for 144-220 MHz preamplifiers having circuits directly extendable to 432-MHz. Fig. **7** is a typical vhf preamplifier layout that can be adapted to most, if not all, of the preamplifiers discussed in this article.

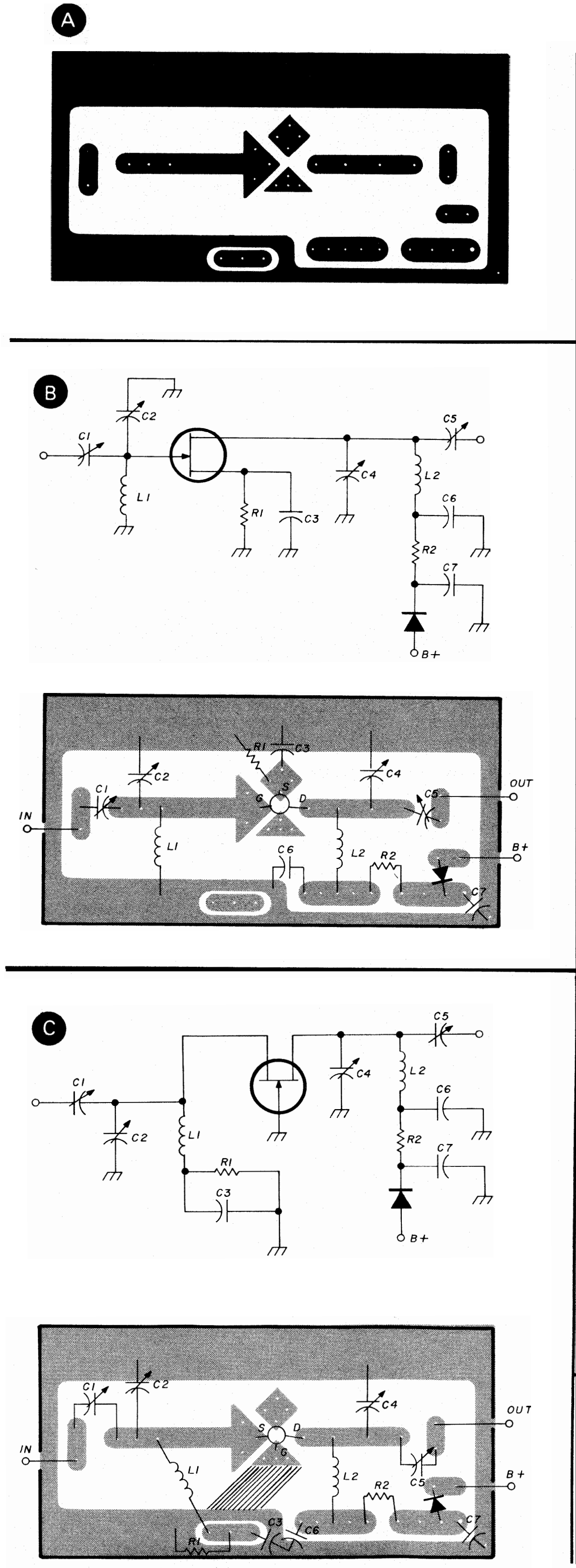
further construction hints

I've tried both point-to-point and PC-board construction (fig. **8**). Because full shielding of the preamplifier is desirable in addition to bias-lead filtering, PC technique is more costly and time consuming. Furthermore, if a double-sided PC board is used, with the unetched ground side as part of the shielding enclosure, the microstripline impedances of the circuit traces will have unexpected effects, particularly at 432 MHz. Unless the preamplifier is specifically designed to use microstripline (which will be considerably larger than a point-to-point wired preamplifier at 432 MHz), the low noise figure of modern-day transistors won't be realized.

Microstripline design generally assumes matching at least the resistive component of the input impedance to the data-sheet-specified device impedance. Because the device you use may have a slightly different impedance, optimum match can only be approached but never fully achieved. The illustrated, discrete-component, input-matching circuits allow adjustment for a range of resistive and reactive components of antenna and device impedances.

in summary

This article illustrates that truly low-noise vhf receiving preamplifiers can be built at the present time. Many of the preamplifiers built as part of this program have been tested not only on the lab bench



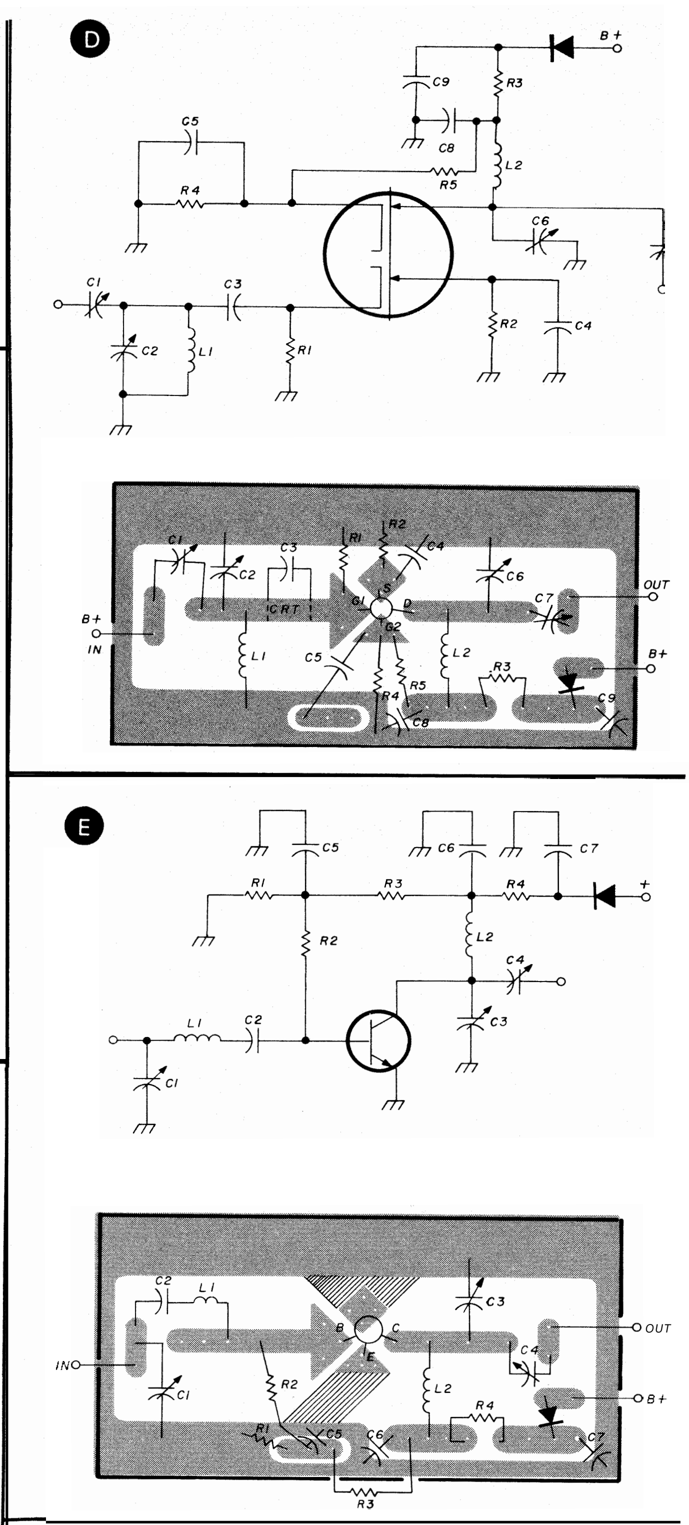


fig. 8. A "universal" PC board for vhf rf preamps, (A) usable to225 MHz. Stripline effects above about 300 MHz may seriously affect operation of 432-MHz preamps if a two-sided board is used (reverse side as a ground plane). (B) through (E) show, respectively. circuits for a **jfet,** common source; **jfet,** common gate; dgfet, common source; and a bipolar transistor. common emitter. (Note that the arrow on the board points from input to output.)

but also under contest conditions by W2SZ/1 in various vhf and uhf contests over the past two years. Additional preamplifiers are constantly being built as new devices are received. And, depending on the results eventually derived from such new preamplifiers, an update may be forthcoming. Preamplifiers for 1296 and 2304 MHz are also being built and evaluated An article covering these preamplifiers will be written when a sufficiently large number have been evaluated.

It should be kept in mind that all noise-figure measurements, especially those below about 2 dB, are not absolute but are *relative* indications of noise performance of the devices tested. The same test setup with circuits designed and built by one person may or may not result in the same answers obtained by others using identical circuits and procedures.

acknowlegments

I'd like to thank everyone who made devices available to me for construction and test of the preamps in this program. Special thanks are due the RPI Radio Club, W2SZ, for contest evaluation and, to Dick Frey, WB2BXP, for his help and encouragement.

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ham radio

repair bench

George Wilson, W1OLP

variable high-voltage supply

This workbench supply fulfills a need for test voltages in the 50-500 volt range. It can be built using mostly junkbox parts. It's particularly useful in reforming high-voltage electrolytic capacitors.

The title of this article originally was "The Care and Feeding of Electrolytic Capacitors." When it came time to take pencil in hand it appeared that the power supply needed to condition electrolytic capacitors has wider application and, hence, the more general (if less exciting) title. The supply described is meant to be used for light-duty experimental work. Its occasional utility on the workbench makes it a good investment of the construction time and cost for parts.

circuit description

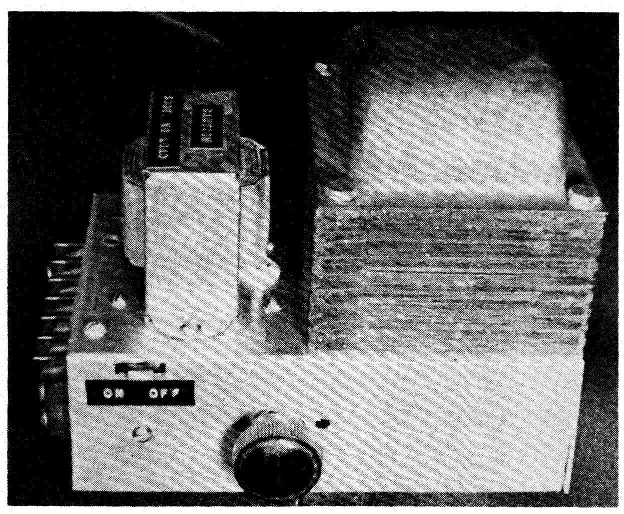
This is a *high-voltage* supply. Variable-voltage supplies for the range between zero and 50 volts are easily built with solid-state devices and have been described in the literature. The supply described here covers 50-500 Vdc. Output voltage adjustment to amounts less than 50 volts is touchy. So if you're interested in a power supply for lower voltages it's best to use one specifically designed for the purpose.

The circuit **(fig. 1)** is an ac voltage regulator (similar to those used in light dimmers and universal motor controllers) followed by a high-voltage transformer, rectifier and filter system. High-voltage silicon rectifiers are used. Vacuum-tube rectifiers would require a separate filament transformer and switch to provide constant voltage to the filaments. An LC filter is used to provide relatively ripple free dc. The choke also helps protect the 450-volt filter capacitor from current spikes. Note that this capacitor will be working close to its voltage rating at times. A 50-ohm 5-watt resistor can be substituted for the choke if you can't locate a suitable choke.

A separate ac switch is used rather than one connected to the voltage-control pot, which allows setting a voltage and turning the supply off and on without changing the set voltage. The bleeder resistor is primarily a safety device to help limit the voltage output (under light or no-load conditions) and to discharge the filter capacitor when the supply is turned off.

The component values in the ac voltage regulator circuit have been selected to provide relatively good voltage control over the 50-500 volt range. The triac circuit does not excel at the lower end of its control range, but its simplicity and low cost make it otherwise attractive.

(A) Author's experimental version of the variable high-voltage supply. The size and layout of yours will depend on the parts you can obtain, Note the separate on-off switch to allow voltage to be turned off and on without disturbing the variable setting. Terminal strips to mount small parts is recommended. A bottom plate will add safety. Don't omit the fuse. A low-value fuse is a good way to protect the circuit you're working on — a panel fuse mount will allow easy replacement.



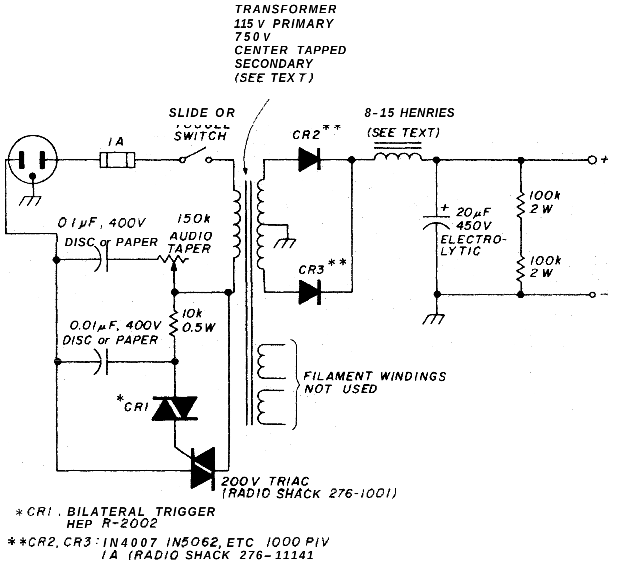


fig. **1.** Variable high-voltage supply.

symbol	part	description
C1	capacitor	0.01 μF, 400 V disc or paper
C2	capacitor	0.01 μF, 400 V disc or paper
C3	capacitor	20 μF, 450 V electrolytic
CR1	bilateral trigger	HEP R-2002
CR2, CR3	diode	1N4007, 1N5062, etc. 1000 PIV 1 Ampere (Radio Shack 276-1114)
F	fuse	1 Ampere
L	choke	8-15 Henries (see text)
Q	triac	200-volt triac (Radio Shack 276-1001)
R1	potentiometer	150k audio taper
R2	resistor	10k 0.5 watt
R3, R4	resistor	100k 2 watt
T	transformer	115-volt primary, 750-volt center tapped secondary (see text)

Except for the transformer and choke, the components should be relatively easy to obtain. The best source for the transformer and choke is an old TV set or a flea market. The semiconductors may be Radio Shack or Motorola HEP devices. An audio taper pot is suggested for smooth low-voltage control; a linear control is acceptable. The maximum voltage may be limited by adding a resistor in series with the wiper end of the pot. Take care not to exceed the filter capacitor rating by more than a few volts.

uses

The supply can be used for experimental tube and transistor circuits calling for voltages within its range. Current output is limited by the transformer or the choke you can obtain. The supply is particularly useful in reconditioning (or reforming) electrolytic capacitors. If you're reactivating an old piece of equipment or using capacitors from the junkbox, it's always best (and frequently necessary) to reform the electrolytic capacitors before applying full voltage. If reforming is neglected, the capacitor may short (completely or partially) internally. This causes heating and promotes further shorting and eventual capacitor destruction. Reforming is accomplished by allowing the voltage across the capacitor to increase slowly to its rated value. This can be done by increasing the voltage slowly by means of the voltage control. Capacitor polarity ***must*** be observed: The positive side of the capacitor must go to the positive side of the supply.

A better method of reforming electrolytics is to allow them to reform through a high resistance. This may be done by connecting the capacitor to the supply through a 100k resistor **(fig. 2)**. The resistor limits the current to less than 5 mA even if a shorted 450-volt capacitor is connected. A 2-watt resistor is recommended when reforming capacitors with working voltages above 300 volts. The current through the series resistor (and the power dissipated by it) will decrease rapidly if the capacitor can be reformed.

To use the setup in **fig. 2,** set the supply voltage (**A** to **C**) to the capacitor's rated voltage using a voltmeter. Then, with a high-impedance voltmeter connected to points B and C, measure the voltage across the capacitor. This voltage should increase slowly until the capacitor reforms completely. At this time the voltage across the supply (A to **C**) will be essentially equal to the voltage across the capacitor (**B** to **C**).

As in all high-voltage devices, be careful when

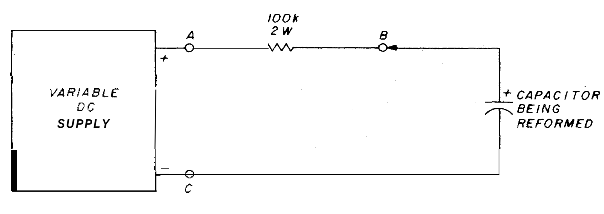


fig. **2.** Circuit for reforming electrolytic capacitors. Procedure is described in the text.

assembling and using this device. ***Dangerous voltages are present.*** Make sure you use well-insulated clip leads in experimental setups and follow the old adage: "Keep one hand in your pocket when working with high voltages."

ham radio

any-state ni-cad charger

The advantage of constant-potential and constant-current charging techniques are incorporated into this circuit

Have you ever wanted to connect your hand-held to a charger and walk away, knowing the charger will charge the battery as quickly as possible, then automatically become a trickle charger to keep the battery charged until you need it again? Here's a circuit that does exactly that.

ni-cad battery charging

Before describing the charger, a brief review of ni-cad battery-charging technique is in order. There are two basic charging methods: constant potential and constant current.

Constant-potential charging. This is the most rapid method and requires no adjustment to the charger during the charge period. This method requires a charger capable of delivering high currents, because at the start of the charge, the battery has a very low internal impedance. This high-current charge tapers off exponentially as a function of charge time **(fig. 1)**. The constant-charge voltage should be set at 1.55 times the number of cells in the battery.

A fully discharged ni-cad can be completely recharged by this method in one hour, although the charge should be continued for three hours or until the current stabilizes for one-and-a-half hours.[1]

The major disadvantage of the constant-potential charger is the high initial charge current. Not only does this require a voltage supply with high-current capability, but it's also possible to damage the battery being charged because of the high initial power being dissipated by the individual cells.

The usual method used to prevent cell damage with constant-potential charging is to monitor the temperature of a cell of the battery being charged and reduce the current to keep the temperature at some predetermined value. This is the system used with Motorola's "rapid charge" batteries and charger. The Motorola NLN-6900A rapid-charge battery has a thermistor that gives the charger battery temperature information.

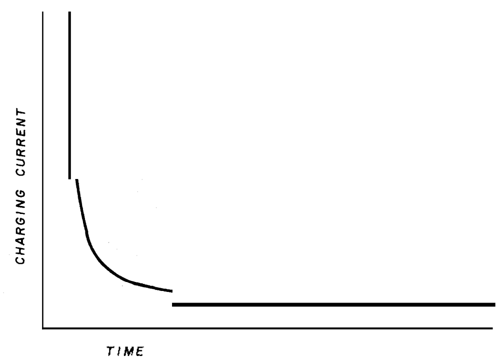


fig. 1. Constant-potential charging of ni-cads at 1.55 volts per cell. This method requires a charger capable of delivering high current.

Constant-current charging. The constant-current method requires a charging source of dc with a voltage of at least 1.8 times the number of cells in the battery. A simple constant-current charger is shown in **fig. 2.** To maintain constant current, the rheostat will require adjusting during the charge as the battery counter emf increases.

Practical values of charging current are usually

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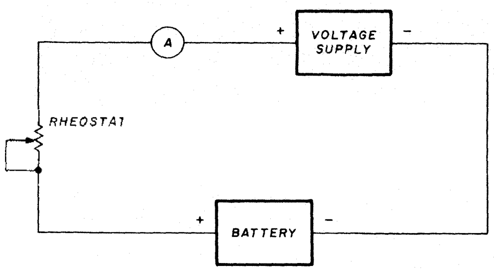

fig. 2. Simple constant-current charger. Constant current is maintained by adjusting the rheostat during the charge period.

selected to accomplish a full charge in three to seven hours. A l-Ah battery might thus be charged at any current rate between 500 and 200 mA, since it is necessary to put more energy back into the battery than was taken out. A good working figure is to adjust the charging time so that the battery receives at least 30 per cent more energy than the discharge, measured in ampere hours. Cell voltage versus time is shown in **fig.** 3.

The major disadvantage of constant-current charging is that if the charge current is not terminated when the battery reaches full charge, the cell must dissipate power no longer needed for charging. This causes cell temperature to rise and also causes loss of electrolyte. The most common method of circumventing this problem is to charge the battery at a constant current of 10-20 per cent of its ampere-hour capacity for a minimum of twelve hours. Overcharging the battery at 10-20 per cent is not usually harmful (although not good) to the battery as long as there is ample electrolyte in the cell.

design

The charger shown in **fig. 4** is a constant-current, constant-potential battery charger that has the advantages of both the systems described above and the disadvantages of neither. With this charger, the current is initially constant at a reasonably high level until the battery nears full charge, at which time the current decreases and the charger becomes a constant-potential charger.

R3, R1, Q1, and CR1 form a constant-current source with R1 controlling the amount of current flowing in Q1 emitter. This is the major source of charging current for the battery. When the potential across the battery reaches the point where CR2 and Q2 turn on, Q2 starts pulling away Q1's supply of base current, which reduces the current from the current source, so that the potential across the battery is constant.

A practical charger for a 200-mAh, 12-volt nickel-cadmium battery is shown in **fig. 5.** This circuit charges the battery at 75 mA until the battery is charged, then reduces the current to a trickle rate. It

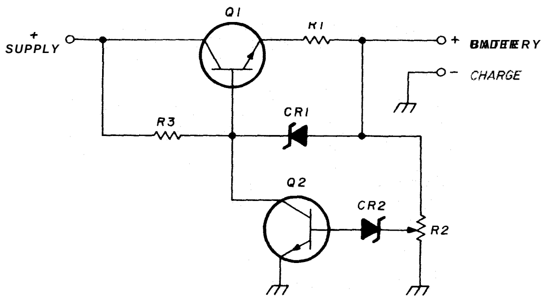

fig. 4. Basic design of the constant-current, constant-potential ni-cad battery charger. which has all the advantages of both techniques and the disadvantages of neither.

will completely recharge a dead battery in four hours and the battery can be left in the charger indefinitely. To set the shut-off point, connect a 270-ohm, 2-watt resistor across the charge terminals and adjust the pot for 15.5 volts across the resistor.

The circuit shown in **fig. 6** is built into the base of my HT-220 Omni charger. This circuit is built around the original Motorola power transformer and furnishes a constant-current charge of 200 mA (45 per cent of capacity). The charging lamp becomes a simple pilot lamp for the charger. The original "charge-trickle" switch now functions as an on-off switch for the charger. The pot is set for a voltage of 18.6 volts with a 240-ohm, 2-watt resistor across the charge terminals.

To modify this circuit to furnish different charge currents, the value of R1 may be determined by:

$$\frac{RL}{\text{current desired}} = 5 \text{ volts}$$

The constant-potential voltage can be determined by

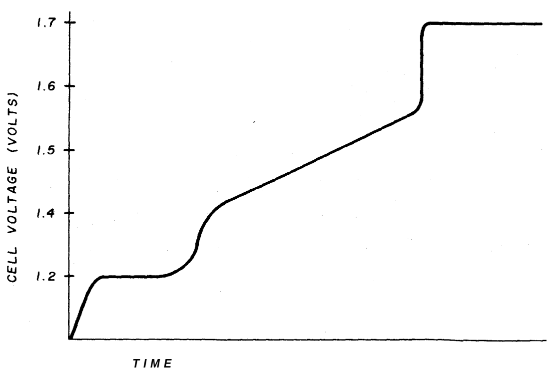

fig. 3. Cell voltage as a function of time using the constant-current charging method.

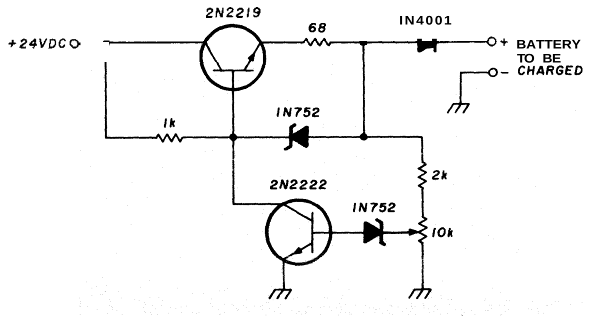

fig. 5. Practicai charger for a 200-milliampere-hour, 12-volt ni-cad battery.

muitiplying rhe number of cells by 1.55. The constant-potential voltage can be set by putting a load on the charger that pulls about one-half the desired maximum current and adjusting the pot for the desired voltage across this load. The supply voltage to the charger should be approximately two times the desired constant potential. This circuit will work for

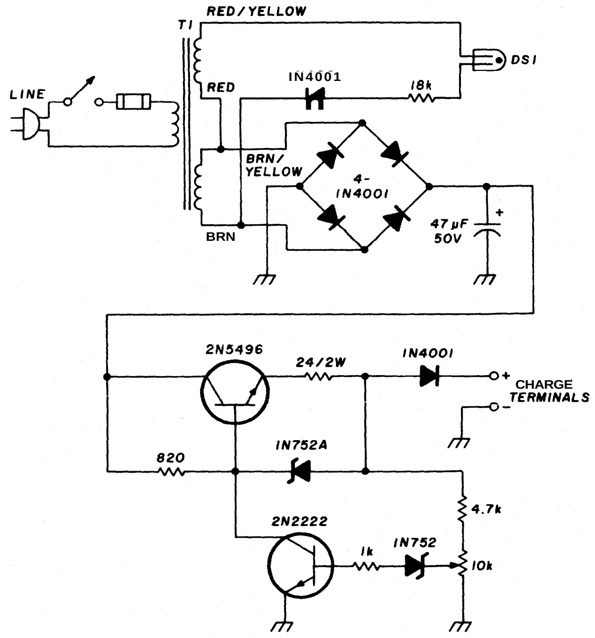

fig. 6. Circuit of author's charger, which is built into the base of an HT-220 Omni charger. Transformer T I is the original Motorola power transformer.

batteries in the 6-18 volt range with no changes. The maximum current available is 250 mA; however, this could be increased by adding a Darlington connection to Q1.

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