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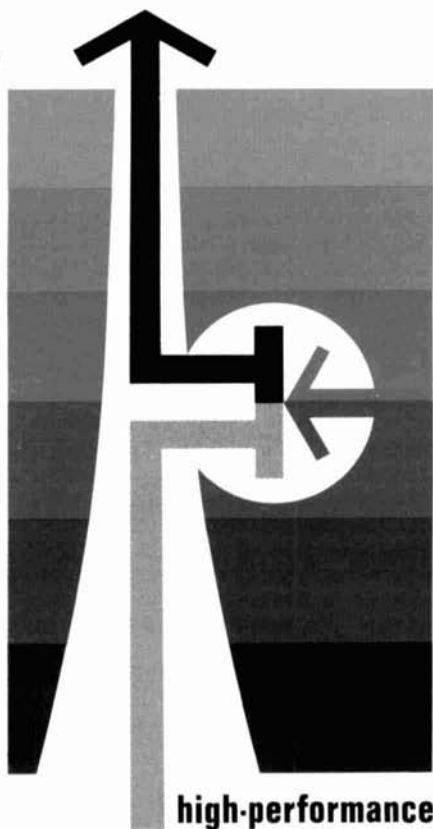
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# ham radio

*magazine*

AUGUST 1970



**high-performance**  
**FILTER / PREAMPLIFIER**  
for vhf/uhf receivers

## *this month*

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- tunable audio cw filter 34
- stable solid-state vfo 36
- tunable six-meter converter 50
- cubical-quad antenna design 55



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February 7, 1969

Mr. Irving Strauber, Sales Mgr.  
Hammarlund Mfg. Co.  
20 Bridge Ave.  
Red Bank, New Jersey 07701

Dear Sir:

I am enclosing a couple of glossy prints, showing our frequency measuring setup, in which the Hammarlund HQ 180-AX receiver is a key instrument. We use it to receive WWV on 2.5- 5- 10 and 15 MHz against whose signal we beat the output of our GR Calibrator for a zero beat. Signals to be measured are then received on the receiver, and in turn beat against the GR Calibrator output.

We measure the frequencies of two other television stations as well as those of some of the local broadcast stations. This includes the color burst frequency, the 4.5 MHz difference beat between visual and aural carriers and by use of Bessell functions we check the calibration of the aural modulation monitors.

I have drawn a chart of the locations of the various instruments in the rack and am also enclosing it.

The man in one of the pictures is Stan Allison of our transmitter staff. I forgot to mention that we also can check our tuning of the transmitter multiplier stages, receive time checks from WWV, and use the receiver as a standby in case of failure of our EBS receiver.

Sincerely yours,  
*Gerald King Allison*  
Gerald King Allison  
Transmitter Supervisor

**EAC**



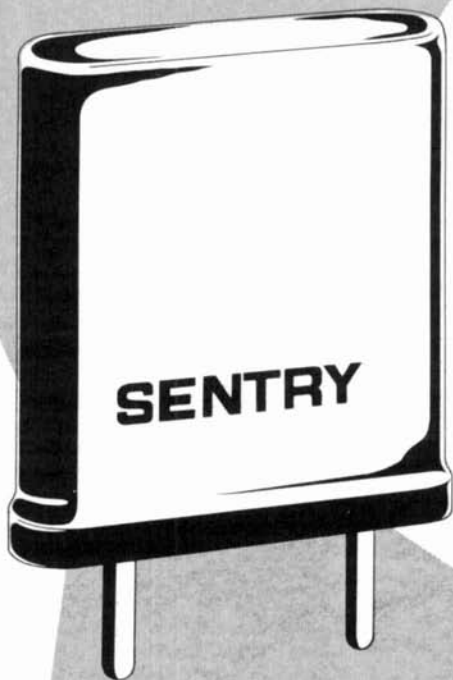
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## a second look

by jim  
fisk

**LSI, or large-scale integration**, is a term that is becoming more and more commonplace. An outgrowth of the transistor, and more lately, integrated circuits, LSI promises to benefit all of us by providing low-cost complex electronic systems for home and industry.

As production yields of less sophisticated devices such as linear integrated circuits have increased, so have the semiconductor manufacturer's efforts to put more and more electronic components on the same chip. Improved semiconductor technology, advanced manufacturing techniques, and new processes have resulted in LSI devices that defy imagination. The LSI chip shown in the photo, for example, contains 1,191 p-channel enhancement-mode mosfet transistors. This device, the Motorola MC1141, is a triple 66-bit shift register that is designed to operate over the frequency range from 10 kHz to 1 MHz, and is packaged in a long-lead version of the TO-100 metal can.

Much more complex circuits are in the works, and some devices are already on the market. In the new Boeing 747 jets,

LSI has made it possible for passengers to have armrest control of their own lights, music and movie soundtracks. A small box containing an LSI chip is located on every third seat—this chip multiplexes the codes for the seat controls. Without LSI, 35 individual wires would have to be run to each seat.

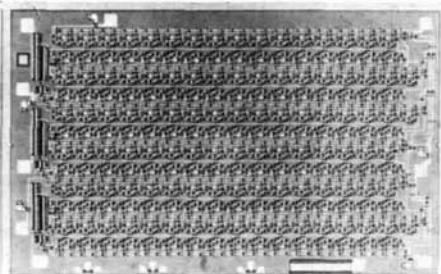
LSI is also being used in advanced computers and home calculators. Consider the computer built for NASA by RCA that is completely contained on a semiconductor chip one-seventh of an inch square, and performs all the arithmetical functions of a medium-sized computer. Or, how about the miniature calculator built by Canon (Japan), soon to be marketed in the United States, that adds, subtracts, multiplies and divides 12-digit numbers to four decimal places? The Canon calculator, which will be priced below \$200 and weighs less than 2 pounds, contains three LSI chips, each an eighth of an inch square and containing 4,000 transistors.

Within a few years LSI may make possible such products as inexpensive home computers, trouble-free electronic controls for stoves and dishwashers, television sets no thicker than a picture frame, and telephones with built-in memories. With these products will come complex electronic instruments and equipment for the amateur that is now too complex or expensive to be practical.

If thinking miniature appeals to you, consider the prospect of 12 billion electronic circuits in a three-pound package—that's the approximate complexity and density of the human brain. With LSI, such an accomplishment may be a reality within the next decade or two.

Jim Fisk, W1DTY  
editor

Motorola MC1141 triple 66-bit shift register contains 1,191 mosfets on a single chip.



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Or ask Information Operator for Varian Electron Tube and Device Group.



# So we moved ahead of them.

# interdigital preamplifier and combine bandpass filter

## for vhf and uhf

High-performance  
filter/preamplification  
for vhf receivers  
that features  
low cross modulation,  
low noise  
and excellent  
unwanted-signal  
rejection

Bob Cooper, W5KHT, 6221 Norman Road, Oklahoma City, Oklahoma 73122

With the proliferation of vhf radio and television signals in virtually all metropolitan areas, weak-signal vhf reception that requires large antennas and extensive rf amplification has posed ever mounting problems to the receiver designer. Unfortunately, the requirements for weak-signal work are 180 degrees out of phase with an overload-proof receiving system.

You simply cannot erect a 16-dB antenna system for 50.1 MHz, follow it with a 20-dB preamplifier and install the whole works within five miles of a 100 kW channel-2 television transmitter—that is if you want the system to operate as if the 55.25 MHz signal was not on the air.

During the past five years or so, operation on 144 MHz has become equally challenging. In addition to the multiple transmitters located in the various public safety and business radio bands above 150 MHz, and the aircraft beacons and air-to-ground links in the 120 to 135 MHz range, we have high powered amateur repeaters within the band and MARS repeaters just outside. Also, activity on the 220-MHz band has always suffered in areas where channel 13 is in use, and channel 13 is allocated in practically all major metropolitan areas.

It all boils down to this: receiver desensitization—voltage overload of the receiver—is a serious problem that is slowly getting worse. As the occupancy of frequency allocations adjacent to our vhf amateur bands slowly builds up, the performance of our own communication systems becomes progressively more inadequate. Equally bothersome are the images and spurious beat products produced when two or more out-of-band carriers mix within our converters and generate new carrier products that fall

only way to cure the problem is to keep one or more of the mixing carriers out of the receiver stage where the mixing is taking place. It's common practice in many vhf converters to install a series-resonant circuit to trap undesired signals that cause spurious mixing products. However, this approach is not too effective, especially in strong-signal areas.

The complete solution is simple enough: make sure that the first stage of the vhf receiving system sees only that portion of the rf spectrum assigned to radio amateurs! However, solution is one thing, implementation another. Until W2COH presented an amateur adaptation of interdigital and combline bandpass filters in *QST*<sup>1,2</sup> the usual amateur practice (and commercial as well) was to laboriously build a re-entrant cavity. The bandpass filter, when placed in front of the receiver, presents the receiver with an rf window a few MHz wide and severely attenuates all rf carriers that fall outside the window. The combline filters described by W2COH offer simple construction, relatively narrow bandpass windows, steep bandpass skirts and extremely simple tuneup.

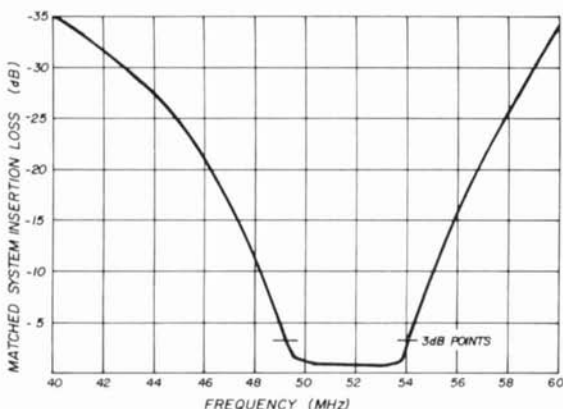


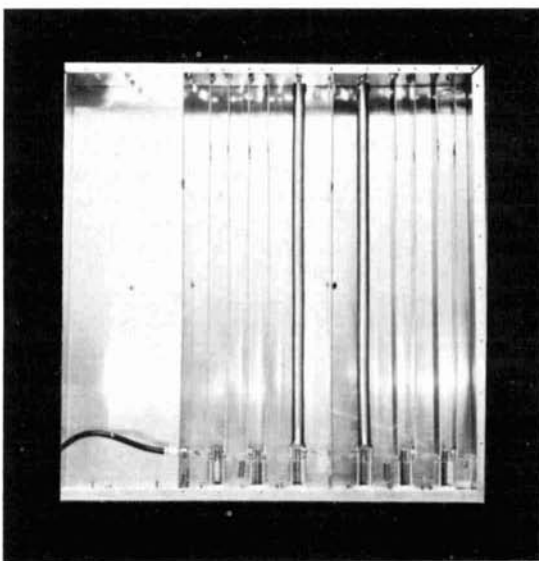
fig. 1. Insertion loss of a 3-element combline filter centered on 52 MHz. The 3-dB bandwidth of this filter is 2.2 MHz.

within the amateur band we want to work.

These problems may be a constant annoyance (such as the beat product of a channel-3 video carrier at 61.25 MHz and a channel-12 video carrier at 205.25 MHz that produces buzz and hash on 144.00 MHz) or they may be intermittent. Carriers may add or subtract to produce the undesirable in-band product, or two mixing carriers may produce a spurious signal outside the band that mixes with a third carrier to produce an in-band birdie. The mathematical possibilities are endless and defy computation.

However, all these problem signals have one thing in common: they are created at some point in the receiver. The

Interdigital preamplifier for 50 to 100 MHz. Input is to the right, output on the left.



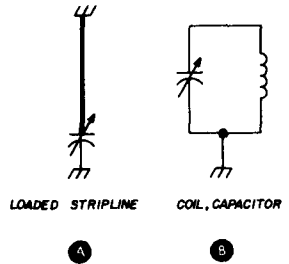
Comblines filters use strip lines for the L portion of the traditional LC network, loaded with variable capacitors to resonate the device to the desired center frequency. The 3-dB bandwidths of a three-element device are from 3 to 8 percent of the design center frequency. The bandpass window of a typical 3-element combline filter at 50 MHz is shown in fig. 1. Insertion loss of these filters is quite low; all the 3-element filters that I've built have had 1.0 dB or less insertion loss at the center of the passband.

### the preamplifier

The small-signal rf transistor has come of age as the radio spectrum has grown more and more crowded. The overload and cross-modulation problems associated with early transistors began to look less

conix\* in 1968 was a major breakthrough for the receiver designer working above 250 MHz. While the 2N4416 created a boom in high-performance converters for the 144 and 220 bands (and occasionally 432 where it was derated), the 2N5397 offered super-simple circuitry. The 2N5397 grounded-gate circuit recommended by Siliconix for vhf applications, described in *ham radio* by K6HCP<sup>3</sup>, results in as simple a preamplifier circuit as we could hope for.

fig. 3. Capacitively-loaded stripline in A is the same as the parallel LC circuit in B.



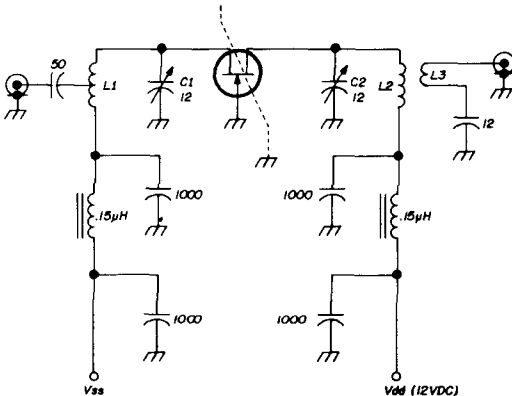
In describing the design parameters of the 2N5397, J. B. Compton notes<sup>4</sup> that "... this device has a typical operating transconductance more than double (the 2N4416) .. and this permits construction of common-source common-gate performance amplifiers that take advantage of the fet characteristics and give performance comparable to bipolar amplifiers over the frequency range from dc to 800 MHz."

The neutralization requirement of the common-source amplifier is eliminated with the common-gate arrangement. By eliminating neutralization you can design a tunable wide-range amplifier with amplification/bandpass characteristics limited only by the resonant characteristics of the input and output tuned circuits.

The common-gate approach is especially suitable because:

1. The input impedance is always lower than a common-source design,

\*Siliconix, Inc., 1140 E. West Evelyn Avenue, Sunnyvale, California. The 2N5397 is available from W. Pat Fralia Company, Inc., P. O. Box 12625, Fort Worth, Texas 76116; \$8.50 each.



- L1 1.5" long no. 16 copper wire
- L2 1.2" long no. 16 copper wire
- L3 2.0" long no. 16 copper, loosely coupled to L2 (3/4" spacing)

fig. 2. 450-MHz common-gate 2N5397 test circuit designed by Siliconix. In this configuration gain is 12 dB, noise figure, 4 dB.

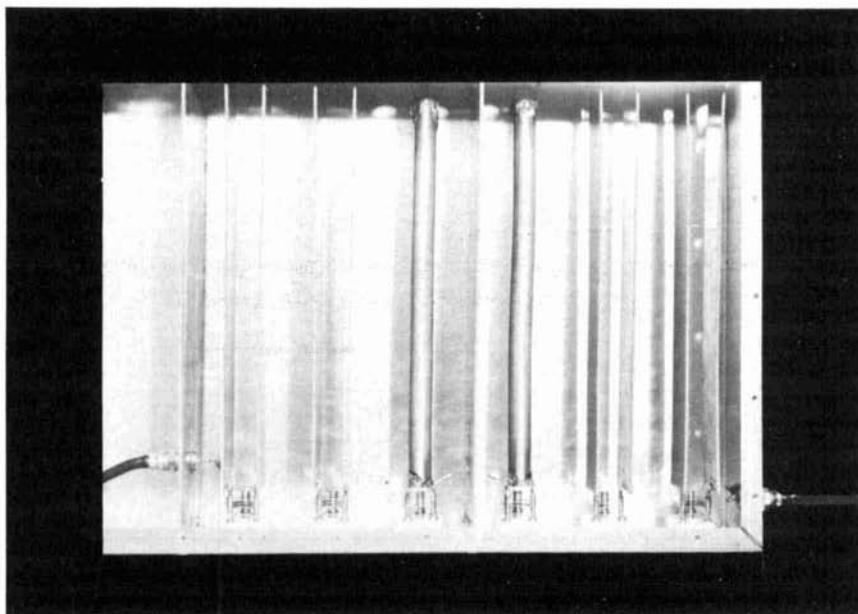
ferocious when the junction field-effect transistor was introduced. Early fets promised improved performance under taxing receiving situations that cause conventional transistors to quit, but early fet noise figures and gain made them less than useful above 100 MHz or so.

The 2N5397 jfet introduced by Sili-

although the output impedance is the same.

2. With no mismatch at the input a vhf noise figure of 2.0 dB or better is possible with the common-gate 2N5397. (Both K6HCP and I have found no difficulty in obtaining 1.5-dB noise figures at 144 MHz, and on 220 MHz I have never had a noise figure greater than 2.0 dB, nor greater than 1.0 dB at 50 MHz.)

without some loss in performance when compared to the common-source circuit. In 450-MHz test circuits developed by Siliconix for the 2N5397, a typical common-source amplifier produced 15-dB gain with a noise figure of 3.0 dB. A common-gate circuit developed a power gain of 12 dB with a noise figure of 4.0 dB. Slight retuning of the common-gate's input circuit lowers the noise figure to 3.3 dB and lowers the gain to 9 or 10 dB (fig. 2).



Interdigital-filter/preamplifier for 120 to 240 MHz. The only differences between this unit and the 50-to-100 MHz design are the physical size and variable capacitors.

3. The amount of amplifier-contributed cross-modulation and intermodulation distortion is proportional to the amplitude of the gate-source voltage. Since input power is proportional to input voltage and inversely proportional to input impedance, best amplifier performance is obtained with the lower impedance design of the common-gate circuit. A common-gate design offers approximately 10 dB more resistance to rf stage overload and cross modulation than a common-source arrangement.

However, the common-gate design is not

### filter/preamp combination

When a 3-section combine filter is cascaded with an identical filter, the 3-dB bandwidth points are essentially cut in half.\* In other words, a 3-section

\*The information presented here covers relatively narrow bandpass devices suitable for amateur applications. They are not suitable for television signal processing where 6-MHz bandwidths and 0.25-dB or less ripple is required. Both the units shown here and wider bandpass units suitable for television signal processing are the subject of patent applications filed by the author in March, 1970. Amateur construction of the units shown here for personal use will not violate the validity claims of the pending patent. editor.

5-MHz-wide combline filter at 50 MHz becomes a 2.5-MHz-wide filter at the output of the sixth section (two 3-section filters in cascade). A 5-MHz-wide filter at 144 MHz also becomes a 2.5-MHz bandpass in a two-unit system, while a typical 6-MHz bandpass on 220 is reduced to 3 MHz.

The factors that determine the width of the bandpass window are beyond the scope of this article. Suffice to say that the 3-dB bandpass can be changed within certain limits, both plus and minus, from

lowest resonant frequency of a given filter is 100 MHz with the loading capacitor fully meshed, the highest resonant frequency is approximately 200 MHz. By combining the flat characteristics of the combline filter with the flat performance characteristics of the 2N5397 common-gate amplifier you can build an amplifier/filter that exhibits essentially the same characteristics over a very wide band of frequencies.

Fig. 4 shows how a pair of 3-section combline filters are combined with a

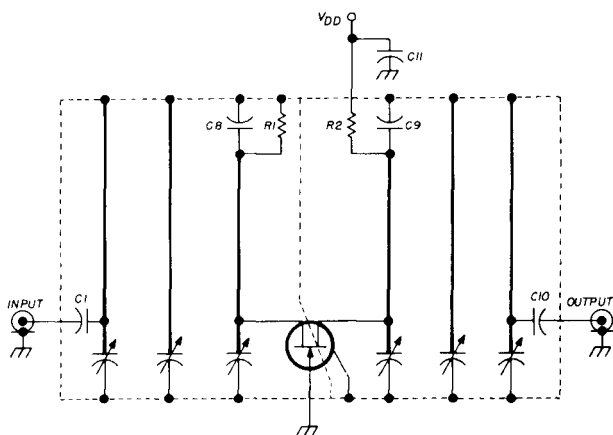


fig. 4. equivalent circuit of the interdigital filter/preamplifier.

approximately 10 MHz at 50 MHz to less than 500 kHz at 220 MHz. The information in this article should be suitable for most amateur applications.

## the filter

A single capacitively-loaded stripline represents a resonant circuit. Electrically, the loaded stripline shown in fig. 3A is identical to the LC circuit in fig. 3B. The efficiency of the loaded stripline is more dependent on the L/C ratio than the standard LC network where L consists of a coil. However, in the stripline system the Q can be made to track (with nearly constant Q factor) over a much wider range than the LC network.

Although W2COH didn't mention it in his article<sup>2</sup> a typical 3-section combline filter is a one-octave device. That is, if the

single common-gate amplifier. In this circuit the third strip in the *first* 3-section filter becomes the tuned input circuit for the 2N5397, while the *first* strip in the *second* section becomes the output tuned circuit.

In the standard combline filter all strips are grounded at the cold end and tuned with the loading capacitors connected from the hot end to ground. In the filter/preamplifier, strips three and four are lifted above ground at the cold end. Strip three—the 2N5397 input circuit—is re-coupled back to ground through R1, a low value resistor chosen to provide the proper current drain for the fet. Strip number four is isolated from ground with a bypass capacitor (C9) and voltage fed through an isolating resistor. This circuit is nearly electrically

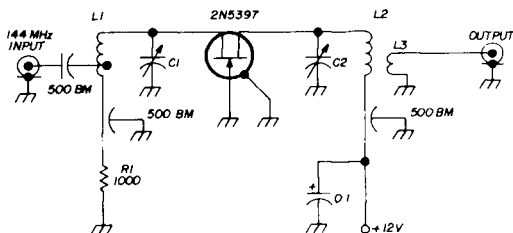


equivalent to K6HCP's 144-MHz preamplifier shown in fig. 5.

By combining the common-gate 2N5397 with the combine bandpass filter we have a low-noise moderate gain preamplifier with an input bandpass of 500 kHz or more.

### operational characteristics

The devices shown here were originally developed in an effort to solve an extremely difficult adjacent-channel tv interference problem at a catv antenna-



- C1, C2** 1.7-14.1 pF air variable
- L1** 5 turns no. 16, 3/8" diameter, 3/4" long, tapped at 1/2 turns from cold end
- L2** 4 3/4 turns no. 16, 3/8" diameter, 1/2" long
- L3** 2 turns no. 20 insulated around cold end of L2

fig. 5. Two-meter common-gate 2N5397 preamplifier described by K6HCP.

receiving site. As a consultant in the catv field I had been called upon to make a 30- to 50-microvolt vhf tv signal appear free of noise and hash with broadcast quality color—not too tough until you consider the 100,000 microvolt adjacent-channel signal that came from a station less than five miles away!

Anyone who has fooled with fringe-area television is all too familiar with the broadness of tv frontends. The typical receiver has no more than 20-dB adjacent-channel rejection; the relationship between 50 microvolts and 100,000 microvolts is on the order of 66 dB. The problem was solved with an interdigital preamplifier similar to the ones shown here.

The common-gate 2N5397 preamplifier used in the filters operates just as

table 1. Operational characteristics of 6-strip interdigital-preamplifiers suitable for 50, 144 and 220 MHz.

frequency (MHz)	3-dB bandwidth (MHz)	20-dB bandwidth (MHz)	gain (dB)	noise figure (dB)
50	0.5	±4	10	<1.0
144	1.0	±5	12	1.5
220	1.5	±6	12	2.0

expected. Stability is excellent. The unit tunes evenly with no instability from the lowest resonant frequency to the highest. Intermodulation and cross modulation within the preamp is as good or better than any solid-state device you can use. The noise figure of this device, as measured on working models, is just under 1.0 dB at 50 MHz, increasing to 2 dB on 220 MHz and rising to 4 dB at 450 MHz and 6 dB at 800; a noise-figure curve is plotted in fig. 6.

The gain of a single stage of amplification depends a great deal upon the L/C ratio of the striplines and load capacitors, as well as coupling between striplines and the match to the transistor. In the early stages of development I built nearly a dozen filter/preamplifiers to determine the proper balance of L and C. The performance characteristics were carefully measured as L and C were juggled,

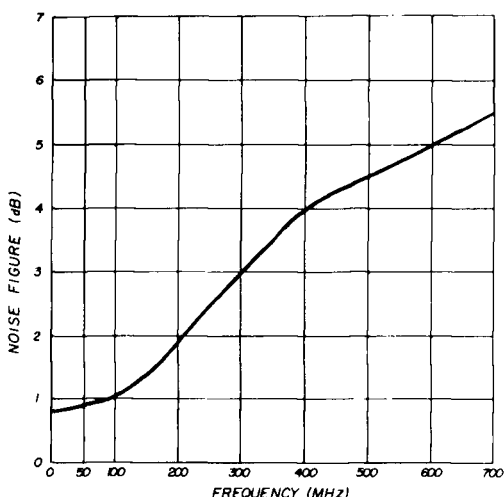
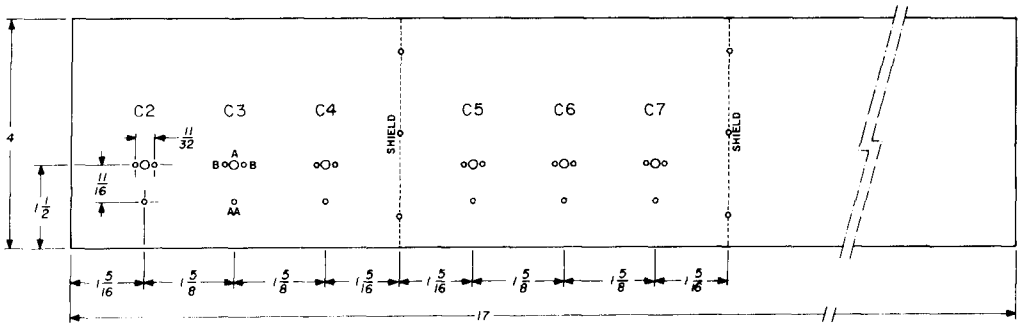
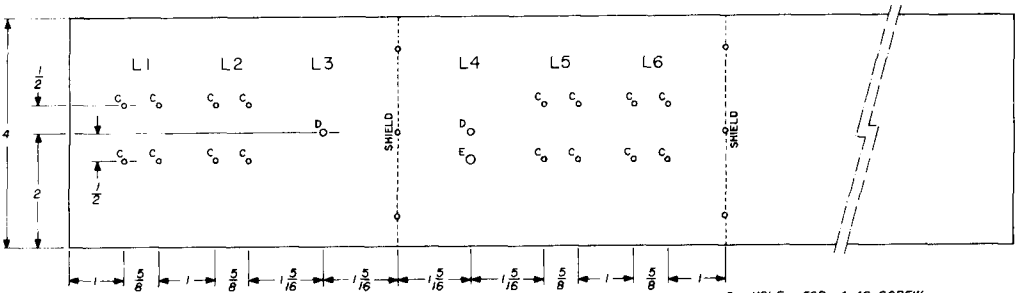


fig. 6. Noise-figure curve for the grounded-gate 2N5397 preamplifier.



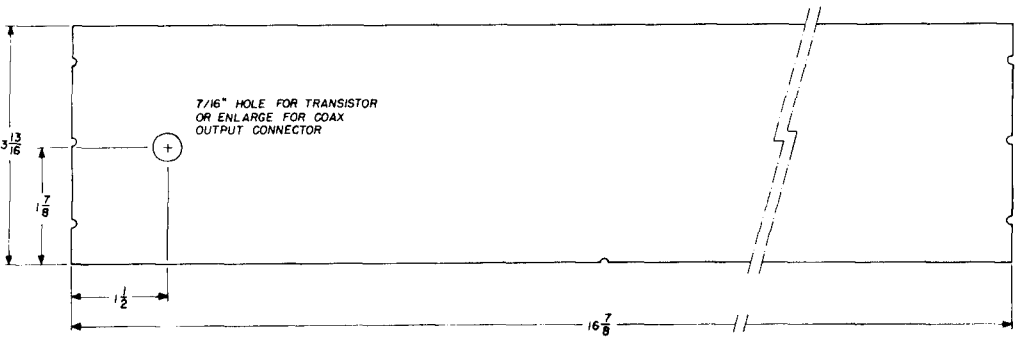
CHASSIS FRONT (FROM OUTSIDE)

A HOLE FOR CAPACITOR SHAFT  
 B CAPACITOR MOUNTING HOLE  
 AA HOLE FOR ROTOR-GROUNDING SCREW



CHASSIS REAR (FROM INSIDE)

C HOLE FOR 4-40 SCREW  
 D HOLE FOR FT-1000 FEEDTHROUGH  
 E HOLE FOR 1/4-WATT RESISTOR



SHIELDS (2 REQ'D)

NOTCHES FOR 4-40 x 1/4" BRASS SCREWS

chassis 17 x 17 x 4" (BUD AC-1431)

shields 16-7/8 x 3-13/16"; cut from double-sided copper-clad printed-circuit board (Kepron P2-1212G)

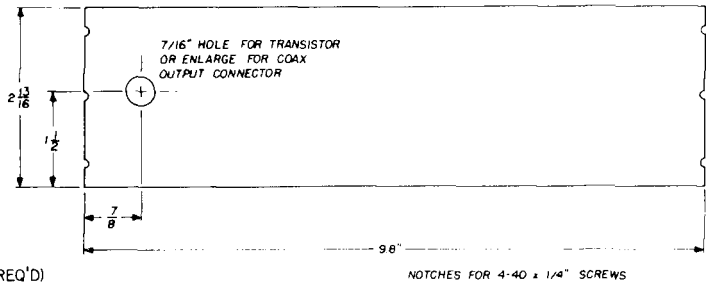
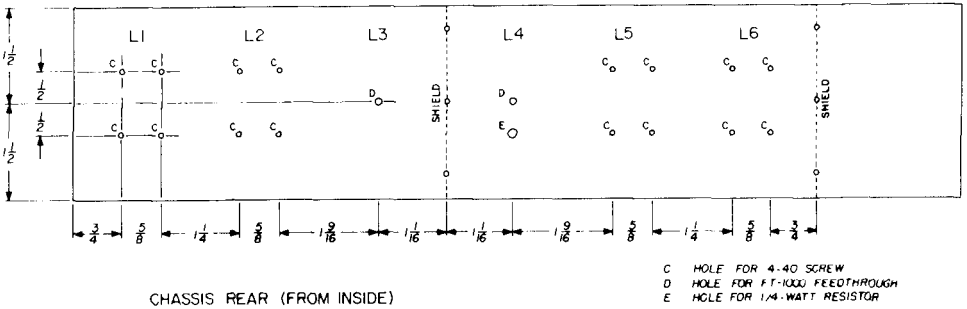
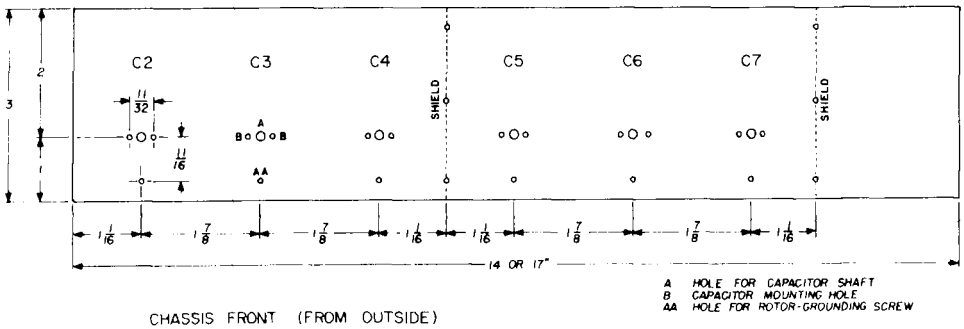
cover 17" x 17" (BUD BPA-1597)

fig. 7. Construction details of an interdigital filter/preamplifier that tunes from 50 to 100 MHz. Striplines are spliced as described in the text. Hole for input coax connector is centered, 1 1/2" from bottom of chassis.

strip mounting and strip-to-strip spacing varied, and input/output coupling changed. As in most repeatable designs, predictable patterns developed that provided the basis for design formulas.

The information presented in this article will be limited to units that are

especially suitable for the vhf amateur bands. The performance characteristics of 6-strip interdigital-preamplifiers for 50, 144 and 220 MHz are shown in table 1. Two identical interdigital-preamplifiers may be cascaded for any of the bands. This results in narrower 3-dB bandwidths,



chassis 10 x 14 x 3" (BUD AC-414) or 10 x 17" (BUD AC-416). Required interior dimensions are 10 x 11 3/4", a non-standard chassis size; required size is provided by an interior end plate

cover 10 x 14" (BUD BPA-1524) or 10" x 17" (BUD BPA-

shields 9.8 x 2-13/16"; cut from double sided copper-clad printed-circuit board (Kepra P2-1212G)

fig. 8. Construction details of an interdigital filter/preamplifier that tunes from 100 to 200 MHz. To raise the resonant frequency to 120 to 240 MHz remove one plate from C2 and C7; remove 2 plates from C3 and C6; remove 2 stator plates and 3 rotor plates from C4; and remove 3 stator plates and 3 rotor plates from C5. Hole for input coax connector is centered, 1-1/16" from bottom of chassis.

considerably steeper 20-dB bandwidth skirts, and twice the gain of a single unit with the same noise figure.

### construction

Two versions of the interdigital-preamplifier are shown; one tunes from

50 to 100 MHz and the other covers from 120 to 240 MHz. Both are based on standard BUD aluminum chassis with bottom plates. The striplines are probably a little different than anything you've seen before—they consist of two pieces of copper-clad printed-circuit board, sup-

ported on the variable capacitor at the hot end and held by two 1/4-inch 4-40 brass machine screws at the other.

Unlike the W2CQH combline, the striplines in the interdigital-preamplifier hang vertically. This effectively gives you twice as much surface area for a given

from the output circuit by a piece of double-sided copper-clad circuit board that runs the full width and full height of the chassis (see fig. 7). The 2N5397 is mounted in a 7/16-inch hole drilled in the shield; the case lead of the fet is soldered to the side of the shield facing stripline

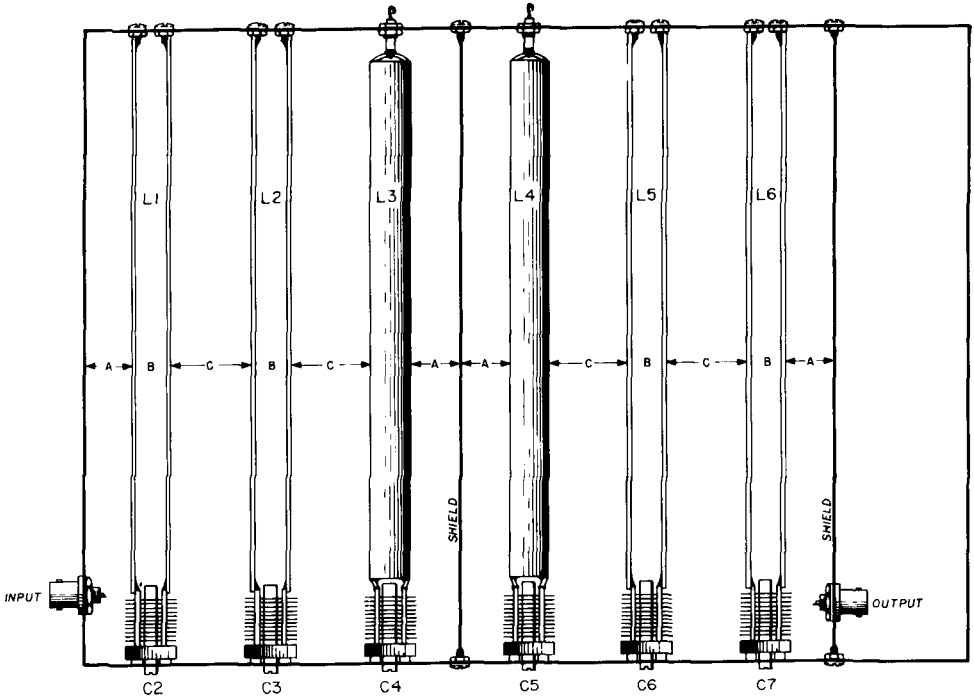


fig. 9. Mechanical layout of the filter/preamplifier. Dimensions are determined by frequency range as shown below:

frequency (MHz)	A	B	C
50-100	1-1/32"	5/8"	1"
100-200	3/4"	5/8"	1-1/4"

resonator length and directly affects the efficiency of the LC networks, particularly at the lower operating frequencies. Striplines 1, 2, 5 and 6 use this technique; striplines 3 and 4, because of impedance matching to the 2N5397, use equivalent lengths of copper tubing. This design approach is several dB more efficient than all copper-tubing resonators, single horizontally suspended striplines or combinations of horizontally suspended strip and copper tubing.

The 2N5397 input circuit is shielded

number 3, the gate lead is soldered to the side of the shield facing stripline number 4, the source lead connects directly to a pigtail from C3, and the drain lead is soldered to a pigtail on C4.

Input coupling to stripline number 1 and output coupling from stripline number 6 is accomplished with ceramic capacitors; values are given in fig. 10 for either 52- or 75-ohm operation.

Striplines 3 and 4 are supported by Centralab FT-series feedthrough capacitors. Sprague BH-series stud-bypass capacitors could be used here, but construction is more difficult because of their small size. Stripline number 3 is resistance

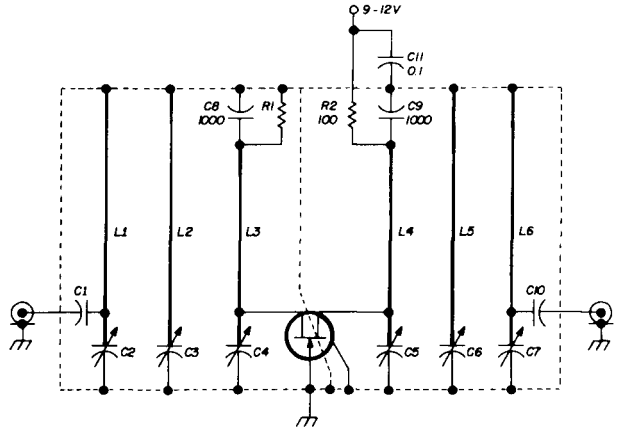
coupled to ground through a resistor that is chosen to allow the 2N5397 to draw 5 mA; typical values range from 100 to 470 ohms with 330 to 390 ohms quite common. Stripline number 4 is resistance coupled to the power supply through a 100-ohm resistor with additional bypass-

tion is installed after stripline number 6. This serves as the outer wall and holds the output coaxial connector.

### how to build it

Before starting construction the chassis should be laid out with each mount-

fig. 10. Schematic diagram for the interdigital filter/preamplifier.



<b>C1,C10</b>	<b>30-pF ceramic (75 ohms); 35-40-pF (50 ohms)</b>
<b>C2,C3,C4, C5,C6,C7</b>	<b>5-100-pF air variable (Calectro A1-226)</b>
<b>C8,C9</b>	<b>Centralab FT-1000 feed-through capacitors</b>
<b>C11</b>	<b>0.1-μF ceramic, 50 volts working</b>
<b>L1,L2, L5,L6</b>	<b>8 strips (2 per stripline) cut from 1 oz. (0.00135) single-side copper-clad printed-circuit board (Kepro P1-1212G). Each strip 15¼" long, 1½" wide</b>
<b>L3,L4</b>	<b>5/8" OD copper tubing, 14¾" long</b>
<b>R1</b>	<b>100 to 470 ohms (see text)</b>
<b>R2</b>	<b>100 ohms, ¼ watt</b>

#### Components for 50 to 100 MHz unit

<b>C1,C10</b>	<b>6.8-pF ceramic (75 ohms); 10-pF ceramic (50 ohms)</b>
<b>C2,C3,C4, C5,C6,C7</b>	<b>2.9-30-pF air variable (Calectro AL-225)</b>
<b>C8,C9</b>	<b>Centralab FT-1000 feed-through capacitors</b>
<b>C11</b>	<b>0.1-μF ceramic, 50 volts working</b>
<b>L1,L2, L5,L6</b>	<b>8 strips (2 per stripline) cut from 1 oz. (0.00135) single-side copper-clad printed-circuit board (Kepro P1-1212G). Each strip 8.9" long, 1.5" wide.</b>
<b>L3,L4</b>	<b>5/8" OD copper tubing, 8.4" long</b>
<b>R1</b>	<b>100 to 470 ohms (see text)</b>
<b>R2</b>	<b>100 ohms, ¼ watt</b>

#### Components for 100 to 200 MHz unit

ing on the outside of the chassis through a 0.1 μF ceramic capacitor.

Part of the BUD chassis is not used in both models of the filter/preamplifier shown here. In addition to the shield between striplines 3 and 4, another parti-

ing hole center punched so each hole is drilled where it should be. Small errors in strip-to-strip spacing are tolerable, but anything over 1/16 inch is *not* small.

Sheet-metal screws are used to fasten the top plate to the chassis. It's manda-

tory that this plate fits very tightly because, as well as a shield, it serves as a ground plane for the striplines. The sheet-metal screws are located at the end of each strip (on the cold end) and directly above the tuning capacitor (on the hot end). If you skimp on the number of screws you'll have problems with erratic tuning and low gain.

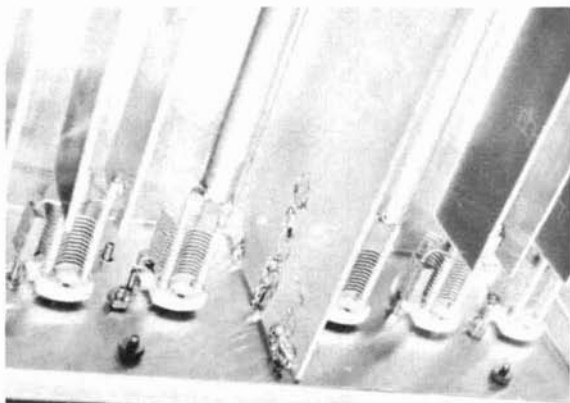
The striplines are centered between the top and bottom plates of the chassis. On the capacitor end of the resonator, the stripline is soldered to the capacitor stud with the copper foil facing in (all striplines are single-sided copper and mount on the outside of the studs so copper faces copper). At the cold end of the stripline brass grounding screws are located inside the strips directly opposite the capacitor studs. The copper-clad strips must run exactly parallel to each other or efficiency will drop.

The input and output ceramic capacitors must have short leads. Small holes drilled just above the center line on the input and output striplines allow you to feed the capacitor pigtail *through* the strip and solder it to the copper side.

The two shields are made from double-sided copper-clad printed-circuit board and are cut to form a snug fit top and bottom, with small clearance on either end. Brass 4-40 machine screws,  $\frac{1}{4}$ -inch long, are mounted in the chassis wall at either end of the shield, and the

shield is soldered to them. Make sure the brass screws are soldered to *both sides* of the shield. Small notches in the board allow the screws to seat snugly into the shield.

In the 50- to 100-MHz model of the filter/preamplifier an additional brass screw is installed through the chassis in the exact center of the shield. This holds the center of the long shield in place and helps to maintain the proper location of



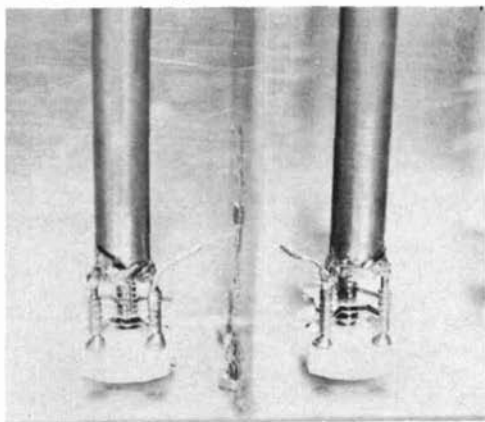
Mounting of center shield and striplines (50 to 100 MHz unit).

the shield in the long chassis.

Kepron printed-circuit board is available in pieces up to 12 inches square; one 12 x 12 inch section will furnish 8 strips, 12 x  $1\frac{1}{2}$  inches. For the 50- to 100-MHz filter, the required  $15\frac{1}{4}$  inch length is provided by splicing a  $3\frac{3}{4}$ -inch piece from another printed-circuit board. To bridge the splice joint, use thin strips of copper,  $1\frac{1}{4}$  inches long, sweating the printed-circuit board and copper strip together with a 40-watt soldering iron. Be sure the  $3\frac{3}{4}$ -inch section joins the 12-inch strip flush on so the resulting  $15\frac{1}{4}$ -inch strip has no bends or kinks.

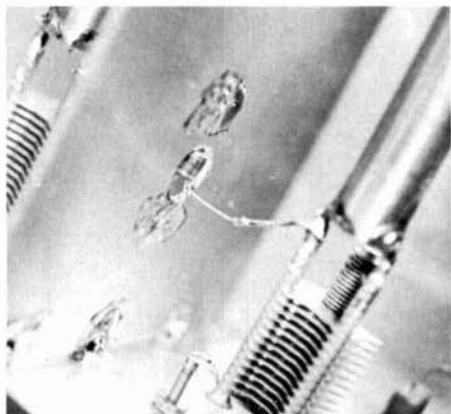
Striplines 3 and 4 are made from common copper tubing. Make sure the tubing is straight—you'll have to straighten the pieces you need carefully since it's usually sold in rolls. The hot end of the tubing fits nicely onto the capacitor studs. The cold end of the tubing is

Transistor installation in 100 to 200-MHz filter/preamplifier.



attached to the mounting stud (FT feedthrough capacitor) with a short loop of number 16 or 18 copper wire that is soldered across the end of the tubing; the wire is soldered to the FT capacitor.

The rotors of the Calectro variable capacitor are not automatically grounded when the capacitors are mounted, so a 1/2-inch 4-40 brass machine screw is used to solder the rotor tab to ground. (Calectro variable capacitors are commonly



Closeup of 2N5397 position in the center shield.

available at the Calectro displays found at most electronic parts distributors.)

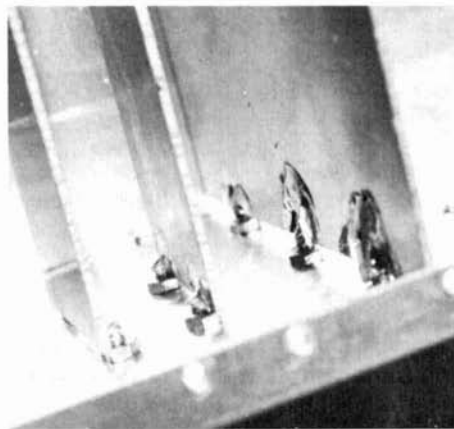
When you get down to actually building the filter/preamplifier it is recommended that you follow the step-by-step instructions given below.

1. Mark and drill all holes. First lay out the top plate, center-punch the holes and drill holes for no. 6 sheet-metal screws. Then put the top plate in position over the chassis and drill corresponding holes in the chassis lips using the top plate as a drill template.
2. Install all of the 4-40 brass screws.
3. Install both shield sections. (If you wait until after you install the other components you won't be able to get them into the chassis.)
4. Install the variable tuning capacitors, grounding the rotor tab to the 1/2-inch 4-40 brass screws provided for

this purpose. Install C8 and C9, the two FT-1000 feedthrough capacitors.

5. Install both coax connectors and the input and output coupling capacitors (C1 and C10). The stripline end is not connected yet.

6. Install the striplines—two per capacitor—starting with lines nearest the end of the chassis. Use a 25- to 40-watt soldering iron when installing the striplines because very little heat is required with the copper-clad strips. Solder the capacitor end first, holding the other end in position so that the weight of the stripline is not on the newly soldered capacitor stud. (If you mark the center lines of the strip you can align the centerline with the stud.) The cold ends of the striplines are held snugly by friction against the 4-40 grounding-screw nuts. Remember that for top performance the copper-clad strips must be parallel *and* centered between the chassis top and bottom ground planes.



Cold ends of striplines are soldered to 4-40 screws.

7. Now install striplines 3 and 4. You'll need a 150-watt iron (or more) to solder the copper tubing. Don't rest the weight of the tubing on the capacitor studs—block the tubing up until both ends are firmly in place.

8. Install R1—try 330 ohms to start—and install R2. Resistor R2 feeds through a small hole in the cold end of the chassis just under the FT bypass capacitor.

9. Solder the loose ends of the input and output coupling capacitors to striplines 1 and 6, pushing the pigtail through the small hole in the strip and soldering it to the copper foil.

10. Carefully install the field-effect transistor with the tab pointing toward the input side of the filter. Solder the case lead to the input side of the shield and the gate to the output side of the shield. The transistor leads can be preformed as shown in fig. 11; the source lead will point toward C4 (stripline 3), and the drain lead will point toward C5 (stripline 4). Short pigtailed of no. 18 copper wire are soldered to the variable capacitor studs, and then to the free drain and source leads. When soldering the transistor leads use heat sinks and no more than a 40-watt soldering iron.

11. Install the power supply decoupling capacitor C11 *outside* the chassis.

### construction tips

If you are interested in duplicating the units shown in this article, there are a few tips that are worth passing along. If you experience problems with gain, and are satisfied that you have used proper layout and construction, remember that output is affected by power supply decoupling, output circuit loading (the device won't work well if too heavily loaded), and saturation. Power gain drops rapidly when the transistor is saturated—the 1-dB gain-compression point at 450 MHz, for example, is coincident with 1-mW input.

In addition, any inductance in the gate lead will severely affect performance. This indicates that the shield between stripline 3 and 4 *must* be well grounded to the chassis. If the side of the shield where the gate lead is soldered is floating above ground it will introduce considerable in-

ductance to the gate circuit.

If any performance problems develop check your construction. The spacing between resonators, between the top and bottom plates, and the shield between striplines 2 and 4 *must* be uniform. The layout shown here was developed for optimum performance; if you deviate much, or use sloppy construction, gain will quickly drop off to 2 to 3 dB. Use extreme care when laying out the component mounting holes; variances of 1/16th inch will deteriorate performance.

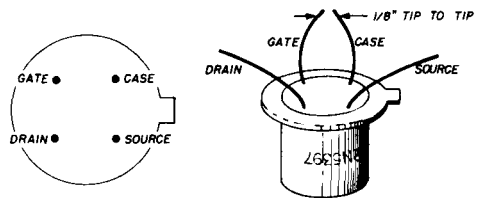


fig. 11. Method of pre-forming the leads to the 2N5397 fet before installing it into the preamplifier.

### tuneup

With the top plate removed from the chassis you can apply voltage to the unit. A meter in series with either the negative or positive power supply lead will allow you to monitor the current drain of the 2N5397. Correct current drain is 5 mA.\* If you have more than this with a 330-ohm resistor at R1, increase R1 to the next standard value. If you read less than 5 mA lower the value of R1 (rule of thumb only—variables between 2N5397 transistors can reverse this suggestion). Keep adjusting the value of R1 until current drain is as close to 5 mA as possible.

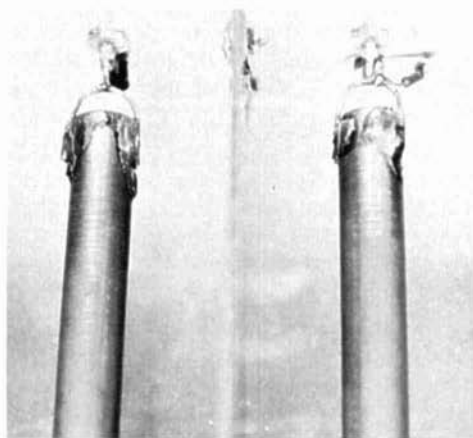
To tune or test the unit the top plate

\*Although this discussion is based on the 2N5397 fet, the less expensive 2N4416 can also be used in this preamplifier. However, the 2N4416 will have considerably higher noise figure, particularly above 100 MHz, and slightly lower gain than the 2N5397. The only change concerns the value of R1—it should be adjusted so that the current drain of the 2N4416 does not exceed 2.5 to 3 mA. The 2N4416 base diagram is the same as the 2N5397.



must be installed firmly in position with all of the sheet-metal screws. Connect a signal source (50 to 100  $\mu\text{V}$ ) to the input and a receiver or converter to the output. Slowly rotate C5 (stripline 4) for an indication of signal. At 50 MHz C5 will be about two-thirds meshed, on 144 MHz it will be half meshed and on 220 MHz it will be nearly wide open. If you can't find the signal (be sure your receiver is on the correct frequency) tune C3 (stripline 2) to the same position suggested for C5.

Once you find the signal, tune C2 and



The cold ends of tubular striplines 3 and 4 are soldered to FT-1000 feedthrough capacitors.

C7, then C6 and C4, in that order, for maximum signal. Peak the capacitors again, in the same sequence, starting with C5. (Crank down the output of the signal generator so you keep the S-meter in a more linear range.) With the unit tuned up for maximum gain, you will have an rf bandpass window similar to that shown in fig. 12. Fortunately this occurs at the same point as maximum gain.

Set the signal generator output so the S-meter reads S9. Disconnect the filter/preamplifier and connect the signal generator directly to the receiver. If the preamplifier is working properly—and the signal generator has a 50-ohm output—the S-meter should drop 10 to 14 dB, depending on the band you're working with.

## precautions

Although I haven't had any trouble with high-power rf from my transmitter, you don't want to blow out the \$8.50 transistor. Regardless of how you handle the potential problem—diodes, coax relay, etc.—do it *outside* the filter/preamplifier chassis. Any additional stray parts mounted inside the chassis only degrade performance by upsetting the balance of the tuned circuits inside the box.

## typical performance

Fig. 13A shows how a 10,000  $\mu\text{V}$  channel-4 television signal appeared on a tv receiver tuned to channel 3. All of the video information on the screen is from channel 4; overload from channel 4 to channel 3 is virtually complete. Fig. 13B shows the same receiver, still tuned to channel 3, with an interdigital-preamplifier ahead of the receiver. The weak snowy picture, all 0.9 microvolts of it, is from a channel-3 transmitter 153 miles away. The 10,000- $\mu\text{V}$  channel-4 signal has been completely eliminated. Fig. 13C shows the same signal passed through a pair of channel-3 interdigital-preamplifiers. With such a weak signal to work with there's not much hope for a snow-free picture. The signal input to the receiver at this point is 32  $\mu\text{V}$ . (In this photo a 20-dB post amplifier was installed between the tv set and the second

fig. 12. Bandpass curves for single and double interdigital filter/preamplifiers.

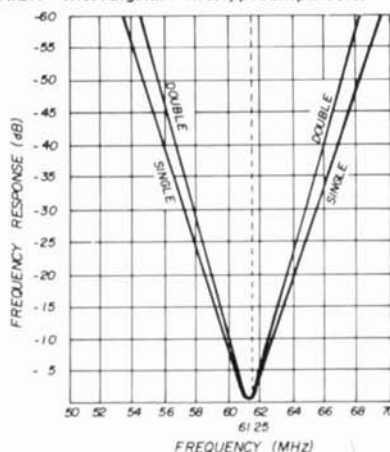




fig. 13. Adjacent-channel performance of the interdigital filter/preamplifier. In A the channel-3 programming is completely obliterated by channel-4 signal. With one interdigital filter/preamplifier in the line, channel-4 interference is eliminated (B). Two filter/preamplifiers improve the quality of the channel-3 signal (C).

interdigital-preamplifier.)

Fig. 14 shows the performance of the unit as a preamplifier without adjacent-channel interference. Fig. 14A shows an off-the-air signal from a channel-12 transmitter 170 miles away. If you look closely you may be able to see the frame bar. Fig. 14B shows the same signal with a single interdigital-preamplifier ahead of the receiver: now we know there's a signal present. Adding another interdigital-preamplifier results in the picture shown in fig. 14C.

It should be noted that the television receiver used for these photographs is a special catv model, the Conrac AV-12E. Comparing this receiver to the typical set you have in your home is like comparing a 75A4 to an S-38!

#### sweep tests

The bandpass characteristics of the interdigital-preamplifier are shown in the

oscilloscope displays of fig. 15. For these displays the output from a Jerrold 601D sweep generator, sweeping from 15 to 100 MHz, is connected through a 1-dB step attenuator to the input of an interdigital-preamplifier tuned to 61.25 MHz. The display is 10 dB from the baseline (top) to the bottom. The 62.25-MHz marker (1 MHz off resonance) on the right is 5 dB down the curve. Fig. 15B shows the same 62.25-MHz marker after passing through a pair of interdigital-preamplifiers; it is now 8 dB down the curve.

Fig. 15C shows the same curve with the marker moved down to 61.75 MHz, 500 kHz off the resonance point of the interdigital-preamplifier; it is 3 dB down the curve. The 61.75-MHz marker moves 6 dB down the curve when a second interdigital-preamplifier is added to the line (fig. 15D). From these measurements you can plot the bandpass curves of single

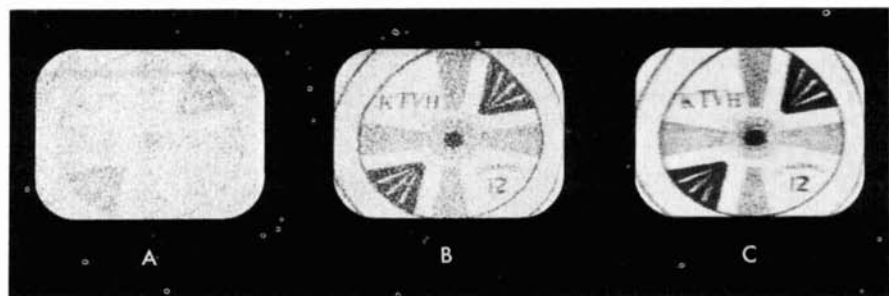


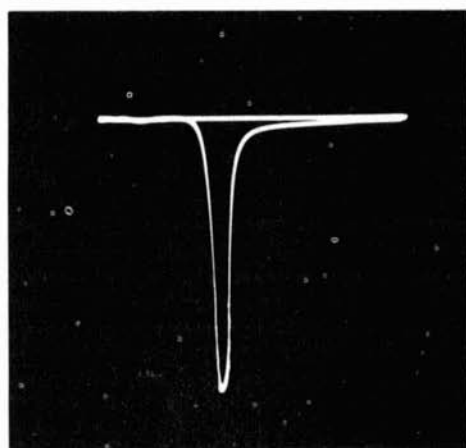
fig. 14. On-the-air performance of the filter/preamplifier. The channel-12 signal in A without the preamplifier in the line is almost covered with snow. One filter/preamplifier results in the signal shown in B. Second preamplifier ahead of the tv set results in the signal shown in C.



A. With single stage 1-MHz off-resonance marker is 5 dB down.



B. With two stages the 1-MHz marker is 8 dB down.



C. 500-kHz off-resonance marker is 3 dB down with 1 stage.



D. 500-kHz off-resonance marker is 6 dB down with 2 stages.

fig. 15. Bandpass characteristics of the filter/preamplifier.

and double interdigital-preamplifiers as shown in fig. 12.

### acknowledgements

I would like to acknowledge the cooperation I received from Charlie Williams of Siliconix, and Harold Cobbs of the W. Pat Fralia Company for their assistance in developing the 2N5397 parameters for this application. I am also indebted to Jay Liebman, W5ORH, who observed, "... state of the art is strange... Just when I had my two-meter preamplifier and converter down to a microminiature package I have to build a new two-meter

preamplifier that is bigger and takes more rack space than my two-meter kilowatt final!"

### references

1. Reed E. Fisher, W2CQH, "Interdigital Bandpass Filters for Amateur VHF/UHF Applications," *QST*, March, 1968, p. 32.
2. Reed E. Fisher, W2CQH, "Combine VHF Bandpass Filters," *QST*, December, 1968, p. 44.
3. Kenneth E. Holladay, K6HCP, "Super Two-Meter Preamp," *ham radio*, November, 1969, p. 72.
4. J. B. Compton, "High-Frequency Characterization and Application of Junction FETs," Siliconix Applications Note.

ham radio

# practical vxo design

An interesting  
approach to  
frequency stability  
in oscillator  
circuits

Gus Gerke, K6BIJ, Box 143, Weimar, California 95736

You're on the air having an enjoyable conversation. You switch over to the other station and the fellow says, "Sorry, missed most of that. Someone drifted onto your frequency." Sound familiar? The "someone" is usually a combination of unstable vfos and receiver drift.

The drifting signals one hears today suggest that vfo stability isn't really as good as claimed by equipment manufacturers and authors of vfo articles in the ham magazines. The best answer I've found to this problem is the variable-frequency crystal oscillator, or vxo.

The vxo circuits described in this article combine the flexibility (within limits) of a vfo with the inherent stability of crystal frequency control. Frequency can be varied between 2-720 kHz depending on the crystal frequency and other considerations, which I'll discuss. Many amateurs I've talked to never heard of varying a crystal's frequency over such a wide range.

Very little information has been written about the vxo. One article<sup>1</sup> describes a circuit that can pull down the frequency of an 8-MHz crystal about 4-5 kHz before the circuit becomes "a rather inferior vfo." With this circuit (fig. 1) as a starting point, I designed the circuits of fig. 2 and 3 using FT-241 crystals in the 450-kHz region and the circuit of fig. 4 using 3.5-8.5 MHz crystals.

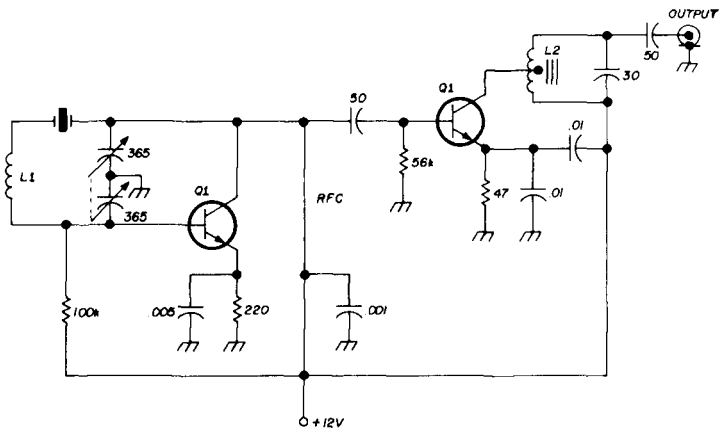
## circuit development

The vxo shown in fig. 2 is a modification I made to a BC-604 fm tank transmitter. The vxo output goes through a stage of amplification and several frequency multipliers to obtain output on 21 MHz. I've used this vxo on 7 and 21 MHz cw with excellent results. The circuit has also been used to operate a 2-meter transmitter. Eight crystals were needed to cover the entire 2-meter band.

The only addition to the BC-604 was L1, C1. Capacitor C1 is used to pad the crystal frequency over a certain range, in this case 2 kHz. With an increase in padding range, the effects of temperature, vibration, and hand capacitance become

more pronounced; and the same precautions in building vfos must be used. These effects are small, however, and the crystal is still the frequency-controlling element. If you don't exceed the padding

the BC-604 are less than 2 kHz apart, continuous coverage to the next lower-frequency crystal is possible. Stable 2-kHz padding was obtained with the circuit of **fig. 3**.



L1 16-24  $\mu$ H for 8-9 MHz crystal (J. W. Miller 4507)

Q1 2N706, 2N2219, 2N3662 or RCA 40237

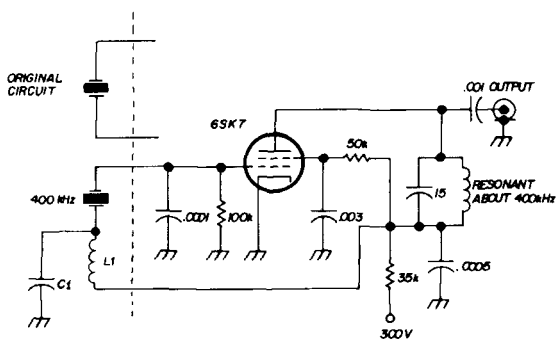
L2 40 turns no. 36, tapped at 16 turns

fig. 1. Circuit described in reference 1. An excursion of 4-5 kHz is claimed for an 8-MHz crystal.

range, the vxo won't become an "inferior vfo."

The circuit of **fig. 3** seems to work well with the same low-frequency crystals used in the vxo of **fig. 2**. The solid-state version shown was also used with the BC-604. Since the crystals furnished with

A transistor vxo that produces stable 50-kHz padding is shown in **fig. 4**. This vxo can also be used with a crystal in the 8-MHz region for 6- or 2-meter operation. Doubling will produce a padding range of 100 kHz on 14 MHz, 150 kHz on 21 MHz with tripling, and 200 kHz on 28 MHz with quadrupling. To cover the entire 2-meter band, you'll need 8 crystals (500-kHz padding range).



C1 Broadcast radio variable with both sections in parallel

L1 Broadcast variloopstick antenna or similar

fig. 2. Oscillator modification made to a BC-604 transmitter using low-frequency crystal.

**Table 1** gives recommended padding ranges for the FT-241 crystals when used in the circuits of **figs. 1** through **3**. If you're interested in a particular frequency range (as for net operation), try to use a crystal that will cover the first 25 percent of the padding range—then you'll have crystal stability.

The transistor circuits will start oscillating with 2.4 V; for more output, up to 12 V can be used. Unless followed by a frequency-multiplier, a buffer amplifier will be needed, as in **fig. 1**.

### a vxo for exciter use

Suppose you want to design a vxo

covering the entire 40-meter band, and you have an exciter such as the Central Electronics 20A using a 9-MHz crystal.

Higher than 9-MHz injection frequency is preferred to avoid unwanted mixer products. Therefore the injection fre-

quency will be from  $7 + 9 = 16$  MHz to  $7.3 + 9 = 16.3$  MHz. Crystals in this range are overtone types and won't operate in these circuits. The solution is to use an 8.150-MHz crystal and operate it on its second harmonic, 16.3 MHz. Padding 50 kHz on the crystal fundamental frequency will produce 100-kHz shift in the output. This will give you full coverage of the 7-MHz phone band. An 8.1-MHz crystal will cover the next 100 kHz, and another crystal at 8.05 MHz will extend coverage to 7 MHz.

Crystals with frequencies of 8.125 and 8.075 MHz will be useful if you want extra stability and don't wish to pad

### tuning capacitor considerations

Referring to fig. 5, capacitor C1 is

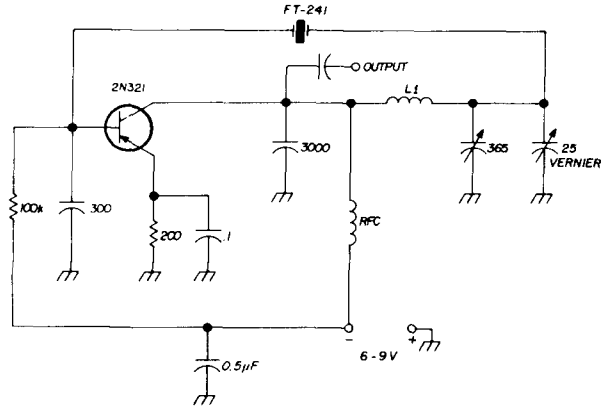


fig. 3. Solid-state version of the vxo in fig. 2.

used to bring the crystal frequency within the range of C2. Both capacitors should have a straight-line frequency response as a function of angle of rotation of the rotor plates. This capacitor characteristic is important for vxo calibration and tuning. For example, the tuning capacitors shown in the circuits of figs. 1 through 4 are common broadcast-band variables. When these are used, frequency decreases slowly at first as the capacitor rotor is turned. Then the frequency change becomes faster, until finally a hairline change in rotor position will produce a 1-kHz jump. This, of course, is very inconvenient at the lower frequen-

L1 40 turns no. 32 closewound on 3/8" slug-tuned coil form

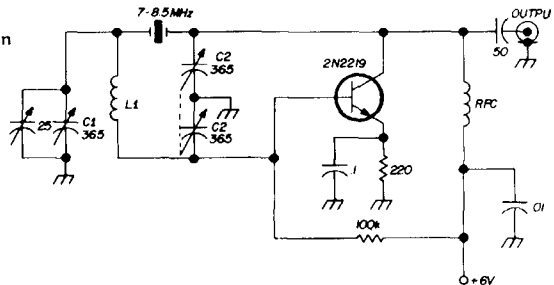


fig. 4. Solid-state vxo that produces stable 50-kHz padding on 7 MHz. It can be used for 6 or 2 meters also.

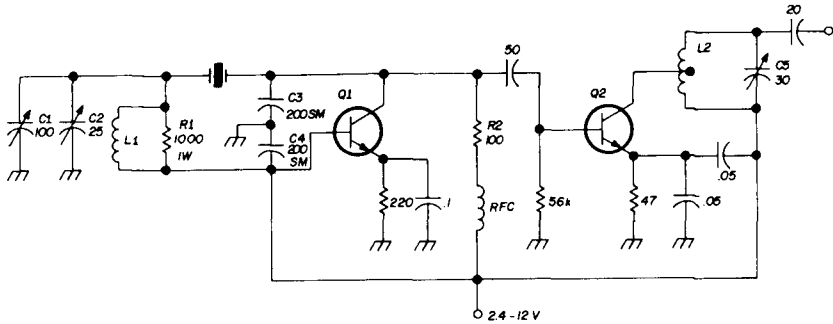
cies. The sketch of **fig. 6** illustrates the geometrical relationship of the stator plates in these two versions of variable capacitors.

In the circuit of **fig. 5**, capacitor C2 should be of good quality, otherwise

Battery voltage may be 2.4-12 V. Higher voltage may result in drift due to heating. I use 6 volts in my vxo.

### in closing

As far as I know, the vxo designs



**fig. 5.** Vxo-doubler circuit for a typical exciter. Three crystals are required for full coverage of the 40-meter band.

contact-scraping noise will be heard in the receiver; small jumps in frequency may also occur. A capacitor with an insulated rotor is recommended for C2.

### circuit description

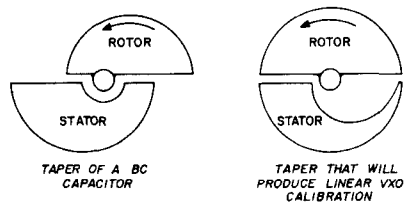
The purpose of R1 in **fig. 5** is to lower the Q of L1. This allows a larger padding range and more stable operation near the low end of the range. If the frequency changes when touching the rf choke, the choke is too small. Resistor R2 prevents oscillation at the rf-choke resonant frequency.

Use a two-section bc variable capacitor to find the exact value of C3 and C4. Then replace the bc capacitor with two silver micas. A value of 200 pF seems right for this circuit.

**table 1.** Padding ranges for various crystal frequencies.

	padding range	
	fig. 1 (kHz)	figs. 2 & 3 (kHz)
FT-241		
0.45 (fundamental)	0.2000	2.00
4.00 (9th harmonic)	2.00	20.00
8.00 (18th harmonic)	4.00	40.00
144.00 (324th harmonic)	72.00	720.00

described in this article have never been published before. The circuit for the 20A exciter has been used on 40 and 15 meters in both the cw and ssb mode. All reports were crystal quality, and all operators asked for the circuit diagrams; so

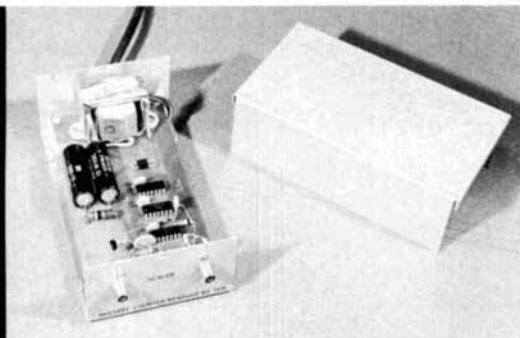


**fig. 6.** Mechanical configuration of straight-line wavelength capacitor used for bc band and a straight-line frequency capacitor.

I've presented them here to share with others. My old vfo has since drifted into the junk box.

### reference

1. J. R. Fisk, W1DTY, "73 Useful Transistor Circuits," 73, March, 1967, p. 20A.



## divide-by-ten frequency scaler

An accessory  
that will increase  
the range  
of your  
frequency counter  
by a factor of ten

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

It's easy to extend the range of your frequency counter with this scaler. Four inexpensive JK flip-flops and a simple power supply comprise a circuit that will divide input frequency by ten. Thus, frequencies beyond the range of most home-built counters can be scaled down to a frequency the counter can handle. The basic measurement accuracy of the counter won't be affected by the scaler. If your counter has an upper measurement frequency limit of, say, 100 kHz, this scaler will extend its range to 1 MHz.

The maximum range of the scaler will depend on the IC's and construction techniques. The model shown here has an upper frequency limit of 106 MHz. Therefore, to realize the maximum potential of the scaler, your counter should operate to at least 10 MHz.

### applications

A frequency measuring system with a maximum upper range of 100 MHz will cover all amateur bands through six meters; however, the instrument isn't limited to these bands. Two-meter transmitters, for example, can be checked at a lower-frequency stage.

While exploring its possibilities, I used this scaler to check transmitters operating in the 900-MHz region by picking up rf at a frequency-multiplier stage. Transmitters operating between 144 and 160 MHz were also checked similarly. An insulated wire probe inserted near the plate coil was used to obtain energy for the scaler. Very little power is needed, and care should be taken not to overcouple the circuit. The IC's have a high-impedance input, and only 0.8 volt p-p is required to toggle them.

### circuit description

The scaler schematic (fig. 1) is a standard circuit for a clocked counter with binary-coded decimal output for decoding. The bcd output is incidental.



It's not used in the scaler, but merely means that the scaler is compatible with other IC's for decimal readout.

The IC's are from Motorola's MECL II family (meaning emitter-coupled logic). The MC1013P and MC1027P are identical, except the latter has lower internal

such as the MC770P, for those interested in using an MECL as the first stage of a counter.

### construction

The main objectives in building the scaler were wide frequency response, sta-

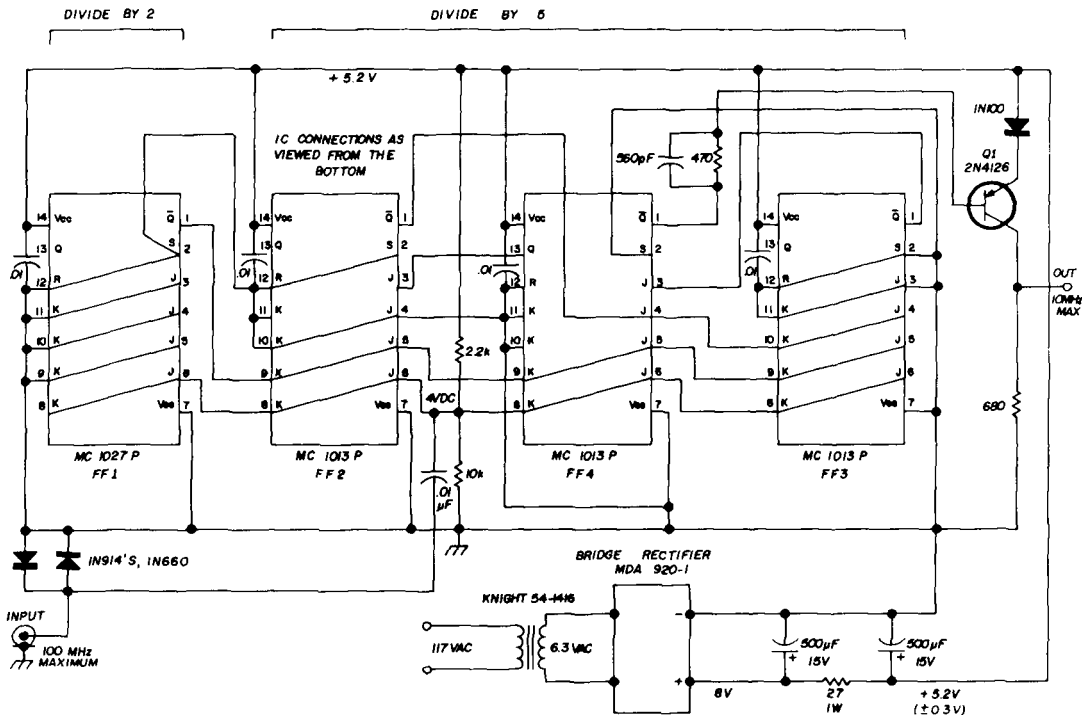


fig. 1. Schematic of divide-by-ten scaler and power supply. Unit delivers 4 volts p-p to drive a digital frequency counter.

resistance values, twice the current drain, and a higher operating frequency. I used the MC1027P in the divide-by-two section as a compromise between cost and performance. You can use all MC1013P's or all MC1027P's, since pin connections and logic are identical. Motorola technical data<sup>1</sup> may be consulted for details.

An amplifier stage, Q1, increases scaler output to 4 volts p-p. It will also translate the MECL logic level to saturated logic levels, so four identical circuits would be useful as an inexpensive interface between the bcd output and a resistor transistor logic (RTL) decimal decoder,

bility, simplicity and low cost. I tried several other circuits using linear IC's, but found these JK's could be driven directly without gate drivers. The elimination of a preamplifier simplifies the power supply and reduces cost.

You are urged to use the board layout as shown in fig. 2, with no changes. Other layout arrangements will work, of course, but I arrived at this one after etching several circuit boards, and I know it will work.

At 100 MHz short leads or circuit trails are imperative, and bypassing is equally important. Note the 0.01-pF min-

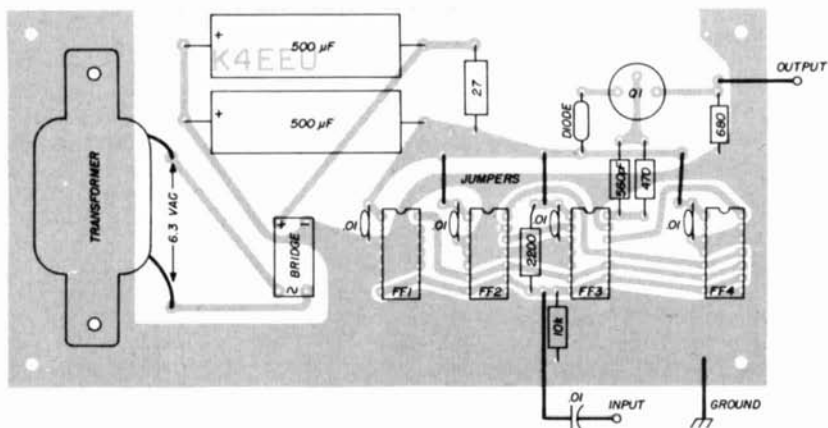
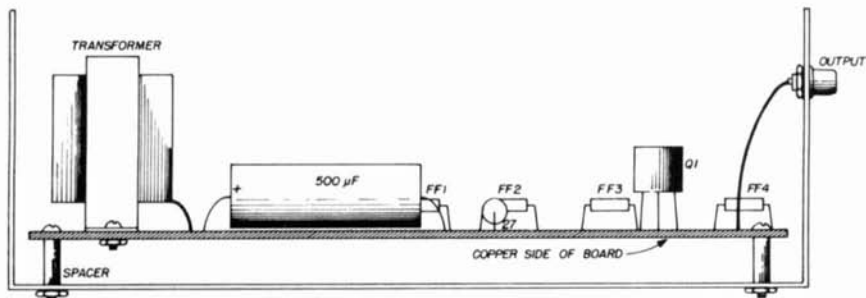


fig. 2. Parts layout. Enclosure is an LMB 138 box 6¼ x 3½ x 2-1/8 inches. Components are shown as viewed from the bottom of the board. Circuit boards are available from Stafford Electronics, 427 South Benbow Road, Greensboro, North Carolina 27401; order HR-7, \$2.50.

ature discs across the  $V_{CC}$  terminals of each IC. Also it's important that power-supply output be close to 5.2 volts. The voltage at the junction of the 2.2k and 10k resistors must be 4.0 volts. Change the value of the 27-ohm filter resistor or the 2.2k divider until you obtain these voltages within 10%.

### wiring

When wiring the IC's be sure they're properly oriented and solder is flowed around each pin. Use a magnifying glass to examine each soldered connection to make certain that no pins are shorted. Use a 22-watt iron on the connection only long enough to ensure a good joint. A hotter iron can be used if you use some kind of device to transfer excess heat from the connection. Spring-loaded heat-sink tools are available for this purpose,

or you can make an acceptable substitute from a strip of aluminum. Most device manufacturers provide maximum pin temperatures, and it's well to heed their recommendations.

If, for some reason, you must remove an IC you'll need something to remove the molten solder before the IC can be removed without damage. The best way is to use a vacuum device, such as a Sol-dapult,\* to pick up molten solder. I've used this device when removing hundreds of solid-state devices and have yet to damage any from heat.

A complete parts list is provided in table 1. A conservative estimate for the total cost is \$25.00—not bad for a simple circuit that will increase the range of your counter by a factor of ten.

\*Edsyn, Inc., 15954 Arminta Street, Van Nuys, California 91406.

## operation

When the scaler is completed, check the power supply and examine the scaler circuit carefully. The JK flip-flops seem to operate with any input from 1 MHz up, so the scaler can be checked at a low frequency with a signal generator and inexpensive oscilloscope. If you obtain a signal at the output jack, the scaler is probably working all right. If not, it's easy to signal trace with the scope, progressing through flip-flops 1 through 4. Output should be found at pins 1 and 13 of each IC, at about 0.8 volt p-p. At the active pins (those not grounded or at  $V_{cc}$ ), you should be able to measure 4.0 volts dc.

If scope checks lead you to suspect an IC, you can substitute another device to see if anything changes. With such a simple circuit, it's unlikely you'll have trouble; however the advice given above is included "just in case."

The two back-to-back diodes across the input provide some overload protection, but it's possible to zap all the IC's by applying a very high-level input signal. Therefore, with the scaler connected to the counter, increase scaler input until the counter suddenly starts reading; that's

table 1. Parts list.

1	3 x 6-inch printed circuit board, copper one side, etched
1	enclosure, LMB 138 6¼ x 3½ x 2-1/8 inches or equivalent
4	spacers, 3/8 to ½ inch long
1	power transformer, Knight 54-1416 6.3V @ .6A (Allied Electronics)
1	single-phase bridge, or 4 diode equivalent (Motorola) MDA920-1
2	500 $\mu$ F @ 15V (CD 500/15)
1	27 ohm 1 watt resistor
1	2.2k ½ W resistor
1	10k ½ W resistor
1	680 ohm ½ W resistor
1	470 ohm ½ W resistor
1	560 pF disc capacitor
5	0.01 $\mu$ F 100V disc ceramic or equivalent
2	1N914 high speed diodes
1	1N100 diode
1	2N4126 pnp silicon transistor
1	integrated circuit (Motorola MC1027P)
3	integrated circuits (Motorola MC1013P)
2	phono chassis connector jacks

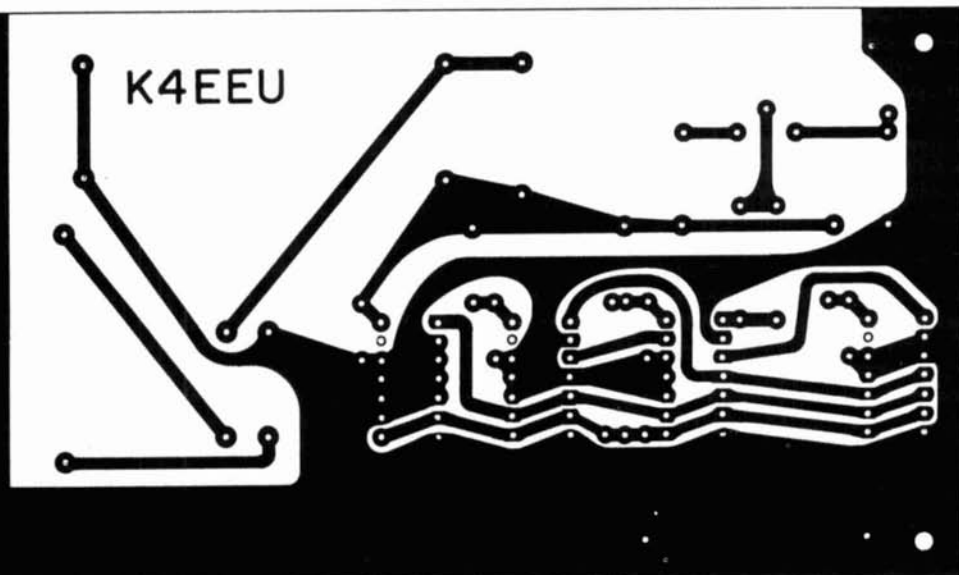
all the input signal necessary. A one-turn loop at the end of a short length of shielded cable makes a good coupling device.

## reference

1. "The Integrated Circuit Handbook," Motorola Semiconductor Products, Inc., P.O. Box 20912, Phoenix, Arizona 85306  
Also, Application Notes AN257, AN224, and AN277.

ham radio

Circuit board for 10:1 frequency scaler.



# computer-aided circuit analysis

A powerful tool  
that eliminates  
trial and error  
in circuit design

The number of unusual tasks that can be more accurately and economically performed by a computer than by manual methods is steadily increasing. One of the more useful applications of the computer is the automatic analysis of electronic circuits.

A number of different circuit analysis programs are available. One of the first and still popular programs, developed by IBM, is ECAP (Electronic Circuit Analysis Program).<sup>1,2</sup> This program is widely used by circuit designers and has been adapted for use with machines other than those made by IBM.

## features

ECAP can be used for dc, ac, or transient analysis by making slight changes to the input data. One of the more useful features is an automatic frequency response analysis, with machine plotting of the waveshape. This is done by inputting the proper data for an ac or transient analysis and asking for an output plot.

As you would expect, a simple dc analysis is less complex than an ac or transient analysis of the same circuit. The dc analysis program for ECAP provides the steady-state solution for linear circuits composed of resistors, fixed current sources, fixed voltages, and dependent current sources. All other circuit components must be replaced with the equivalent circuit formed by these elements. The program also prints out diagnostics to inform the user of input errors and gives suggested remedies.

## application

To see how the program works, let's make a dc analysis of a simple two-stage transistor amplifier. We start with the schematic shown in **fig. 1**. The first step is to convert the schematic into a dc equivalent circuit so we can describe the circuit in the language of the computer.

It's necessary to follow a few simple rules to draw the equivalent dc circuit correctly. Capacitors must be replaced

M. A. Ellis, K10RV, 61 Marlboro Road, Sudbury, Massachusetts 01776

with an open circuit and inductors with a short circuit, since this is how they appear to dc. The new circuit will consist of a network of junctions (nodes) and branches, each of which will have an input and output terminal. We call a node any point where two or more branches meet. (They are generally referred to as voltage nodes, since voltage calculations are made for these points as part of the mathematical solution.) A branch must contain at least one passive element—in our case a resistor. The branch may also contain voltage or current sources, or a dependent current source.

Each node and branch must be numbered. Any order will do, but we must begin with number 1 and not skip any numbers. Ground will be called zero ( $V_0$ ). Current-flow direction must be selected and may be arbitrary except for dependent current sources (the transistor collector current) and their controlling branches (base current branch); then the direction of flow must be consistent.

For example, the dependent current

source in the transistor model must show current flowing from emitter to collector ( $h_{FE}I_B$ ) when controlling current ( $I_B$ ) flows from emitter to base. Refer to fig. 2, which illustrates the equivalent circuit model for the transistors in our circuit.

Fig. 3 shows the equivalent dc circuit, with the two transistors (T1,T2) replaced by their models. The capacitors and inductors have disappeared; and the voltage nodes,  $V_x$ , and current branches,  $B_x$ , have all been numbered.

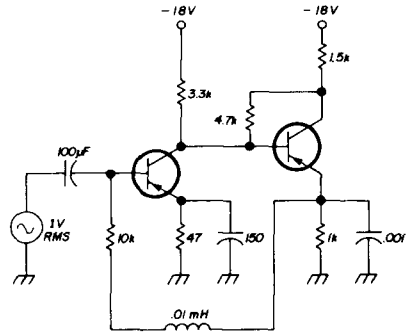


fig. 1. Two-stage amplifier used in the circuit analysis. Transistor gain (beta) = 50, output impedance = 50k, input impedance = 200 ohms.

Many readers will wonder why an article of this type is presented, since most amateurs just don't happen to have a digital computer handy. The fact is that the program described here is available through numerous software companies, who maintain remote terminals in many large cities linked by telephone line to a central computer facility. The computer is used by these companies and others on a time-sharing basis; thus cost to a terminal subscriber is prorated. Libraries of programs for many scientific and business-oriented problems are available.

The analysis program described here was developed for the IBM 1620 computer about eight years ago. It's only recently, however, that the cost of using the program has been within the means of the average user by time sharing. Typically, the cost of running this program would be about \$15.00 an hour, including terminal occupancy and computing time. Within ten years or so this will be substantially lower, so you could have a time-share terminal in your home for perhaps the same price as a telephone subscription. **editor.**

## program analysis

We now start writing our input listing as pictured in table 1. We bypass the usual preliminary instructions to the computer concerning items not important here. "DC" (first line) indicates a steady-state dc analysis request. The second line tells the computer that in branch 1, current flows from node 2 to node 1 through a resistance of 200 ohms, and the resistor is connected to a battery voltage of -0.5 volt. (This represents the base-to-emitter voltage.) The third line describes current flow in branch 2 from node 0 (ground) to node 2 (emitter terminal) through a resistance of 47 ohms. Line four shows that, in branch 3, current flows in the dependent-current source from node 2 (the emitter-collector junction) to node 3 (the collector terminal), through a resistance of 50k ohms. E3 in computer language means multiply by  $10^3$ ; i.e. three decimal places

table 1. Computer input listing for the electronic circuit analysis program.

```

DC
B1 N(2,1),R=200,E=-0.5
B2 N(0,2),R=47
B3 N(2,3),R=50E3
B4 N(3,0),R=3.3E3,E=18
B5 N(1,4),R=10E3
B6 N(4,3),R=200,E=-0.5
B7 N(5,3),R=4.7E3
B8 N(0,4),R=1E3
B9 N(4,5),R=50E3
B10 N(5,0),R=1.5E3,E=18
T1 B(1,3),BETA=50
T2 B(6,9),BETA=50
PRINT,NV,BP
EXECUTE
    
```

should be added to the number 50. Line five depicts current flow from node 3 to node 0 through a resistance of 3.3k ohms and a battery voltage of 18 volts.

If the battery polarities seem confusing, remember that current flow from negative to positive through the battery represents an *increase* in voltage. Therefore  $E = 18$  is a positive voltage. The opposite occurs in branch 1 when emitter-base current represents a decrease in voltage; hence, the battery is shown with current flowing into the positive terminal and out the negative terminal. Consequently the battery is represented by  $E = -0.5$  volt.

The remaining branch descriptions can be read down through the line beginning B10. Line T1 shows that the transistor currents flow through branches 1 and 3, and that the transistor current gain is 50. In this line, the controlling branch, B1, must be listed first followed by the

dependent branch, B3. The line beginning T2 is read in the same way.

We now tell the computer that we want it to print out node voltages and branch power losses by the PRINT, NV, BP line. The word EXECUTE, which follows, tells it to perform the calculations and print out the results.

## output

The printout is shown in table 2. Node voltages are printed first, with the node numbers shown at the left. The voltages for nodes 1 through 4 are listed on the

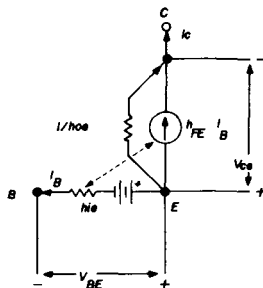


fig. 2. Equivalent circuit model for the transistors in fig. 1. The dashed arrow indicates the interdependence of the emitter-collector branch with the emitter-base branch.

first line and that for node 5 on the second. Note that the E 01 listed after the value for node voltage 3 means that the decimal point is to be moved one place to the right, indicating -2.64+volts. For node 5, E 02 states that we should move the decimal two places to the right.

table 2. Printout giving node voltages and branch power losses.

node voltages	voltages			
nodes				
1- 4	-0.83987243E-00	-0.31386716E-00	-0.26482825E-01	-0.21401336E-01
5- 5	-0.11750809E-02			
element power losses	power losses			
branches				
1- 4	0.33813746E-05	0.20960125E-02	0.15285775E-01	0.71416734E-01
5- 8	0.16906791E-03	0.33202385E-06	0.17628932E-01	0.45801718E-02
9-10	0.21426438E-01	0.26034927E-01		

Branch power losses are listed next in the same order as the node voltages. E-05 trailing the number for branch 1 tells us to move the decimal point 5 places to the left for a 0.00000338+ watt loss. Branch 2 shows a power loss of 0.002+ watt.

## modeling

You've probably guessed that the greatest difficulty with circuit analysis (or design) via the computer is coming up with realistic models for the active components, such as tubes and semiconductors. A good model allows useful results

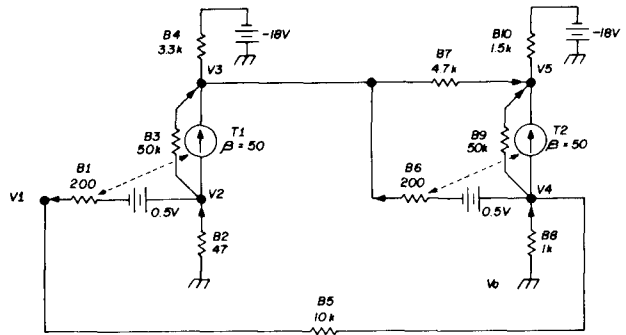
node voltages. Obviously, these two outputs would be valuable to the circuit designer by eliminating many hours of manual calculations and the need for breadboarding.

## conclusions

Computer-aided circuit analysis or design is a powerful tool for the engineer or technician. With it he can design on paper, check out the circuit, and modify and recheck without breadboarding.

Those who have access to a computer may find it worthwhile to try their hand

fig. 3. Complete model of the two-stage amplifier with all the voltage nodes identified for the program input listing.



to be obtained, whereas a poor model will give questionable results or none at all. Almost as many different basic models are available as circuit analysis programs. Books have been written on modeling (references 3 through 6), therefore no attempt is made to go into that subject.

## other outputs

The circuit we've examined is simple. We have only asked for simple outputs. We could have asked for more profound information, such as voltage sensitivities or a worst-case analysis. Sensitivities refer to whatever change may occur in a node voltage for a one-percent change in the branch parameter. This allows the designer to take precautions and use closer tolerances with components that have the greatest effect on circuit operation. A worst-case analysis means to sum all the positive (or negative) tolerances in a parameter and compute the resulting

at circuit analysis with ECAP or one of the other available programs. Anyone with a good understanding of electronic circuits can become proficient in computer-aided design and analysis techniques after a few practice sessions.

## references

1. "IBM 1620 Electronic Circuit Analysis Program (ECAP)," document 1620-EE-02X, User's Manual, IBM, Inc., New York.
2. R. W. Jensen and M. D. Lieberman, "IBM Electronic Circuit Analysis Program," Prentice-Hall, New York.
3. J. F. Pierce, "Semiconductor Junction Devices," Charles E. Merrill Books, Inc., Columbus, Ohio.
4. "Transistor Manual," 7th edition, 1964, General Electric Co., pp. 52, 53; 69-77.
5. F. F. Kuo and W. G. Magnuson, Jr., "Computer Oriented Circuit Design," Prentice-Hall, New York.
6. "The Semiconductor Data Book," 4th edition, 1969, Motorola Semiconductor Products, Inc., pp. AN-3 to AN-15.

ham radio

# a tunable audio filter for cw

A selective audio filter can really help to pull cw signals out of heavy interference. What will help more is a tunable unit that will allow you to select the tone you want.

This circuit uses two cascaded Raytheon RM709 linear operational amplifiers\* in an active filter that can be tuned while still maintaining essentially constant bandwidth at the 3-dB points of its response curve.<sup>1</sup>

## design

Fig. 1 shows the filter response at 1000 Hz using the circuit of fig. 2. Gain at center frequency is approximately zero dB (gain of one), and the tuning range is from 750 to about 1600 Hz. The 3-dB bandwidth is 140 Hz.

The gain of one means that when the filter is switched in, audio blasting won't occur at the tuned frequency. Maximum input signal is about 5 volts before clipping occurs at the output.

## power supply

The ICs require a dual-polarity power supply. My supply makes use of the trick known as "zero" or "common" reference to the IC.<sup>2</sup> A single 9-volt transistor battery with two resistors is used, as shown in fig. 3. While this supply is adequate, slightly higher voltages will allow the filter to handle larger input signals.

## construction

Careful parts layout and the usual construction practices for assembling and wiring integrated circuits are a must for

\* Directly replaceable with the Motorola MC1709. Both devices are packaged in a TO-99 case. Editor.

this filter. The ICs have high gain and wide bandwidth. Short leads and bypassing at IC terminals will ensure against internal oscillation, which could destroy the devices. Overall shielding, as well as input-lead shielding, will keep transmitter rf out of the filter. The parts can be mounted on a small perforated board, which can then be installed in a Minibox or similar enclosure.

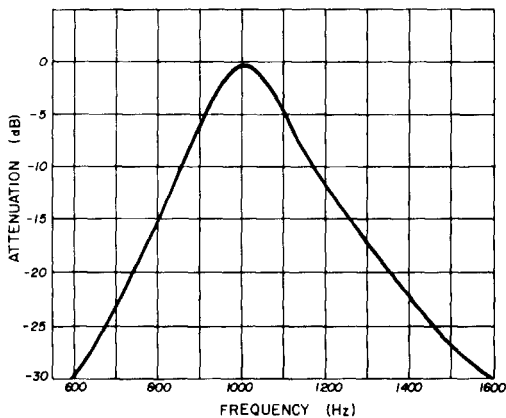
## operation

After checking filter and power supply wiring, set the filter switch to OUT, insert the filter phone plug into your receiver phone jack, and connect your headphones to the filter output. Tune in a cw signal tone of your choice, then switch the filter to IN and adjust the filter tuning for maximum volume. Select a tone within the filter tuning range. With practice this will be easy.

## performance

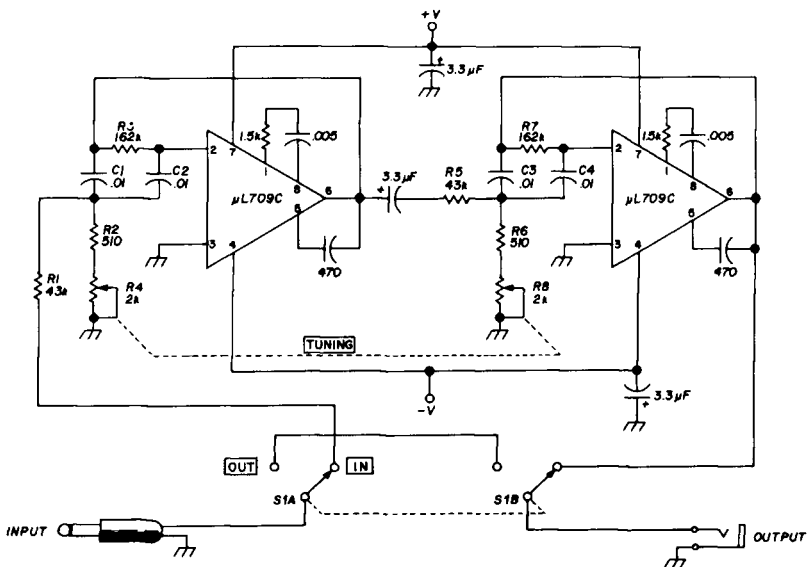
Although the filter skirts are not the best I have seen, the unit performs remarkably well. If your ssb transceiver

fig. 1. Filter response with a center frequency of 1000 Hz.



N. J. Nicosia, WA1JSM, 85 North Street, North Reading, Massachusetts, 01864





- C1, C2, .01 μF, matched to 5% or better  
 C3, C4  
 R1, R5 43k, matched to 2% or better  
 R2, R6 510 ohms, matched to 2% or better

- R3, R7 162k, matched to 2% or better  
 R4, R8 2k ganged potentiometers (IRC no. PQ11-110 and M11-110)

fig. 2. Schematic of the tunable audio filter. Circuit features selective audio tone for cw reception.

has a fixed receiver bandwidth (approximately 2.5 kHz) you'll be able to receive cw with considerable ease with this filter. The filter tuning control literally scans the receiver passband and picks out the wanted tone. Using the filter with my Swan 500C has made many enjoyable cw sessions possible.

### possibilities

Because the filter can handle high-level signals, it's not restricted to headphone use. With some thought it could be used for speaker operation.

For higher gain, a different audio range, or change in bandwidth, the component values can be adjusted in accordance with the following formulas:

$$R_a = \frac{1}{2\pi\Delta GC}$$

$$R_b = \frac{1}{\pi\Delta C}$$

$$R_c = 2\pi C (2f^2/\Delta) - \Delta G$$

where

$\Delta$  = desired 3-dB bandwidth per stage

G = nominal stage gain at center frequency\*

C = C1, C2, C3, C4

f = midrange frequency

R<sub>a</sub> = R1, R5

R<sub>b</sub> = R3, R7

R<sub>c</sub> = R2 + R4; R6 + R8

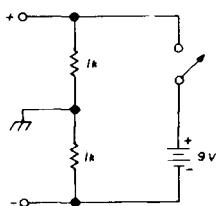
\*Total filter gain = gain product of stage 1 and stage 2.

### references

1. Leslie Robinson, "Active Bandpass Filter with Adjustable Center Frequency and Constant Bandwidth," *EEE*, February, 1968, p. 124.
2. John J. Schultz, W2EEY, "Using Integrated circuits with Single-Polarity Power Supplies," *ham radio*, September, 1969, p. 35.

ham radio

fig. 3. Transistor-battery power supply that provides dual-polarity output for the filter.





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Vackar oscillator

C. E. Galbreath, W3QBO, 8326 Still Spring Court, Bethesda, Maryland

Amateurs who have built and used very low power solid-state transmitters have marvelled at their ability to carry on solid contacts over normal propagation distances, even on 80 and 40 meters. Most of these transistor transmitters, however, are designed around crystal-controlled oscillators. The limitation of operating within the frequencies of one's supply of crystals is not relished by most amateurs accustomed to moving freely over the bands. In this age of the vfo, crystal control of operating frequency represents a step backward in amateur communications practice.

Having been bitten by the transistor transmitter bug but not wanting to be crystal bound, I began searching the literature for solid-state vfo construction articles. Fortunately there are a number of helpful articles on various transistorized vfo circuits (see especially reference 1), and I will not discuss these circuits here. Unfortunately a dearth of *construction* articles exists for the amateur wanting to build a stable, easy-to-

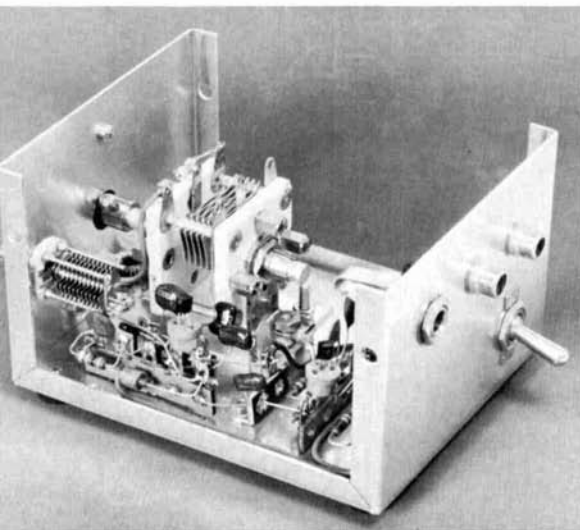
construct solid-state vfo. Most amateurs do not have the time or technical knowledge to design, work out the construction details, and conduct the tests necessary to build such a unit. These roadblocks limit the interest of many amateurs in transistor cw work. This article will help to fill a gap in the literature and provide an interesting construction project.

The vfo is easy to reproduce, produces a high-quality chirpless note when keyed, and is relatively inexpensive to build, even when all parts are purchased new.

Two vfos were constructed, one for 80 meters and one for 40. The 80-meter unit, however, can be used on 40 if a doubler stage is used in the transmitter. For those wishing to construct a 40-meter unit, appropriate oscillator component values are given. All other construction details are the same for both. In fact, the circuit can be used from 10 through 160 meters with appropriate changes in values of the frequency-controlling components.

The circuit is uncomplicated and easy

**Stable high-frequency vfo uses a double-bearing Millen variable capacitor.**



to follow (fig. 1). It uses a Vackar oscillator with a toroidal inductor, a buffer stage for isolating the oscillator and minimizing the oscillator load, and a second buffer for further isolation as well as impedance matching. Inexpensive jfets, MPF102s, are used in the oscillator and in the first buffer stage, and a bipolar transistor, an HEP 55, is used in the impedance-matching stage.\* The HEP 55, operating in the common-collector configuration, matches the high impedance output of the MPF102 buffer to the low-impedance input of the transmitter.

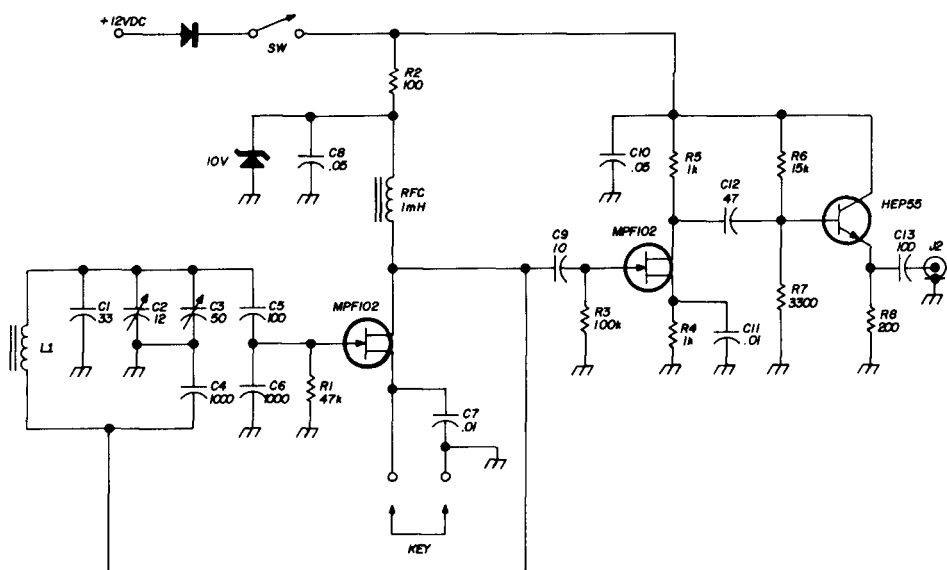
### construction

The vfo is constructed in a 3 x 4 x 5-inch aluminum minibox. Rubber mounting feet on the minibox help reduce mechanical shock and vibration. Also, a heavy aluminum plate is used to reinforce and strengthen the base of the minibox. The main tuning capacitor—a double bearing type—and the toroidal inductor are mounted on a 1½ x 2¾-inch platform of plexiglas or polystyrene supported on spacers on the base of the minibox. This method of mounting the capacitor and coil minimizes vibration and increases stability. It is an improvement over the conventional practice of direct mounting on a heavy metal base. As someone has pointed out, heavy metal, as in a bell, vibrates vigorously when even lightly tapped. This is not true of plexiglas or polystyrene. The oscillator coil is wound on an Amidon T-50-2 toroid core and needs no shielding. It is mounted on a 1 x 1-inch piece of perforated board and is held firmly by coil dope and short leads soldered to lugs bolted to the board. The perforated board, in turn, is bolted to a small standoff insulator mounted on the plexiglas platform be-

\*The MPF102 is listed at 90 cents and HEP 55 at \$1.20 in the current Allied Radio Company catalog.

tween the main tuning capacitor and the front of the minibox. The other oscillator components are mounted on terminal strips along one side of the plexiglas platform. A 10-volt zener provides excel-

Transistor sockets are used in preference to soldering the transistor leads into the circuit. This not only avoids possible damage to the transistors from the heat of the soldering iron, but also permits



**L1** (80 meters) 48 turns no. 30 enamelled on 1/2" toroid core (Amidon T-50-2)

(40 meters) 25 turns no. 30 enamelled on 1/2" toroid core (Amidon T-50-2)

**C1** 33 pF N750 negative temperature coefficient ceramic

**C3** (80 meters) Millen 23050MKF or 19050

(40 meters) Millen 19025

**RFC** Millen J300-1000

For 40-meter operation C4 and C6 should be 680 pF

fig. 1. Schematic for solid-state vfo and buffer stages.

lent voltage regulation for the oscillator.

All components in each buffer stage are mounted on a single five-tie-point terminal strip, center terminal grounded. Each buffer stage is assembled as a unit before attaching to the chassis. Copper-clad circuit board is not recommended. Not only is it difficult and time-consuming to use, but it offers no performance or constructional advantage over conventional terminal-strip wiring in this application. In fact, circuit boards frequently cause ground-loops and other problems.<sup>2</sup>

easy substitution of transistors if and when necessary. The latter feature paid off for me, as I found a weak transistor in the circuit when I first tested the vfo. In mounting each socket, two of its terminals are soldered to the appropriate tie points on the terminal strip.

The need for temperature compensation to reduce frequency drift of the oscillator is minimal so long as the vfo is kept away from heat-generating equipment. A 33 pF N750 negative-temperature-coefficient capacitor is sufficient to take care of small temperature-induced

effects.

Note that a subminiature iron-core choke is used to provide the load for the drain of the oscillator transistor. The choke replaces a resistor that was first tried. I could not get the circuit to oscillate on a 12-volt supply using the load resistor because of the voltage drop. The low value of the inductance (1 mH) of the choke not only provides low dc resistance but also the proper load impedance for the MPF102.

For adjusting the main tuning capacitor, an inexpensive 2-inch Japanese-made vernier dial is used. This works very well. Band edge is adjusted by a 12 pF variable capacitor on the front panel; a slotted shaft for screwdriver adjustment is recommended, as this capacitor rarely needs attention. The bandspread over the main tuning dial is excellent.

On the back of the minibox are the key jack, two phono jacks (one for rf output and one for battery or power supply leads) and an on-off switch for B+. To avoid damage to the transistors from reversed polarity, a semiconductor diode is inserted in series with the B+ lead between the B+ jack and the on-off switch.

The value of capacitor C12 coupling the vfo to the transmitter will depend on the amount of drive required for the transmitter and the input impedance of the crystal oscillator stage, whose crystal the vfo now replaces. Although the schematic shows a 100 pF coupling capacitor, the value might well vary from 50 to 300 pF.

## checking and testing

Before connecting the vfo to its power supply and before inserting the three transistors, use your vom to check for any possible short between the B+ line at the "on" side of the switch and the chassis. Still without inserting the transistors, connect the power supply to J1. Check with the vom for correct voltage between the "off" side of the switch and chassis. If the voltage is not correct, the diode is either inserted backwards or is

defective. Next, insert an MPF102 in the oscillator-stage socket and connect a key to the key jack. Turn on the switch and, while keying, listen for the oscillator signal in your receiver. With vfo tuning dial set at 90 you should find the signal toward the low end of the cw portion of the 80-meter band; the exact spot will depend on the setting of C2.

When you have ascertained that the oscillator stage is functioning normally, turn off the switch and insert the other two transistors in their sockets. Also connect a 0–500 ohm or a 0–1000 ohm potentiometer or variable resistor as a load to the output at J2. Turn on the switch and with the key depressed measure the rectified output voltage, using a

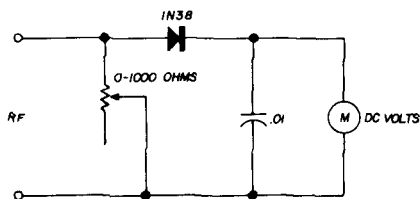


fig. 2. Test probe for checking output of vfo.

probe such as that in fig. 2. The output should peak about  $1\frac{1}{2}$  as the potentiometer is adjusted—probably around 400 ohms. One other voltage should also be checked: measure the regulated voltage at the zener; it should remain unchanged while keying.

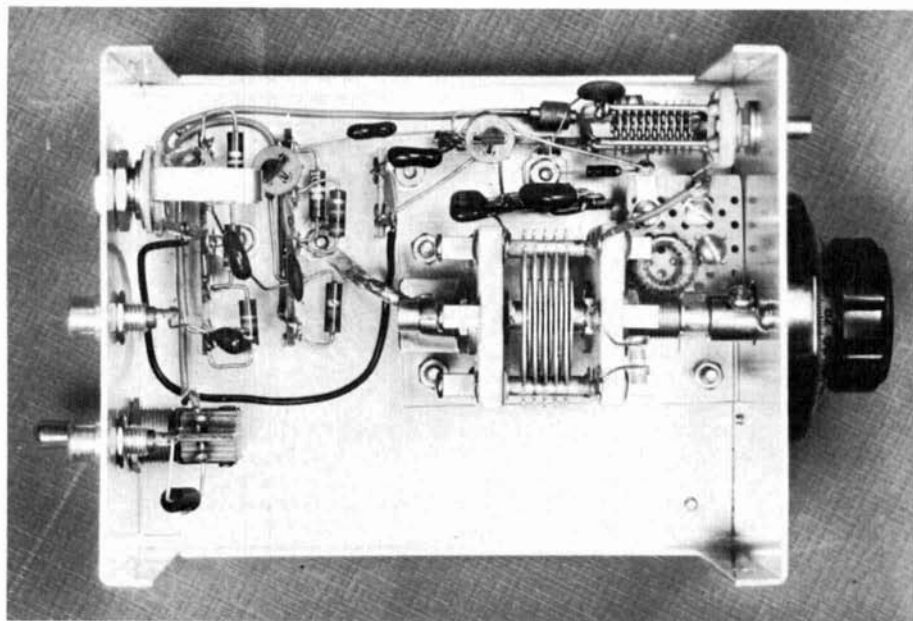
When these tests have been successfully completed, you are ready to connect the vfo to the transmitter.

## connection to transmitter

A short piece of coax or shielded wire can be used to connect the output of the vfo to the transmitter. The center of the coax is tied to the terminal of the crystal socket that is connected to the input of the oscillator (usually the base). Be certain that this connection is made correctly! Do not connect the other crystal socket terminal to the chassis unless it is already connected that way. The braid of the coax, of course, is tied to chassis

ground. It is recommended that a phono jack be installed in the transmitter for the connecting cable.

The only other adjustment is to set the low-band edge on the main tuning dial. This is done by turning the main tuning



Parts layout for the stable solid-state vfo.

You may have to change coupling capacitor C13 to obtain the required voltage at the input of your transmitter.

dial to 100 (full capacity) and adjusting C2 while checking with the crystal in your receiver.

The vfo is powered by a 12-volt lantern battery or eight D-type flashlight cells. Be sure to connect a 100- $\mu$ F electrolytic capacitor across the battery terminals.

I'd be happy to hear from those who build and use this vfo.

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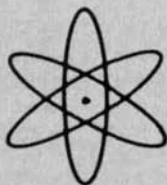
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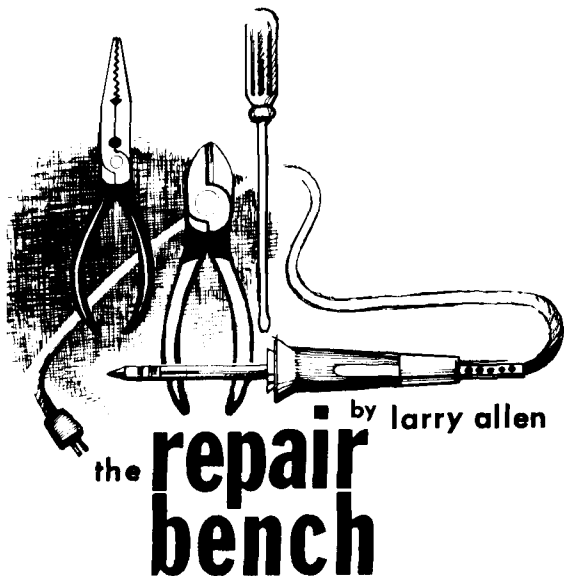
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by Larry Allen

# the repair bench

## how distortion creeps in

"Whaaatsch at cherr saoyynn naaiir, ohhhymm?" the speaker sputters. "Cauyynn unnderrschtaunnnd moy kewwayyssaooh?"

Heck no, you can't understand his QSO. He sounds like his mouth's full of spaghetti. What's happening is the speech amplifier in that fine new high-power transmitter has a gremlin. He's being bugged by distortion.

When distortion hits you in your receiver audio amps, you at least have a sporting chance. You can hear it. But audio distortion in your transmitter is embarrassing. Everyone else knows it before you do. (They do unless you're a lucky one who has a scope modulation monitor and watches it.) Whichever place it hits, the sooner you get at fixing it, the sooner you'll start enjoying your hobby again.

### the unlinear amp

Today's speech amps and audio amps are mostly transistor. A good percentage of the faults that happen pop up sudden-

ly. You're working along and—bang, there's trouble.

Not so, when the trouble is distortion. Garbled sound has a way of slipping up on you. At first the sounds you (or your buddies on the net) hear are only vaguely "not right." But they get steadily worse over the next few hours or days—sometimes over weeks or months. Then one day you wake up to how hard it is to understand anybody on your once-hot receiver. Or people ask you to repeat so often it's annoying.

Preventive maintenance, particularly in the transmitter, may help some. If you're running ssb, it's awful easy to have distortion just from misadjustment. So it's a good idea to check out the speech amps whenever you feel the urge to make sure your signals are shipshape.

The chief cause of audio (or speech, or voice, or sound, whichever you like to call it) distortion is an amplifier that has become nonlinear. A class-A amp should operate on the linear portion of the base-collector (or grid-plate) transfer characteristic (**fig. 1**). Let the bias shift, and operation moves onto one of the knees. The resulting distortion can be bad. See the sine-wave comparison in **fig. 2**.

Some speech drivers operate outside class A. They may run with bias nearer cutoff, like a class AB<sub>1</sub> (**fig. 1**). Major distortion is avoided there by carefully limiting the amplitude of input signal and by using transformer output coupling. A transformer helps restore the roundness of a sine waveform that has been flat-



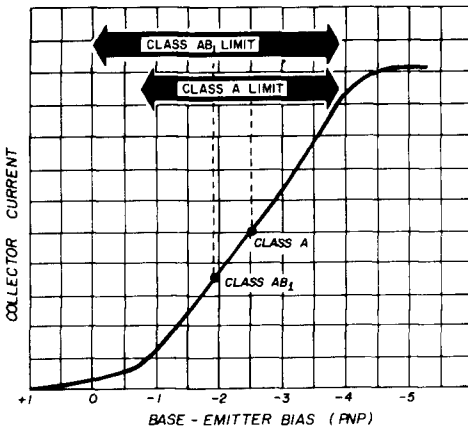
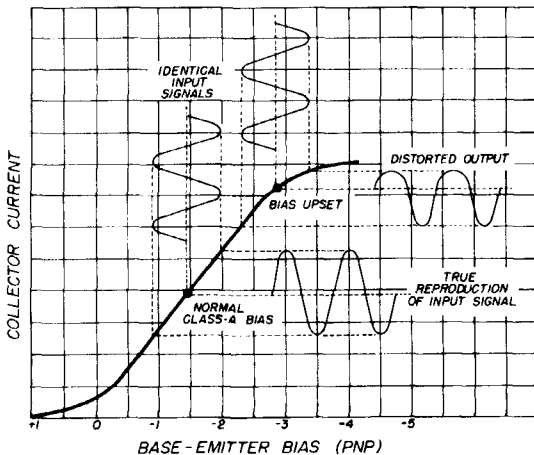


fig. 1. Base-collector transfer curve of a typical pnp transistor. Class-A operation is entirely linear. Class AB1 has bias nearer cutoff, but signal input is limited so it never actually reaches the cutoff or saturation portions of the curve.

tened at the knee of an amplifier curve.  
**the square test**

Your best instruments for distortion-chasing are the square-wave generator and oscilloscope. The units used for this article are illustrated in fig. 3. The generator is a kit-type and inexpensive—plenty okay for ham testing. (You can use it, and the techniques I'm about to describe,

fig. 2. A class-A amp distorts input waveforms if the bias shifts. Here, severe compression of the positive half-cycles (of the output) is caused by the curve's knee.



for hi-fi testing, too.)

Why not a use sine wave, you wonder? The reason is simple. It's hard to spot slight distortion in a sine wave. In a bad case, yes, the twisting or flattening is visible. But you just can't see subtle alterations except by superimposing input-output waveforms with a dual-beam scope or an electronic switcher. And that's more expense and complication.

So use a square wave. Feed a square wave into your receiver and transmitter voice stages and look at the waveforms you get with your own scope. Become familiar with them.

If you have a Polaroid camera, take pictures of the scope patterns when the equipment is operating normally. (Do it in the dark, and use ASA-3000 Polaroid film.) On the back of each picture, mark the brand and model of the equipment, what test point the scope is connected to, and the normal amplitude of the waveform. Keep the photos with the service manual and schematic.

There are two nominal square-wave frequencies for testing a communications audio amp. Bandpass of the stages is limited, remember. Low frequencies seldom go below 100 Hz, and highs seldom much above 3000.

A square wave stays square on the *trailing* edge if low frequencies are okay down to one-tenth of the fundamental frequency. Therefore, a 100-Hz waveform should stay square at the trailing (right-hand) corners if the stage it's going through has normal response down to 100 Hz. The normal input waveform in fig. 4A takes on a tilt like fig. 4B if the bass response is limited more than it should be.

On the other hand, the *leading* edge of the square wave tells about the high frequencies. The leading corner of the waveform stays square if frequencies are okay up to ten times the fundamental frequency. With a 3000-Hz upper limit in communications gear, a 1000-Hz square wave would show rounding at the leading edge—even in a normal amplifier. You can see this rounding in fig. 4C.

What you do is move the generator frequency to 300 Hz. The *leading* edge of the square wave is the one to watch. Just ignore the trailing edge, because it now tests down to 30 Hz—well below the amplifier's normal response. It'll be rounded. The waveform in **fig. 4D** shows the amplifier being tested is good to 3000 Hz. The leading edge is square.

### distortion on the square

A few unusual distortion waveforms are pictured in **fig. 5**. Two of them

tack" to set the stage oscillating. Voice signals were plenty sharp enough.

Bias distortion in an amplifier stage created the terrible waveform in **fig. 5C**. This one was a fooler. Tracing a square wave through the stages with a scope uncovered fairly quickly the driver that was distorting the signal. But the actual defect turned out to be a bad transistor in an earlier stage. Coupling was dc. The leaky transistor fouled up the bias on the driver, yet the scope showed the square wave going through the defective transis-



**fig. 3.** Scope and sine-square generator used to test the amplifiers for this article. Both are service-type instruments—not lab equipment.

illustrate faults that cause distortion but are not necessarily from a nonlinear amplifier. In their cases, the amplifiers were self-oscillating.

In **fig. 5A** you can see a parasitic oscillation modulated on the square wave. Also obvious, once you know to watch for it, is the rounding on the leading edges—the sign of poor high-frequency response. This speech-amp trouble was traced to a bad bypass capacitor. It was letting the stage oscillate and also overload on below-normal signal levels. You should have heard the crazy sound this transmitter produced!

The waveform in **fig. 5B** reflects an extreme case of self-oscillation. The square wave was triggering the oscillation, one burst for each cycle of square wave. Voice signals were unintelligible. Yet, a sine wave showed hardly any distortion. It took a waveshape with a sharp "at-

tor okay. It just happened that bias for the driver was more critical than in the earlier stage.

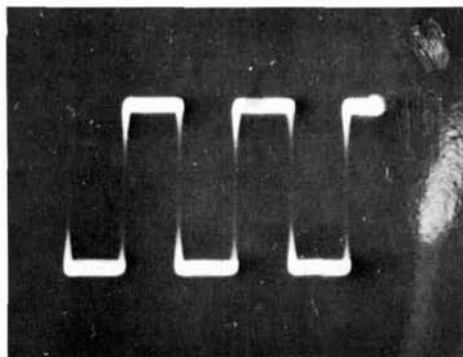
A bad capacitor created the severe overload waveform in **fig. 5D**. You could duplicate this waveform by simply feeding too much signal through any stage. But in this one the input signal was well below the natural overload level. The stage just couldn't handle much signal. A leaky coupling capacitor had drastically altered bias on the transistor.

You'll see plenty of other distortions as you test squawky-sounding amplifiers with square waves. You don't always have to recognize the trouble from the waveform, although practice will make you pretty good at doing that. What the generator and scope do is identify the stage where the distortion is. Then, you track the defect down by specific troubleshooting within the faulty stage.

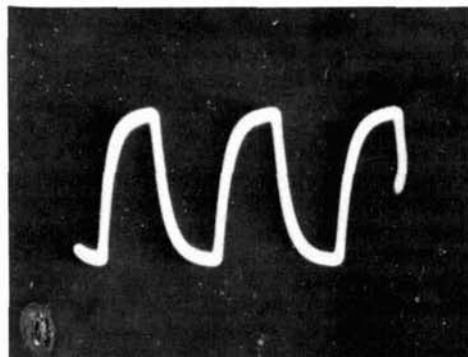
Of course, the end goal is to find the defective part. Once you've identified the stage that's bad, there aren't usually many parts to test. But it's also generally quicker to be scientific about it, rather than haphazard.

Using the typical speech amp in **fig. 6** as an example, I'll tell you the steps I go

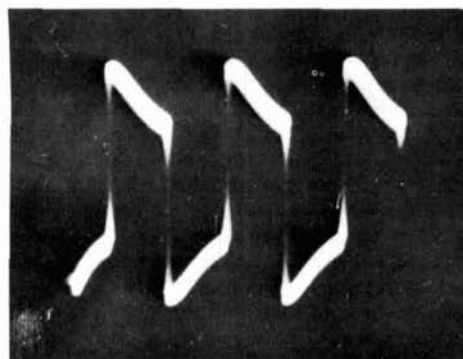
the mike in place. Say "ahhhh" into the mike at close range and notice the amplitude of scope trace. Then unplug the mike and substitute the generator. Turn up its output control until the scope display is the same height it was with the mike. The scope should show you a waveform like **fig. 4A**.



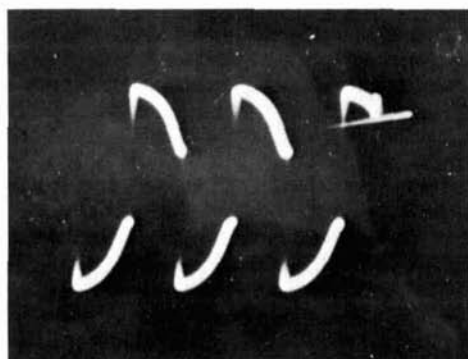
**A. Normal, generator directly to scope**



**C. Rounded leading edge, poor high response**



**B. Tilted trailing edge, poor low response**



**D. Normal, when 300-Hz square wave is used; amplifier doesn't go down to 30 Hz, hence downturned trailing edge**

**fig. 4. Square waves are informative for testing.**

through to find the bad part. It works no matter what the type of distortion, so just what the waveform looks like isn't that significant.

First disconnect the mike. Connect the square-wave generator in its place. Turn the generator up high enough to simulate the mike input. The average crystal or ceramic mike puts a millivolt or two into a high-impedance input.

If you want to be sure, connect the scope across the microphone input, with

Then move the scope from test point to test point. Keep in mind what I already told you about the frequency response of the speech amps. The further along you get, the more rolloff you find above 3000 Hz. It's especially pronounced after Q2 (C8 bypasses a lot of highs). Check both ends of the response; use 1000 Hz for the low end and 300 Hz for the high end. If that sounds strange, reread the paragraphs that describe "the square test."

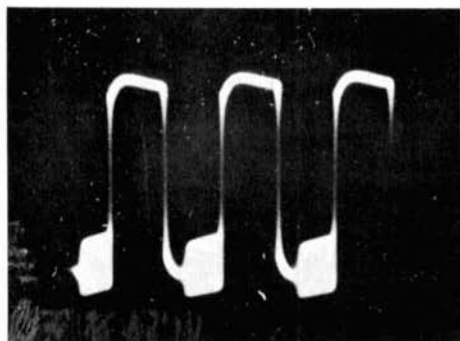
Make the tests in sequence. A bad waveform at A tells you R1 or C1 is in trouble. Trouble at B is most likely the fault of C2 or C3—or perhaps R2. However, a defect in Q1 could reflect back. For example, a base-emitter short could load down the waveform at point B.

A bad waveform at C points to Q1 or

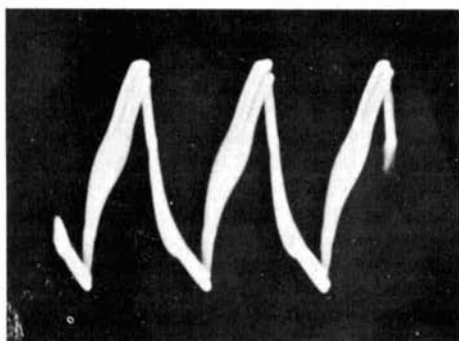
They don't impart any voltage gain to the square-wave signal. However, Q2 does, and you should see a considerable increase in amplitude between points E and F.

### botched-up bias

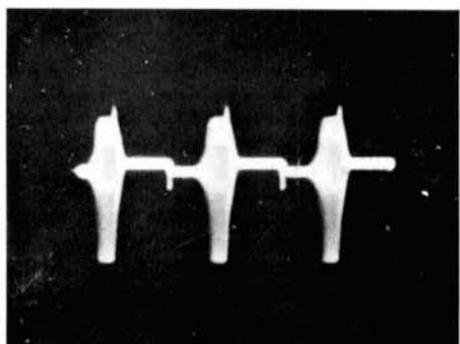
When you find trouble around a tran-



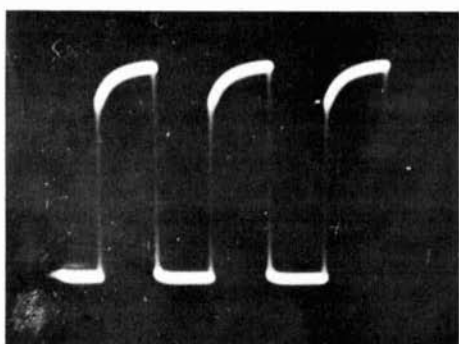
A. Parasitic oscillation



C. Distortion makes sawtooth



B. Oscillation, amplifier blocked



D. Overload distortion

fig. 5. Waveforms that reveal distortion.

C4. At D it could mean C5 or R5 is bad. Don't forget to try turning the shaft of R5, though, to make sure it's not just turned down.

And so on. Each test point tells you about the components connected there and those between there and the test point just preceding. You narrow down the fault to just a few parts, which you can then test individually without much wasted time.

Pay attention to the kinds of stages. You might be misled if you don't. Q1 and Q3 in fig. 6 are common-collector amplifiers (or emitter followers, if you prefer).

sistor, don't just jump in and replace it. Measure the dc voltages first. Distortion, as you read earlier, is commonly the fault of wrong bias on a transistor. True, it may be due to transistor leakage, but it could just as likely be a bad resistor or a leaky capacitor.

For example, suppose in a system like that diagramed in fig. 6 you touch your scope probe to point D and find the square wave still normal. At point E there's a tiny change but not enough to worry you yet. But the waveform at F is lousy. The trouble seems isolated between E and F. Is Q2 defective? That's

usually your first thought.

If you have a transistor tester, great. Use it, and save yourself a lot of measuring and figuring. But you probably haven't, so you go ahead with voltage measurement.

Collector voltage is wrong, but so is base voltage. Strangely, the base voltage changes as you turn modulation level control R5. That alone should tell you what's wrong. But maybe you don't happen to try turning R5. You check the two most likely parts: R6 and R7. Both measure okay with your ohmmeter. You disconnect the transistor, just in case the base is shorted to emitter or collector; normally there is hardly any base current to drag the base voltage down.

When you disconnect C6 to test it, the base voltage goes back to normal. The capacitor is very leaky. You can test it

the scope connected at the output (toward the balanced modulator). Just bridge each capacitor with a good one of the same value while you watch the square wave. If the waveshape changes, the one you bridged is open. Watch the corners, particularly at the leading edges. These bypasses affect high-frequency response.

An emitter bypass like C7 has more effect on low frequencies. With the scope connected as just mentioned, bridge the capacitor with a known good one. If the trailing corner of the square wave alters much, replace the capacitor. It's probably open. If it were leaky, it would change the emitter voltage drastically.

A sensible combination of bias checking and square-wave tracing should solve almost any distortion trouble for you. At least, it gets you so close to the root of

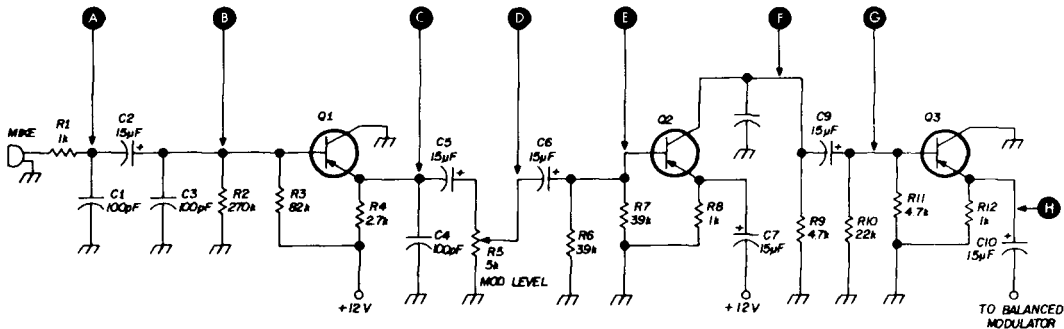


fig. 6. Points for square-wave testing in a typical speech amplifier. Voltage or substitution tests find the defective part.

with your ohmmeter, or by the "open end" test (disconnect the end nearest ground and measure for voltage leaking through from the base-bias circuit). The leakage didn't affect the waveform between points D and E, but it sure did play heck with the bias on Q2.

Watch out for bad bypass capacitors. They're hard to spot, especially if they're open. Capacitors like C1, C3, C4, and C8 show definite trouble when they short, or even if they get very leaky. The fault may be a little harder to spot if one of them is open.

The best way to check them is with

the trouble it leaves only a few parts to test.

### on the griddle

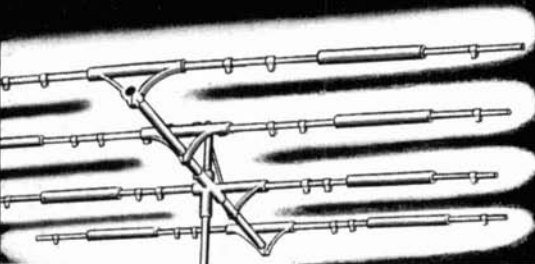
For my next column I've cooked up some solutions to a problem that aggravates even experienced professional servicers. Troubleshooting rf and i-f stages has peculiarities all its own. Yet, you'll be surprised how simply it can be done. All you need is a little advance knowledge of what to expect, and some hints for what tests to make. Both are in repair bench next time.

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The electrical principals employed in a trap type multi-band antenna are quite simple. The "trap" is a parallel resonant circuit consisting of a high Q inductance with a coaxial sleeve capacitor connected in parallel. At resonance the impedance across the trap is very high, and its effect is the same as inserting an insulator at that point. Thus, the electrical length of the antenna element can be altered by insertion of a parallel tuned trap. Each element in a Swan Triband Beam has two traps for 10 meters, and two traps for 15 meters. (None are required for 20 meters since the elements are full length on this band). It is vitally important that the traps are tuned to exact frequency, or antenna performance will not be optimum. Normal manufacturing tolerances in the coil and capacitor assembly create considerable variation in resonant frequency. Unless the trap has some means for precise adjustment, its resonant frequency will be pretty much a hit or miss affair. The exclusive precision tuned traps in Swan antennas\* explain why they give consistently superior performance.

**2000 watts P.E.P. power rating:** All models of the Swan Triband Beams are rated at the full legal power limit.

**Low SWR.** Swan Triband Beams are designed for a near perfect match on each band with 52 ohm coaxial cable. Standing wave ratio will be down as low as 1.2 at band center, resulting in extremely low transmission line loss. There are no gimmicks or gadgets in the feed system. It's very simple, and works very well. In fact, we offer no magic whatsoever in our antennas, except that they are designed to provide maximum performance on each band, and mechanically built to withstand severe weather conditions for a long long time. Ask any ham who owns one. The Swan antenna is an improved Hornet design, manufactured formerly by Hornet Antenna Products. Compare the Swan Tribander feature for feature with other brands. Ask your dealer to open up the shipping carton and let you inspect the Swan antenna. We think you'll like what you see, and like its performance even more.

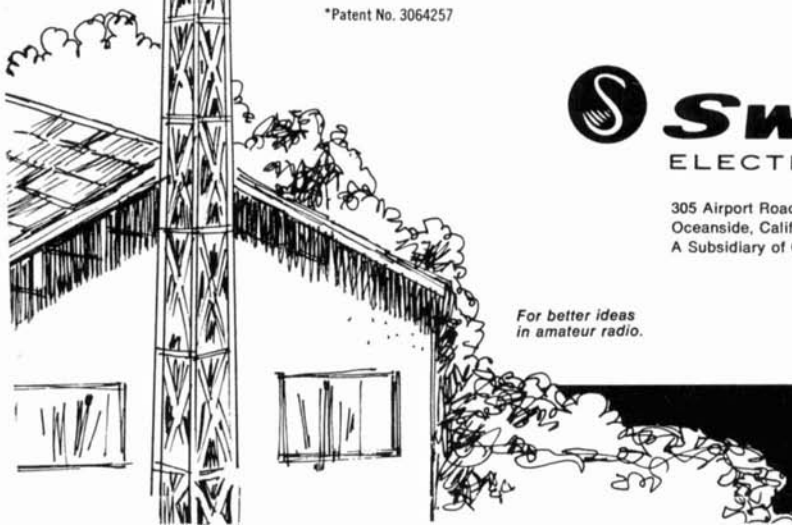
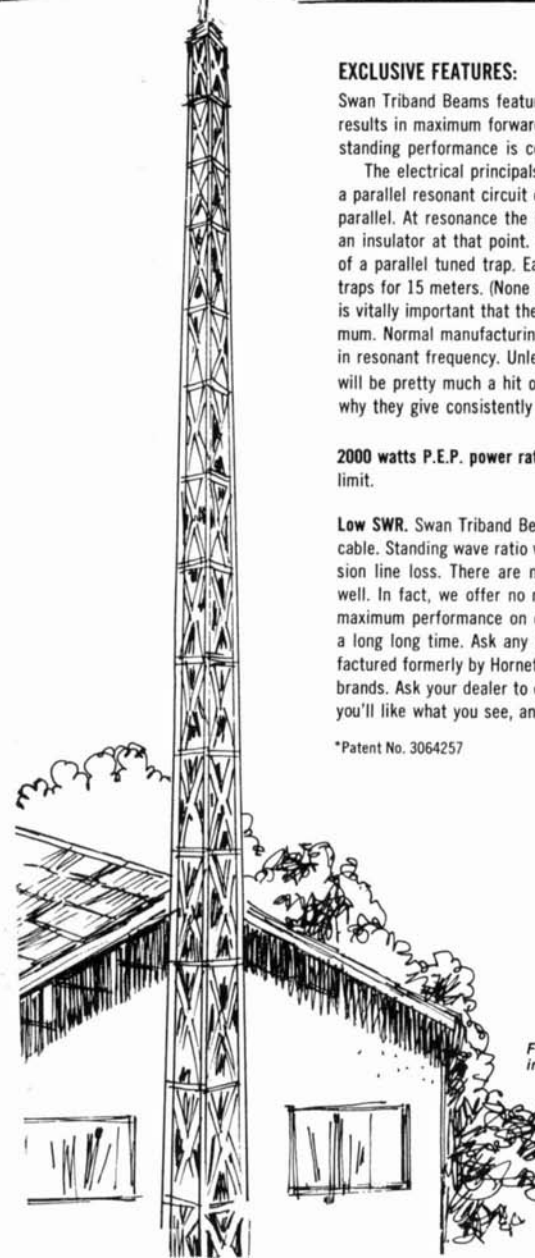
\*Patent No. 3064257



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patented \* tuneable traps

# TRIBAND ANTENNAS AND 20 METERS

## 4 ELEMENT MODEL TB-4H

The Swan 4 Element Heavy Duty Triband Beam gives you 4 working elements on each band: 10, 15, and 20 meters. A 24-foot boom permits optimum spacing for maximum forward gain and front-to-back ratio. All traps have been precision tuned and weather proofed. The Heavy Duty mechanical design of the TB-4H means that it will easily take winds up to 100 mph, and provide years of reliable service in any kind of climate from the arctic to the tropics.

- Forward Gain: 9 db average
- Front-to-Back Ratio: 24-26 db
- Power Rating: 2000 watts P.E.P.
- Weight: 64 lbs.
- Wind Load at 80 mph: 148 lbs.

Price: \$169

## 3 ELEMENT MODEL TB-3H

Same Heavy Duty design as the TB-4H, but with 3 elements on a 16-foot boom.

- Forward Gain: 8 db Average
- Front-to-Back Ratio: 20-22 db
- Power Rating: 2000 watts P.E.P.
- Weight: 44 lbs.
- Wind Load at 80 mph: 110 lbs.

Price: \$139

## 3 ELEMENT MODEL TB-3

Of somewhat lighter construction and shorter boom length, the TB-3 is adequate for many installations. Wind survival rating is 80 mph compared to the 100 mph rating of the TB-3H. Its lighter weight permits easier erection, a lighter weight tower, and lighter duty rotator, all resulting in lower overall cost. The same precision tuned, weather proofed traps are used, so power rating and reliability are the same as in heavier duty models. Boom length is 14 feet.

- Forward Gain: 7.5 db average.
- Front-to-Back Ratio: 18-20 db
- Power Rating: 2000 watts P.E.P.
- Weight: 37 lbs.
- Wind Load at 80 mph: 92 lbs.

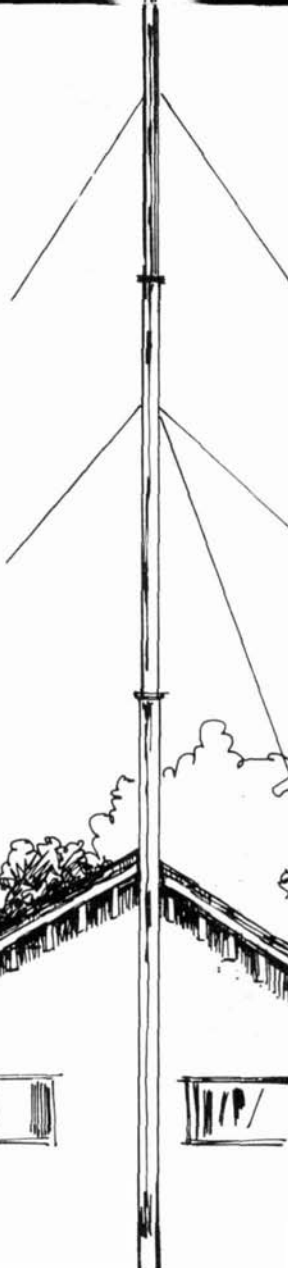
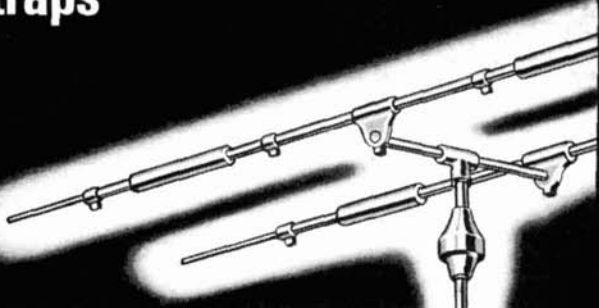
Price: \$119

## 2 ELEMENT MODEL TB-2

Same design as the TB-3 but with 2 elements on a 6½-foot aluminum boom. Weighing in at only 15 pounds, this model can be a real surprise. An inexpensive telescoping mast and TV rotator will easily get it 60 feet or higher off the ground, and at that height it will out perform a 3 or 4 element beam at lesser height. If your choice is putting up a TB-4H at a 30 to 40 foot height, or this 2 element model at 60 feet, by all means put up the TB-2. We won't make as much money, but you'll put out a terrific signal, and maybe we'll sell more TB-2's. Of course, if you can put the 3 or 4 element model up 60 feet, or more, there's no argument. Just don't under estimate the TB-2. It's a little bomb.

- Forward Gain: 5 db Average.
- Front-to-Back Ratio: 16-18 db
- Power Rating: 2000 watts P.E.P.
- Weight: 15 lbs.
- Wind Load at 80 mph: 60 lbs.
- Shipping Carton: Just 6 ft. long.

Price: \$99



# an improved six-meter converter

A new approach  
to portable  
vhf converters  
using fets  
and a tunable  
local oscillator

Rick Littlefield, K1BQT, 209 Marin Street, San Rafael, California 94901

Many converter circuits have appeared in print for those who wish to use an inexpensive transistor broadcast set in a solid-state receiver or transceiver. I've built a few of these converters, and in many cases they turned out to be a compromise between simplicity and performance. So I designed one that's simple but free of a few traditional flaws. The resulting circuit uses inexpensive fet's in rf and mixer stages and a Vackar tunable oscillator for frequency control (fig. 1). This approach represents an improvement over the new autodyne and crystal-controlled bipolar circuits.

## design

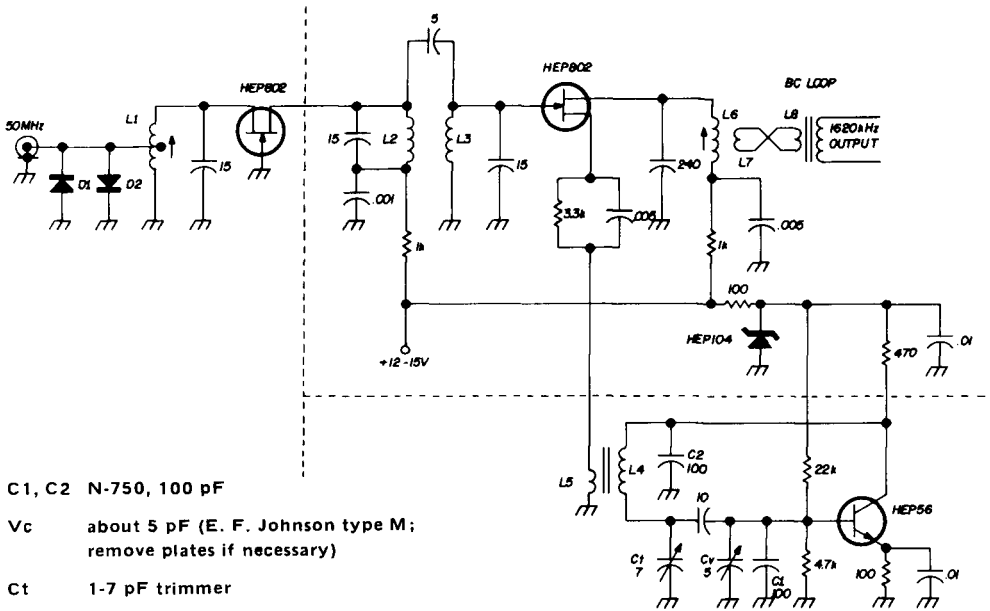
The tunable-oscillator-type converter offers the best design approach, because the transistor radio can be used to maximum advantage as a fixed-frequency i-f amplifier. By setting the bc receiver to a frequency just above the broadcast band, 1620 kHz, image rejection is held to the highest possible level over the six-meter band, and local broadcast-station feed-through is eliminated. Also the mixer output can be peaked with a high-Q circuit for higher gain. None of this is possible when the transistor radio has a tunable i-f.

The most obvious drawback to this approach is the requirement for a stable, tunable oscillator in the 50-MHz region. To solve this problem I chose a solid-state Vackar oscillator circuit. The version in



this converter uses an easily obtainable toroid inductor,\* and provides more stability than any simple circuit I know of. If good vfo construction practice and the prescribed temperature compensation are

weakest signals. The circuit is very simple to build and adjust, since it tunes quite broadly and is not prone to oscillation. If more gain is required, the optional amplifier shown in fig. 2 provides a hotter



C1, C2 N-750, 100 pF

Vc about 5 pF (E. F. Johnson type M; remove plates if necessary)

Ct 1-7 pF trimmer

L1, L2, 0.55-0.80 mH (Millen 69054-0.68)

L3

L4 9 turns no. 22 on T50-10 core (see text)

L5 2 or 3 turns test probe wire (depending on oscillator output) on cold end of L4

L6 20-50  $\mu$ H (Calectro D1-854)

L7 20 turns no. 30 scramble-wound slightly above L6 on same form

L8 5 or 6 turns no. 22 on cold end of bc loop antenna, connected to L7 with twisted hook-up wire

L9 10 turns no. 24 tapped 3 turns from gnd on 1/4" slug-tuned form (Miller 4500-4)

fig. 1. Schematic of the fet six-meter converter.

used, you'll have no trouble with instability.

The rf and mixer have a low noise figure and a much wider dynamic range than is possible with bipolars. Similar circuits can be found in the more recent editions of the ARRL handbook. The grounded-gate rf stage, while not an extremely high-gain circuit, does provide sufficient gain for reception of all but the

\*A kit of two toroid forms and wire is available from Circuit Specialists Co., P. O. Box 3047, Scottsdale, Arizona 85251 for \$1.25. Order: VFO-6.

front end. More care must be used in layout and tuning procedure, because the alternate circuit tunes quite sharply and requires neutralization. The ARRL handbook provides tuning procedures for grounded-source amplifiers, should you run into a problem.

The double-tuned circuit between the rf and mixer is a worthwhile investment. It improves out-of-band signal rejection. The mixer is a standard grounded-source circuit, with oscillator injection via a grounded link. The output is tuned slight-

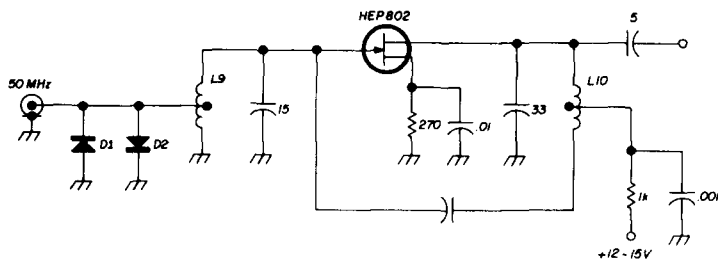
ly above the edge of the broadcast band and is fed to the transistor receiver loop antenna by a simple balanced line.

## construction

Component arrangement is somewhat flexible, depending on the size of your circuit board and your needs. My board measures a scant  $1\frac{3}{4} \times 3\frac{1}{4}$  inches; even this is too small unless you enjoy building ships in bottles.

generator should be used to tune the mixer, since the mixer pulls the oscillator frequency just enough to make signal-generator peaking a frustrating experience. I used a regenerative receiver, placed across the room as a noise source.

Once properly tuned, the unit's sensitivity should produce a noticeable increase in background noise when an antenna is connected. The signal-to-noise ratio may be optimized by tuning in a



L9 0.55—0.80 mH  
(Millen 69054-0.68)

fig. 2. Alternate 50-MHz rf amplifier provides more sensitivity than the grounded-gate stage used in fig. 1 but must be neutralized to avoid oscillation.

A few considerations should be kept in mind when laying out and building the oscillator. I mentioned earlier that some form of temperature compensation is needed. This can be provided by using N750 capacitors for C1 and C2. Also, to prevent overcoupling between oscillator and mixer, the grounded link is made from rubber-insulated clip-lead wire. The insulation thickness allows fairly loose coupling, yet provides a snug support for the link. The coil assembly is then secured with epoxy cement.

Finally, if you operate the transistor radio and converter from the same battery, be sure to include the zener regulator. Otherwise, the varying load of the bc radio's class B output stage will cause the oscillator to drift.

## adjustment and operation

By listening on any good six-meter receiver, you can find the oscillator signal. Then set the signal for the tuning range you wish. Mine tracks from 51.5 to 52.7 MHz, which covers only the populated low end of the band. A noise

weak signal and adjusting the rf-input coil for the desired characteristic. With the grounded-gate amplifier, little difference between maximum gain and maximum s/n will be noticed; however, with the grounded-source circuit, the significant improvement can be achieved.

## performance

The performance of my unit is more reminiscent of a modest communications receiver than of the shoestring converter/bc set configuration. Sensitivity is excellent, and you may want to broaden the i-f response of the transistor radio if its selectivity seems too sharp for your operating needs.

Although overall gain won't blow you out of the room, the low noise figure allows you to hear weak signals that would be buried in noise of a similar bipolar setup. The circuit is well worth considering if you're planning a portable rig or are tired of fighting the drift problems and low sensitivity of older designs.

ham radio

# improving the intelligibility of communications receivers

The sound reproduction clarity of communications receivers can be improved significantly by simply improving the speaker enclosure. Common utility speakers are poor performers at best and can be improved (or impaired) by the baffles they are mounted in. Much of the acoustic distortion is caused by either speaker resonance, cabinet resonance, cabinet reflections, back waves or poor frequency response, and these can be easily corrected with a good enclosure. We have all become accustomed to the sound of a small speaker in a metal cabinet or the boomy sound of a large speaker in a box just big enough for it, but these systems leave a lot to be desired.

Sound reproduction can be improved by using a small high-fidelity speaker or hi-fi headphones. The small hi-fi speakers I use on ssb and cw have resulted in a big improvement in communications effectiveness.

## how to do it

Use a small round or oval speaker, 5-inch or 5x3-inch, mounted in a wooden box made from 3/4-inch pine or plywood. The box can be just big enough to contain the speaker with 2 or 3 inches of depth. Use 1/8- or 1/4-inch Masonite or plywood for the front panel. Cut the speaker opening out completely—don't

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drill a few small holes in a cute pattern and hope for the best. Cover the front panel with screen, an open grill with very small reflecting surfaces or plastic grill cloth designed for hi-fi speaker systems.

Mount the speakers and grill so there is no vibration of loose parts. Then fasten thick sound-absorbent material such as felt or foam rubber on the inside surfaces of the box. When the speaker enclosure is completed, put it near the operating position, preferably with the center axis of the cone directed toward your head. If your receiver has an internal speaker it should be disconnected.

## how it works

The thick wood of the enclosure reduces cabinet vibrations, and the unrestricted speaker opening practically eliminates reflections between the cone and grill. The absorbent material helps to damp speaker response and reduces reflections between the inside walls, resulting in smoother frequency response. The small speaker and small box reduce low-frequency response; boominess is reduced by the cancellation effect of the back-wave.

With the speaker placed close to the operating position the volume control can be turned down. This reduces undesirable sound reflections from nearby objects and reduces distortion caused by any non-linearity in the speaker mechanism.

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Bill Rogers, WA5RAO, 1121 Southwest 53rd Street, Oklahoma city, Oklahoma 73109

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# quad-antenna

## design

### parameters

The extensive use of parasitic antennas over the past thirty years or so has resulted in much published data. Design information is available that allows the amateur to easily construct two- or three-element monoband parasitic antennas with little or no experimenting or tuning, except perhaps for feed-point matching.

The quad antenna, which is a special type of parasitic array, unfortunately has been neglected as far as detailed design information is concerned. However, when it is recognized that the quad exhibits the characteristics of conventional parasitic arrays except for differences in element length caused by the folded wire elements, quad design is very similar to that of conventional parasitic arrays and is not particularly mysterious.

This article presents a design approach that allows construction of a quad antenna with no special element tuning or fussing with element lengths, tuning stubs, or loading coils.

#### characteristics

Before proceeding with details of the quad design approach, it is well to discuss some facts regarding the quad and dispel some misunderstandings.

**forward gain.** Quad antenna gain has been underrated in some of the literature. If a quad loop is considered as one element, its gain can be shown to match or exceed that of a monoband Yagi on an element-for-element basis.

**front-to-back ratio.** The quad really shines here. Its front-to-back ratio will exceed that of a Yagi with the same design criteria by up to 10 dB (based on element spacing, element length, and related factors).

**vertical radiation angle.** The vertical radiation angle for any horizontally polarized antenna is a function of its height. Therefore, any half-wave dipole, Yagi, or quad will have the same vertical radiation angle if they are all at the same height above ground. The quad's performance as an excellent band opener and closer is due to its directivity characteristics, not to its vertical radiation angle.

**element interlacing.** The minimal mechanical and electrical complications of accommodating three bands on one antenna structure with no apparent loss in efficiency is a definite advantage of the quad.

**bandwidth.** The bandwidth of an antenna can be based on (a) the frequency range beyond which the feedpoint *vswr* exceeds a specified value, usually 1.7 to 1; (b) the bandwidth over which front-to-back ratio or gain does not drop below a certain value; or (c) the point at which either the director or reflector closely approaches resonance. Because the director and reflector are electrically close to the driven-element length (much more so than with a tubing Yagi), the point at which the director or reflector approaches resonance will govern the bandwidth of a quad designed for optimum performance.

**rain static.** Compared with the Yagi, the quad is relatively immune to rain or snow static.

**cost.** Prices of aluminum tubing in the United States are low compared with most countries. In addition, a large variance exists in prices of aluminum bought from surplus sources versus those of new aluminum bought from metal suppliers. The price of a home-constructed Yagi therefore depends on how good a shop-

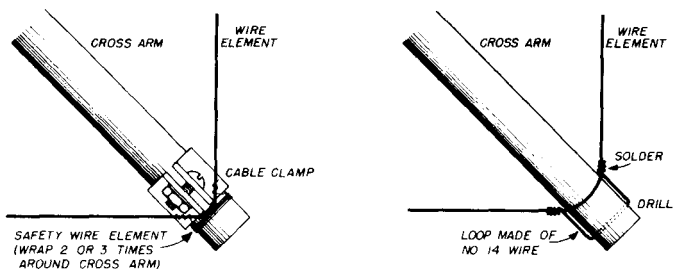
W. R. Hillard, K60PZ, 28700 King Arthur Court, Palos Verdes Peninsula, California 90274

ping job is done. It is usually found that a well-built, home-constructed quad with fiberglass cross arms will cost more than a monoband Yagi. This is also true of purchased quads. However, quads will tolerate a lighter rotator and tower than a comparable Yagi because of the quad's lighter weight. Therefore, the quad is usually a less expensive total antenna system.

A word of caution: aluminum cross arms are not satisfactory for a quad, since they degrade antenna performance and absolutely ruin front-to-back ratio. Breaking up the aluminum cross arms with insulators helps, but this is difficult to implement.

with two or three loops of safety wire to anchor the wire to the quad arm, as shown in **fig. 1**.<sup>\*</sup> Start with the element too long, with a slight droop and a temporary splice at the bottom. Note that quad parasitic elements not using stubs or loading coils are a continuous wire loop. The element length can be adjusted for resonance, maximum gain, or front-to-back ratio. Once the proper length is obtained, the loop splice can be permanently soldered, coated with silicone grease, and the loop run outboard (but equally spaced) to tighten the loop into a perfect square. Another method is to drill a hole in the arm several inches outboard from where the element would

**fig. 1. Two methods of securing quad elements to cross arms.**



### stubs and loading coils

Since the quad is a closed loop, adjustment of element lengths is difficult. Adjustment of element lengths is usually done with tuning stubs or loading coils. Stubs are unsightly, and the problem always arises of how to anchor them so they don't move in a wind. Both stubs and loading coils destroy the electrical symmetry of a quad loop. All in all, stubs and loading coils are best avoided.

Alternatives are to (a) use element lengths established by criteria such as those in the following paragraphs and not attempt to tune the quad elements, or (b) select the optimum element-loop length on a trial basis.

### element length adjustment

There are two ways to adjust element-loop length. Instead of drilling holes in the bamboo or fiberglass arms, small nylon or metal cable clamps can be used,

normally be anchored. Then use a wire loop to hold the quad loop in place. Slack in the quad loop created by element adjustment can be accommodated by the small loop.

### boom length

Factors that govern boom length for a Yagi apply also to a quad. Most quads are built with a boom too short to realize maximum performance on 20 meters, which is a highly competitive band. To achieve a good balance between practicality and performance, a 20-meter quad should have a boom length as follows:

Two-elements	10 feet
Three-elements	18 to 20 feet
Four-elements	24 feet

As with a Yagi, placing more than one

<sup>\*</sup>The safety wire should be copper to avoid galvanic action, which results in corrosion.  
**editor**

**table 1. Element dimensions for a quad with a 14-foot boom.**

	20 meters	15 meters	10 meters
reflector	2.8% long (1.028)(70.7) = 72.68 ft	2.1% long (1.021)(47.25) = 48.24 ft	2.1% long (1.021)(35.1) = 35.84 ft
driven element	$\frac{1004}{14.2} = 70.7$ ft	$\frac{1004}{21.25} = 47.25$ ft	$\frac{1004}{28.6} = 35.10$ ft
director	1.4% short (.986)(70.7) = 69.71 ft	1.2% short (.988)(47.25) = 46.68 ft	2.1% short (.979)(35.1) = 34.36 ft

director on a 20-foot boom just doesn't give a satisfactory return for the effort *on 20 meters*. Definite advantages can be obtained with quads that have more elements on 10 and 15 than on 20. For example, an effective quad is one with a 20-foot boom having three elements on 20 and four on 10 and 15 meters.

### boomless quads

Three-band, two-element quads using no boom and the same element spacing on 20, 15 and 10 meters have only one real advantage over conventional quads: all can have the same feed-point impedance. A conventional quad has its widest element spacing on 10 meters, which helps to broadband the antenna on this band.

### design graph

I developed the graph shown in **fig. 2** expressly for high-frequency quad arrays. Using the graph and the approach shown in the following design example, a quad can be built with no stubs, loading coils, or tuning. For those who wish to avoid the math involved in the design example, **table 1** was developed from **fig. 2** for popular quads that can be purchased or homebuilt.

Design graphs similar to that in **fig. 2** have been published for Yagis. Because of its wire elements and the fact that the loop is, in effect, two half-wave dipoles folded at the ends and stacked a quarter wavelength, the quad has characteristics that make its element dimensions much different from those of a tubing Yagi array. These factors, plus others, were considered in the analysis used to develop

**fig. 2**, which expressly applies to a quad designed for 20, 15, and 10 meters.

### design example

As an example of using **fig. 2**, suppose it is desired to design a three-element quad for 20, 15, and 10 meters on a 14-foot boom with 6-foot director spacing and 8-foot reflector spacing. The desired center frequencies are 14.2, 21.25, and 28.6 MHz. Proceed as follows:

1. Establish percentages for the desired band excursion within which the antenna must be effective. To cover 20 meters, the excursion from 14.2 to 14.0 MHz is  $200/14.2 = 1.41$  percent, and the excursion from 14.2 to 14.35 MHz is  $150/14.2 = 1.056$  percent. For 15 meters, the excursion is 1.175 percent to reach 21.45 MHz, with a 21.25-MHz center frequency.

Ten meters represents a problem requiring a compromise. To cover the entire band, the excursion is 2.1 percent to reach 28.0 MHz and 3.84 percent to reach 29.7 MHz. The parasitic elements must be so far from optimum to avoid resonance that performance is severely compromised. A good alternative is to settle for  $\pm 2.1$  percent, or 28.0 to 29.12 MHz. The phone man may elect to choose 29.1 MHz as center frequency for 10 meters.

2. Establish element spacings in wavelengths for each band using the formula:

Element spacing is  $\lambda = xf/984$ , where  $x$  is the element spacing in feet and  $f$  is the design-center frequency in

MHz. Using the formula, element spacings are:

band	director (6 ft)	reflector (8 ft)
20 meters	0.086	0.115
15 meters	0.13	0.172
10 meters	0.175	0.234

3. Identify element spacing (fig. 2).

4. Select parasitic-element length as a percentage of driven-element length, bearing in mind the reactance excursion of the element across the band. Note that a reflector spaced 0.1 wavelength from the driven element should be 2.8 percent longer than the driven element. A director spaced 0.15 wavelength from the driven element should be 1.4 percent shorter than the driven element. Shown in fig. 2 are the choices I made using this technique (circled values).

It's not always possible to choose optimum element lengths at the design center frequency and avoid parasitic-element resonance at the band edges. In this regard, it's sometimes better to provide wider spacing for the director than for the reflector.

5. Calculate element dimensions. The driven element length equals  $1004/f$  in MHz. This is 70.7 feet for 20, 47.25 feet for 15, and 35.1 feet for 10 meters. The reflector for 20 meters should be 2.8 percent longer than the driven element, or  $(1.028)(70.7) = 72.6$  feet. The reflector for 15 meters should be 2.1 percent longer than the driven element, or  $(1.021)(47.25) = 48.24$  feet. Calculation of each element results in the data listed in table 1.

When the element lengths are established using the above approach or by direct use of table 1, all that remains is to construct the array and decide on the feed method. The data in table 1 are applicable to a two-element quad with 8-foot reflector spacing by simply ignoring the director data. The data are

applicable to a four-element quad on a 20-foot boom. Make the first and second directors the same length.

## tuning

The above design approach should not be considered an absolute cure-all that eliminates benefits from tuning the array.

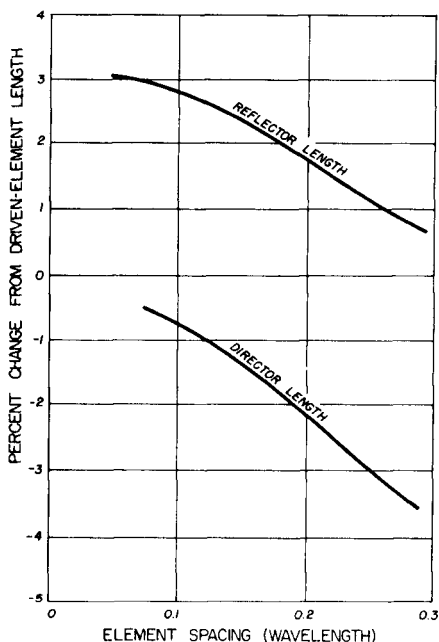
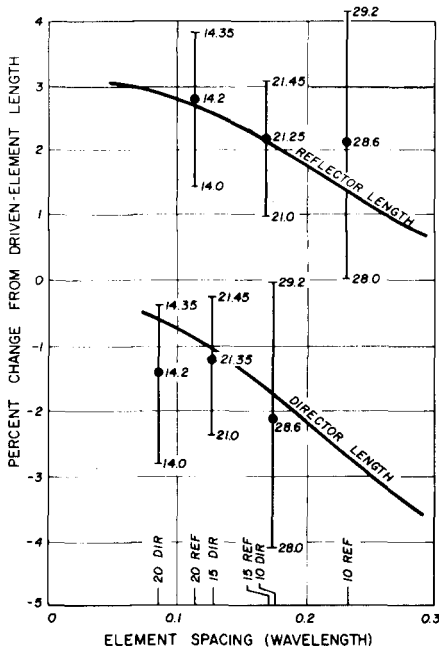


fig. 2. Optimum director and reflector lengths as a percentage of driven-element length as a function of element spacing.

However, few have the talent, equipment, facilities, or perhaps patience to properly tune an array, particularly one with three or more elements. For these amateurs, the "no tuning approach" will be most useful, and those who wish to tune by experimenting with loop length will find that this article defines an excellent starting point. If you wish to tune the quad, using the preceding design data as a starting point, some comments are in order.



The curves in **fig. 3** are recommended element dimensions. At design center frequency, maximum gain will usually be found with a dimension closer to the driven element length than that shown in **fig. 3**. Correspondingly, maximum front-to-back ratio will usually be found with a dimension shorter for a director and longer for the reflector than shown in **fig. 3**. The design must be compromised to achieve best overall performance consis-



**fig. 3.** Plot showing the use of **fig. 2** in establishing quad element lengths for the design example discussed in the text.

tent with gain, front-to-back ratio, and bandwidth.

Tuning for maximum gain at the design center frequency will usually cause narrowband response. If the array is tuned for maximum forward gain, check the response at the band edges to get the best overall gain across the band. As an example of how dangerous forward-gain tuning can be, a director spaced 0.1 wavelength from the driven element will theoretically give maximum gain when it

antenna	quarter-wave matching transformer	main feedline
<b>2-element quad</b>		
10	none (direct feed)	RG-63/U
10	two pieces RG-63/U in parallel (electrical length $\frac{1}{4} \lambda$ ea)	RG-8/U RG-58/U
15	none (direct feed)	RG-59/U RG-11/U RG-62/U
15	one piece RG-11/U or RG-59/U (electrical length $\frac{1}{4} \lambda$ )	RG-8/U RG-58/U
20	none (direct feed)	RG-8/U RG-58/U
<b>3-element quad</b>		
10	none (direct feed)	RG-11/U RG-59/U
15	none (direct feed)	RG-11/U RG-59/U RG-8/U RG-58/U
20	none (direct feed)	RG-8/U RG-58/U
<b>4-element quad</b>		
10		
15	none (direct feed)	RG-8/U
20		RG-58/U

is the *same length* as the driven element, and a reflector spaced 0.25 wavelength from the driven element will theoretically give maximum gain when it is the *same length* as the driven element! Such conditions are obviously bad, because a small frequency excursion can convert the director to a reflector, or vice versa. If you tune for forward gain, try for good response across the whole band, or at least over the band segment of interest.

Tuning for front-to-back ratio is safer and usually much more effective. Such tuning is, however, very touchy. Because of nearby objects it can be frustrating and sometimes leads to inconsistent results.

Don't attempt to tune an array unless the services of a nearby amateur are available, or you can locate a temporary transmitter site several wavelengths and preferably at least a mile from the array. Use the nearby amateur or your remote site for transmitting, and use a receiver with an s-meter or field-strength meter at the site of the array being tuned.

## feed-point matching

Several techniques for feeding the quad can be used. I prefer the gamma match with separate feed lines, but any of the following methods are effective.

The simplest method is to connect all three driven elements together, using a single insulator across the 15-meter driven element. The system can be fed with a single 52-ohm line. This is an effective feed method despite published statements to the contrary. Measurements with an impedance bridge have shown a good match on all three bands.

A second approach is to use one of the methods in **table 2**. These are also effective, as proved by countless users. Some mismatch will occur, however.

## the gamma match

The preferred matching system for quad antennas is the gamma match. It's simple, easy to use, and reliable. Referring to **fig. 4**, the gamma match is implemented and tuned as follows:

1. The coax cable shield connects to the exact center of the driven element. No center insulator is used.
2. Capacitor C is housed in a plastic box. A drain hole should be drilled in the bottom. The box should be sealed with cement. It can be secured to the driven element with a clamp.
3. The matching stub is made of the same wire as the quad element, usually no. 14 copper wire. Spacers can be pieces of polystyrene or any good plastic rod.
4. The values of capacitor C and dimension A are as follows:

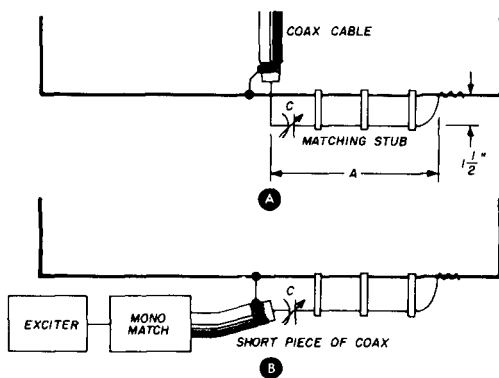
band	capacitor C	dimension A
	(pF)	(inches)
10	50	12 ±6
15	100	15 ±8
20	140	25 ±10

5. Tune as follows: connect a Moni-match or similar instrument, driven by your exciter, to the antenna feed point

(**fig. 4B**). Adjust dimension A and vary capacitor C for minimum standing-wave ratio.

## conclusion

In the final analysis, antenna selection should be based on many factors, including the antenna's adaptability to your location and operating habits. There is no "miracle antenna." The size of a well-designed antenna is a good gauge of its expected performance. In general, if you



**fig. 4.** The gamma match, and typical instrumentation for tuning.

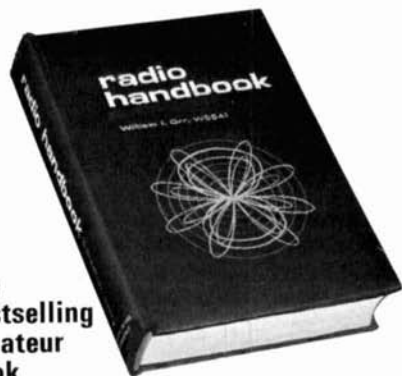
desire to operate on 20, 15, and 10 meters with high performance on all three bands, the quad definitely qualifies as a front runner. If, however, you wish to operate exclusively on one band, the monoband Yagi is probably first choice. Trap-element Yagis are in a class by themselves, and although they certainly have their place, their efficiency isn't as high as the quad or monoband Yagi.

## acknowledgements

I'd like to express my sincere appreciation to Karl Scharping, W6KWF, president of the Cubex Company, Altadena, California, for his assistance in the preparation of this article. Also, thanks are certainly due my wife, whose support and indulgence over the hours of experimenting made it all possible.

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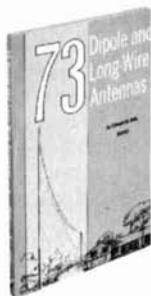
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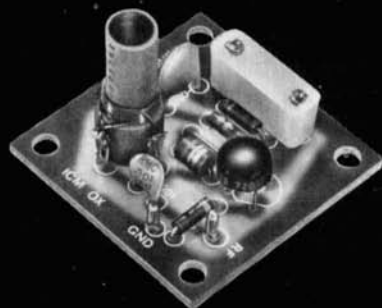
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# modular modulus

Mother  
and daughter  
printed-circuit  
boards are used  
to increase  
IC counter-circuit  
versatility

Ronald M. Vaceluluke, W9SEK, 17W540 Hillcrest, Wood Dale, Illinois

The many articles on integrated circuits published during the past few years have helped these little plastic bugs gain wide acceptance by those who like to build their own equipment. The construction projects that didn't specify printed circuit boards usually turned out to be quite difficult because of the wiring required on the closely spaced pins of the ICs. IC sockets are really no help except when frequent replacement is necessary; besides, they're an added expense.

Let's face it—the PC board route is the only way to go when working with ICs. They are designed for PC board mounting, and any other mounting method just doesn't result in a neat, professional appearance. Of course, IC projects have been published specifying PC board construction. The only trouble here is that the circuits are restricted to the function for which they were originally designed.

## modular modulus

With these limitations in mind, I'd like to present what I call "modular modulus." The word "modular" is familiar to most readers, but "modulo" may not be, so a word of explanation is in order.

A modulo, or mod, is a circuit that

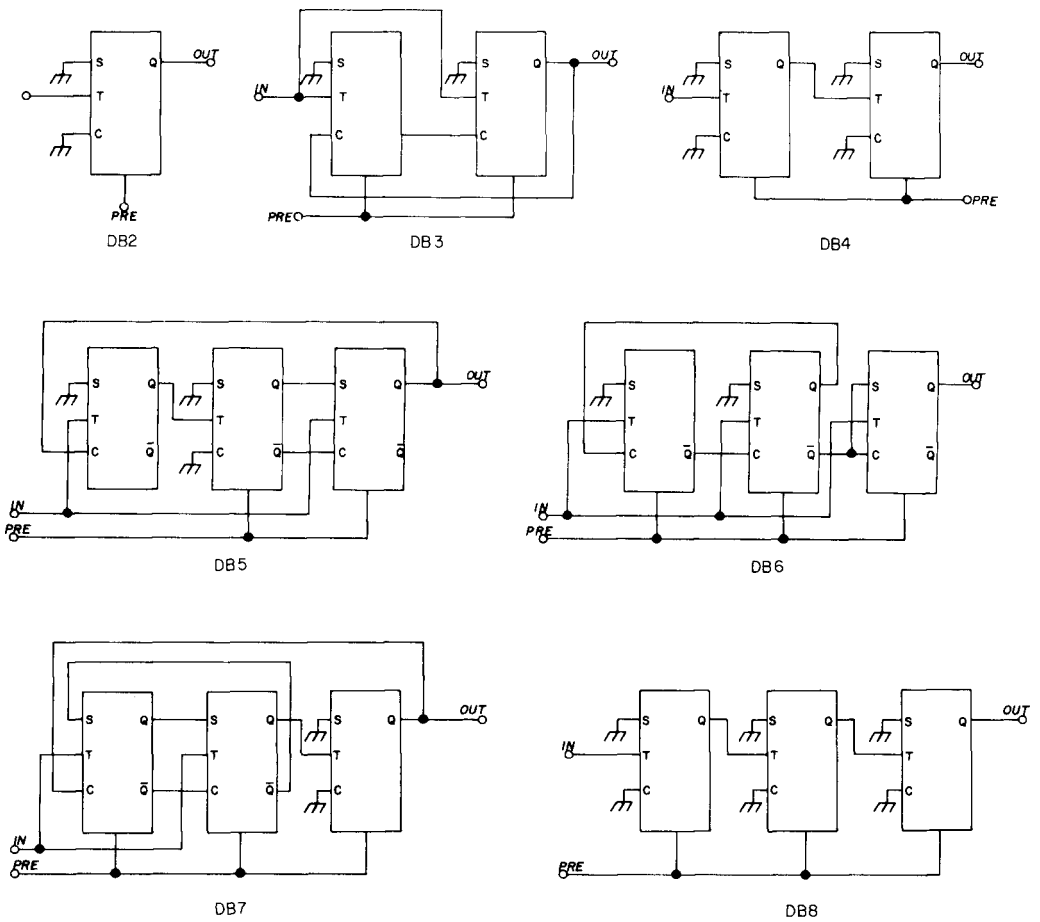


fig. 1. Modulo circuits. Each circuit mounts on a daughter board, ten of which can be mounted on a mother board for a mix of counter functions. DB corresponds to "divide by;" the DB-3 circuit divides by 3.

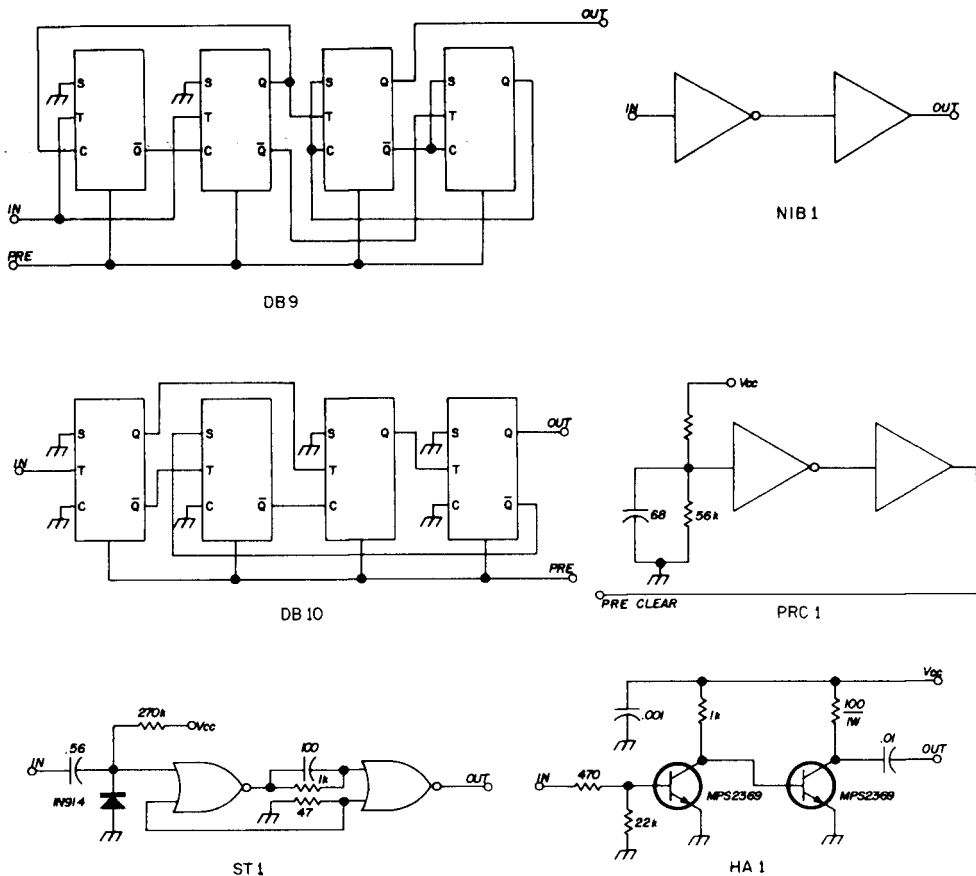
counts; or as some prefer, divides. A mod is designated by how many counts occur before an output is produced. A mod 7, for example, will count to 6 and the output will remain unchanged. When the 7th input occurs, the output will change to indicate that the count is complete.

The resistor-transistor logic (RTL) ICs used here are edge-triggered by a sharp negative transition from logical 1 (approximately  $V_{cc}$ ) to logical 0 (approximately ground). During a count cycle the modulo output will go from logical 0 (its reset condition) to logical 1.

The part of the count at which this action occurs will vary between modulos; that is, mod 5 will go high at some point in

the count different from, say, mod 10. The transition from 0 to 1 will have no effect on following counters. However, when the count is complete and the output goes from 1 to 0, the following counter will be triggered.

This output on a scope will appear as a train of square waves. If an accurate frequency source were connected to a number of modulos wired in tandem, a division of this frequency will give accurate submultiples from which time pulses can be obtained. These can be used to calibrate receivers, oscilloscope time bases, etc. The modulos described here were used for this purpose in the frequency and time standard shown in the photo.



The transistorized oscillator used in the standard is extremely stable. It should be remembered that no matter how many frequency divisions are made, the output is only as accurate as the input.

These modulos can also be used as a basis for a TV synch generator count-down chain (ATV'ers please note), or to obtain the gating pulses in a frequency counter. By dividing the 60-Hz power-line frequency by 60, 1-Hz pulses can be obtained for an electronic clock. Other uses will undoubtedly come to mind for these versatile circuits.

### construction

Each circuit (fig. 1) is mounted on a small PC board. These boards, called daughter boards, mount on a mother board (fig. 2) by Varicon connectors

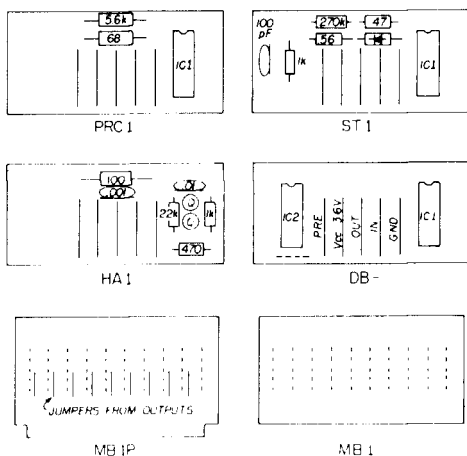


fig. 2. Top of PC boards showing parts layout. Dashed jumper on DB board is required only for DB7 and DB9 circuits.

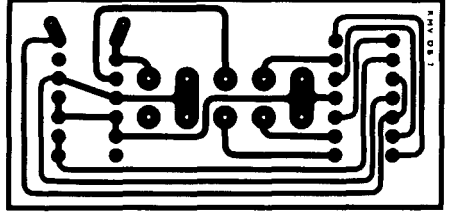
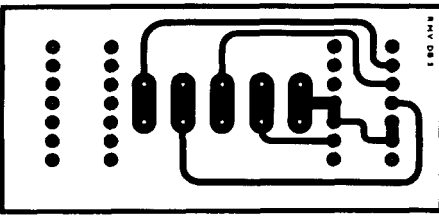
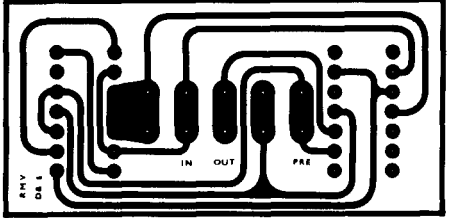
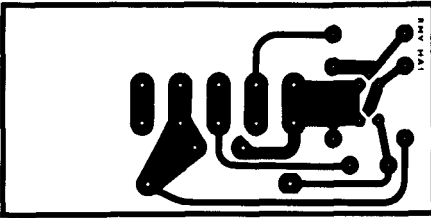
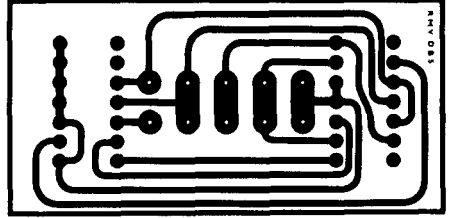
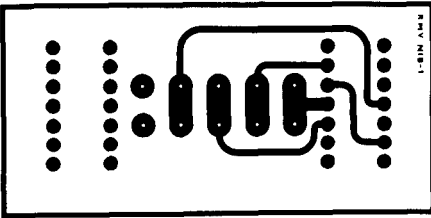
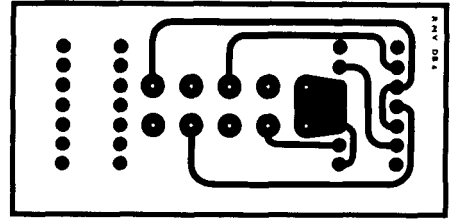
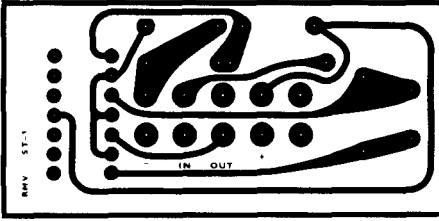
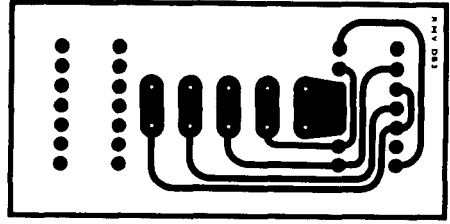
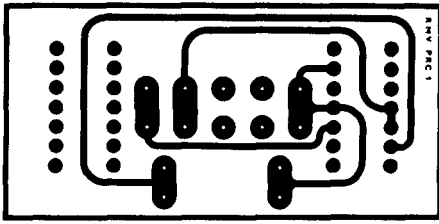


fig. 3. Etched side of mother and daughter boards.



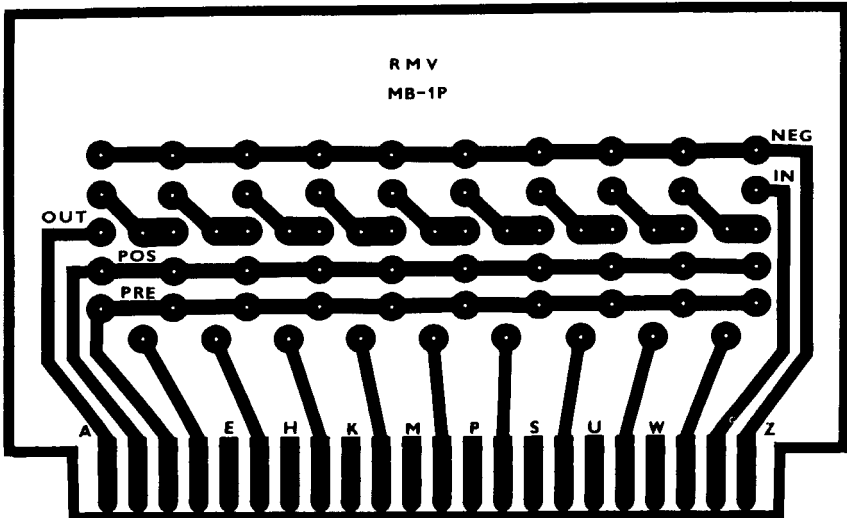
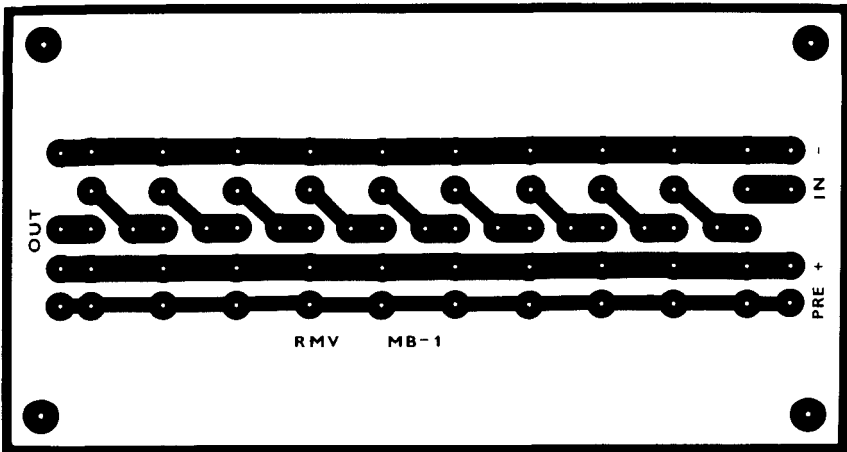
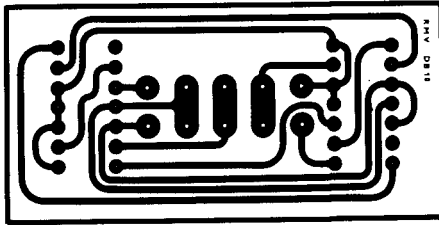
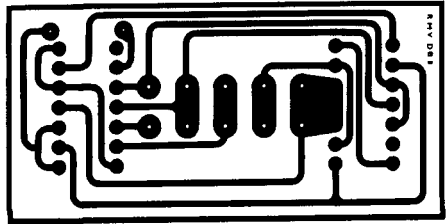
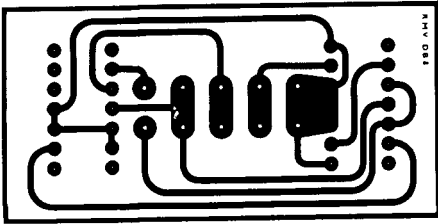
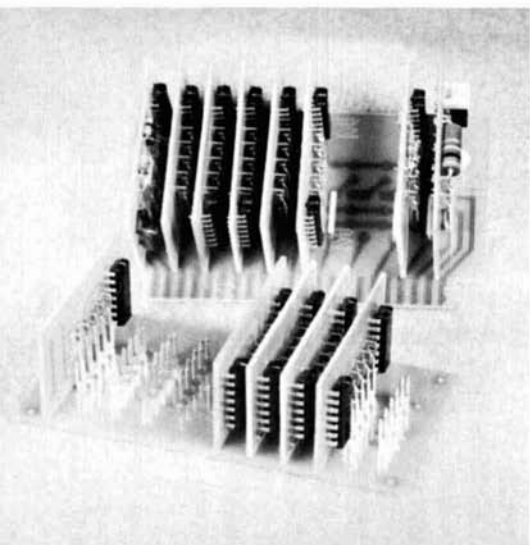


table 1. Loading factors for integrated circuits used in the counters and associated circuits.

board	in	out	clear	IC1	IC2
DB-2	2	5	1	MC726P	—
DB-3	4	4	2	MC791P	—
DB-4	2	5	2	MC791P	—
DB-5	4	4	3	MC791P	MC791P
DB-6	6	5	3	MC726P	MC790P
DB-7	4	5	3	MC790P	MC726P
DB-8	2	5	3	MC726P	MC790P
DB-9	4	5	4	MC791P	MC790P
DB-10	2	5	4	MC791P	MC790P
ST-1	-	5	-	MC724P	—
PRC-1	-	25	-	MC799P	—
NIB-1	2	25	-	MC799P	—

made by Elco. The daughter-board connector is no. 02-030-013-5200 and mates to the mother board with connector no. 02-030-113-6200. Thirty connectors are mounted on a plastic strip. They should be broken into segments of five connectors each. The PC etched board layout is shown in fig. 3; however, if it isn't convenient to etch your own boards, they may be purchased.\* In any event, the use of a small-tipped, twenty-five-watt iron is recommended. It's best to mount the

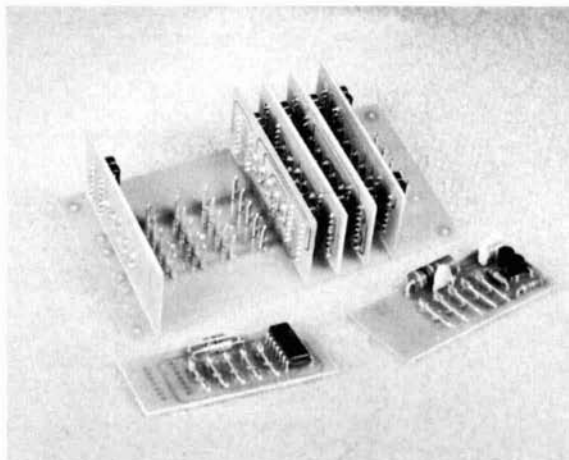
The two styles of mother boards. The plug-in type is used in the frequency and time standard. The other (below) can be mounted on a small chassis with outputs brought out to banana jacks.



connectors first using the plastic strip to hold the connectors in alignment until soldered in place. After soldering, the plastic can be broken away.

The mother boards are shown in two styles, and both will hold up to ten

Mother board can be cut to shorter length if ten daughter boards aren't needed. Shown below is the pre-clear (left) and harmonic amplifier modules.



daughter boards. One mother board can be mounted on a chassis with stand-off insulators over a cutout. The other plugs into a socket, such as an Amphenol 143-022-01. This socket, as well as the Elco Varicons, can be purchased from Allied Radio in Chicago.

The output side of the daughter boards feeds the input to the next stage. The output connectors appear on a separate pad on the board as well. Leads can be connected to these points to give different ratios of division throughout the chain if desired. The plug-in-style mother board must have these points jumpered so they appear at the main connector. See

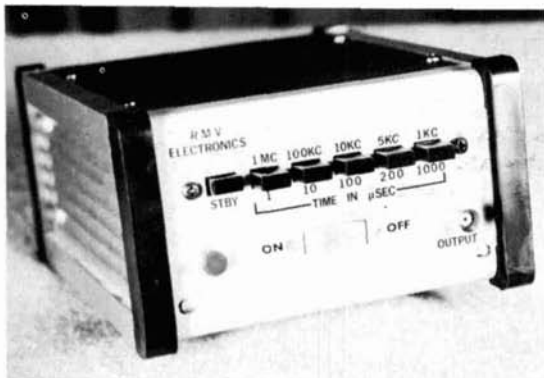
\*Available from RMV Electronics, P. O. Box 283, Wood Dale, Ill. 60191. Daughter boards are \$1.25 each postpaid in quantities of 1 to 5; \$1.10 each in quantities of 6 and up. Mother boards are \$2.00 each postpaid. All boards are tinned, drilled, and ready for components.

fig. 2 for the location of these jumpers.

### rules for use

The most important thing to observe is the loading factor, **table 1**. Do not let the input loading factor exceed the output loading factor (fan-out) of the preceding stage. A higher output loading factor than is needed is permissible. If the required input is greater than that available, use a buffer stage (NIB 1). Since the flip-flops are edge-triggered by a sharp drop from logical 1 to logical 0, frequencies below approximately 100 kHz must be squared-up with a Schmitt trigger (ST 1). This circuit can be used to the subaudible range: 7 Hz or less. If a high-harmonic output is desired (as in a frequency standard) the harmonic amplifier (HA 1) is recommended.

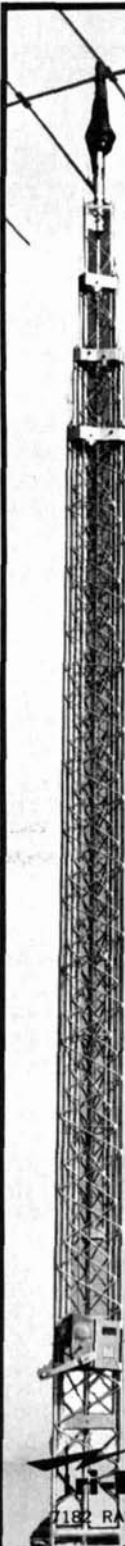
It is desirable to clear all stages before starting to count. This is done by first applying Vcc (3.6V) to the pre-clear line



Frequency and time standard using modular modules.

and then grounding this line when counting. This can be done manually or by the PRC 1 board, which will pre-clear the stages automatically as power is applied to the mother board. The voltage required by these circuits is 3.6 volts dc. A well-regulated and filter supply will provide the best service.

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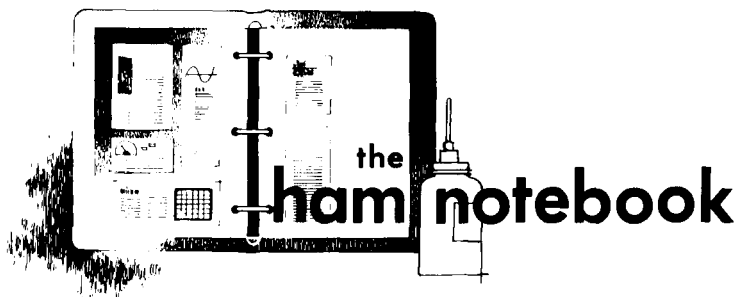
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## perking up your HW-17A

The Heath HW-17A certainly can qualify as one of the more popular 2-meter transceivers, judging from the many in use on this band. However, as with most equipment of this type, it's always possible to improve on the original design as new devices become available after the fact.

The broadband preamplifier described in this article will improve sensitivity, increase rejection of fm and tv images, and provide better agc action in the HW-17A. Improvement in agc is not to be passed over lightly, because it makes the S-meter much more useful and compensates for some of the loss in the automatic noise limiter circuit. An additional feature of the preamp modification is that no new holes need be drilled to accommodate the circuit.

Also included are changes to improve transmitter modulation and increase modulator bass response.

### preamp circuit

Nothing exotic or unproven is included in the preamplifier circuit (fig. 1). This circuit (with a 3N140 mosfet) was used by many 2-meter enthusiasts in a

converter described in an earlier issue of *ham radio*.<sup>1</sup> Although not shown in the schematic, the 40673 has internal protection for the gates. This eliminates the need for external diodes, which are found in many semiconductor front-end circuits to eliminate static burnout.

### preamp construction

A physical layout compatible with the HW-17A is shown in fig. 3. A thin copper U-bracket chassis bolts to PC board 85-205 in the HW-17A with a single machine screw, which is also used to secure the tuner to this board. All resistors are mounted on the opposite side of the bracket. Feedthrough capacitors are used for bypassing.

### modulation improvements

Not all HW-17A transceivers are fully modulated, despite several circuit modifications incorporated by the manufacturer. One change was very helpful in my case, although it did decrease my power output slightly. By increasing the value of the screen resistor in the final amplifier, input power is decreased, which allows higher modulation percentage and improved upward modulation. The exact value of the resistor is best determined by observing the signal on an oscilloscope. My experience indicated

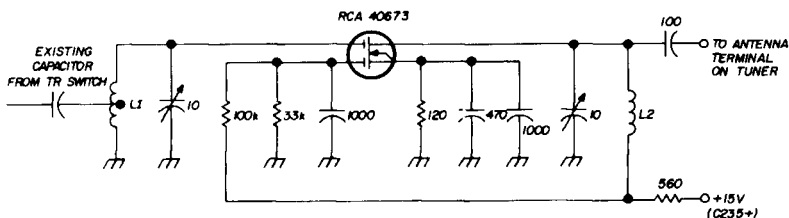


fig. 1. Low-noise preamplifier using a tetrode mosfet for improving the Heath HW-17A receiver front-end response.

that resistor R115 should be increased to 27k.

### modulator bass response

A final touch to improve the bass response of the HW-17A modulator will bring compliments on your audio quality. The emitter of Q11, a 2N3391, should be

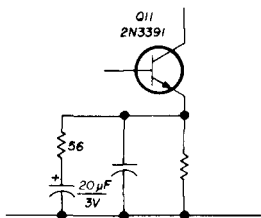


fig. 2. Additional components for improving the HW-17A modulator bass response.

bypassed with a capacitor larger than the 2- $\mu$ F unit in the original circuit. Unfortunately, the stage will motorboat with direct bypassing. My solution was to use a 56-ohm resistor in series with a 20 $\mu$ F capacitor, as shown in fig. 3.

### conclusion

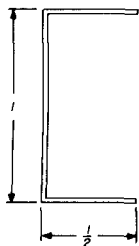
The simple changes described above will more than repay you in increased performance and operating pleasure for the time and effort involved. When I made these modifications I was able to realize the best features of this fine transceiver. Now when DX is there, I work it!

### reference

1. D. Nelson, WB2EGZ, "High-Performance Mosfet Converter—the Two-Meter Winner," *ham radio*, August, 1968, pp. 22-29.

Don Nelson, WB2EGZ

fig. 3. Suggested parts layout for the preamp. Simple U-shaped copper chassis bolts to an existing pc board in the HW-17A.



## removing IC's

I was accumulating parts for a project using ICs and needed a couple of JK flip-flops. I had some in my junk box, but the little rascals were firmly embedded in one of those bargain PC boards obtained from a surplus outlet. The board had been wave-soldered, which means the parts were there to stay put.

I didn't have a vacuum desoldering tool handy, so I used a trick known as wicking. I flattened a short piece of shield braid and held it against the pins of the IC. Applying the soldering iron to the braid causes it to absorb the solder, and the IC comes out easily. Everything must be clean: soldering iron, braid, and PC board.

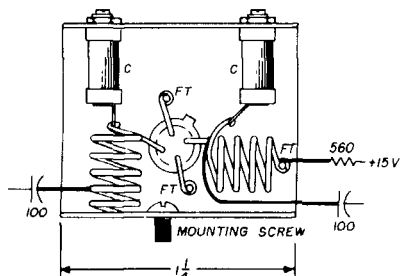
A vacuum desoldering tool is best if you have to remove many parts, but the wicking method is okay for one or two devices. I made a vacuum desoldering tool out of an old ear syringe, but it's slow and the tube tends to clog with solder. The best device is a spring-loaded tool designed for this purpose.

Alf Wilson, W6NIF

## two-meter converter

The converter shown in fig. 4 receives on two frequencies. I use it with an auto radio to receive my vhf club stations on 145.35 MHz and an fm repeater on 146.94 MHz.

Most of the parts were salvaged from uhf television converters and transistorized bc receivers. Crystal frequencies were chosen to put the converter output into a clear spot in the broadcast band.



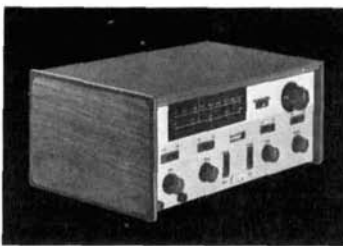


# he worked the world.

Mr. John H. Thompson, W1BIH/PJ9JT, recently packed his Ten-Tec Power-Mite PM 3A transceiver into a suitcase and headed for the Coral Cliff Hotel, Curacao (Netherlands Antilles). From there he worked the world.

"Final tally on the PM 3A results at PJ9JT are 261 QSOs on 14 MHz and 41 QSOs on 7 MHz for a total of 302. This includes 32 different countries in 5 continents. I operated only with the PM 3A on 7 and 14 Mc. CW. No contacts were set up first on high power, nor was any auxiliary receiver used. It was all done with the PM 3A. Of course I had a FB location and the PJ9 call didn't hurt. Among the DX worked were five VKs, a ZL, VU and 4X4. Only Africa was missed and I did get a PJ? response from an EL. The batteries, a pair of 6V lantern batteries in series, lasted the entire operation and showed no signs of failing. Some comments from stations worked:

- W8KIT:** 'Congrats on that signal with real QRP'
  - W00PK:** 'Unbelievable'
  - W5IUW:** 'Ur really busting my ears'
  - W3KR:** 'Boy, ur 5 watts FB here on my attic antenna'
  - W4KC:** 'Did you say 5 watts?'
  - W2GA:** 'Boy, ur rig doing FB'
  - W4YWX:** 'Unbelievable — if I didn't know you I'd swear you're pulling my leg because ur hitting 20 DB'
  - K3CUI:** 'Are you really running only 5 watts? FB'
  - OK1AOR:** 'Sigs 589 FB'
  - K6IC:** 'Your 5 watts sure good here'
  - UK2KAF:** (ex UP2KNP): 'Ur low power sure doing FB'
  - K4ZA:** 'Ur sig has real punch'
- I did other hamming, making some 400 contacts on the other bands, both CW and SSB using high power equipment. Could have made many more QSOs in the same time using the high power rig but it wouldn't have been half the fun."



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**PM 3:** 20-40 meter bands CW only — power input 5 watts power required 12 V d.c. — 500 ma. High impedance output for headphones.

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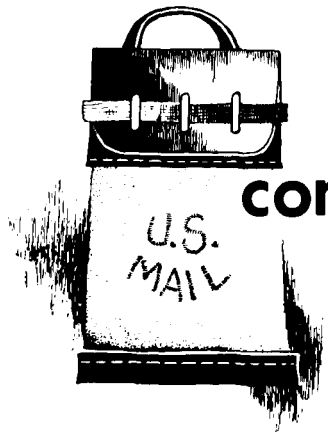
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## comments

### Dear HR:

During a recent QSO with a well-known amateur, the conversation turned to the quality and content of the general run of QSO's heard on the bands. What had started out to be just a casual contact turned into a lengthy discussion of the directions in which amateur radio is headed.

It seems that, more and more the prevailing idea is to contact a fellow, exchange signal reports and equipment types, and then cut and run to the next one. This situation is aggravated by the almost continuous stream of contests sponsored by one group or another, the purpose of which seem to be the maximum number of contacts in the minimum length of time.

I will be the first to acknowledge that amateur radio is a hobby open to all types and persuasions. Those whose main interest lies in the compilation of an impressive list of contacts and a wall full of certificates attesting to their ability in multiplication, are and should be free to pursue their aims. However, must we all be forced into this type of operation? Listen in any evening on the DX bands and count the number of genuine QSO's going on.

Some amateurs, in self-defense, have grouped together into small regional

groups and don't even attempt, nor in some cases welcome, outside contact. The foreign amateurs seem to be getting more reluctant to answer calls, and I suspect it is for precisely this reason.

If we wish to retain our bands, perhaps it is not too early to examine the purposes for which they were granted in the first place and the utilization that is being practiced today. A QSL card used to be something a fellow sent to another station to acknowledge a pleasant contact, not something to confirm a statistic.

Perhaps we need to call *The Old Man* back from his well-earned rest to keep watch on us. His ever-ready Wouff-hong and threat of the Rettysnitch struck fear into the hearts of those who were guilty of "rotten" practices.

This station is open anytime I'm on the air to contacts with others that want to talk about something more meaningful than, "ur RST 5NN eqpt hr Foghorn II tnx qso diddley-bump-de-bump."

**Jim Crouch, K4BRR**  
Jacksonville, Florida

## mnemonics

### Dear HR:

Here's a mnemonic for you which I learned from Dr. Nolde, my first Electric- al Engineering Professor at Berkeley in 1956.

ELI the ICE man.

ELI voltage leads (comes before current in an inductor

ICE current leads voltage in a capacitor.

Thanks for the article—I can use some of the ideas with my students.

**Les Hamilton, PH.D., K6JVE/3**  
Prince Georges, Maryland



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**J. Pat Hawker, G3VA**

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# new products

## antenna rotator



Radio Shack's new antenna rotator, the *Servo-Rotor*, although designed primarily for television antennas, will handle a three-element 20-meter beam according to the manufacturer. The *Servo-Rotor* uses a modern solid-state amplifier/control circuit that compares actual antenna direction with the heading at the control box, and corrects any errors. Pointing accuracy is claimed to be  $\pm 0.5\%$ . Special clutch action overcomes the effects of ice and wind, and a built-in brake eliminates overshoot. Indicator lights show turning direction. Heavy-duty V-shaped clamps attach the rotator to masts from 1½ to 2-1/8 inches in diameter. \$39.95 from your local Radio Shack store.

## electronic keyers

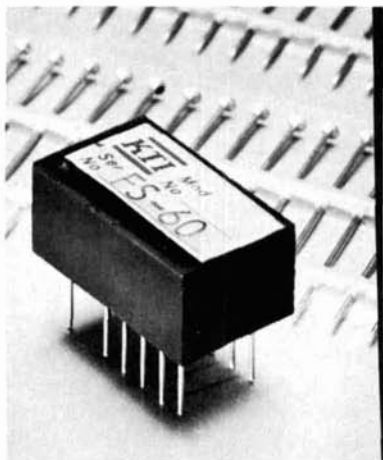


Curtis Electro Devices has announced two new integrated-circuit keyers that incorporate iambic-mode squeeze keying into the basic *Electronic Fist*. These new designs provide iambic-mode character generation (alternate dots and dashes) when both key levers are closed. The EK-39 Iambic Deluxe model includes dot memory, self-completing dots, dashes and spaces plus instant-start circuits that ensure easy and accurate character generation. The *weight* control, which is completely independent of the *speed* control can be used to lengthen characters according to individual preference.

A built-in monitor oscillator and speaker include an external volume control and internal pitch control. Provision is made for solid-state or relay switching. The unit will key grid-block rigs up to -150 volts at 50 mA with output transistor provided. A back-panel jack allows connection of a straight key. Three popular color combinations allow you to match your other station equipment. All cables and connectors are provided.

The EK-39 is available as a kit for \$87.95, or factory wired and tested for \$97.95. The model EK-38, which has all the features of the EK-39 except the *weight* control, is \$72.95 as a kit, \$82.95 wired and tested. A reed relay option is \$4.00 additional. For more details, write to Curtis Electro Devices, Box 4090, Mountain View, California 94040.

## active filter



Kinetic Technology, Inc., has introduced a new hybrid integrated-circuit active filter that features many interesting characteristics. The low-power FS-60 filters require only 0.3 mW of power at  $\pm 2$  volts, making it particularly suitable for battery-operated equipment. The filter works in the frequency range of dc to 10 kHz and has multi-loop negative feedback for high stability. Q range is from 0.1 to 500. The voltage gain of the unit is adjustable to 40 dB. Bandpass, high-pass and low-pass outputs are available simultaneously. The center frequency and Q can be tuned by adding external resistors.

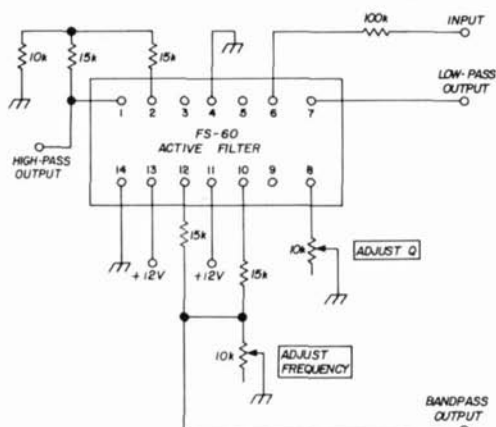


fig. 1. Use a series resistor (100 to 1000 ohms) in the chosen output line to isolate the FS-60 possible capacitive or inductive loads. Device will not drive less than 10k load without reduced output swing.

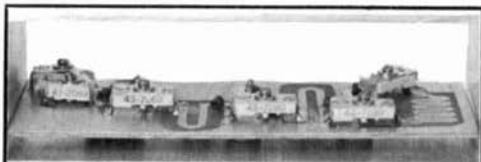
A typical amateur application of the FS-60 is shown in fig. 1. This circuit tunes from 200 to 800 Hz and makes a good outboard CW filter. Practically all the active moonbouncers are using this type filter and report that they have nothing better. It's even been rumored that Sam Harris, W1FZJ/KP4, has his chained to the bench! In the near future we expect to have more applications information on this interesting new device in *ham radio*.

For complete specifications and pricing information, write to Kinetic Technology, Inc., 3393 De La Cruz Boulevard, Santa Clara, California 95051.

## general-coverage receiver



Radio Shack has introduced a new solid-state general-coverage receiver that tunes the entire range from 535 kHz to 30 MHz. The new DX-120 receiver features an fet in the front end to minimize cross-modulation from strong local stations, but still provides good sensitivity for weak signals. Four color-coded tuning scales and a logging scale provide easy readability; fine tuning is performed with a separate bandspread control. An S-meter, variable bfo and automatic noise limiter are provided; a built-in regulated power supply stabilizes receiver power supply-voltage drift and battery aging. It can be plugged into your local 117 Vac line, or used with 12 Vdc battery power for portable operation. Front panel design is a modern black with brushed-aluminum trim; the wrap-around cabinet is steel. The DX-120 can be operated with it's own built-in speaker, or with external headphones. \$69.95 from your local Radio Shack store.



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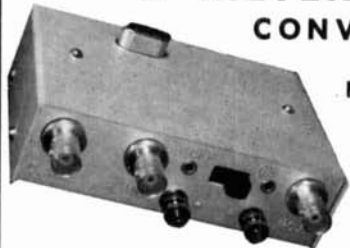
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## VANGUARD LABS

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## how to use test instruments in electronics servicing

This new book by Fred Shunaman, former editor of *Radio-Electronics* is different from the usual books of this type. Instead of a series of "how-it-works" descriptions, this manual describes specific tests and troubleshooting techniques for the electronic experimenter. The first two chapters explain new ways to use an oscilloscope in tv troubleshooting, the next two chapters present tricks that can be done with vom's and vtvm's. Signal generators are covered in the next chapter, and the text describes signal-injection troubleshooting, alignment techniques, and how to measure inductance and capacitance with a signal generator.

Other chapters cover capacitor checks, test probes, tube and transistor checks, crt testers and rejuvenators, cathode-current measurements, field-strength measurements, signal tracing, impedance and frequency measurements, and maintaining and calibrating test equipment. One whole chapter is devoted to color-tv test gear, revealing a number of important pointers (and pitfalls) facing the technician. 256 pages. \$4.95 from your local electronics distributor, or write to TAB Books, Blue Ridge Summit, Pennsylvania 17214.

## variable power supplies



Blulyne Electronics Corporation has announced a new line of economical, continuously-variable low-voltage power supplies that are ideal for semiconductor work. The units feature excellent regula-

tion and ripple characteristics, and are short-circuit proof. Single, dual and triple units are available.

The single unit (PS-61C) provides  $\pm 15$  Vdc output with a load current of 700 mA (usable up to 1 ampere over 10 Vdc). Regulation is 0.0005% Vdc per mA load current; ripple is less than 0.005 at full load. Price: \$49.95.

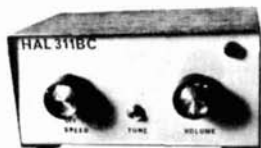
The double unit (PS-62C) is ideal for working with operational amplifiers and has the same electrical specifications as the PS-61C at each output. Price is \$74.95. The triple unit (PS-63C) is three completely independent power supplies that may be used in any combination. Each output has the same electrical specification as the PS-61C. Price is \$99.95.

All power supplies are furnished in modern walnut-finished cabinets. The grounding plug, single-switch operation, chassis-ground terminals and isolation from the ac line provide complete safety. For more information, write to Blulyne Electronics Corporation, 3 Sand Springs Road, Williamstown, Massachusetts 01267.

## multi-antenna coupler

The new multi-antenna coupler from Antennalabs is designed to couple several different antennas to a single transmission line. The coupler separates incoming signals according to frequency, and routes signals in each band to a separate terminal. It can be used with any 52-ohm unbalanced load (72-ohm models available), and may be located any convenient distance from the antenna since feedline length is not critical. The coupler is encapsulated in a durable weatherproof epoxy housing; power rating is 2000 watts, insertion loss 0.5 dB maximum, isolation between terminals is 20 dB and vswr is less than 1.5:1 on all bands. Models are available for all amateur bands from 3.5 to 144 MHz. Prices range from \$24.95 to \$34.95 depending on frequency. For more information write to, Antennalabs, Post Office Box 458, Ocean Bluff, Massachusetts 02065.

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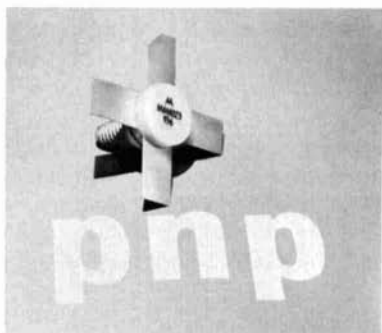
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## pnp rf power transistors



Motorola Semiconductor has added four new transistors to their line of pnp rf power devices. With the addition of the new transistors, types MM4020 through MM4023, the line now covers the power range from 0.5 to 40 watts at 175 MHz. Each of the transistors features balanced-emitter construction for maximum safe operating area, isothermal design for flat power output vs temperature, and low inductance stripline packaging.

Balanced-emitter construction uses thin-film nichrome resistors in series with each of the multiple emitters to distribute power evenly throughout the semiconductor chip. Isothermal design results in exceptional output power stability with changing operating temperature. This insures even generation and flow of heat from the chip so that the power slump usually encountered in an rf power device operating near its maximum frequency is nearly absent. The stripline package provides minimum series lead inductance, and when combined with the inherently higher input impedance of pnp transistors, simplifies the design of interstage matching circuits; it also results in higher gain-bandwidth products.

At 175 MHz, the MM4020 provides 3.5 watts output with 11.5-dB gain; the MM4021 provides 15 watts (7.0-dB gain), the MM4022 provides 25 watts (5.5-dB gain), and the MM4023 provides 40.0 watts at 4.5-dB gain. For complete specifications, write to Technical Information Center, Motorola Semiconductors Products, Inc., Box 20924, Phoenix, Arizona 85036.





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**THE PUGET SOUND** Council of Amateur Radio Clubs will again issue a Washington State Operating Achievement Award Certificate signed by Governor Daniel J. Evans, for contacts made during Washington State Amateur Radio Week, September 6th through 13th, 1970. Out-of-state amateurs must contact ten Washington hams, and in-state amateurs must contact 20 other Washington hams during this week. Send list of stations worked, their locations, dates of contacts, your call and name and address to: The Puget Sound Council of Amateur Radio Clubs, 12306 80th Avenue East, Puyallup, Washington 98371.

**QSL'S — BROWNIE W3CJI** — 3111-B Lehigh, Allentown, Pa. 18103. Samples 10¢. Cut catalogue 25¢.

**SING** a September song! 1970 ARRL National convention September 25-26-27, Statler-Hilton Hotel, Boston. Free brochure & discount coupon from W1VRK, 28 Forest Ave., Swampscott, Mass. 01907.

**WARREN OHIO AMATEUR RADIO ASSOCIATION (W.A.R.A.)** 13th Annual Hamfest, Sunday, August 16, 1970 at New Location, New Trumbull County Fairgrounds, 2 miles north state Route 305 on Bazetta Road just west of Mosquito Creek Reservoir. Registration begins at 1000 hrs. Mobil talks on 3940 KHz and 145.3 MHz.

**"DON AND BOB"** guaranteed goodies. Monarch dual-meter SWR bridge (reg. \$24.95) \$15.95; Ameco preamplifier, PT, \$49.95; Amphenol PL-259, SO-239 49c; Astatic D-104, G stand \$29.95; HAM-M \$99.00; TR-44 \$59.95; Goldline 2 position coax switch \$5.95; Hygain quad (reg. \$129.95) \$109.95; RCA 6LQ6/6JE6C \$3.50; 6146B \$4.45; Cetron 572B \$15.95; TRI-EX W-51, prepaid \$330.00; Drake W-4 wattmeter \$55.00; 1000KHz crystal, HC6U \$3.95; 100KHz xtal \$7.00; Telex HMY-2K stethoscope headset \$3.95; Jennings vacuum variable 300PF/7.5KV \$30.00; 3000MFD/30V computer capacitor 35¢; 100 ft. #14 antenna wire \$1.95; many more items. Write for needs. Write for cash quotes major line gear. BankAmericard, MasterCard, GECC. All items guaranteed, FOB Houston. Don K5AAD, Bob WA5UUK, Madison Electronics, 1508 McKinney, Houston, Texas 77002. (713) 224-2668.

**SPECIAL OFFER** — Ham Radio Binders regularly \$3.95 each now three for \$10.00 postpaid. Ham Radio, Greenville, N. H. 03048.

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**ORIGINAL EZ-IN DOUBLE HOLDERS** display 20 cards in plastic, 3 for \$1.00, 10 for \$3.00 prepaid. Guaranteed. Patented. Free sample to dealers. Tepabco, John K4NMT, Box 198R, Gallatin, Tennessee 37066.

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**FORCED SALE:** Collins 75S-3B receiver with 500 cycle filter \$600.00; 32S-3 transmitter with 516F-2 AC power supply \$725.00; 30L-1 linear \$400.00; Drake MN-2000 \$100.00; Waters 334A \$85.00. All equipment new March 1970. Write: Box D, Ham Radio, Greenville, NH 03048.

**TECH MANUALS** — R-390A/URR, BC-639A, OS-8C/U, \$6.50 each; R-388/URR, \$7.50; TS-323/UR, SP-600JX, \$5.50 each. S. Consalvo, 4905 Roanne Drive, Washington, DC 20021.

**APR-14 RECEIVER WANTED.** I am looking for a APR-14 surplus VHF - UHF receiver. Write or call Leroy Sparks, W6SYC, 924 W. McFadden Avenue, Santa Ana, Calif. 92707. Phone: (714) 557-8122.

**THE ENGLEWOOD AMATEUR RADIO ASSOCIATION, INC.,** invites all amateurs the world over to take part in the Eleventh New Jersey QSO Party from 1900 GMT Saturday, August 15 to 0600 GMT Sunday, August 16 and from 1200 GMT to 2300 GMT on Sunday, August 16. For complete details please contact the Englewood Amateur Radio Association, Inc., 303 Tenafly Road, Englewood, New Jersey 07631.

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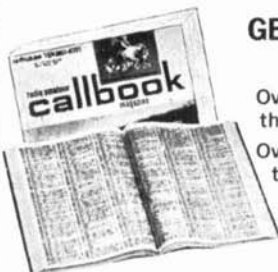
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**SAROC, JANUARY 7-10, 1971,** Flamingo Hotel Convention Center, Las Vegas, Nevada. Sponsored by Southern Nevada ARC, Inc., Box 73, Boulder City, Nevada. Advance registration \$14.50 per person accepted until January 4, regular registration at door, includes Flamingo Hotel Late Show and drinks, Sunday Breakfast, Cocktail Parties, technical seminars and meetings, ARRL, DX, FM, MARS, QCWA, WCARS-7255, WPSS-3952 and WSSBA. Ladies Program. Flamingo Hotel SAROC room rate \$12.00 plus room tax, per night, single or double occupancy January 3 thru 12, 1971. Mail accommodations request to Flamingo Hotel. Mail advance registration to SAROC. W7PRM, Club President. W7PBV, SAROC Convention Chairman.

**IT HAS BECOME NECESSARY** to change the date of the low 75 meter phone net picnic from August 9th to August 16. The picnic will be held in Marshalltown, Iowa at the Riverview Park. Festivities will begin around noon with the frequency of 3970 being monitored.

**CIRCUITS** for 32 electronic projects, R.F., audio and gadgetry, complete plans \$1.00. P.M. Electronics, Inc., Box 46204, Seattle, Wash. 98146. Dealer inquiries invited.

**BUILD YOUR LIBRARY!** Be sure to see the many excellent RSGB and other publications advertised elsewhere in this issue of Ham Radio (check advertiser's index for exact pages). RSGB publications are available only from us (in North America) and are the finest amateur reference material ever offered. Comtec, Box 592, Amherst, N. H. 03031.

**SIERRA HAMFEST** will be held August 22, 1970 at Mormon Station Historic State Monument at Genoa, Nevada just off Hwy. 395 south of Carson City. For details contact K7ZAU.

**SWAN 350** with ACPS, VOX, 510X, \$325.00. Swan 250 with ACPS, 5el Beam, \$225.00. Drake L-4, \$375.00. No shipping. W2EOS, AC315-457-7751.

**PRINTED CIRCUIT BOARD KITS.** Very easy method includes 6" x 10" circuit board, one pint etchant, tape, etc. Full instructions. Many sold at Hamfest. \$4.75 includes shipping. Electronic Circuits & Design Co., 33 E. Chestnut St., Alliance, Ohio 44601.

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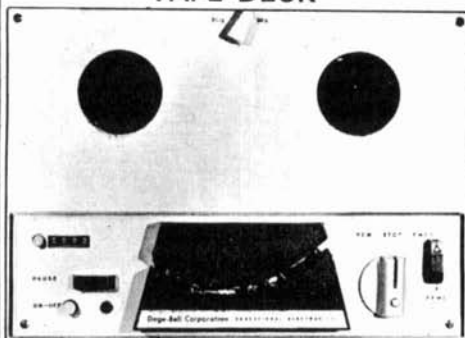


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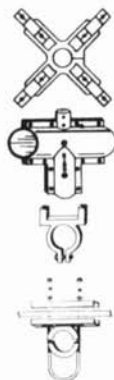
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**CINCY STAG HAMFEST:** The 33rd Annual Stag Hamfest will be held September 27, 1970 at Stricker's Grove, Compton Road, Mt. Healthy, Cincinnati, Ohio. Door prizes each hour, raffle, lots of food, flea market, model aircraft flying, and contests. Identify Mr. Hamfest and win prize. \$5.00 cost covers everything. For further info, contact, John Bruning, W8DSR, 6307 Fairhurst Avenue, Cincinnati, Ohio 45213.

**WORLD QSL BUREAU** — see ad page 91.

**THE WINNIPEG DX CLUB** announces a new award: The award is a personalized presentation case containing a genuine new Canadian Silver Dollar issued in honour of Manitoba's Centennial. All contacts must be made after January 1, 1970. 31 contacts are required, 5 each from Africa, Asia, Europe, North America, South America, Oceania and 1 contact from any Antarctic Station. I.A.R.U. continental boundaries apply. The 5 North American contacts must be with members of The Winnipeg DX Club. Contacts can be made on any band or mode but a station can only be counted once. QSL cards must be in the applicants possession, but need not be submitted if a verified list is sent. Amateurs worldwide are eligible. Canadian, American and Mexican applicants must be a member of DXCC, and must submit the number and date of their certificate. The cost of the award is 15 International Reply Coupons. If I.R.C.'s are not available in a particular country the applicant may send mint stamps of that country to an equivalent value. Members of The Winnipeg DX Club are VE4's — AA, AE, AS, BJ, CJ, IM, MP, TJ, RP, SA, SD, SK, XJ, ZX. Award Custodian is VE4AE, 22 Sweetwood Bay, Winnipeg 17, Manitoba, Canada.

**1970 INTERNATIONAL FIELD DAY.** The Burlington Amateur Radio Club, Inc., invites you to the 1970 INTERNATIONAL FIELD DAY to be held at THE OLD LANTERN, CHARLOTTE, VERMONT, on Sunday AUGUST 16. Come one day early on Saturday and join the crowd with trailer, camper, or tent. Featured will be the popular flea market, net meetings, antenna raising contest, junk box amplifier contest, hidden transmitter hunt on 2FM, mobile and portable RF output contest, bingo for the ladies, door prizes and raffle drawing. Talk in frequencies include 3909, WIKOO and WIABI repeaters. Food, refreshments, and camping supplies available all day. Registration \$3.00 at the gate or send \$2.50 for early bird to Bill Fake, W1FS, 30 South Hill Dr., Essex Junction, Vt. For further information write or call Bob Hall, W1DQO, General Greene Rd., Shelburne, Vt. 05482, phone 802-985-2235.

**COMMERCIAL LICENSE EXAMS:** Second \$18.00; First \$24.00; Sample questions; Price list \$1 refundable. Edco Enterprises, P. O. Box 432, Sparks, Nevada 89431.

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**WANTED MAY 1945 CQ.** Will trade November 1945. George Goldstone, W8AP, Northland Towers East, Southfield, Michigan 48075.

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**NEW THIS MONTH,** 3rd edition, Amateur Radio Techniques by J. Pat Hawker, G3VA. Latest edition of this popular book has 48 more pages and much valuable new material. New deluxe binding. Only \$3.50. Order today from Comtec, Box 592, Amherst, N. H. 03031.

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August 1970

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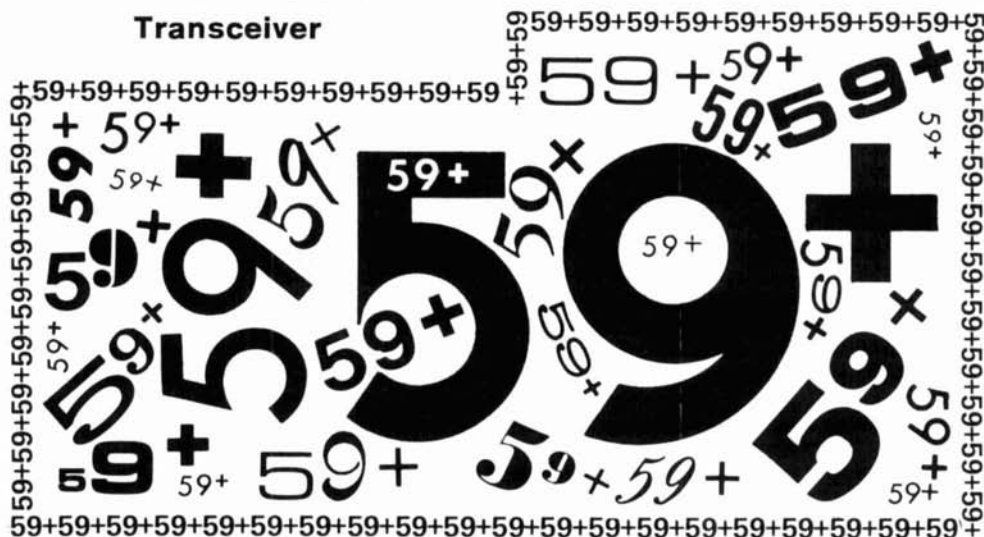
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**SPECTRONICS EAST**

Box 1457, Stow, Ohio 44224 / Telephone: (216) 923-4567



# home station performance

## with a mobile rig?

### NOTHING COMES CLOSER THAN THE SWAN 500CX

The performance of the Swan 500CX in your car may not quite equal your home station, but you will be amazed at how often the difference is very small. Many Swan mobile operators have established enviable DX records. QSL cards on file here at the factory verify solid contacts from all continents, including stations in Antarctica.

Old timers who have fought QRM with old style AM equipment have discovered a new world of mobile communication thanks to the power and punch of the Swan 500CX transceiver. The efficiency of the SSB mode coupled with the convenience of transceiver operation and a power level unmatched in the field represent a combination well worth looking into. So if you haven't heard ZL, DU or JA say station," then put a Swan 500 CX in your car this summer!

The 500CX installs conveniently under the dash of most cars, and the power supply goes under the hood. Add the Swan Model 35 high Q single band mobile antenna, or your choice of either the Swan Model 45 or 55 band switching mobile antenna, and you can enjoy mobile operation to the fullest. Home station performance with a mobile rig? Nothing comes closer than a Swan transceiver.

#### ACCESSORIES:

Model 500CX .....	\$565
Model 14-117 DC Supply .....	\$130
Mobile Mounting Kit .....	\$12
Model 35 Single Band Antenna, from ..	\$20.90
Model 45 5 Band Manual Switching Antenna .....	\$79.50
Model 55 5 Band Antenna (Remote control band selection) ....	\$129.50



See your Swan dealer today.

## SWAN

ELECTRONICS

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