

# ham radio magazine

# 15<sup>th</sup>

## ANNIVERSARY ISSUE

- 15-meter SSB transceiver
- precision noise bridge
- bobtail curtain follow-up
- capacitively coupled hybrids
- GaAs FET evaluation

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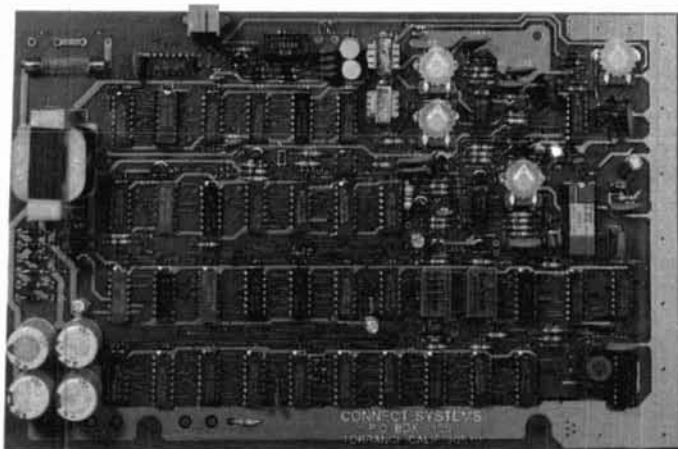
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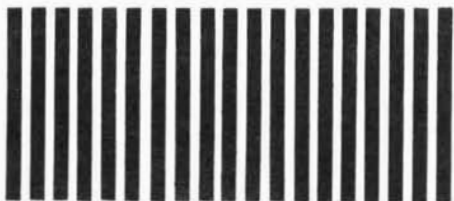


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# REFLECTIONS

Happy Birthday ham radio!

This issue, on our fifteenth anniversary, marks the completion of fifteen continuous years of providing the Amateur community with important technical information. We have attempted, and we hope succeeded, in bringing you a consistent flow of data: past techniques, still useful and important today; present techniques for solving today's problems; and looks into the future technologies that so rapidly are becoming a part of today.

This issue also signals the start of our sixteenth year of publishing in the same tradition, with articles that clearly emphasize the ham's true investigative spirit: always improving on existing techniques, making gear smaller, better, and less expensive. This month, that is best brought out by the articles by WA4ZXF and K2BT. Douglas Glenn (WA4ZXF) provides more than just another detailed design (and construction information) on a 15-meter transceiver. He shows, for example, how different operating functions can ingeniously be combined in the same circuit, thereby economizing on the number of components. A typical example is the use of the first mixer as a converter in the receive mode and as a balanced modulator during transmit.

Just because you're working with inexpensive components from the junkbox doesn't mean you have to put up with limited performance. Doug designs around them, taking full advantage of their characteristics. And the best part is that even if you don't build this unit, an understanding of a complete SSB transceiver is yours just for the reading.

Forrest Gehrke, K2BT, closely examines rf bridge operation — its inherent inaccuracies and limitations — in his article "Precision Noise Bridge." He shows, in his design, how real world impedances can be confidently and accurately measured. Forrest has used the noise bridge in building one of the most competitive 75-meter directional arrays in the world: the "4 square" (much more about that in several upcoming issues). With his bridge, he is able to accurately determine self and mutual impedances that unfortunately don't always come close to a resistive 50 ohms.


On a separate tack, Woody Smith, W6BCX, continues where he left off (in his Part 1, last month) with construction details for the half-square and Bobtail curtain. He closely examines feeding methods and locations, showing their influences on pattern and on the frequency bands that are usable with the Bobtail.

Dennis Mitchell, K8UR, provides additional insight into the GaAs FET technology. His circuit, designed around the Alpha Industries ALF1023 device, incorporates features that most hams can readily use in achieving higher-gain, low-noise preamplification inexpensively on 2-meters. This is especially significant since the trend, industrially, has been to use GaAs FETs at microwave frequencies (where their impedances are closer to 50 ohms), avoiding designs on the lower frequencies. Dennis evaluates seven different FETs for noise figure, gain, 1-dB compression, and cost.

Ernie Franke, WA2EWT, explores capacitively coupled hybrids, devices useful at any power level. A little bit of thinking shows several applications where a continuously variable phase shift at rf could be very useful, be it in the laboratory or out on the range (antenna range, that is).

These, as well as the other articles in this issue, are but a sampling of what we at *ham radio* have in store for you in 1983. Consider what you would like to see in the following issues; it will probably be there. Our crystal ball has recently been calibrated to accurately focus in on your needs. Nevertheless, we still welcome all of your communications. Let us know your views!

While on the subject of what *ham radio* has in store for the upcoming year, we take this opportunity to thank the people who've made *ham radio's* continued existence possible. First, of course, are our readers, without whose interest in excellence in radio communications there would be no *ham radio* magazine. We thank our advertisers, who have chosen our magazine for presenting you with the latest in communications gear — be it transceivers, antennas, or accessories. They are worthy of your support, for it's all a complete cycle — readers — advertisers — *ham radio* magazine. Let the advertisers know your feelings and needs. Tell them "you saw it in *ham radio*."

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Rich Rosen, K2RR  
associate publisher/technical editor

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## comments

### debunking myths

Dear HR:

The vertical versus dipole comparison by Bill Orr in "Ham Radio Techniques" (*ham radio*, October, 1982) should help debunk the myth that a quarter-wave vertical over a good radial system is automatically a low-angle radiator. An earlier article by John Belrose (*ham radio*, September, 1981, page 36) first got me wondering whether verticals are automatic guarantees of a DX edge over dipoles, inverted vees, and other simple antennas. Belrose showed that a quarter-wave vertical over very poor ground but with an excellent ground system has a radiation angle of 30-35 degrees, about the same as a half-wave dipole erected a half wave above ground. This has apparently been confirmed by Orr in his extensive 10 MHz listening comparisons.

What makes or breaks a typical vertical's low angle of radiation is the conductivity of the ground, out to 50 or more wave lengths from the antenna. (This factor has a much smaller influence on the take-off angle of a dipole of a half-wavelength or more above ground.)

Where does this leave the vertical which has become so popular for high frequency communications? I'd say that for 160 meters, 80 meter, and 40 meter long-haul QSO's, the vertical is still the best of the simple antennas for the average ham. A dipole must be a half wavelength above ground to give the 30 degree take-off angle the vertical will give at ground level. Most city-dweller hams can't

easily get a dipole that high. However, as Bill Orr demonstrates, the dipole probably becomes the better simple antenna from 30 meters on up. Exceptions would be the ham on a boat, in a salt-marsh, or possibly in an agricultural setting where continual use of fertilizers has given the soil superior conductivity out to 50-100 wavelengths from the antenna. Verticals in these settings will have take-off angles of 10-15 degrees. For Amateurs fortunate to be operating in these settings, the vertical still reigns supreme among the simple antennas for DXing from 30 meters on up.

Mark Bacon, WB9VWA  
Decatur, Illinois

### the 432 Yagi is alive and well

Dear HR:

The excellent article, "Requirements and Recommendations for 70-cm EME," by Joe Reiser (*ham radio*, June, 1982) mistakenly notes that "the K2RIW (432 Yagi) has gone out of production."

The K2RIW Yagi, the original RIW 432-19, is still in production by RIW Products. In fact, to assist the homebrewer, kits for the 432-19 are also available that include all the hard-to-get parts at a modest price.

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BOTH THE "NO-CODE" LICENSE AND AMATEUR EXAM ADMINISTRATION PROPOSALS are out, after the Commissioners voted to adopt the appropriate Notices of Proposed Rule Making on January 20. The "No-Code" is PR Docket 83-28, and the exam proposal is PR Docket 83-27. Both will have comment due dates of April 29. While the texts of both these crucial NPRMs had still not been released at press time, much of their content was already available.

The "No-Code" Proposal Is Very Much As Predicted in last month's Presstop. It offers two alternatives. The first is simply removing the Element 1A (CW) requirement from the present Technician Class exam, granting the successful applicant full privileges above 50 MHz. If such a licensee later wished to operate below 50 MHz, he could then be granted access to Novice frequencies after successfully demonstrating his CW ability to a higher-class licensee. A record of that "certification" would have to be kept in his station records, but there would be no notification to the Commission.

The Second Alternative Would Be The Establishment of a new "Digital Operator" class of license. As proposed, the "Digital Operator" would have full access to all Amateur bands above 144 MHz. Commissioner Dawson expressed a great deal of concern that restricting the frequencies available for this license would make it too unattractive. A new exam, "Element 5," perhaps incorporating present exam elements, would be written for the "Digital Class."

The FCC's Decision To Go Ahead On The "No-Code" License was made in spite of heavy opposition by the ARRL, and the item actually received relatively little discussion at the agenda meeting. League President Vic Clark, W4KFC, had urged the Commissioners, both in personal letters and visits, to hold off on no-code for at least 18 months to give the Amateur-administered exam program a chance to get rolling, but the Commissioners have been quite interested in seeing a no-code license established and were not receptive to seeing it delayed at this time. It's likely the League's position is not set in stone, either. In a Westlink interview, League Counsel Chris Imlay, N3AKD, noted that neither membership nor director opposition to no-code was unanimous and that the League position would certainly be up for review at the April Directors' meeting. In view of the strong sentiment in favor of no-code apparent within the Commission, it seems likely the League's final position could be one of tempering the no-code license to make it more acceptable to the majority of its members rather than continuing to oppose no-code outright.

The Proposal To Set Up A Volunteer-Administered Licensing Program generated much more discussion among the Commissioners and staff than did no-code. A good deal of the NPRM came from the ARRL's Petition for Rule Making, but with many changes from both within the Commission and from comments filed by the QCWA and others. One concern was that there must be more than only one Volunteer Examination Coordinator group; QCWA, IEEE, and OMICK (a national group of black Amateurs) were all mentioned as other possibilities along with the League. Another concern was that membership in a given group should not be required for examiner certification.

Specifics Of The Exam Proposal include the use of three-person exam teams, with the team chief being an Extra Class Amateur and team members either Advanced or Extra. CW exams for General or above would be administered by an Extra Class, and exams for the Extra Class license would require an all-Extra team. Exam questions would be drawn from lists made up by the FCC from submissions by individuals or groups. Teams could issue Interim Permits, just as Field Offices do now. Team members would have to be over 18, and could not be related to the applicant. No one working for a manufacturer or distributor of Amateur equipment, or for a publisher of training material, could serve as an examiner.

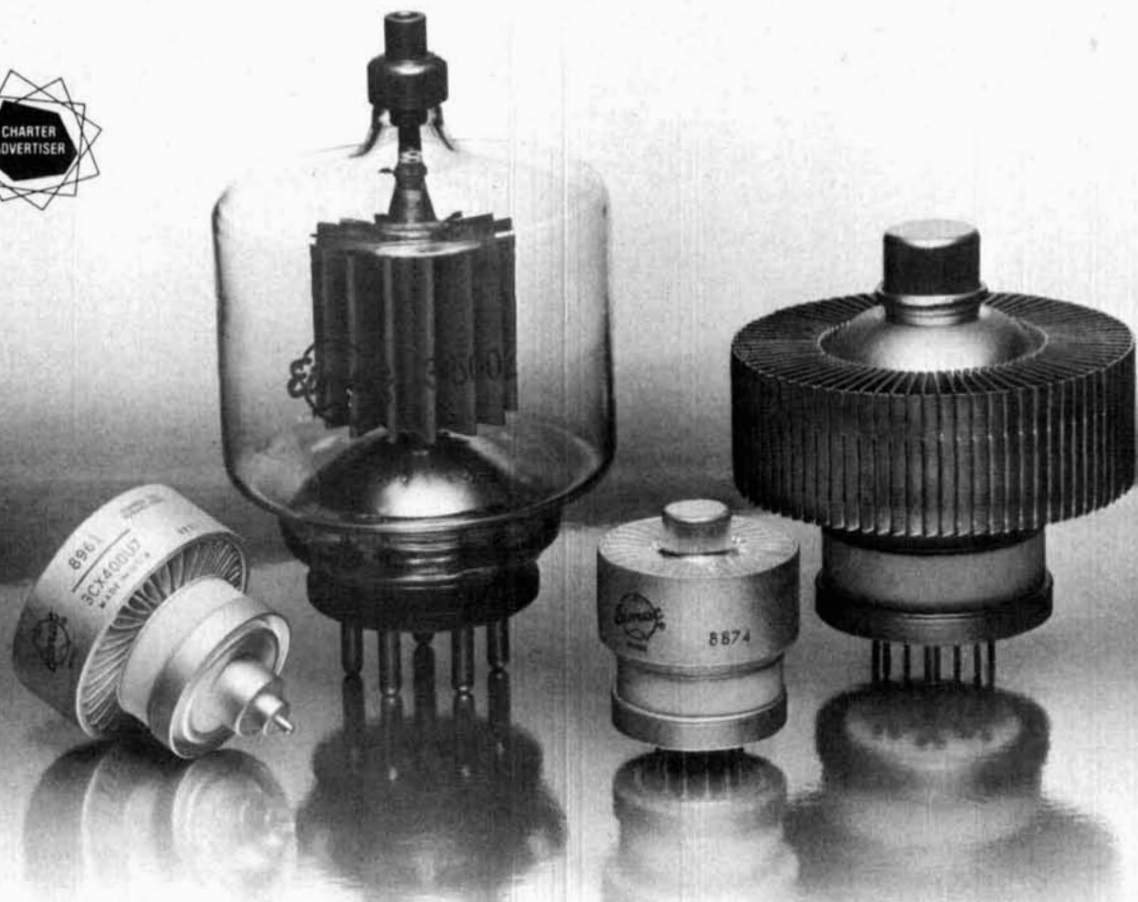
Though The ARRL Had Pushed Hard To Have The Novice Exam Included in this new exam proposal, the Commissioners elected to leave it as proposed late last year in PR Docket 82-727. Under the simpler procedure proposed in that NPRM, a Novice applicant could receive his license a few weeks after taking the exam. Under the more cumbersome procedures of the proposed higher-class exam program, Novice license processing would be considerably slower.

THE "AMTOR" DIGITAL SYSTEM WAS APPROVED for general Amateur use on 80 through 10 meters by the FCC on January 27. AMTOR is a synchronous RTTY technique, similar to the marine service's SITOR, which verifies message content and accuracy, thus permitting communications with much lower signal levels than RTTY otherwise requires. A number of Amateur stations have been using AMTOR under FCC Special Temporary Authority, and W1AW should begin using it within a few weeks.

GE'S PROPOSAL FOR A NEW 900-MHZ PERSONAL RADIO SERVICE WAS ADOPTED as an NPRM by the FCC on January 20. Located just below 900 and above 937 MHz, it would use self-identifying transmitters, and have automatic access to telephone service via the user's own base station or commercially operated repeaters. A similar service, though without repeaters, has just begun operation in Japan at 903-905 MHz (Region 3 did not get the 902-928 MHz Amateur band).

This Proposal Is Significant To The Amateur Service for several reasons. First, as a much more useful version of CB it's very likely to attract the interest of sophisticated would-be users who would otherwise be good candidates for Amateur Radio, with or without a no-code license. Second, the technology that such a service could support would spill over into Amateur Radio, particularly for the new 902-928 MHz band.

Comments On The Proposed New Service Will Also Be Due at the FCC on April 29. It has been assigned General Docket 83-26 by the FCC.



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# 15-meter sideband transceiver



A construction project  
of solid design  
with step-by-step explanations  
of each circuit function

With some careful shopping, it's possible to build a 15-meter sideband transceiver for less than \$100. The design presented here is for a basic rig at minimum cost, but with excellent features:

**Receiver:** 0.5  $\mu\text{V}$  sensitivity; 1.0  $\mu\text{V}$  agc threshold; 0.5-watt audio output; 2000-Hz selectivity; agc; 80 dB compressed to 4 dB; CW audio filter.

**Transmitter:** 10 watts PEP; fully adjustable output; CW break-in keying; 50-dB carrier and LSB suppression.

I designed this rig for 15-meters because propagation conditions on this band put low-power operators at less of a disadvantage than on the other bands. If you prefer, the rig could be built for the 20-meter band by using a different offset crystal frequency and by adjusting several tuned circuits. With an output that is fully adjustable from a pot, it makes a good exciter for a linear amplifier.

## theory of operation

Some of the circuit details may appear unfamiliar or unusual, but I will provide enough of an explanation for you to understand how the circuit operates. Component substitutions may be made to hold the cost down.

Let's start by reviewing the receiver signal path, beginning at the antenna jack (see **figs. 1 and 7**). No antenna relay is used. Instead, the input signal is coupled from the antenna low-pass filter to the receiver rf amplifier through a series-resonant circuit, C1-L1. This series-resonant circuit is used both as a T-R switch during transmit and as an rf attenuator during receive.

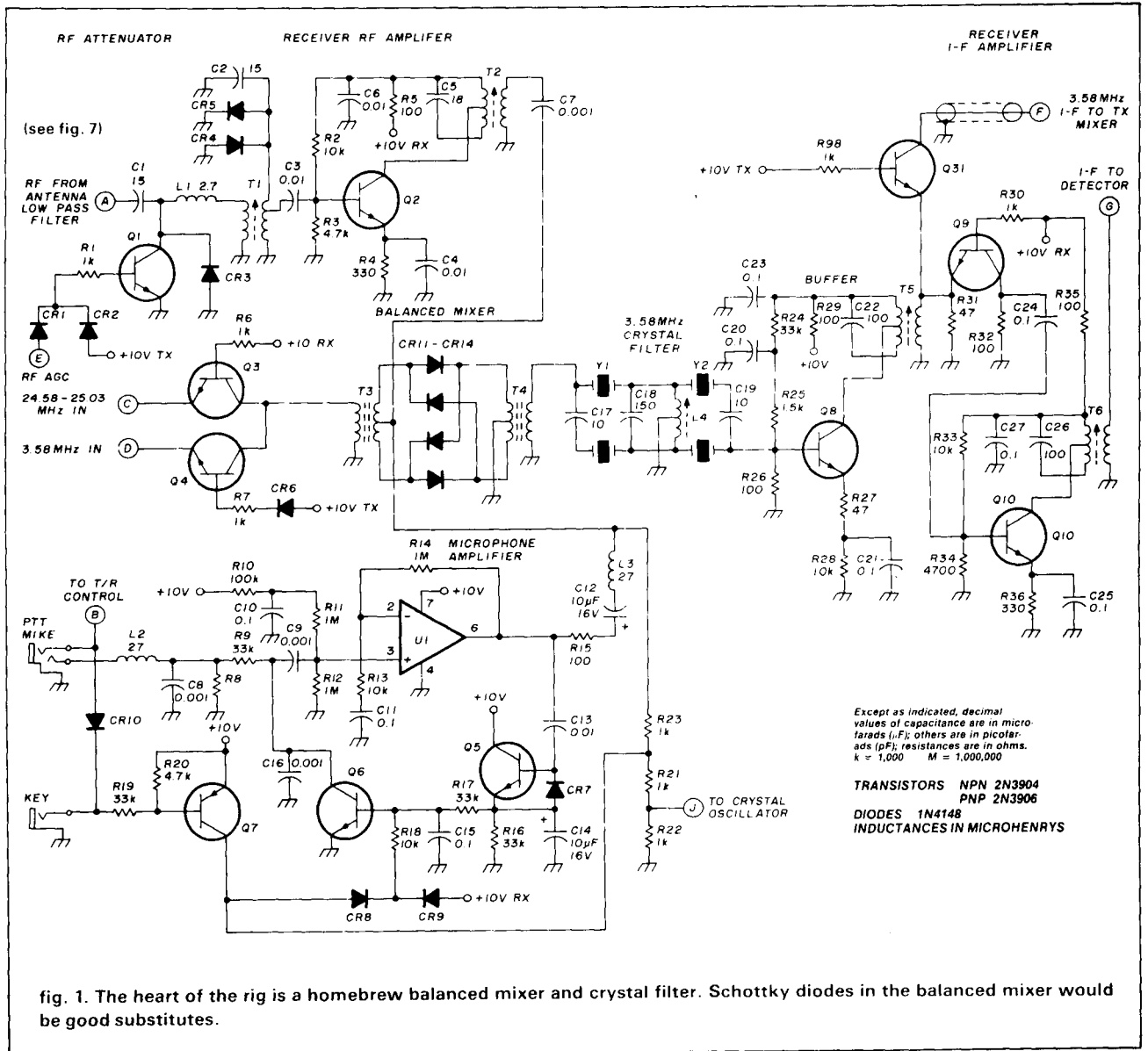
A series L-C circuit presents a low impedance at resonance, while the center junction is at a high im-

pedance. If you put a variable resistor from this junction to ground, you create a variable attenuator. But instead of a resistor, use the collector impedance of a transistor. The transistor looks like a small capacitor (junction capacitance) in parallel with a variable resistance. The resistance value is controlled by the dc base current, as shown in **fig. 2**, as long as the signal amplitude doesn't forward-bias the transistor junctions. A typical value is less than 10 ohms with 1 mA of base current, and it increases to nearly infinity as base current reduces to zero. The effect of the junction capacitance must be compensated for, so avoid transistors with large capacitance. A diode across the transistor protects the receiver's front-end from other transmitters feeding antennas on the same tower. The circuit can also be used for filtering if desired, but it doesn't have to be. In this case, filtering was not a concern.

There is a trade-off that must be made. As the L/C ratio gets larger, you get higher  $Q$  and more attenuation range. But as the capacitor value approaches the value of the shunt capacitance from the transistor and diode, the capacitive voltage-divider effect significantly attenuates the input signal. The L1-C1 values were chosen for a good compromise, and they have low  $Q$  so that fixed components with normal tolerances can be used.

The input signal from the rf attenuator is coupled to an rf amplifier, Q2, with about 20 dB of gain. Tuned transformers are used at the input and output of this stage to prevent image and i-f feedthrough. The transistor that I used was an inexpensive general-purpose type, the 2N3904. This transistor's input impedance is about 50 ohms, so the transformer has unity impe-

By Douglas Glenn, WA4ZXF, 715 Fairlane Drive, Lewisburg, Tennessee 37091



dance ratio from input to base. The circuit *Q* is not allowed to get too high, which would hurt sensitivity at the band edges. Diodes across the high impedance winding give extra protection from high-level signals.

The output of the rf amplifier is coupled to the first mixer by a capacitor, C7. This same mixer is used as the balanced modulator during transmit. The mixer is a double-balanced, diode-ring-type, selected because it provides good performance, is inexpensive, and is easy to build. The output transformer is wound to provide the impedance needed by the crystal filter instead of the usual 50 ohms. Good carrier suppression requires matching of the diodes. But this is easy, as explained later.

The mixer input that is used for the signal from the receiver's rf amplifier is also used as the audio input

during transmit. To avoid the need for a switch of some kind at this mixer input, the high frequency signal in receive is coupled through capacitor C7, blocking the low-frequency audio during transmit. Similarly, coil L3 in series with the audio blocks the rf.

During receive, the local oscillator input to the mixer is from the high frequency VFO, but in transmit this same input is from a lower-frequency crystal oscillator. Two transistors (Q3 and Q4) provide the mixer input switching function. A relay could be used but it would probably cost more, take up more board space, and draw more power-supply current. This analog switch doesn't provide the high isolation of a relay, so you may notice a birdie 30 kHz above the top edge of the band ( $6 \times 3.58$  MHz) in receive. To minimize this, a diode is placed in series with the base of Q4, increasing the turn-on threshold.

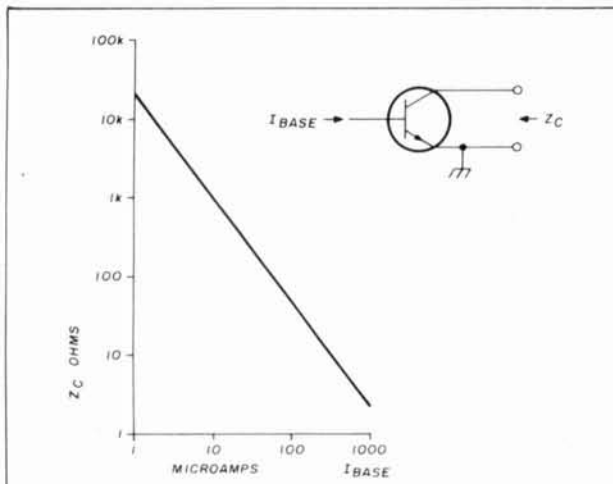


fig. 2. Resistive part of small-signal collector impedance versus dc base current. At high frequencies, the collector junction capacitance will appear in parallel with the resistance.

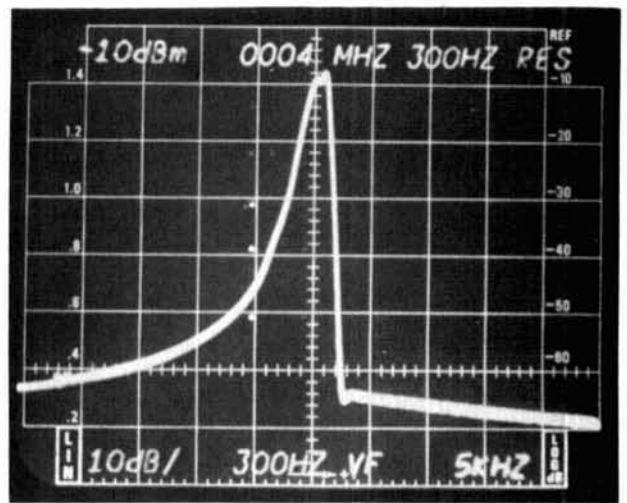


fig. 3. Crystal filter response with 1000 ohm source and load. Insertion loss is 5 dB.

### crystal filter

An inexpensive crystal filter requires inexpensive crystals. At present, the most readily available and inexpensive crystals are TV color-burst crystals at 3.58 MHz, which are used here. The circuit is a standard four-crystal lattice. This requires two sets of two crystals with a small frequency difference. Since

color-burst crystals all come at the same frequency, small capacitors are inserted in series with the crystals on opposite corners of the lattice to provide the needed frequency difference. The value of the capacitors establishes the filter's bandwidth. I used capacitor values of 10 pF, which results in about a 2-kHz bandwidth. Smaller capacitor values will give wider bandwidth. You should not try to make it too

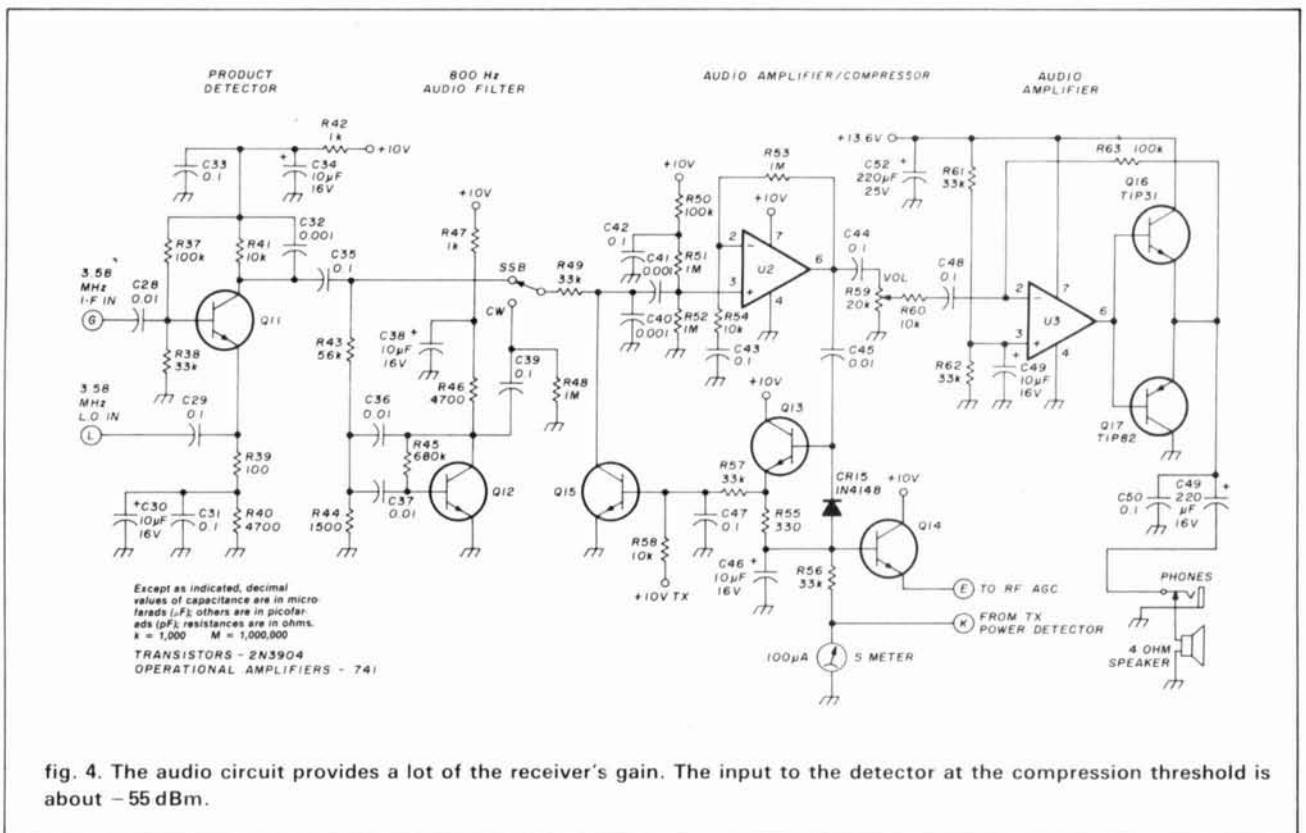


fig. 4. The audio circuit provides a lot of the receiver's gain. The input to the detector at the compression threshold is about -55 dBm.



much wider or you will have trouble adjusting the 3.58-MHz oscillator later. I used the small, round, axial-leaded, titanium-dioxide capacitors found in many TVs and fm tuners, but only because I had some handy; other ceramic or mica 5-percent-tolerance types will work. **Fig. 3** shows the shape of the filter. The shape obviously dictates use for the lower sideband. Since the upper sideband is commonly used on 15-meters, high side injection must be used in a mixer to invert the sidebands.

The filter's output is terminated by a buffer amplifier, Q8. The 1000-ohm filter load is supplied by the base bias resistors in parallel with the transistor's input impedance. Part of the amplifier's emitter resistance is un-bypassed to keep the transistor's impedance high. This sacrifices some gain, but allows for accurate control of the filter's termination impedance by the fixed resistors.

The output of the buffer amplifier is coupled to an i-f amplifier, Q10, in receive, or to the second conversion mixer in transmit. Transistors Q9 and Q31 are again used as analog switches. In receive, the i-f signal is amplified and sent to the receiver's detector. The receiver i-f amplifier is turned off during transmit to prevent the possibility that the oscillator signal might bypass the crystal filter.

## receiver audio circuits

The receiver's detector (see **fig. 4**) is a single transistor, Q11, configured as a mixer for product detection. Signals feed the base while the local oscillator is injected at the emitter. The collector load of the detector is bypassed for rf, leaving only the recovered audio.

The audio is coupled to an active unity gain band-pass filter, Q12, centered at 800 Hz. Since neither high gain nor high  $Q$  is needed, I couldn't justify using an op-amp when a single transistor would do. The formulas that you would use with an op-amp don't work with the low-input impedance of the transistor. The filter  $Q$  and center frequency will be less than predicted. As shown, the filter is centered at 800 Hz with a 300-Hz bandwidth. There's no need for precision components in the filter because the filter's performance can be easily adjusted using the two voltage divider resistors at the input (R43 and R44). The small resistor to ground, R44, will vary the center frequency (less resistance for higher frequency) and the large resistor in series with the input, R43, is then changed to set the gain again to unity. A SPDT switch bypasses the audio filter for phone reception.

The wide or narrow audio from the bandwidth switch is coupled to the receiver's audio amplifier/compressor. This circuit is configured as a current-controlled attenuator followed by a 40-dB gain amplifier. I used 741-type op-amps because they are inex-

pensive and readily available. The op-amp is not being used near its maximum gain or slew rate, so you can get by with a 10-volt supply. The attenuator is ahead of, not inside, the feedback loop and consequently loop gain remains high at all input levels. This arrangement provides tight control of the audio level. You'll never have to dive for the gain control or claw at your headphones when a strong signal comes on frequency.

The attenuator is a voltage divider that consists of a fixed resistor, R49, and the variable impedance of Q15, which has the characteristic shown in **fig. 2**. This controls the signal level at the op-amp input. A transistor peak detector, Q13, at the amplifier's output provides drive to the attenuator in the form of base current to Q15. The compression threshold occurs when the peak-to-peak output of the amplifier exceeds the base-emitter diode drop of Q13. This corresponds to 210 mV rms for a silicon transistor, or 2 mV at the compressor input. The circuit features a fast-attack, slow-decay characteristic compatible with speech signals. The circuit also limits sharp noise pulses. A small resistor, R55, is placed between the emitter of the peak detector transistor and the large storage capacitor, C46. Short noise pulses are coupled directly to the attenuator, but only slightly affect the charge on the storage capacitor. A smaller capacitor, C47, is placed across the base-input of the attenuator transistor; it serves only to eliminate the audible pop that would occur when the input signal crosses the compression threshold. Once in compression, the voltage across this capacitor is fixed at the potential of a base-emitter diode drop and will not limit the response to the noise pulses or the fast attack characteristic. During transmit, the receiver's audio gain is killed by forcing the attenuator transistor to saturate. This doesn't charge the storage capacitor, so the receiver is at full gain and ready to go as soon as the transmit bias is removed.

The voltage on the storage capacitor is fed to the receiver's front-end attenuator as rf agc. An emitter follower, Q14, is used to reduce the loading on the storage capacitor and a diode is placed in series with the emitter follower output to protect it from reverse bias damage during transmit. The result is that the rf agc voltage is reduced by two diode drops. This provides delay in the activation of the rf attenuator so that the input is not attenuated until the signal is well above the noise.

The voltage on the storage capacitor is also used to operate the receiver's S-meter. The S-meter current helps to control the decay time of the receiver agc and it ensures that the capacitor discharges completely when there is no signal received. If you don't use a meter, replace it with a short; don't omit the discharge resistor. I used a surplus meter that I

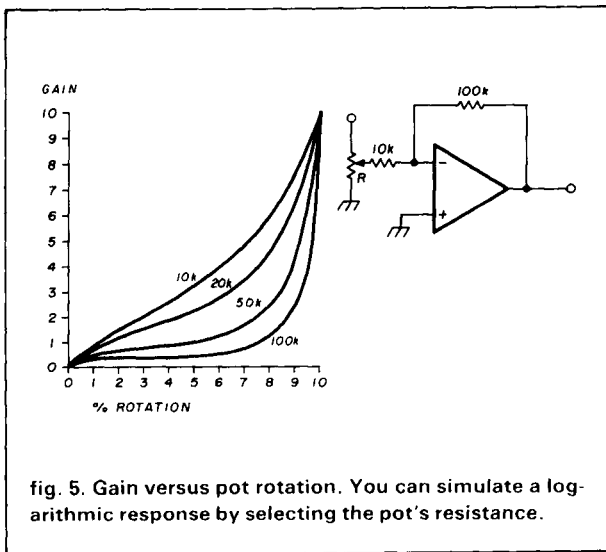


fig. 5. Gain versus pot rotation. You can simulate a logarithmic response by selecting the pot's resistance.

picked up at a hamfest for half a dollar. I painted over the face and drew a new scale. The meter doesn't read until the received signal is above the agc threshold at 1 microvolt. This corresponds to an S-level of about 3.5, so the first mark on the meter is S-4. You can add capacitance across the meter's terminals to

increase the damping of your meter if you want. You'll have to experiment to find the right value.

The audio output amplifier uses an op-amp followed by a high current buffer. The buffer transistors are not biased into the linear range, but they are included inside the feedback loop. The slight amount of crossover distortion won't be noticed under normal conditions, and it's a small price to pay for circuit simplicity and the saving in quiescent bias current. A capacitor was found to be necessary across the output to bypass rf pickup that had the effect of making the audio amplifier squeal during transmit, especially when I was using an external speaker.

The output amplifier is operated directly from the external 13.6-volt supply. Power supply hum is not a problem because of the op-amp's good power supply rejection ratio and a heavily bypassed reference (non-inverting) input. Operating from a higher supply voltage also increases the maximum output power available. In addition, stability of the audio circuits is improved because the supply voltage for this high level circuit is isolated from the supply for the low-level circuits.

One interesting effect of this circuit is that a linear volume control gives a non-linear response to the cir-

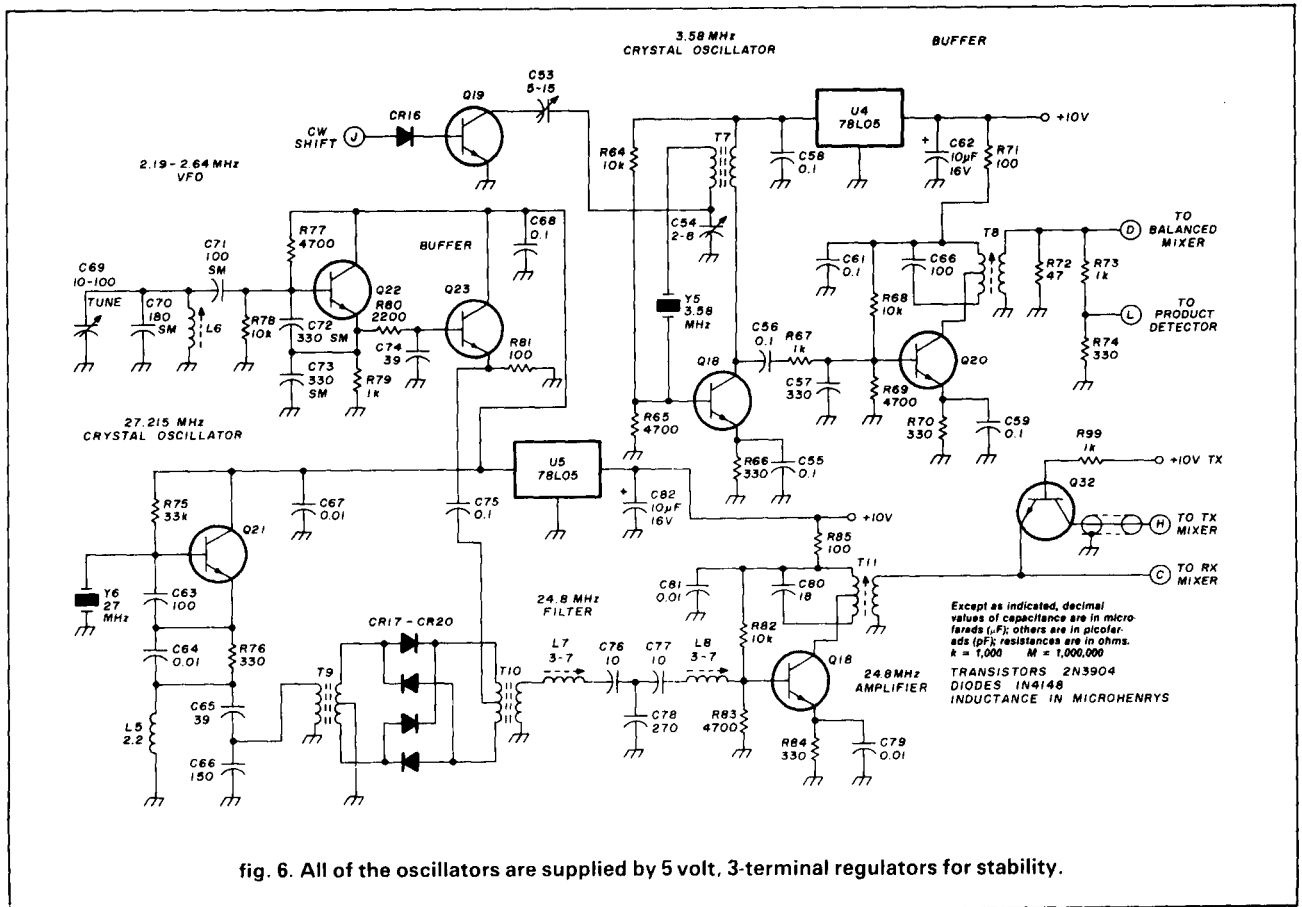


fig. 6. All of the oscillators are supplied by 5 volt, 3-terminal regulators for stability.

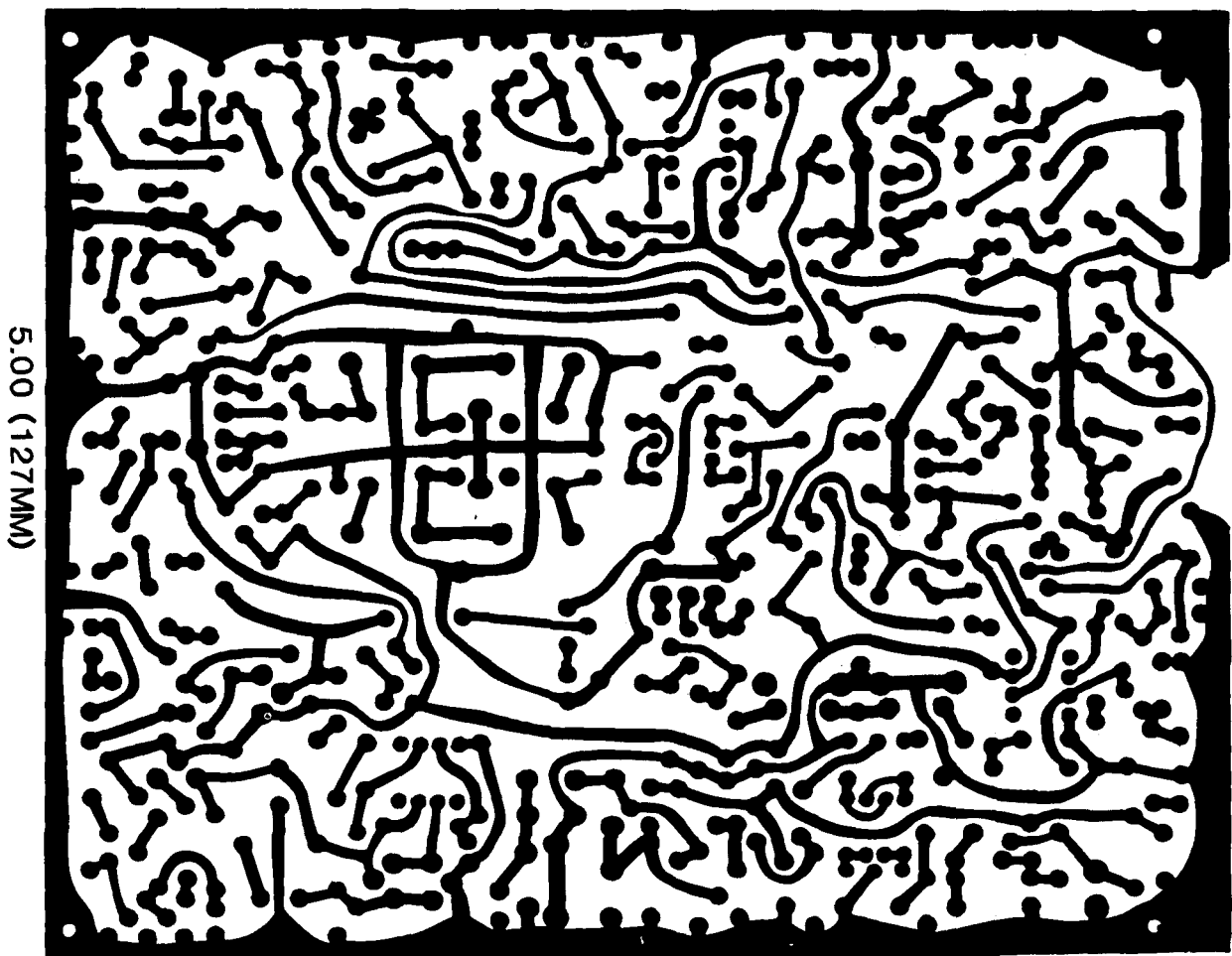
cuit. The shape is not the true logarithmic curve of audio controls, but the overall effect is similar. Part of the pot's resistance adds to the resistance in series with the op-amp input and so changes the circuit's gain slope as the pot is rotated (see fig. 5). The effect is more pronounced for pots with larger resistance values. If you use a pot that is significantly more than 20 kilohms, you may want to proportionately increase the input and feedback resistors.

### transmitter circuits

Transmitter circuit review begins at the microphone input. (see fig. 1) The microphone amplifier uses a compressor amplifier (U1, Q5, Q6, etc.) nearly identical to that used in the receiver. By using a compressor that maintains a constant audio level, the other circuits can be optimized for this level. The only circuit difference (from the receiver's compressor) is one missing resistor. The resistor is omitted since the microphone amplifier doesn't have to contend with receiver impulse noise.

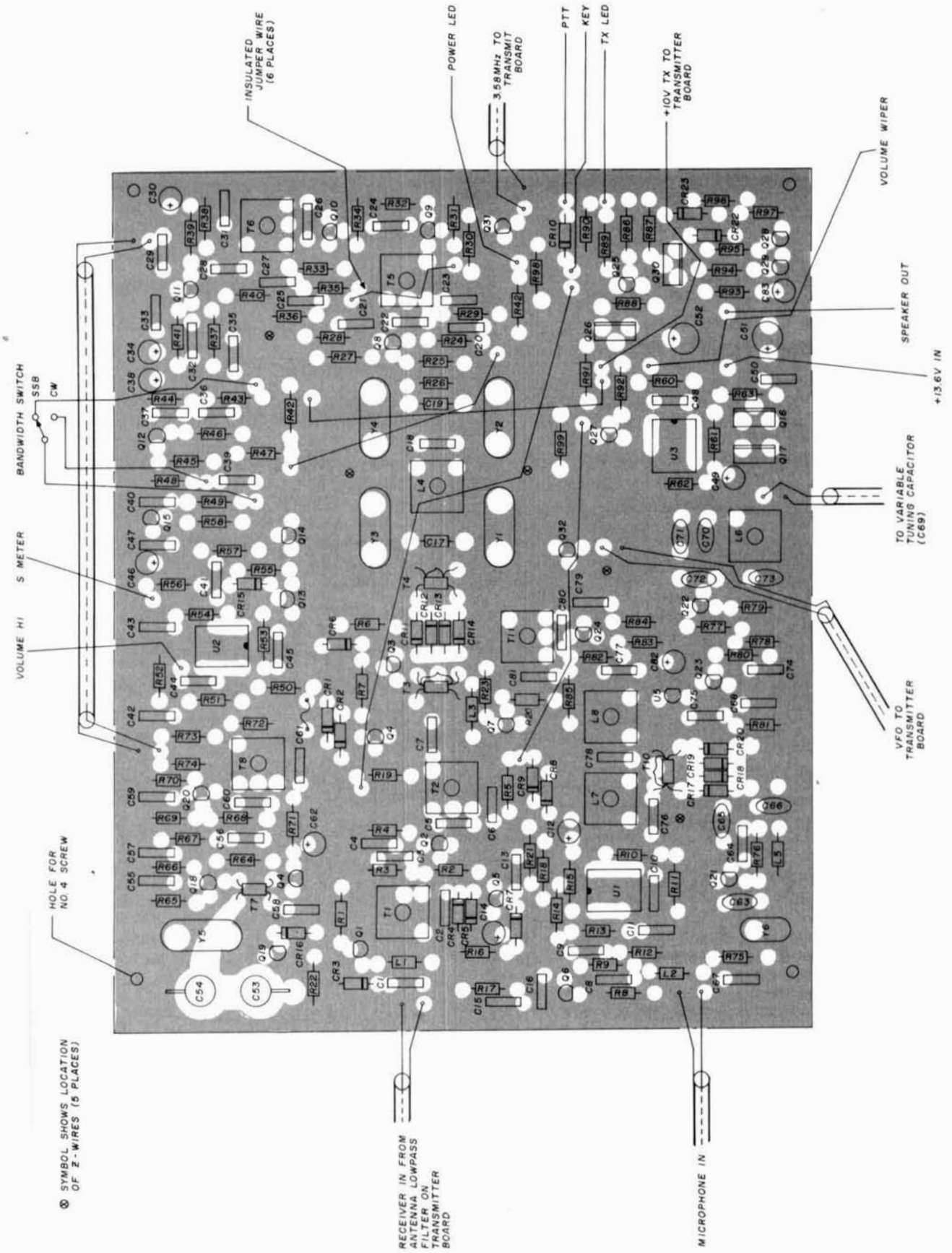
The amplifier input is terminated for a dynamic microphone. This is because the inexpensive, readily available microphones for citizen band sets are usually of this type. A filter consisting of a coil and capacitor keeps rf pickup out of the radio. A higher input impedance, for another type of microphone, would require some redesign of the input.

The output of the microphone amplifier is coupled to the balanced mixer. The carrier signal is suppressed at least 30 dB by the balanced mixer. The signal then passes through the crystal filter, which removes the upper sideband and suppresses the carrier another 20 dB. When keyed for CW, the mixer balance is upset intentionally and the carrier comes up to full level. A transistor switch, Q7, supplies current to the dc-coupled input of the mixer to unbalance the mixer. At the same time, this switch sends current to the current-controlled attenuator of the microphone amplifier. This kills the gain of the microphone amplifier and prevents audio from being superimposed on the CW signal.



6.25 (159MM)

Main circuit board PC layout.



Main circuit board, groundplane side with components layout superimposed.

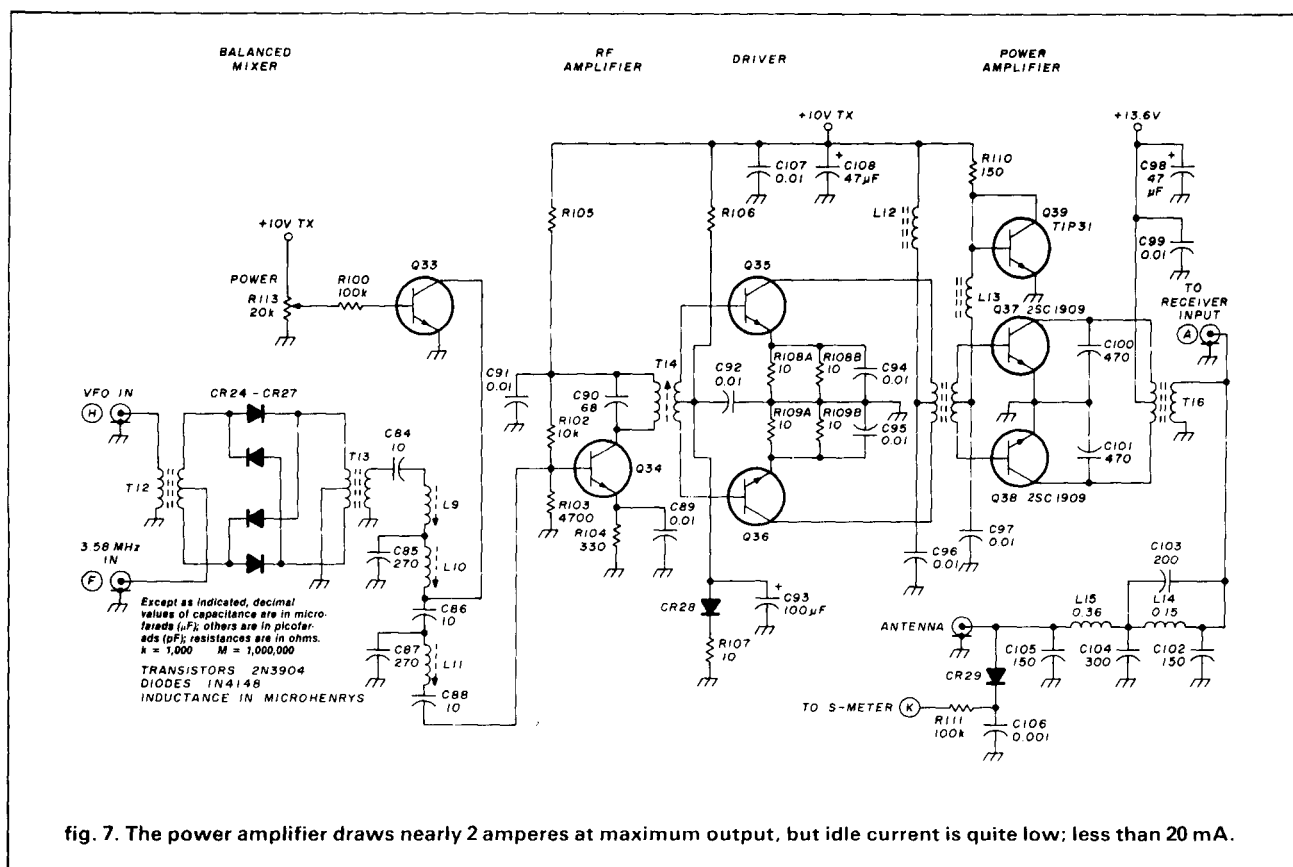


fig. 7. The power amplifier draws nearly 2 amperes at maximum output, but idle current is quite low; less than 20 mA.

### carrier oscillator

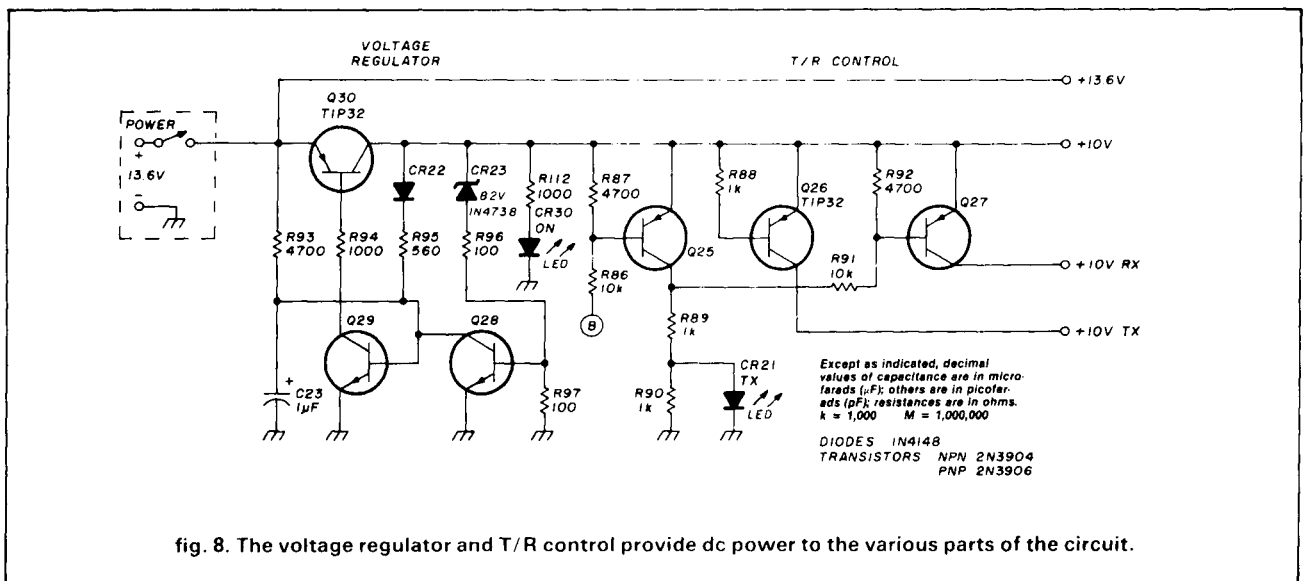
The carrier oscillator (see fig. 6) also uses a 3.58-MHz color-burst crystal. The oscillator circuit configuration was selected for its very good frequency stability. The phase inversion, from collector to base of the transistor, is provided by a transformer rather than some form of resonant circuit. This helps stability by removing all frequency-sensitive components except the crystal and a small capacitor. The capacitor shifts the series resonance of the crystal-capacitor combination to the desired frequency. The transformer is constructed by winding magnet wire on a ferrite bead. The capacitor can be a fixed type that's selected for the proper frequency, or a variable if it's a stable type.

The capacitance is located in the ground leg of the series circuit so that another capacitor can be switched in parallel to shift the frequency. This is done when the rig is keyed for CW. The purpose is to shift the frequency by 800 Hz. When properly adjusted, the output in transmit will be the same frequency as a received signal centered in the CW audio filter pass-band. A transistor is used as the switch. A low capacitance transistor is absolutely necessary in this location.

The output of the oscillator is coupled to a buffer stage, Q29, through a low-pass R-C section. The R-C section does three things: The series resistor isolates the oscillator from the rest of the circuit, the low-pass section filters the signal providing a sine-wave output, and the signal is adjusted for optimum drive to the mixer. The value of the series resistor, R66, or shunt capacitor, C57, can be changed, if necessary, for more or less mixer drive. The best drive level would be +7 dBm (a 0.5 V rms). However, this level is too high for the transistor analog switch. I set the drive level to approximately 0.35 V rms (+4 dBm) with satisfactory performance. A voltage divider across the output of the buffer sets the proper drive level for the receiver's detector, about 60 mV. This level is not critical. Any value from 40 to 220 mV will work, but you should keep it toward the low end of the range. At the higher drive levels, the receiver's detector is more susceptible to stray signals such as the ever-present 60 Hz hum.

### variable frequency oscillator

The high-frequency VFO that tunes the transceiver is critical to the rig's performance. It should be stable and not change frequency when the transmitter is



keyed. The design uses a low-frequency variable oscillator which is mixed with a high-frequency crystal offset oscillator to arrive at the needed injection frequency.

High-side injection is necessary to invert the sideband, as explained above. To cover 15 meters, the injection frequency is 24.58-25.03 MHz. The least expensive and most readily available crystals, after color-burst crystals, are 27 MHz units intended for multichannel citizen band HTs. One of these is used for the offset oscillator. They are usually sold in T-R pairs for less than \$5. I used a channel 21 transmit (27.215 MHz) crystal. The exact frequency is not very important since the VFO is aligned to set the final frequency. If you have a choice, use a higher channel crystal since this will improve spurious responses. The output of the crystal oscillator is coupled to the offset mixer by a capacitive voltage divider that couples the proper level to the mixer.

The low frequency variable oscillator, Q22, is a standard Seiler type that tunes 2.19-2.64 MHz to cover 15-meters. Silver mica capacitors are used in the frequency determining circuit for improved stability. The variable capacitor used is from the oscillator stage of a two-section unit salvaged from a five-tube ac-dc broadcast receiver. If you use a variable capacitor that is significantly different, you may have to size some of the fixed capacitors proportionately. This is particularly true if your variable has more capacitance, because the oscillator quits if the L/C ratio gets too low. Increasing the value of C71 usually cures this problem. A 5-volt, three-terminal regulator, U5, operated from the regulated 10-volt supply, provides a stable, doubly regulated supply that is also used by the offset crystal oscillator. Frequency stability is good. Don't use just a zener diode; use the IC regulator. The variable oscillator output is coupled to

the offset mixer through an R-C section and emitter follower, Q23, that isolates the oscillator, reduces signal harmonics, and sets the signal to the proper level for the mixer.

The mixer is a doubly balanced, diode-ring type with a two-pole, series-resonant type filter at its output. By using a balanced mixer, the oscillator feed-through at 27 MHz is suppressed reducing the following filter requirements. The mixer design is for a nominal 50 ohms at each port, and a commercial mixer can be substituted here. The spurious responses that result from the undesired mixer outputs, as well as the transceiver image, are above the 15-meter band so that the antenna low-pass filter aids the resonant circuits of the transceiver in minimizing spurious signals. A buffer amplifier, Q24, follows the filter that provides +3 to +4 dBm drive level to the mixers.

### transmitter rf circuits

The rf circuits of the transmitter (fig. 7) are activated by the microphone PTT or key. The modulated 3.58-MHz signal is applied to a balanced mixer where it mixes with the variable frequency oscillator injected signal. The balanced mixer is the same type as the one that mixes the two oscillator signals and is at the normal 50-ohm impedance level. (A commercial mixer could be used instead.) The mixer output is passed through three poles of series-resonant filtering. This is the same type of series-resonant circuit used elsewhere. A shunt transistor, Q33, is used at the second pole. The base bias of this transistor is controlled by a pot and it becomes the transmitter power control. The fixed resistor in series with the transistor's collector determines the maximum power reduction. The control can vary the transmitter output at least 30 dB (10 watts to 10 milliwatts). The

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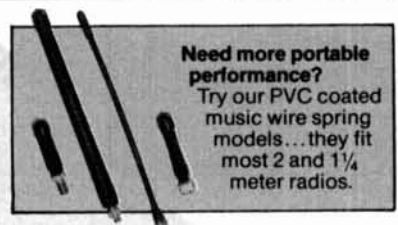


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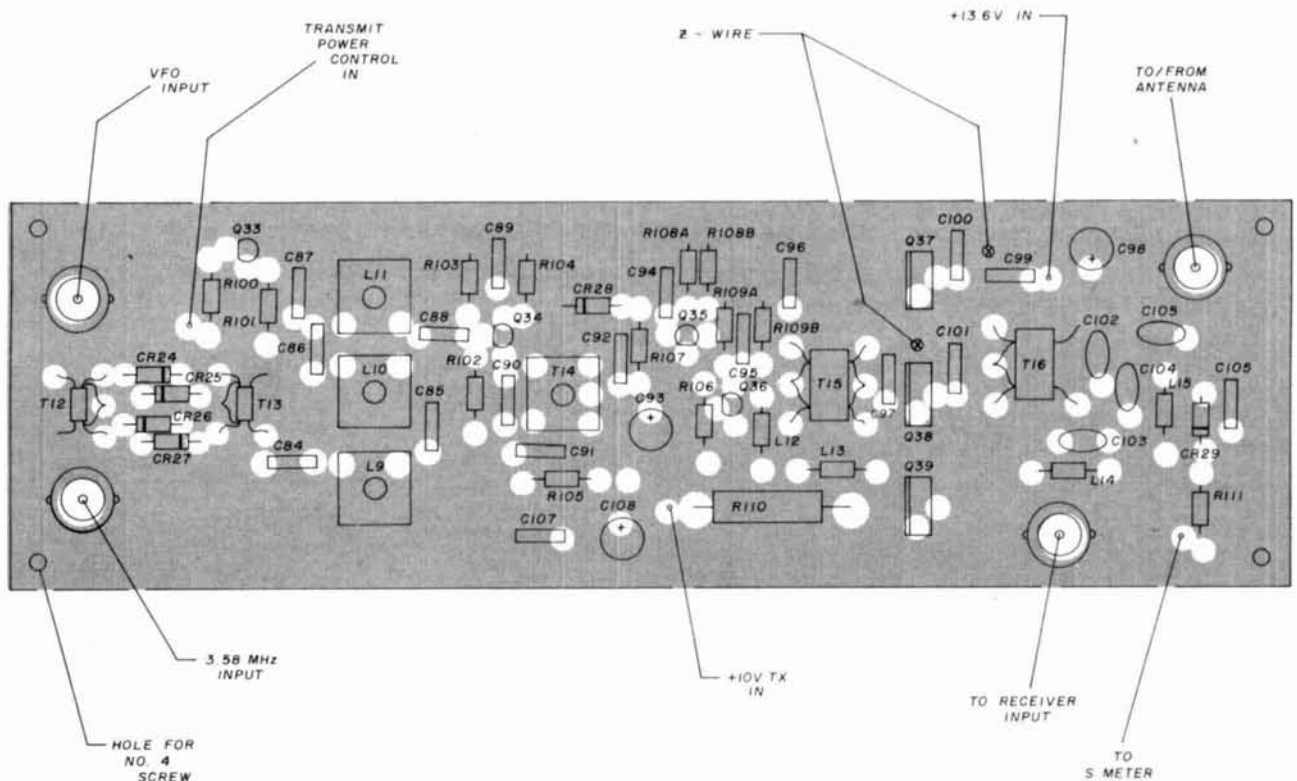


Transmitter board PC layout.

attenuator is placed after all of the mixing stages to attenuate the low level spurious signals along with the desired output.

The mixer output, after filtering, is amplified in turn by a class A stage and a push-pull driver stage running class AB, prior to the final. The final amplifier, also running class AB push-pull, uses a pair of transistors intended for the output stages of 5 watt citizen band transmitters (2SC1909 from Radio Shack). Base bias current is controlled by transistor

Q39 connected as a current mirror. The bias transistor is a general purpose TO-220 type that is attached to the same heatsink as the finals for temperature compensation. The input and output transformers of the final are wound on the two-hole ferrite balun cores that are found at the VHF input of many TV sets. The ones I used were salvaged from the coupler of a TV game. The final output is passed through a low-pass filter. The filter was designed to put a notch at the mixer image frequency, which is helpful in



Transmitter circuit board, groundplane side with components layout superimposed.



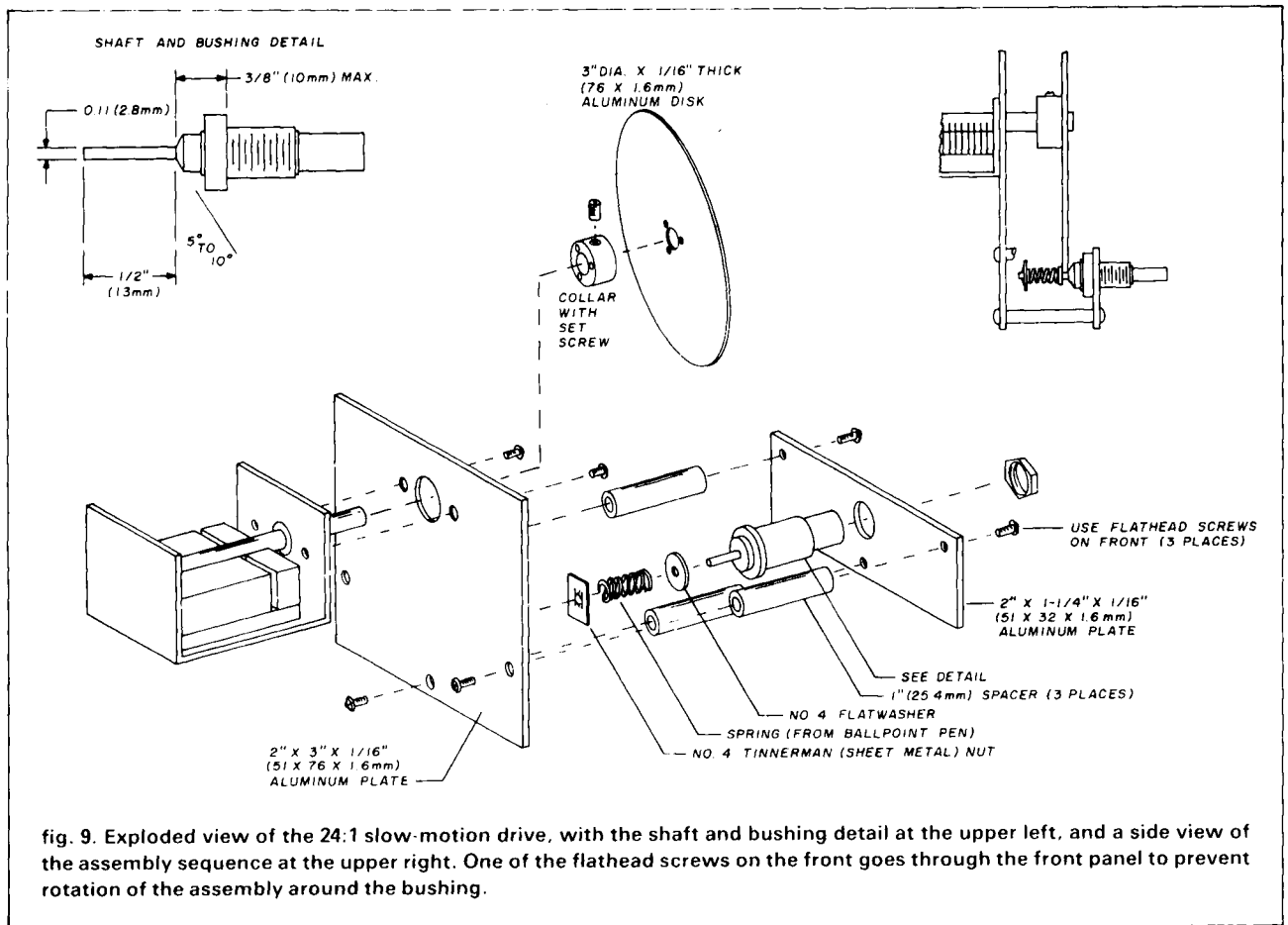


fig. 9. Exploded view of the 24:1 slow-motion drive, with the shaft and bushing detail at the upper left, and a side view of the assembly sequence at the upper right. One of the flathead screws on the front goes through the front panel to prevent rotation of the assembly around the bushing.

both transmit and receive. A diode detector at the antenna output supplies a readout of transmitter power on the front panel S-meter.

### power and control circuits

The dc power for all circuits, except the transmitter final and the receiver audio output (fig. 8), is supplied by a 10-volt regulator. The design, which was adapted from one in a Fairchild Semiconductor application note, provides good regulation even when the input voltage drops close to the regulated level. The schematic shows an 8.2-volt zener diode, but a 7.5-volt or 9.1-volt diode also works if R96 is adjusted to compensate. Don't use R97 to adjust the voltage; its purpose is to set the zener current. The circuit also has a coarse form of current limiting. This occurs when Q29 saturates and the pass transistor, Q30, gets no more base drive. The point of current limiting can be set by adjusting the value of R94.

The dc power is switched to the various sections of the transceiver by a T-R control circuit. When activated by the PTT switch or CW key, power is applied to the transmitter circuits and removed from some receiver circuits. Current is also supplied to the several transistors that act as variable impedances throughout the transceiver. Most of the receiver cir-

cuits do not have the dc power removed in transmit. This enables quick recovery of the receiver for break-in keying.

### construction

All of the circuits are contained on two doubled-sided printed circuit boards. The transmitter's mixer and amplifier chain are on one board, and everything else is on the second. This separates the high-level transmitter circuits from the other low-level circuits. You'll notice from the board layouts that the top (component) side is devoted to ground plane. Jumper wires are used where necessary rather than disturb the continuity of the ground plane. If, like me, you don't have the capability of plated-through holes on circuit boards, solder a number of component leads on top and bottom to link the ground plane. I usually did this only on resistor leads and the few Z-wires where access to the top of the lead is easy. This is particularly important on the transmitter board and especially for the emitter leads of the P.A. output transistors which need a good solid ground.

The transceiver was housed in a cabinet (Radio Shack #270-270) that provides plenty of space. The main PC board is mounted on the bottom of the cabinet using metal standoffs, which grounds the circuit

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to the cabinet. The transmitter board is mounted on the rear wall the same way. The layout of the transmitter board includes provisions for four PC mount phono jacks, that are used for the rf connections. This was done to simplify removal of the board, but a soldered connection is the obvious way to go if you can't find PC phono jacks. The heatsink for the transmitter's final is a 2.5 x 4 inch (60 x 100 mm) piece of 1/8-inch (3-mm) aluminum plate that is secured to and insulated from the three power transistors that stand in a line across the PC board. The heatsink is not grounded and not supported in any other way. The speaker is mounted on a bracket behind perforations in the cover since the cover has to slide on when installed. The material that forms the front panel of the cabinet is soft aluminum, but it is secured at all four corners when the cover is in place. To increase the rigidity of the front panel, I used a piece of 0.062-inch (1.6-mm) aluminum cut to fit the front panel. It is held in place by the bushings of the various panel mounted components. It is also easier

to paint and letter this flat plate and it covers four holes in the cabinet's panel intended for handles.

Because of the wide variations in tuning capacitors that might be used, mine wasn't mounted on the circuit board. Instead, it was mounted on the front panel of the cabinet and connected to the PC board with a short length of shielded wire. The details of the tuning mechanism are shown in fig. 9. This design evolved during my search for a simple slow-motion drive that is easy to duplicate. I salvaged the panel bushing and shaft from an open-style, rotary-wafer switch that had a standard 1/4-inch (6-mm) shaft. The detent mechanism and switch sections are removed, leaving only the bushing and shaft. The rear end of the shaft is then turned down to a diameter of 0.11 inch (2.8 mm). Notice the angled bevel where the shaft is turned down just behind the bushing. A 3-inch (76-mm) diameter disk is fashioned from 0.062-inch (1.6-mm) aluminum and secured to a collar which can be mounted on the tuning capacitor's shaft and held in place with a setscrew. Slide

### COIL WINDING CHART

REF	BOTTOM VIEW	WINDING INFORMATION
T1		PRI - 16 TURNS NO. 40, TAP AT ONE TURN, 1.5 - 4.5µH SEC - 1 TURN NO. 40
T2		PRI - 16 TURNS NO. 40, TAP AT TWO TURNS, 1.5 - 4.5µH SEC - 2 TURNS NO. 40
T3, T9 T10, T12 T13		SEE TEXT
T4		SEE TEXT
T5, T8		PRI - 36 TURNS NO. 34, TAP AT 6 TURNS, 10 - 30µH SEC - 3 TURNS NO. 34
T6		PRI - 36 TURNS NO. 34, TAP AT 6 TURNS, 10 - 30µH SEC - 3 TURNS NO. 34
T7		PRI - 7 TURNS NO. 34 SEC - 1 TURN NO. 34 CORE - FERRITE BEAD AMIDON FB-43-101, FERROXCUBE K500100/3B, ETC.
T11		PRI - 16 TURNS NO. 40, TAP AT TWO TURNS, 1.5 - 4.5µH SEC - 2 TURNS NO. 40

REF	BOTTOM VIEW	WINDING INFORMATION
T14		PRI - 8 TURNS NO. 34, 0.4 - 1.2µH SEC - 2 TURNS, CENTER TAP
T15		PRI - 8 TURNS NO. 26, CENTER TAP SEC - 2 TURNS NO. 26, CENTER TAP FERRITE BALUN CORE
T16		PRI - 4 TURNS NO. 22, CENTER TAP SEC - 14 TURNS NO. 26 FERRITE BALUN CORE
L4		36 TURNS NO. 34, CENTER TAP, 10 - 30µH
L6		36 TURNS NO. 34, 10 - 30µH CUT OFF UNUSED PINS
L7, L8 L9, L10 L11		18 TURNS NO. 34, 3 - 7µH CUT OFF UNUSED PINS
L12, L13		3 TURNS NO. 26, CORE - 2 FERRITE BEADS END TO END AMIDON FB-43-101, FERROXCUBE K500100/3B, ETC.

the disk forward in contact with the beveled part of the shaft and tighten the setscrew. This creates a friction drive with a tuning rate of about 35 kHz per revolution and no backlash. The exact tuning rate depends on the distance from the shaft's center where contact is made with the disk. The hole for mounting the bushing is a little oversized, to allow for some adjustment. Frequency markings are applied to the disk and viewed through a window cut in the front panel. You can use a commercial vernier drive, if you have one, or as a last resort, use a small trimmer capacitor in parallel with a main tuning capacitor as a bandspread control.

### component selection

There is little about the circuit that is critical. You can use substitutes for many of the components with little change in performance. I made an effort to minimize the number of component values used. Where possible, resistors and capacitors were selected from decade values in 1-3.3-10 sequence. The idea was to make it easier for those who had to buy what they didn't have. Most hams will resort to the junk box first. For resistors, watch ratios and keep the dc bias conditions for active devices from varying too much. For coupling and bypass capacitors, check the reactance at the frequency of interest and keep the leads especially short in rf circuits. If you use the circuit board layout provided, all resistors, except the bias resistor for the final, are quarter watt size. I used 1/4-inch (6.3-mm) disc ceramic and radial electrolytic capacitors throughout to minimize board size.

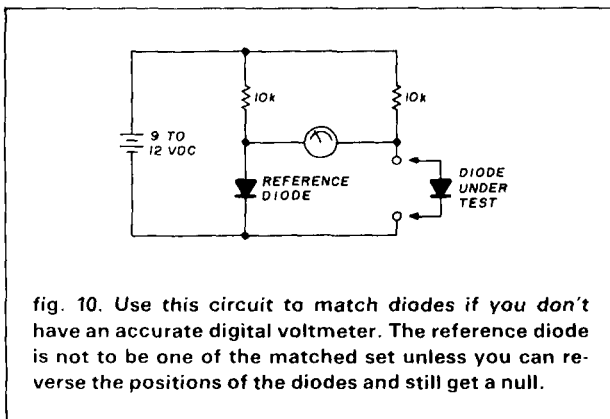


fig. 10. Use this circuit to match diodes if you don't have an accurate digital voltmeter. The reference diode is not to be one of the matched set unless you can reverse the positions of the diodes and still get a null.

You will have to wind a number of coils and transformers, the details of which are shown in the table. The variable inductances use 10-mm i-f cans that I rewound. They were originally 4.5-MHz and 10.7-MHz transformers in a selection from Poly-Paks. These higher frequency units have a plastic bobbin which makes them easy to rewind. This is not true of lower-

frequency (455-kHz) transformers that use a threaded cup rather than a slug. Many of them have internal capacitors, and in some cases, I retained the internal capacitor and added enough capacitance externally to get near the value on the schematic.

Inexpensive, general-purpose transistors are used everywhere except in the final. The device that I used most was a 2N3904 (and its PNP complement, 2N3906) since it met my criteria of good performance, cost, and availability. It has the popular E-B-C lead configuration and there are a lot of substitutes that you can use without sacrificing performance,

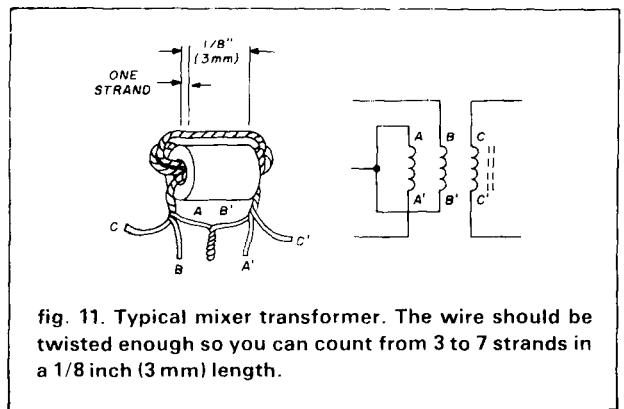


fig. 11. Typical mixer transformer. The wire should be twisted enough so you can count from 3 to 7 strands in a 1/8 inch (3 mm) length.

except possibly in the receiver front-end and transmitter driver. Remember to use transistors with low-output capacitance for Q1 and Q19. I was conservative with transistor ratings and used power-type transistors if there was a possibility of exceeding half the rating of a 2N3904.

The printed circuit layout provided accommodates the mini-dip package version of a 741 op-amp. The same layout can be used with the round metal can version if the leads are spread. Since a lot of op-amps are supplied with the same popular pinout as a 741, there are many substitutes that will work here, but make sure you use an internally compensated op-amp.

### balanced mixers

The transceiver uses three double-balanced mixers. There's nothing hard about rolling your own. To get good carrier suppression, you simply have to maintain good balance. The two places where balance is important are the transformers and the diodes. Transformer balance is achieved by using transmission line techniques. Diode balance relies on matched diodes.

All of the diodes used in the transceiver, except one zener, are the same type, 1N4148 (1N914, 1N4454, etc.). You can get these diodes in quantities from mail-order houses for a nickel or less apiece.

Since you will need a total of twenty-seven for the rig, you'll have enough to find three matched sets of four. Because the highest frequency of interest is only 27 MHz, don't bother matching the reverse capacitance. Just use the same type from the same manufacturer, so the capacitance is close, and match the diodes' forward characteristic as outlined here. Measure and match the forward voltage drop across the diode with about 1 mA of bias current. A 10 K resistor in series with a stable 9 to 12 volt supply works fine. A high-impedance, accurate VTVM should be used (a digital meter is best), but don't despair if you don't have one. The bridge circuit shown in **fig. 10** works with any meter that can distinguish a change of a few microamperes.

The transformers for the transmitter and oscillator offset mixers (T9, T10, T12, T13) are the same. Using small enamel magnet wire, about AWG 34 (0.16mm), twist three strands together uniformly to make trifilar wire. Wind two turns of the trifilar wire on a ferrite bead (Ferroxcube K500100/3B, Ind. Gen. F1650-1-H, Amidon FB-43-101, etc.). Two of the strands are series connected to form the center-tap of the balanced winding (see **fig. 11**). The input transformer of the i-f mixer (T3) is constructed the same way except that two beads, stacked end-to-end, are used for the core. The output transformer of the i-f mixer (T4) is different because it matches the high impedance of the crystal filter. For this transformer, twist six strands of magnet wire together, and use two turns on a two-bead stack. Connect two strands to make the center-tapped winding, as before, then series connect the other four strands to make the high-impedance winding that connects to the crystal filter.

## checkout and adjustment

Using these instructions and the theory of operation, you should have no trouble getting the rig up and running if you have had any experience with building. The first part of the circuit you want to check out is the 10-volt regulator, since this affects almost everything else. Connect the external 13.6-volt supply and monitor the input current, which should be about 100 mA for receive. The regulated output should be near 10 volts. Load the regulator's output and watch the voltage. It should change less than 10 mV for currents up to 750 mA, but go into current limiting at about 1 ampere. Adjust the value of R94 if the current limiting point is very far off.

Check the frequency response and peak gain of the CW audio bandpass filter using the adjustment points described in the theory of operation. A simple way to check out the entire audio circuit of the receiver is as follows: Set the bandwidth switch for **CW** and the **VOLUME** control at minimum, then

connect a jumper from the speaker to the filter input (top of R43). When you turn up the volume, the whole loop will oscillate slightly (50-100 Hz) above the center frequency of the audio filter and the S-meter will track the volume setting.

The adjustment of the crystal filter and low-frequency crystal oscillator are interdependent. This is best done using a sweep generator. It's a slow, tedious job with an ordinary signal generator, but here's the procedure if a sweeper isn't available: Disable the AGC (lift one end of C45) and monitor the audio output, in the SSB mode, with an ac voltmeter. Start with L4 set to minimum inductance. Tune across the passband and you'll find nearly equal audio peaks at 500 and 2000 Hz with a big dip in between. Adjust the tuning of the 3.58 MHz crystal oscillator if necessary to get the frequencies near these values. Increase the inductance of L4 until the variation of audio across the passband is less than about 1 dB from 350 to 2200 Hz and there is a sharp roll-off of the audio at 300 Hz. Don't confuse the low-frequency roll-off of the audio circuits for the filter's corner. To be sure, you can tune the crystal oscillator up a little higher than normal, which will shift the audio frequencies up, so that the filter's corner frequency can be identified, then reset C54 so the corner is at 300 Hz. Set the bandwidth switch to **CW** and center the signal at the peak of the audio filter. Temporarily short the collector of Q19 to ground and adjust C53 for a zero-beat by watching the voltage at the collector of Q11 with a scope.

If you have trouble getting the 3.58-MHz oscillator adjusted properly, the problem is probably a transistor with too much junction capacitance. One possible solution is to bias the junction of Q19 by connecting a high-value resistor (10 K to 100 K) from the collector to +10 volts. You can check the operating frequency of each crystal in the circuit and use the one with the highest frequency in the oscillator.

The other adjustments consist of tuning for maximum output at the center of the band.

On-the-air reports with the rig running barefoot are favorable. A number of operators have commented on the good sounding audio quality that comes through in spite of QRP signal levels. I get questions about what kind of processor I'm using. The passband may be only 2 kHz, but only once did I get a comment about the lack of highs in the signal. My voice tends to be on the low side anyway. I've had no trouble working stations all over North and Central America from my home in Tennessee. Pile-ups on DX stations are another matter however. Of course, a linear amplifier is next on the want list.

ham radio

# Bobtail curtain follow-up: practical DX signal gain

The second part  
of a two-part series  
on this remarkable antenna

**The actual DX signal gain** of any one type of antenna over another, at distances beyond about 2500 miles, does not always correlate well with the theoretical "free space gain over isotropic." After all, antennas do not operate in free space. Surrounding objects, especially ground, are a part of the antenna system.

For distances beyond 2500 miles, angles of signal departure below about 15 degrees are almost always the most effective. This is true regardless of propagation path, whether it's one acute geometric bend near midpoint or a chordal or ducting mode. And it's true regardless of ionospheric tilt. Although it's most noticeable on 10 and 15 meters, it still applies to long-haul 40- and 75/80-meter propagation.

To get the angle of radiation down while still keeping the antenna height acceptable on 40 and 75/80 meters, vertical antennas have long been used. Some verticals do a good job. A few, such as a full-size half-wave vertical, can do an excellent job in all directions. Others seem to radiate "equally poorly in all directions." I'll not take time to go into all the rea-

sons why short vertical radiators that are current fed near ground level are often ineffective.

As noted in Part I of this article, simply turning the antenna upside-down greatly reduces the ground loss problems. For one thing, it minimizes the conduction current flowing to ground at the feed point when the antenna is ground mounted. It also minimizes the losses caused by displacement currents in the near field fighting their way through lossy dirt trying to find a "mirror image" that, in this case, is more theoretical than real. In addition, getting the high-current portion up in the air allows the antenna to radiate somewhat more effectively, and it lowers the angle of radiation slightly.

Users of the Bobtail antenna often report gain improvements on long-haul DX of from 10 to 20 dB over their previous antenna. But only a little of that improvement results from the azimuthal directivity. The sometimes startling effectiveness of the Bobtail for 40- and 75/80-meter DX is the result of its inverted configuration. This is ordinarily more noticeable in a built-up, residential area than in an open field.

The gain attributable to the horizontal directivity of a two-element Bobtail, or half square, is about 4 dB over that of an inverted groundplane using two resonant radials. The half-power beamwidth of each lobe of the bi-directional figure-8 pattern is about 60 de-

**By Woodrow Smith, W6BCX, 2117 Elden Avenue, Apt. 20, Costa Mesa, California 92627**

greens — wide enough to cover some worthwhile geography while still providing useful gain.

The full-size Bobtail (for those that have the room) has a directivity gain of slightly over 5 dB (compared with the same reference). The half-power beamwidth is about 50 degrees. Four or five dB doesn't sound like much, but if your signal is marginal it can make the difference between copy and no copy. On receive, the discrimination you get from the azimuthal directivity can be worth more than the 4 or 5 dB when it comes to what you can copy through noise and QRM.

The slight extra gain of the full-size Bobtail over the two-element should be considered a bonus. The main advantage of the full-size version is that spurious lobes are reduced in amplitude, and therefore end-fire, high-angle pickup on receive is reduced. There is no point in being able to deliver a readable signal 6000 miles away on 40 if you can't hear the other station because of bad off-axis QRM from a station 600 miles away.

### **direct coax feed versus voltage feed**

As noted in Part I, if certain precautions are taken, direct coax feed can be used with a half-square version of the Bobtail. The disadvantage of doing so is that it limits the antenna to one band. Voltage feeding one end of a half square permits use on half frequency, as sort of a drooping, half-wave dipole. At twice the frequency it functions as a combination of two co-phased vertical dipoles a wavelength apart, with the resultant cloverleaf end-fire and broadside pattern having its nulls filled in fairly well by the pattern produced by the horizontal full-wave portion. Voltage feeding the half square at three times frequency produces an interesting multiple lobe pattern that results in unpredictable results. Voltage feeding one end of a half square cut for 40 meters will thus provide a low-angle, "far DX" figure-8 pattern on 40 while also doing a good general-coverage short- and medium-haul job on 80.

On 20 meters the composite pattern obtained results in a good, general-coverage "long" DX antenna that in addition is effective in the 750- to 1500-mile range. Voltage feeding it on 15 meters produces an interesting multiple lobe pattern that will sometimes, as noted above, produce results that often are surprising — and occasionally amazing.

On 10 meters it should be considered simply a random long wire that is capable of providing lots of good clean fun when the band is hot.

Unfortunately there is no really good way to feed a three-element Bobtail directly with coax, even for one-band operation. No matter how the coax is

brought down from the center element there is going to be objectionable unbalanced coupling to the coax from radiating portions of the antenna. Using a balun does not cure the problem.

Connecting the coax to an end radiator junction as described for the half square does reduce the unbalanced coupling, but the current distribution in the three elements no longer is symmetrical. Current will be greatest in the fed element, thereby skewing the pattern and reducing the gain.

---

*"Here's the best place  
to feed  
a  
Bobtail curtain . . ."*

---

The only really good way to feed a three tailer is to voltage feed the bottom of the center element. This permits multi-band operation in pretty much the same fashion as with the half square, though the lobe pattern on bands other than the fundamental will be slightly different. The main difference is that on half frequency the three-tail version makes a much better end-fire "medium DX" antenna than a half square. However, this is at the expense of high angle (short haul) effectiveness, particularly broadside.

### **"Zepp" voltage feed**

Way back in the 1930s, PA0ZN came up with an antenna resembling a two-tail Bobtail (or half square) in appearance except for feed method. It was fed at the center of the horizontal section (a high impedance point) via Zepp feeders. The antenna obviously belongs to the "inverted ground plane" family, with the attendant advantages over a right-side-up arrangement.

However, the mutual impedance between the Zepp feed line and the rest of the antenna is such that strong in-phase currents are induced in the line, thus producing considerable "antenna effect" on the feed line. This is strong enough to produce sufficient distortion of the pattern to "dirty up the nulls" a bit without providing a significant increase in gain. The pattern is cleaner when the Zepp feed line is attached to one end of the radiator. While both radiator and

feed line are unbalanced slightly with such an asymmetrical arrangement, the resulting imbalance is not enough to cause serious mischief.

If you want a dual band (say, 40 and 20 meters) omnidirectional DX antenna and would like to use Zepp feeders all the way to the shack, I would suggest an inverted ground plane with two radials, with the vertical element of the I.G.P. fed at the bottom with one side of the open wire line.

If you would like to Zepp feed a three-element Bobtail, be sure to feed the center element. But remember that the main reason for using a three-element Bobtail instead of a half square is the cleaner pattern, and while Zepp feeding the center element will not unbalance the antenna itself, the inherent unbalance in the Zepp feed line will cause some pick-up by the line when receiving, even if the open line is brought off at right angles out of the near field.

Considerable voltage will be built up across an open feed line at the high impedance points when the line is used as a Zepp feed line on transmit, so be sure to use sufficient spacing if you are running power.

### ground screens and grounding

No antenna that is fed via an unsymmetrical feed system can be 100 percent ground-independent. However, if the impedance between antenna feed point and ground is over about 1000 ohms there will be very little current flowing to ground. In this case not much of an earth ground will be needed at that point for the antenna. A small ground screen, laid on the ground or a flat roof, or suspended or supported near the network matching 50-ohm coaxial line to the high impedance Bobtail antenna feed point, makes an effective rf ground. Such a screen will make a better rf ground than a lightning stake, or stakes driven in the soil. If one of the latter, properly installed, is advisable for lightning protection in your area, it still is a good idea to use a small ground screen as an rf ground.

One ready-made ground screen widely available is a 3 × 5 foot (0.9 × 1.5 m) piece of galvanized hardware cloth, packaged by Sears under the catalog number 44531 and available off-the-shelf at some Sears retail stores. Current catalog price is about \$7.00 a roll.

To see if the ground screen is doing its job, simply touch the "rf ground" while running low power to see if the VSWR or field strength changes significantly. If it does, either shorten the connecting wire between screen and matching network or series resonate the wire with a mica capacitor. The value will not be especially critical. Possibly on 75/80 more screen will be required. Ordinarily if the antenna di-

### 10-MHz Bobtail dimensions

Because adding 10 MHz to a typical tribander is not the easiest thing in the world to accomplish, there is bound to be interest in 10-MHz Bobtails and half squares. Dimensions are not extremely critical. Except for direct coax feed of a half square, slight deviations from optimum can be compensated for in the tuner or matcher at the antenna end of the coax line.

Assuming No. 12 or 14 wire (M2.1 or 1.6) and typical insulators, the following dimensions will be found satisfactory for 10.1 to 10.15 MHz:

spacing of vertical elements:	48 feet 9 inches (14.86 m)
length of vertical elements:	23 feet 7 inches (7.19 m)

For the sake of convenience the voltage fed element can be made up to 5 percent shorter or 8 percent longer if fed by a resonant tank or L network. For Zepp feed, the fed element should be cut to the exact length shown.

mensions are near optimum, not much screen will be required in order to do the job properly.

### guys and metal masts

When using wooden masts to support a Bobtail the usual precautions apply to breaking up adjacent guy wires to avoid resonance. There is nothing wrong with using metal masts of EMT or aluminum tubing as the outside vertical elements to save lot space. This requires a strong base insulator of low loss material, preferably having low shunt capacity. Nonmetallic guys do a better job electrically than guy wires, even if the latter are well broken up. The tubing should be of no greater diameter than what is required for adequate mechanical strength. A tubing element will be slightly shorter physically than a wire element for the same electrical length. Also, the shunt capacity of the base insulator will require shortening the element a bit more. However, if you merely make them 3 percent shorter than a wire element of optimum electrical length, and try to keep the base shunt capacity low, it should be close enough. Tubing joints should make good electrical contact or be jumpered.

Unless a piece of thick-walled fiberglass tubing or the like is used as the base insulator, it should be used only to support the weight of the mast, and not to hold the bottom rigidly vertical. This probably means an extra set of guys.

Keep in mind when planning your installation that the bottoms of the tails are quite hot with rf and can cause a bad burn if you are running much power. One way to avoid this is to slip some small-diameter clear plastic (such as fuel line) tubing over the ends of wire tails closer than 7 feet (2.5 m) from ground. If you decide to use metal tubing for the end elements, slit or saw lengthwise a few feet of PVC plastic pipe and tape it over the bottom of each mast.

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## A simple, automatic negative voltage tracking system

# dual voltage power supply

Anyone who tinkers with integrated-circuit projects will eventually have to provide a negative voltage supply for op amps (operational amplifiers — see **fig. 1**). Batteries may seem a simple solution, however, after purchasing several quality 9-volt alkaline batteries at \$2 each, you soon see the false economy in that.

Deciding my money could be better spent on parts for a power supply, I set out to find a good schematic design. After my initial search it seemed I would have to purchase two voltage regulators, one for positive and one for negative, to obtain a regulated positive/negative supply. I wasn't willing to spend at least \$4 each for voltage regulators and have the hassle of adjusting two knobs to set both voltages. Neither was I excited about watching two meters to be sure both voltages remained equal under changing load conditions. After extensive research, I found a small diagram in a Radio Shack semiconductor guide under the 78XX series voltage regulators using op amps.

A 741 op amp is much cheaper than a monolithic negative voltage regulator, such as the LM337T. A 741 op amp costs 79¢ at Radio Shack (part #276-007). Even if you add the cost of Q3 to the op amp, it is still three dollars cheaper than the negative regulator, plus the 5-kilohm potentiometer required to control it. And, only a single control is needed to set both voltages simultaneously.

### operation

Automatic negative voltage regulation for a fixed-voltage supply is provided by the circuit illustrated by **fig. 2**. It uses a 741 op amp in an inverting configuration. The op amp compares the voltage between pins 2 and 3. It supplies an equal but opposite voltage at its output, pin 6, by sensing the difference between the two input voltages and inverting that voltage at its output. This in turn biases the pass transistor into conduction providing the power supply negative voltage output.

**Fig. 3A** shows how this basic idea is used to build a positive/negative variable voltage supply. R11 and R12 are current-limiting resistors connecting the negative and positive power supply outputs. Since these resistors are of equal value, they will have equal voltage drops with equal applied voltages.

When the negative supply voltage is exactly equal and opposite the positive supply voltage (– 10 volts and + 10 volts, respectively) the two voltage drops across R11 and R12 are equal. They cancel each other and provide zero volts at pin 2 of the op amp, U1, which compares the zero volts on pin 2 with the ground on pin 3 (also zero volts) and finds they are equal. U1 therefore gives no output on pin 6.

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Whiteville, North Carolina 28472



The finished product. The top half of the faceplate contains all controls for the 1-ampere supply. The bottom half has all controls for the 2.5-ampere supply. The face plate was made prior to final testing, and bears the estimated 3-ampere rating.

When adjusting the positive voltage regulator (U2) for a desired positive supply voltage, that voltage drops across R11, raising the voltage sensed by pin 2 of U1. U1 supplies an equal but opposite voltage to the base of Q3, biasing Q3 into conduction. Q3 drives Q2 into conduction, providing a negative voltage at the supply output. There it drops across R12, lowering the voltage sensed by pin 2. When the negative supply voltage lowers to the exact opposite of the positive supply voltage, they cancel each other at pin 2, completing the cycle. The op amp continuously and instantaneously controls the negative supply voltage, keeping it exactly equal but opposite the positive supply voltage.

### circuit description

R13 is a balance control to adjust for differences in the values of R11 and R12. A 1 kilohm potentiometer should be sufficient for R13, since the difference between two 4.7K, 10 percent tolerance resistors shouldn't exceed 940 ohms, even in an extreme case. I used a circuit-board-mounted pot since this should be a one-time only adjustment (when you first calibrate the supply). Simply adjust R13 so the negative voltage is exactly equal to (and of opposite sign to) the positive regulated voltage. No further adjustment should be required. I suggest calibrating R13 in the VOM range, that covers the full power supply voltage range. This will eliminate the need for switching ranges when checking your calibration. Switching from one range to another could give you slightly different readings at the same voltage setting.

R10 is a current-limiting resistor intended to keep U1 from burning itself out should it try to supply too much current to Q3. When testing the circuit, I tried to bias Q2 directly with U1. But Q2 required too much base current and U1 self-destructed. Any small current PNP can be used in the place of the Radio Shack 2025 I used for Q3, as long as it will withstand

the required driving current to Q2. I measured  $19 \mu\text{A}$  on the base of Q2, under *no load*, and 17 mA (nearly 900 times more current) at a 2.4 ampere load at 15 volts output.

The op amp requires a positive/negative voltage supply for its own operation. The regulated supply voltages to run the op amp are obtained by employing 18-volt zener diodes D1 and D2. Since the op amp draws very little current, R3 and R4 are mainly used to limit current through the zeners. Half-watt or one-watt zeners provide a sufficient power rating.

### a new approach

I believe the manner in which I have employed the regulator U2 in biasing the pass transistor Q1 is a novel approach. Other designs using 3-pin adjustable or 3-pin fixed voltage 78XX series regulators use a PNP pass transistor with its base connected to the emitter through a resistor or its base tied to the regulator input, as in **fig. 4**. The regulator output and pass transistor collector are tied together at the output. My approach uses an LM317T 3-pin adjustable positive voltage regulator (U2). It can be used alone for up to 1.5 amperes current, or with a pass transistor to increase the available current. I used this type regulator alone for the 1.2 to 25-volt supply, as shown in **fig. 3B**. This extra supply was built into the same cabinet as the positive/negative supply. U3 is shown in a standard application of the LM317-type regulator. I used  $0.47 \mu\text{F}$  electrolytics for C5, C6, C9, and C10. The recommended capacitors are  $0.33 \mu\text{F}$  tantalums for C5 and C9, and  $0.1 \mu\text{F}$  tantalums for C6 and C10. (See final comments.)

In **fig. 3A** you see that the output of U2 is used to bias the base of the NPN pass transistor directly, as opposed to the bias coming from the same line as the PNP emitter (**fig. 4**). R5 and R6 are used as current dividers. R5 is half the ohmic value of R6, allowing twice as much current through pass transistor Q1 as passes through regulator U2. Therefore, if the regulator is passing its maximum current of 1.5 amperes, the transistor can pass twice that, or 3 amperes, the

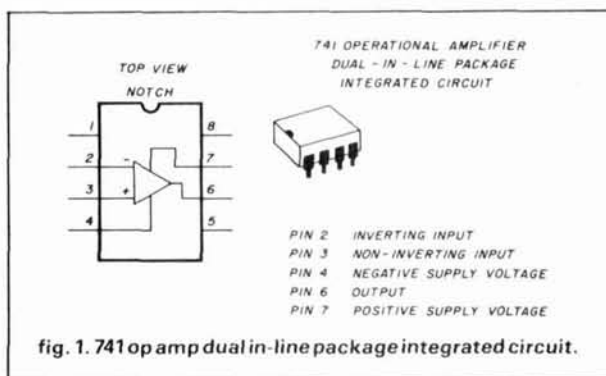


fig. 1. 741 op amp dual in-line package integrated circuit.

oretically providing up to 4.5 amperes of current at the output.

The LM317-type regulator has internally controlled automatic shutdown for overheating and overcurrent protection. Should the regulator reach its safe limit, it will automatically shut down. In this configuration, if U2 shuts down, there is no bias for Q1. This shuts off Q1, and consequently, turns off the negative supply. Q1 is a 115-watt transistor rated at 15 amperes collector current, so it is not likely to overheat while passing only 3 amperes of current. In fact, it should never get past warm to the touch if it is properly heatsinked. Even if the regulator is operating at its extreme limitations, it will always shut down in an overload situation before Q1 can overheat, because Q1 is operating at only one-fifth its rated capacity in that same extreme case. Overdesign was mostly a function of junk box availability.

D4 is used to isolate the Q1 emitter from its base. If D4 shorts out, it shuts down Q1, again preventing any possibility of thermal runaway. Be sure D4 is rated higher than your expected 1.5 ampere current through the regulator, or you may be constantly replacing D4. A 2.5 ampere diode should provide a sufficient safety margin.

As shown in **figs. 3A** and **3B**, I used fused outputs on both power supplies as an additional safety factor.

### additional features

M1 and M2 are two halves of a dual VU meter modified for use as voltmeters. R14 and R15 are actually 100-kilohm resistors in series with 25-kilohm circuit-board-type potentiometers, used to calibrate the meters. I added some LED indicators (D3, D6, and D7). D6 and D7 indicate (**fig. 3B**) to which voltage scale I have the meter set. I use a 0-10-volt scale for better resolution at low voltages, and a 0-30-volt

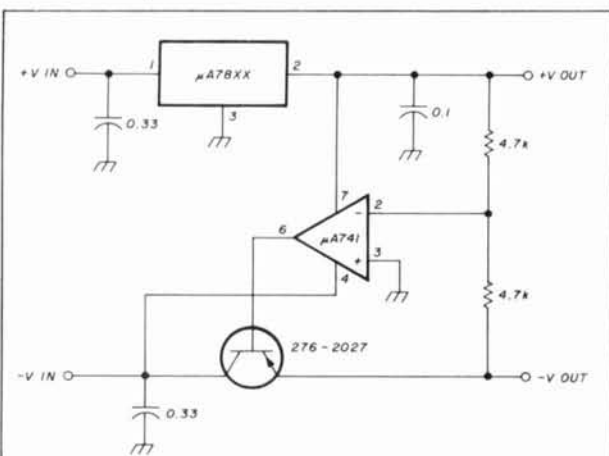
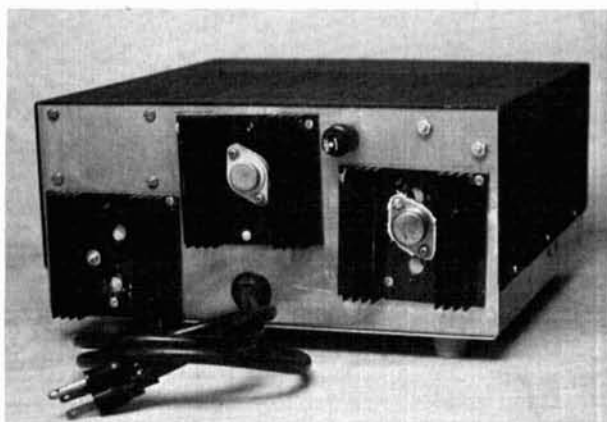


fig. 2. Automatic negative voltage regulation circuit, copyright Tandy Corporation, 1980 *Archer Semiconductor Guide*, reproduced with permission.



This rear view shows heatsinking. Far left heatsink is for the 1-ampere supply regulator, mounted to the sink from the inside. Center sink has the GE 14 (Q1) while the right sink has the 2SB539 (Q2). Be sure to use silicone heatsink compound when mounting transistors to any heatsink.

scale for higher voltages. A microammeter would be preferred over the milliammeter used for M3, but again, I used what was available in my junk box. R20 through R23 are calibrating resistors for M3.

D3 tells me when the output is turned on by switch 2. Since the LM317 regulator can be turned down to no lower than 1.2 volts, such a switch is handy. With S2 open, U2 has no input and, therefore, no output. Likewise, Q1 and Q2 are turned off when U2 has no output, so there is no need for a switch on the negative side.

My design has partially taken care of the minimum voltage problem, even without S2. Due to the voltage drop across D4, and a similar drop across the Q1 base, the minimum output is reduced from 1.2 to about 0.6 to 0.8 volts. That low a voltage forces very little current through anything but a dead short. However, you can program the minimum voltage up to 1.2 volts, or any higher voltage, by adding a fixed resistor between R8 (or R18) and ground. The formula for calculating the needed resistance is covered in the applications data sheet supplied with the regulators.

### a limitation

The idea of using an op amp as the negative voltage regulator has one limitation: the supply voltages for the op amp. The operating voltage limits for op amps are generally from 16 to 18 volts. The op amp has internal voltage drops. With the extra voltage drops across R10, Q3, and Q2, I can get only up to 15.5 volts at the negative supply output (depending on the load). This is quite sufficient for operating op amp test circuits.

The 741 op amp I used was rated at 16 volts maximum supply voltage and I am presently operating at

18 volts. No damage has occurred thus far, but I would not be surprised if the op amp fails one day. Since this is a learning experience for me, I'm willing to chance a 79-cent op amp. But if you want to be safe, use the recommended supply voltages when choosing your zeners.

## performance

My original intent was to operate small experimental circuits which draw no more than 500 mA in extreme cases. The power supplies outlined here are more than adequate for that purpose. I built them to

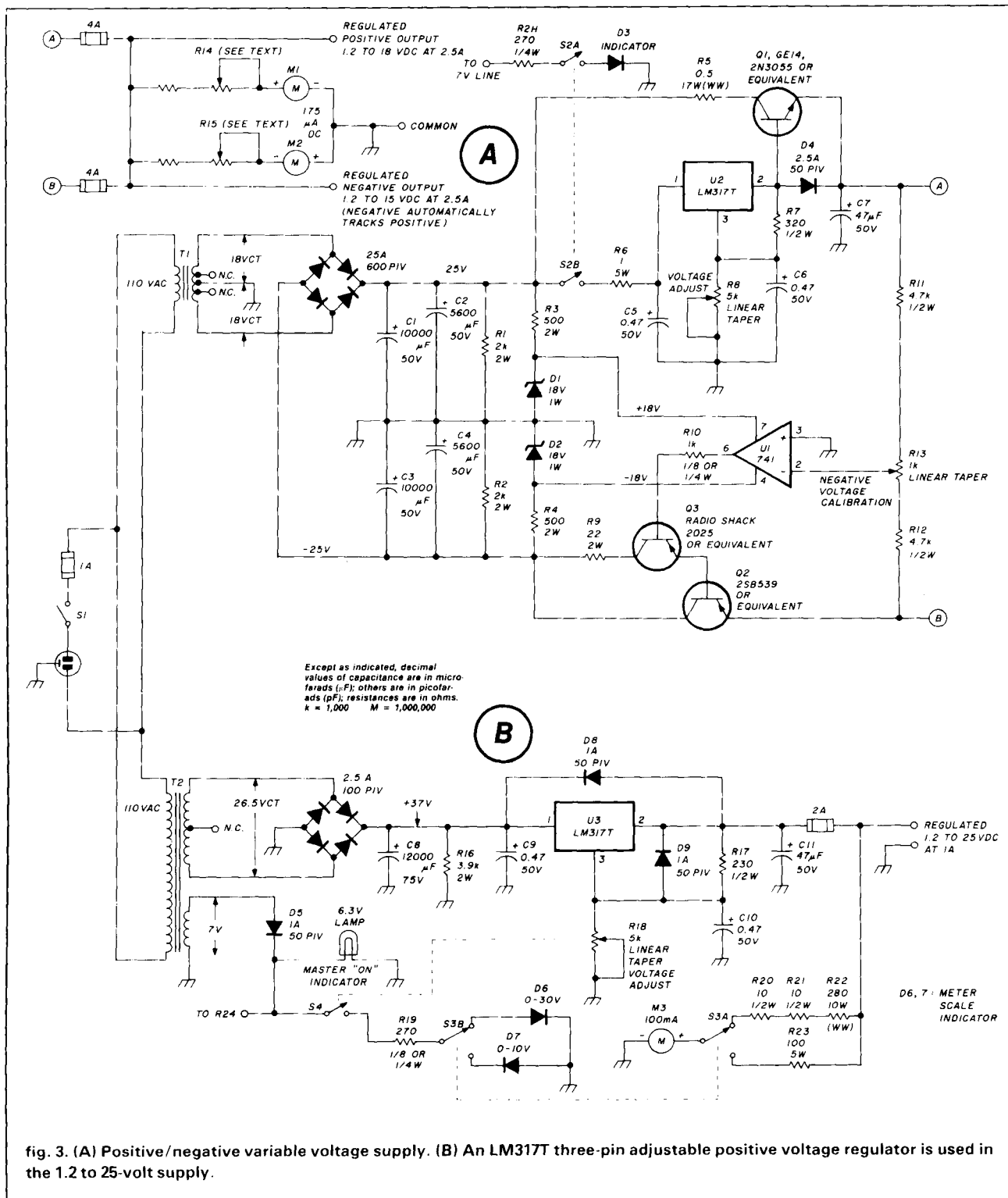
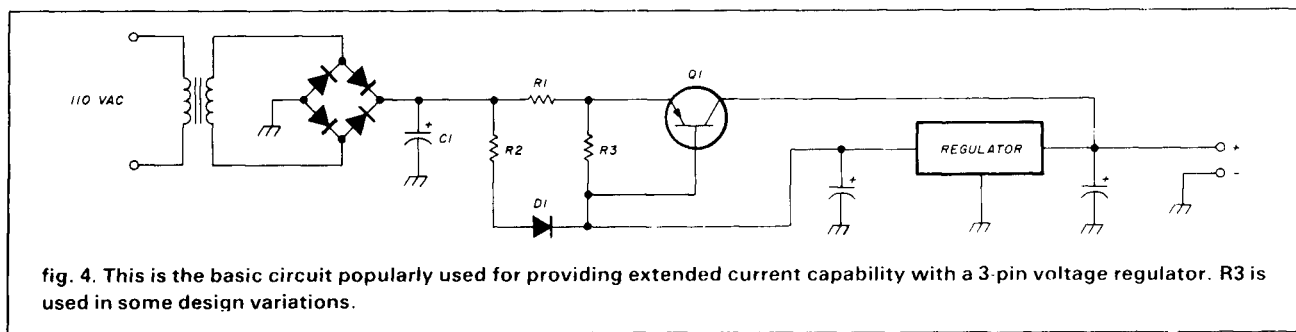


fig. 3. (A) Positive/negative variable voltage supply. (B) An LM317T three-pin adjustable positive voltage regulator is used in the 1.2 to 25-volt supply.



give everything the transformers would provide, and they do.

Voltage and ohmic measurements were made using a Fluke 8024A Digital Multimeter. Ripple measurements were made with a Hewlett Packard 180A 50-MHz scope, with the Hewlett Packard 1801A Dual Channel Vertical Amp and 1821A Timebase.

I overestimated the capacity of T1, a transformer I salvaged from an old TV set. I guessed it would supply 3 amperes, which was about 0.5 ampere too much. T2 was salvaged from a junked receiver and rewound to suit my needs.

The positive supply in **fig. 3A** showed 17.4 volts into a 6.3-ohm load, with less than 10-mV ripple. However, between 17.4 and 17.8-volts output the ripple rose sharply to 900 mV. That gives 2.76 and 2.82 amperes at 17.4 and 17.8-volts respectively. So, a rating of 2.5 amperes for T1 would have been quite reasonable, instead of my estimated 3 amperes.

The negative voltage side of **fig. 3A** delivered 14.6 volts into the same load with less than 10-mV ripple also. Since the negative voltage is limited by the op amp supply voltages, it would not adjust any higher.

Positive and negative outputs were tested simultaneously into 12-ohm loads, and yielded the same performance as their individual ratings into 6.3 ohms. Additionally, the negative voltage tracked positive within  $\pm 0.05$  volt throughout the range 1.2 to 15 volts.

The positive supply was further tested using a 3-ohm load, which put a 4.5-ampere current through the supply. The regulator circuit was monitored with one meter while the total output was monitored on another. The regulator circuit provided 1.6 amperes of the total 4.5-ampere output, with 2.9 amperes flowing through the pass transistor. This is very close to the designed current distribution ratio. The supply was run for a full fifteen minutes at the 4.5-ampere output, which is 2 amperes higher than its rated capacity. My intent was to force the regulator to overheat and shut down. However, the supply performed flawlessly, and ran the whole time with no signs of cutting back or shutting down. The regulator heat-sink was too hot to touch but the pass transistor was

not excessively hot. No doubt forcing the issue would eventually cause the regulator to shut down. Since the unit had already performed far above expectation, I decided to let it go at that.

The supply in **fig. 3B** works equally well. It supplies 24 volts into an 18.3-ohm load with less than 10-mV ripple. At 24 and 25-volts output, with the same load, the ripple sharply rises to 900 mV. The 1 ampere rating for T2 is therefore a fair rating. This supply was also tested with a 31-ohm load and yielded 27.2 volts at less than 10-mV ripple. This is surprising since that is above the transformer voltage of 26.5 volts.

### final comments

A technician friend reviewed my test data and recommended changes in C5, C6, C9, and C10, which should improve the ripple rejection. Due to the internal reactances of electrolytic capacitors, as compared to tantalum capacitors, the desired frequency response can be approximated by using 0.1- $\mu$ F ceramics for C5 and C9, and by using 10- $\mu$ F electrolytics for C6 and C10. You might try this change when building yours.

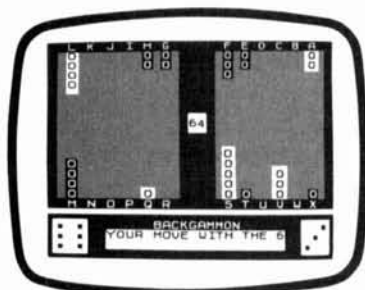
Except for the LM317 regulators, all parts for these two power supplies were obtained from the proverbial junk box. I present these ideas as building blocks for others to improve upon. I am certainly no expert, but I will try to answer any questions you may have if you will enclose an SASE. I suggest some type of overvoltage protection circuit if you plan to use this or any other power supply with expensive or sensitive equipment.<sup>1</sup>

I wish to thank N4BGU for assisting with final testing and test data, and without whose patience I would probably not have learned much of anything about electronics except that required to pass my Novice exam.

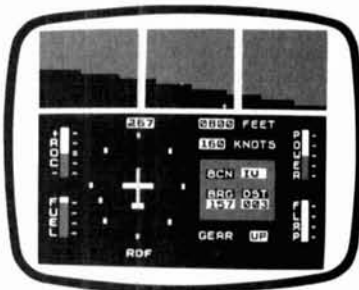
### reference

1. Evert Fruitman, W7RXV, "A Better Overvoltage Protection Circuit," 73, January, 1980, page 140.

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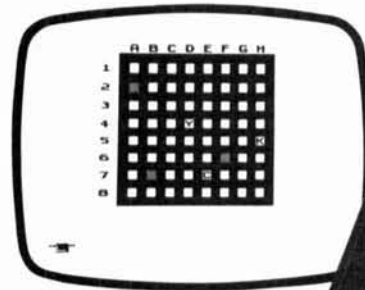
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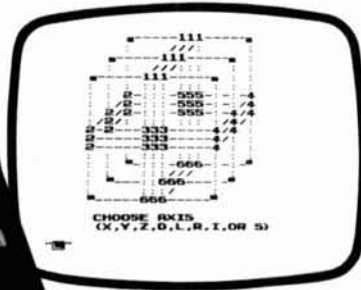
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ham radio magazine presents this latest examination of GaAs FET low-noise operation at lower than normal frequencies.

# GaAs FET performance evaluation and preamplifier application

A look at some popular devices for UHF and VHF

The performance of GaAs FETs in terms of noise figure, gain, and overload capabilities at VHF and above is well known. A new device, the ALF1023 by Alpha Industries, makes it possible to obtain good performance at low cost (\$12 in small quantities). These new GaAs FETs are particularly useful in improving two-meter receivers where front-end overload is a problem.

I tested seven popular GaAs FET devices at 144 MHz for noise figure, gain, 1-dB compression point, and third-order output intercepts. The results are

shown in table 1. All the devices have essentially the same characteristics at 144 MHz.

The third-order output-intercept point was measured under slightly different bias conditions than for low-noise biasing, as was the 1-dB compression-point measurement. A higher drain current (approximately 50 percent higher) was used for these measurements. The third-order output intercept was substantially improved by this bias condition.

This improvement in third-order intercept indicates one of the design trade-offs that one must consider. A sacrifice of noise figure to obtain a higher intercept point is typical. However, when biased for lowest noise, the GaAs FET is already superior to the bipolar device it is to replace.

The improvement in freedom from overload and intermod obtained by using a GaAs FET device to replace a bipolar transistor is pronounced. Fig. 1 shows the schematic of the test circuit used at 144 MHz.

Similar circuits for 220 and 432 MHz are popular and give similar results. At 432 MHz and above, however, some devices start to show slightly higher noise figures than others, even though they are essentially identical at 144 MHz.

The accuracy of noise-figure measurement must be mentioned here, for I have encountered advertisements stating "0.2 dB" noise figure. Noise-figure measurements are relatively easy to make — noise-figure accuracies are inherently uncertain. This uncertainty will not be explored here except to state that the major elements of this error are:

1. Instrumentation error (meter, electronics accuracy)
2. ENR uncertainty (noise-source accuracies)

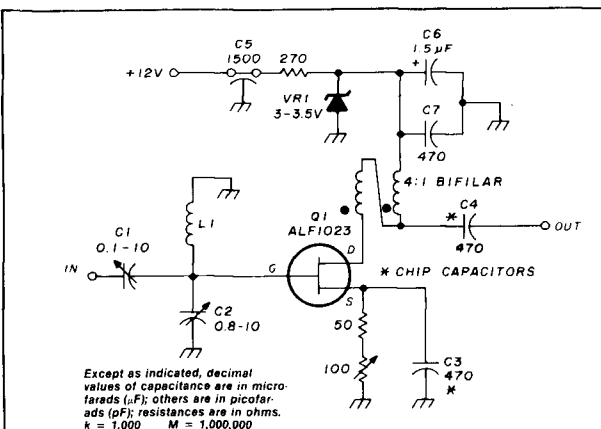


fig. 1. Preamplifier circuit for the ALF1023 GaAs FET. Values shown are for 144 MHz.

The author, Dennis Mitchell, K8UR, is employed as Applications Engineer for Alpha Industries.

By Dennis Mitchell, K8UR, 35 Mt. Pleasant Street, Marlboro, Massachusetts 01752



# LARSEN ANTENNAS TRAVEL IN THE FAST LANE.



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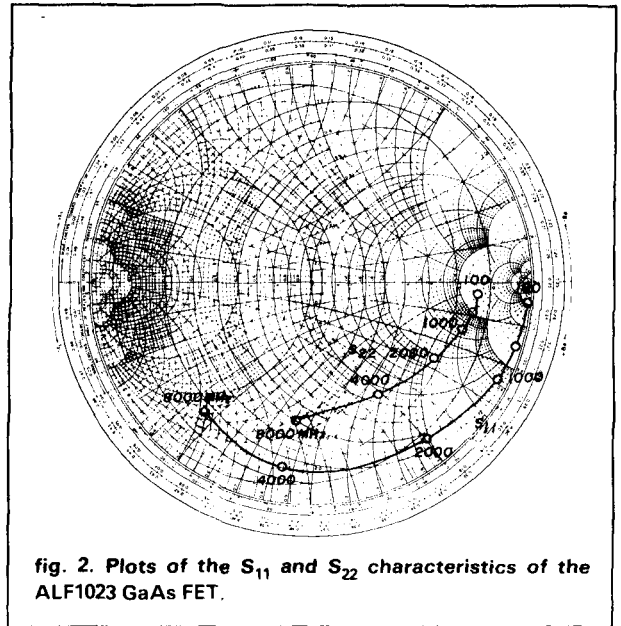
3. Mismatch uncertainties (VSWR between source and device)

4. Second-stage uncertainties

I did not verify the noise-figure claims for many Amateur-band preamplifiers tested.

The system used for the gain and noise-figure measurements listed in **table 1** was an HP 346B Noise source and HP 8970A Noise-Figure Meter with a typical root-sum-of-squares uncertainty of  $\pm 0.23$  dB at 144 MHz.

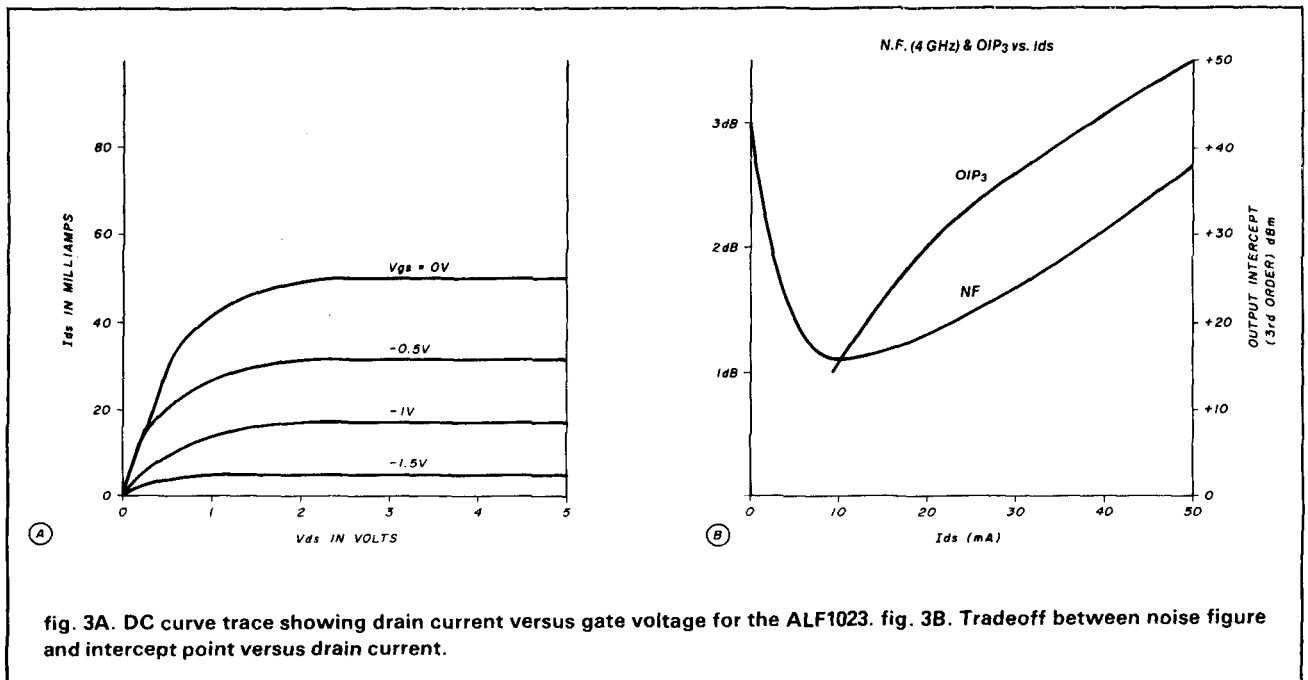
**Fig. 2** shows a plot of the  $S_{11}$  and  $S_{22}$  characteristics of the ALF1023 GaAs FET from 100 to 8000 MHz. **Fig. 3** clearly depicts the ALF1023 dc and rf performance. **Fig. 3a** is a curve trace showing  $I_{ds}$  versus  $V_{gs}$  with an  $I_{dss}$  of 50 mA. **Fig. 3b** illustrates the trade-off possible between noise figure and intercept point for given values of drain current.



**fig. 2.** Plots of the  $S_{11}$  and  $S_{22}$  characteristics of the ALF1023 GaAs FET.

**table 1.** A listing of test results of some popular devices for VHF and UHF preamplifiers, including the new Alpha Industries ALF1023.

device	noise figure dB	gain dB	1-dB compression point, dBm	3rd-order output, intercept, dBm	drain bias	approximate (1-100) cost, \$
Avantek AT-8110	0.6	19.0	+ 18	+ 28	3 V, 20 mA	44
NEC NE21889	0.6	19.9	+ 17	+ 27	3 V, 12 mA	30
NEC NE72089	0.6	19.8	+ 17	+ 27	3 V, 12 mA	12
MGF 1402	0.6	19.8	+ 17	+ 29	3 V, 24 mA	30
MGF 1400	0.6	19.7	+ 17	+ 29	3 V, 25 mA	22
MGF 1200	0.8	19.8	+ 15	+ 30	3 V, 25 mA	15
ALF 1023	0.6	19.8	+ 19	+ 30	3.5 V, 15 mA	12



**fig. 3A.** DC curve trace showing drain current versus gate voltage for the ALF1023. **fig. 3B.** Tradeoff between noise figure and intercept point versus drain current.

ham radio

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HW-5400 Transceiver

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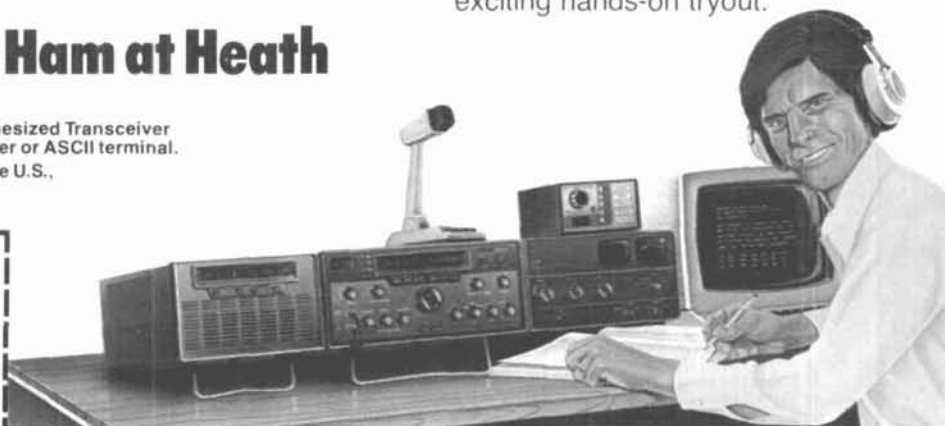
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# hy-gain

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ROTATOR MODEL	ANTENNA WIND-LOAD CAPACITY	
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CD45 II	8.5 sq. ft. (.79 sq. m)	5.0 sq. ft. (.46 sq. m)
HAM IV	15.0 sq. ft. (1.4 sq. m)	N/A
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## technical forum

Welcome to the *ham radio* Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!

In this month's Technical Forum, we are going to begin presenting questions for our readers to answer. Our readers are encouraged both to send in questions of their own relating to Amateur Radio, and to answer the questions of other Amateurs as they appear in this column. The best and most complete answer to each question will be published in a future issue, and the reader who provides the answer will be awarded a prize in the form of a book from Ham Radio's Bookstore. Send all Technical Forum correspondence to Technical Forum, Ham Radio Magazine, Greenville, NH 03048. Now for the first question:

*I am building an eight-turn helical antenna for the 432-MHz band. Every article I have read indicates a 140-ohm input impedance (or something close to that value). My feed line is TV coax cable: 70-ohm solid aluminum outer jacket with solid, copper-coated steel wire for the center conductor and foam insulation.*

*Can anyone tell me the most efficient way to couple the antenna to the 70-ohm line? And can anyone tell me the best way to couple the 70-ohm line to the 50-ohm input of the converter?*

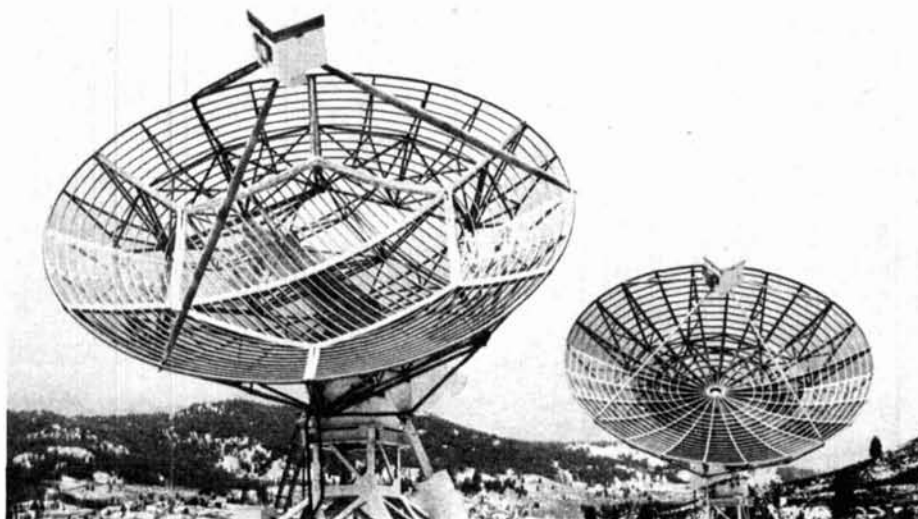
*Joseph Czerniak, W8NWU*

*At the Dominion Radio Astrophysical Observatory, British Columbia, Canada, we have built a synthesis radio telescope, using surplus 28- and 30-foot reflectors (see below). Aperture synthesis is a technique for making big radio telescopes out of little ones, and we have four small dishes spaced along a 2000-foot baseline. This gives our synthesis telescope a beam of 1 minute of arc (one sixtieth of a degree) at 21 cm wavelength (1420 MHz). It is a powerful radio telescope, which produces world-class astronomical research.*

*The four dishes we are using were obtained from surplus. Two 28-foot Kennedy Model 717 antennas came from the Canadian Defense Research Board. Two 30-foot Radiation Systems dishes were found in a scrap metal yard in California.*

*We are looking for more dishes of these two types, so that we can extend our telescope and make it more sensitive. Can any of ham radio's readers help us locate used dishes which might be for sale?*

*Tom Landecker  
Dominion Radio Astrophysical  
Observatory  
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Extended double zepp V Series antennas consist of two stacked .64 wave vertical sections in phase. Two sets of 1/4 wave radials decouple the antenna from the mast and feed line so all RF goes into the antenna and is not radiated by the coax. The feed line connects through the lower section to the center matching coil. This not only provides weather protection for the connector (SO-239 connectors for V2, V3. Type N connector for V4) but also places the entire antenna at dc ground to reduce lightning hazard and QRN.

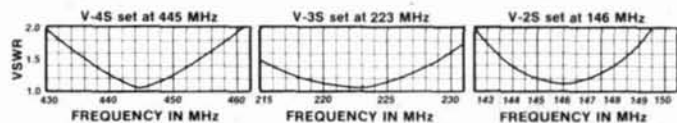
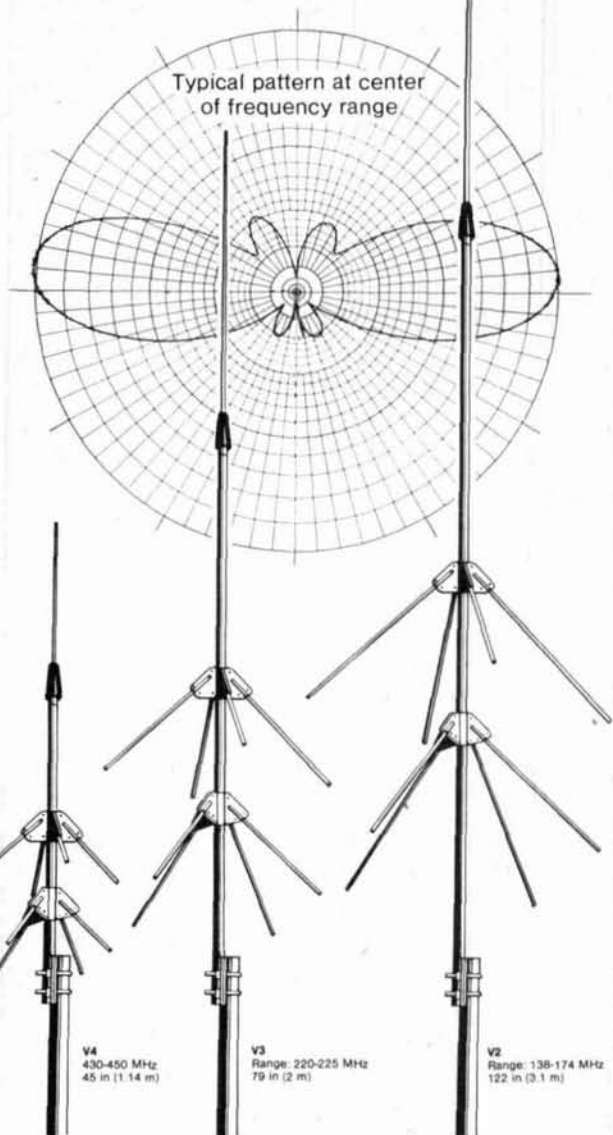
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# MESSAGE PROCESSOR TERMINAL

## MPT3100



Message processing is now available for radio communications systems. The MPT3100 is a complete up-date of the popular HAL DS3100 RTTY terminal, adding the ability to store RTTY messages, edit them, and retransmit them singly or in preset groups. ALL of the previous features of the DS3100 and MSO3100 are retained and new mailbox commands are included. The editor may be used with any file that is stored. The MPT3100 includes ASR (Auto Send-Receive), MSO (Message Storage Option - "mailbox"), and TRO (Traffic Relay Option) modes. The MPT3100 is a new software package that works in ANY DS3100 with MSO3100 circuit board. Some of the features of the MPT3100 are:

### NEW FEATURES OF MPT3100:

- Automatic storage of all received text in files separated by the standard "NNNN" terminator (TRO-REC mode)
- Full editing capability of all files stored by mailbox (MSO) or by TRO storage
- Editor allows insertion or deletion of text in any part of a stored message - 15 keyboard edit commands
- Editor may be used even while receiving, transmitting, or storing messages - even when MSO mailbox is in use
- Files may be renamed, created in the editor, cut into smaller files, and deleted with keyboard commands
- Message files may be transmitted singly or in batches
- Transmitted messages may be serial-numbered automatically
- The full format requirements for NAV MAR COR MARS NTP-8(A) are supported
- New TRO commands include: RXON, RXOFF, DIR, SEND, STOP, RESUME, RESTART, EDIT, CUT, CREATE, QUIT, RENAME, DELETE
- On-screen status indicators show: TRO mode; bytes of memory remaining; file names being recorded, transmitted, and edited
- MSO mailbox .SDIR directory command revised to shorten time required for transmission
- New .DIR [filematch] and .SDIR [filematch] mailbox commands give listing of only file names that include [filematch]
- Programmable "header ID" for each mailbox transmission

### MSO Mailbox Features:

- Programmable MSO call-up command
- Mailbox may be controlled by external station to store message files, read files, delete files, and list the file directory
- DS3100 operator may perform all MSO operations on the keyboard without transmitting
- Mailbox transmissions include user-prompting and automatic CW and RTTY identification
- HELP messages are provided to assist the new user in operation of the mailbox
- All mailbox messages stored may also be edited, renamed, and transmitted using TRO commands
- MSO commands are: .DELETE, .DIR, .DIR [filematch], .ENDFILE, .FILEHELP, .HELP, .KY1ON/OFF, .KY2ON/OFF, .PRINTON/OFF, .QBF, .READ, .RYS, .SDIR, .SDIR [filematch], .WRITE

### DS3100ASR Terminal Features:

- Send and receive ASCII, Baudot, Morse codes
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- On-screen status indicators with real-time indication
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- Current loop or RS232 RTTY input/output
- Positive and negative Morse key outputs
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# ham radio TECHNIQUES

Bill Over  
W6SAI

## RFI revisited

In my September, 1982, column I discussed some of the problems I encountered with severe radio and TV interference generated by my microwave oven. I also told of the subsequent runaround I received from the dealer and the manufacturer's service department. I've received a lot of mail about microwave oven RFI, and I think you might be interested in some of the remarks:

*Art Nichols, W6EVL* — "I returned my first Montgomery Ward microwave oven because it was 40 dB over S9 on 80 meters. The second one I got was better — it was only 20 dB over. I used an external filter until the warranty period was over, then put a pair of 0.1- $\mu$ F, 600-volt capacitors from each side of the line to the case. That did the job fine. Montgomery Ward told me they didn't supply a filter."

*David Cornell, K9BO* — "I live in a medium-signal-strength area outside of St. Louis, Missouri, and own a General Electric Model J-ET015-OY1 oven. The TV shows no interference from the oven, and all I hear is a slight noise on the broadcast band when

the radio's tuned to a quiet spot. This contrasts with your Amana report, suggesting that perhaps some microwave ovens are less objectionable than others in terms of RFI."

---

**"He finally wrote  
the FCC about the  
problem, and  
received a reply  
by telephone."**

---

*Jim Ford, N6JF* — "I was more successful than you in getting a filter for my microwave oven, but it took a letter to the Bureau of Appliance Repairs in Sacramento, California. The filter they installed helped but was no cure. A friend's Sharp microwave oven is clean of RFI. There is no interference to any channel of a nearby TV set that uses a set of rabbit-ears

for an antenna." (Note: Jim's oven is an Amana. The form letters he received from the Customer Specialist at Amana seemed to indicate that they either don't understand the problem or else prefer to pretend it doesn't exist.)

*Francis Stifter*, President, Electronic Specialists, Inc. — "We have had many calls from people with microwave oven interference problems. One of our line-cord filters usually takes care of it. Occasionally it is necessary to fold the oven power cord into a small bank to eliminate the last traces of TVI." (Note: The catalogue of filters and other interference-control devices can be obtained from Electronic Specialists, Inc., 171 South Main St., Natick, Massachusetts 01760.)

*Curt Powell, WB4WAA* — "I had a similar RFI problem with my Montgomery Ward microwave oven Model KSA-8192A." Curt describes a frustrating experience he had with the company. He wrote the FCC about the problem, and received a reply by telephone. The FCC told him that the pulsed power supply energy was creating interference that was radiating down the power cable and being conducted back into the line by the con-

---

***“The FCC may have received brickbats from hams, but here they are strongly on the side of the consumer!”***

---

nection. They informed Curt that the oven had been built by Sharp, and that Sharp was sending “a new transformer.” Eventually the replacement part arrived, was placed in the oven by the Montgomery Ward Service Department, and the problem was solved. As Curt says, “The squeaking wheel gets the grease!” Obviously, the FCC had called both Sharp and Montgomery Ward over this problem. A follow-up letter was sent to Curt by the FCC.

### **the letter from the FCC**

The letter Curt received was from Charles M. Cobbs, Acting Chief, Equipment Authorization Branch, Federal Communications Commission, Office of Science and Technology, Box 429, Columbia, Maryland 21045. The letter said, in part:

“We are very interested in the outcome of cases similar to yours. However, at the time of Type Approval of the model oven you have purchased, a line-conducted specification was not in effect for miscellaneous equipment, as microwave ovens are defined by our rules.

“In an effort to alleviate the problem, we are writing to the manufacturer of your oven and asking them to contact you directly and report to us their solution and specific action they used to bring the oven into a proper operating condition. Please inform us if nothing has been done within a rea-

sonable time.” (signed) Charles M. Cobbs

### **hurray for the FCC**

The FCC may have received brickbats in the past from hams, but here’s an instance in which they are strongly on the side of the consumer! And the Goldwater RFI bill should make the situation a lot better in the future than it has been in the past.

I suggest that if you contact the FCC regarding RFI from your microwave oven or other home equipment, you specify the name, model, and serial number of the equipment and the manufacturer and purchase date. It seems that when the FCC contacts some of these manufacturers, something happens.

### **the ground plane versus the dipole**

In my October column I commented that tests I’d run comparing a ground plane antenna and a dipole seemed to show that the dipole outperformed the vertical both for transmission and reception. That conclusion brought some interesting comments from the readers:

*Harry Hyder, W7IV* — “In the late ‘60s I had both a ground plane vertical and an inverted-V. I was thus able to make direct comparisons. The vertical was a Hustler 4BTV, roof mounted, with four resonant radials, drooping slightly. The inverted-V was 35 feet at the apex and about 20 feet off the ground at the ends.

“At that time, I worked ZS1A frequently on 40-meter CW. I always heard him better on the inverted-V because of the lower noise level. On transmit it was a different story. He always heard me on the vertical (conditions permitting) but when I switched to the dipole my signal was always much weaker and I frequently dropped out completely . . .

“A friend of mine, W7IR, made the first 160-meter WAC from W7-land. He has a 90-foot vertical and a 160-meter inverted-V with the apex at 65 feet. The dipole is better for reception

but is useless for DX transmission; the vertical outperforms the V by a large margin.”

*Warren Amfhar, W0WL* — “From this writer’s experience, when the ZLs are coming through on 80 meters during the early hours and a switchable vertical-to-horizontal antenna is used, the horizontal works better. Why? Because the signal is usually coming via multiple-hop, high-angle propagation.”

*Alan Bloom, N1AL* — “Aren’t you comparing apples and oranges? The radials on a ground plane 12 feet above ground do not act like an infinite ground. This antenna has a vertical radiation pattern similar to that of a dipole unless it has an “infinite” high-conduction ground at its base. Also, could the greater noise on the ground plane have something to do with its location nearby to house wiring? My own pet antenna is a high and long end-fed wire with lots of counterpoise wires. I used to have fun working Europe on 80 CW with my HW-16 with my “invisible” (No. 28 wire) 200-foot end-fed wire.”

*Roy Lewallen, W7EL* — “In an experiment similar to yours, comparing a ground plane antenna with its bottom at 35 feet to an inverted-V dipole at about 40 feet, I found no direction or distance for which the ground plane did as well as the inverted-V on 20 meters.

“On 40 meters, a ground-mounted vertical with a good radial system approximately equals an inverted-V with its apex at 40 feet for signals beyond 500 to 1000 miles; closer stations are better on the inverted-V.

“When comparing vertical and horizontal antennas, signals must be monitored for some time to get accurate results. The antennas may in fact take turns being better than each other by 30 dB, because of polarization rotation of the received signals. On 40 meters, a typical period of rotation is from about one-half minute to a few minutes. The frequently heard question, ‘Now I’m trying antenna number one. Now I’m using antenna number two. Which is better?’ can

lead to results 30 dB in favor of one antenna at one moment, and equally in favor of the other a minute later!

"The effect of trees on vertically polarized signals can be significant. I measured a 10-dB loss through a small stand of fir, pine, and cedar trees which were roughly a quarter-wavelength high. A number of experiments left me convinced that the trees were indeed the culprit.

**"In a few years  
the 24-MHz band  
will be dead . . .  
Let's use the  
band before  
this happens."**

"The absorption of low-angle, vertically polarized signals is well known. One of the best illustrations appears in the Canadian Department of Communications CRC Report 1255 (unfortunately out of print). It clearly shows that absorption is a function of ground conductivity for some distance from the antenna, and it is not affected by the ground radial system as commonly thought (unless, perhaps, the radials extend for many wavelengths from the antenna). Since ground conductivity improves as frequency decreases, because of greater skin depth there is less absorption at lower frequencies. I suspect but haven't confirmed that a vertical is superior at and below 3.5 MHz. When compared with dipoles at heights at which the average Amateur is liable to put them, the vertical is almost certainly superior at the lower frequencies."

W.B. Pretchtl, W3KO — "I must take issue with the statement that it is

difficult to devise a better antenna than a simple dipole. Putting aside the usual trials and tribulations accompanying a new antenna, I ended up with a rectangular loop, 40 feet on top and bottom and 27 feet on each side. The bottom is 7 feet above ground level and center-fed with a 4-to-1 balun and 50-ohm line. The VSWR on 40, 20, and 10 meters is better than 1.2:1. Its performance far exceeds expectations."

W3KO and N5CIE (XYL) point out that the loop functions well with low SWR on three bands. This is certainly a simple tribander antenna that doesn't require tuning and costs very little.

So there you are! It seems that the jury is still out as far as a meaningful decision between the horizontal dipole and ground plane vertical goes. I am interested in readers' comments on their comparison of the two antennas. Join in the fun!

## 18 and 24 MHz

Isn't the 10-MHz band wonderful? Plenty of stations to work, no contests to jar the nerves, and lots of good DX showing up every day. Now it's time to turn our attention to the 18-MHz and 24-MHz bands. I have monitored these bands and notice plenty of European Amateur activity, plus a few signals from South America and the Republic of South Africa. As of this writing (late November), no signals from "down under" or Asia have appeared.

Both bands seem nearly empty in comparison to other services. A few RTTY signals, one or two commercial CW stations — and that's all! The spectrum space seems wasted and I suggest it could be put to better use by permitting the U.S. hams temporary authority to use the bands before they become useless. Remember, the sunspot cycle is declining and in a few years the 24-MHz band will be dead for long-distance communications. It would be nice to use the band before that happens.

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## a precision noise bridge

An rf impedance bridge is an important tool for anyone with more than a casual interest in measuring complex impedances. One typical application for such a bridge is in the design of a phased-array antenna.

Solid-state technology has made it possible to move the rf impedance bridge out of the laboratory and into the average ham shack, but, unfortunately, laboratory accuracy cannot usually be achieved with most of the available units. Although readings within 10 percent are acceptable for most Amateur applications, much better accuracy and discrimination are necessary when taking measurements of antenna-

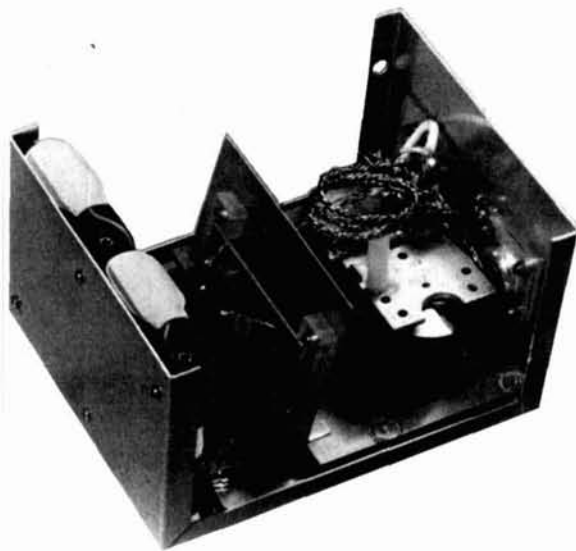
array mutual impedances. We must in fact be able to accurately determine resistive components of impedances which may change as little as only 3 or 4 ohms when a nearby antenna element is mutually coupled. Obviously, most noise bridges can't do the job.

This article explains how it is possible to achieve the needed accuracy in the range of the hf Amateur bands. A multiple-line, distributed-impedance transformer is the critical bridge component. It is not very susceptible to fabrication variations, and it maintains a single calibration over a wide frequency range. With care in adjustment and good construction practice, it is possible to achieve an accuracy of 3 percent over most of the range — with even better results at 14 MHz and lower.

### background

In an excellent article on noise bridges,<sup>1</sup> W6BXI and W6NKU contributed two major innovations for improving noise bridge accuracy: compensating for bridge circuit strays by adjusting inductance in one of the bridge secondary arms; and equalizing primary-to-secondary interactive effects by the addition of a dummy primary wire.

The first idea is one of those insights that seem so obvious after someone else has pointed it out: One wonders why the  $C_p$  calibration "rotation" problem at high frequency ever seemed so difficult. The second suggestion is logical, but I could confirm it only empirically. After winding a dozen different kinds of



Internal view of noise bridge showing coupler and PC board placement.

By Forrest Gehrke, K2BT, 75 Crestview Road, Mtn. Lakes, New Jersey 07046

three-wire and four-wire toroid transformers and comparing results, I leaned toward the four-wire version. I became convinced that control of interactive effects between the primary and secondary was a key factor in achieving the accuracy I wanted.

I used the improvements and suggestions in the referenced article in three of my own bridges, with good results similar to those achieved by the authors: a spread of 2 to 3 ohms (resistance and/or reactance) for a range of accurately known impedances. But further improvement in accuracy was still needed.

W6BXI, in private correspondence, told me that he'd had no particular difficulties with the bridge transformer, either in reproducing it or with circuit adjustment. I wasn't as lucky. But he did discuss a method of matching and balancing the bridge transformer to the noise source. I tried an emitter follower that provided a low-impedance source for the noise amplifier. I also tried a balun toroid transformer, to take advantage of balanced drive to the double-ended bridge circuit from the single-ended noise source. Neither approach worked.

I have probably wound a hundred or more different configurations of the toroid transformer used in this bridge, in an effort to get a better understanding of this key component in the bridge. One possible solution lay in trying to get maximum inductive coupling with minimum capacitance between primary and secondary. So I wound a more classic type of transformer, with the primary on one side of the toroid and a single-wire, center-tapped secondary on the other. Capacitive coupling between windings was very low — 3 pF — but the transformer behaved poorly in the bridge. Calibration barely held over a single Amateur band, and there was no correspondence to the real values of the reference arm of the bridge. Next I wound a second transformer, also with the windings on opposite sides of the toroid, except that now the secondary was a bifilar winding. The bridge worked, and calibration held from 1.8 to 30 MHz. But the noise signal was too low at the high-frequency end, indicating insufficient primary-to-secondary coupling, resulting in poorly defined nulls. Different core materials did not help. This experiment showed me that the required bridge coupler was no ordinary transformer, but something akin to a distributed impedance transmission line.

Most of the trifilar and quadrafilar toroid transformers I tried had been wound with the three or four wires kept carefully parallel. I wound some with the wires twisted together at two or three turns per inch. A stranded bundle of approximately 9 inches (23 cm) was wound on the core. I decided to try connecting up a strand in the usual way, but as a large single-turn loop, without winding it on a toroid core. This worked excellently, without any of the puzzling

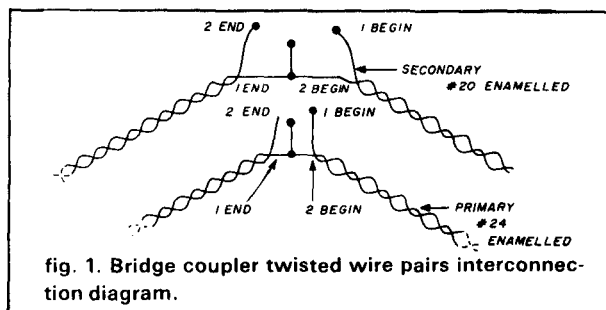


fig. 1. Bridge coupler twisted wire pairs interconnection diagram.

anomalies of the toroid versions. Noise signal was as high as with a toroid, and calibration held constant from 1.8 to 30 MHz. Also, the reference potentiometer settings closely coincided with the value being measured. With this approach there was no question about the four-wire line being superior for holding calibration over a wide range, and my results were reproducible: Reproductions of these lines always produced the same results. I concluded that the toroid core had been contributing most of the anomalies — and that it was not at all essential to this bridge component.

Having eliminated the toroid, I next worked on optimizing line length, wire size(s), and insulation thickness, and on a method of combining the individual wires into a line. The matter of whether adjacent or alternate wires of a single twisted line should be used in connecting up the primary and secondary circuits was also investigated. I tried dozens of variations: No. 24 (0.5 mm) hookup wire with PVC insulation, enameled wire sizes from No. 18 (1.0 mm) to No. 32 (0.2 mm), including mixed sizes, lengths from 4 inches (10 cm) to 16 inches (40 cm), single twisted and compound twisted, and so on. This is what I learned:

1. Close coupling between primary and secondary is absolutely required. This means using enameled wire, twisted as tight as possible.
2. Line lengths between 10 and 12 inches (20 and 30 cm) are best: If too short, noise signal is too low; if too long, a large amount of compensation is required on one secondary half for  $R_p$  adjustment, which is accompanied by a pronounced  $C_p$  calibration shift between 21 and 30 MHz.
3. Compound twist gives the most consistent results (primary and secondary wires are first twisted independently, then these are twisted into a single strand). All four wires can be twisted simultaneously, but they would then have to be interconnected with attention to whether alternate or adjacent wires are being used as the respective primary and secondary. See fig. 1 for winding interconnection detail.
4. Enameled wire size No. 20 (0.8 mm) seems best. Smaller wire sizes are more difficult to connect,

adjust, and compensate because of their fragility. Heavier wire, though effective, resulted in a line that became too stiff to work with. I found a good compromise to be No. 20 (0.8 mm) for the secondary and No. 24 (0.5 mm) for the primary. These lines can be looped into multi-turn coils, to conserve space, with no effect on calibration. In the single-turn calibration no ill effects occur unless the loop is tightly collapsed on itself, and then the most noticeable effect is a resistance calibration shift at low frequencies.

These coupling devices act like distributed-impedance transmission lines. The circuit impedance appears to be low, as it is virtually immune to stray capacitances. As with transmission-line baluns, however, the device displays sensitivity to lead dress of the ends of the lines. Unfortunately, it is not physically possible to keep these ends apart. One end of each pair has to be connected to each other, and the ends connected to the bridge arms must be as short as possible. If, indeed, this is a multiple-wire transmission line, theoretical calculations would point toward a low characteristic impedance. The effect of the re-entrant connection is difficult to assess. I would welcome letters from readers who are able to show the mathematics. The most puzzling aspect, to me, is that a line so short (in terms of wavelength) should be effective at very low frequencies. I have used a bridge calibrated for frequencies between 3.5 and 30 MHz to as low as 150 kHz. The calibration was unchanged and the noise signal level was as high as at 3.5 MHz.

### scale extensions/expansion

The article by W6BXI and W6NKU and subsequently published letter<sup>2</sup> give a technique for measuring impedances beyond the basic range of the bridge by using series or parallel resistances inserted at the **Unknown** terminal. Though this approach is quite effective, the user is warned that, as with any scale multiplier, inaccuracies in the basic range calibrations are also multiplied. Improvement of the base range accuracy should make these range extenders more trustworthy — and using the minimum multiplier possible. For antenna measurements the series extender finds the greatest use; I have several, ranging from 10 to 100 ohms.

For some applications, like mine, there is a need to expand the resistance scale to be able to discriminate between small changes. Since the  $R_p$  range of interest was between 27 and 77 ohms, a 50-ohm potentiometer in series with a 27-ohm fixed resistor was used in the reference side of one of my bridge models. This affords nearly 270 degrees of scale rotation for a translated 50-ohm range. To minimize strays, the potentiometer connections should be as shown in fig. 2.

### transformation equations

A series combination of resistance and reactance can always be found that exhibits the same equivalent impedance as any given parallel combination of resistance and reactance.

The transformations give the relationship between the elements of the series and parallel networks (shown below) when the driving-point impedances are equal. The equations for the respective impedances are:

$$Z_p = \frac{R_p (jX_p)}{R_p + jX_p}$$

$$Z_s = R_s + jX_s$$

Equating the real and imaginary parts of both expressions for network equivalency yields:

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2}$$

$$X_s = \frac{R_p X_p}{R_p^2 + X_p^2}$$

Transforming from series to parallel:

$$R_p = R_s + X_s^2/R_s$$

$$X_p = X_s + R_s^2/X_s$$

### other modifications

The other major changes from prior circuits were to go to a fixed capacitor in the reference arm of the bridge and to place the variable capacitor,  $C_p$ , across the **Unknown** terminal. Operation of the bridge remains the same, except that the meanings of the direction of rotation from "0" on the calibration scale become reversed. That is, an indicated null which *increased* the capacitance above its value at "0" means the measured impedance has an *inductive* reactance component. Conversely, a decrease in capacitance from "0" indicates a capacitive reactance. This change ensures that the reactive component being measured is in direct correspondence with that of the unknown impedance. The difference is a subtle one but, in effect, at nulled conditions the reference arm is always "seeing" the identical situation, the fixed capacitance across the potentiometer.

### adjustment

Two adjustment procedures are necessary, both requiring a bit of care and patience. One is done at low frequency to enable the unknown load resistance to track with the reference potentiometer's nulled resistance. The second adjustment is done at the high frequency end of the range to enable the load resistance to continue to track with the potentiometer and to reduce the  $C_p$  calibration shift. The adjustments are performed in the order given and must be iterated since they are interdependent. They are best done with a good commercial load termination (of minimum reactance), though 50 to 100 ohm quarter-

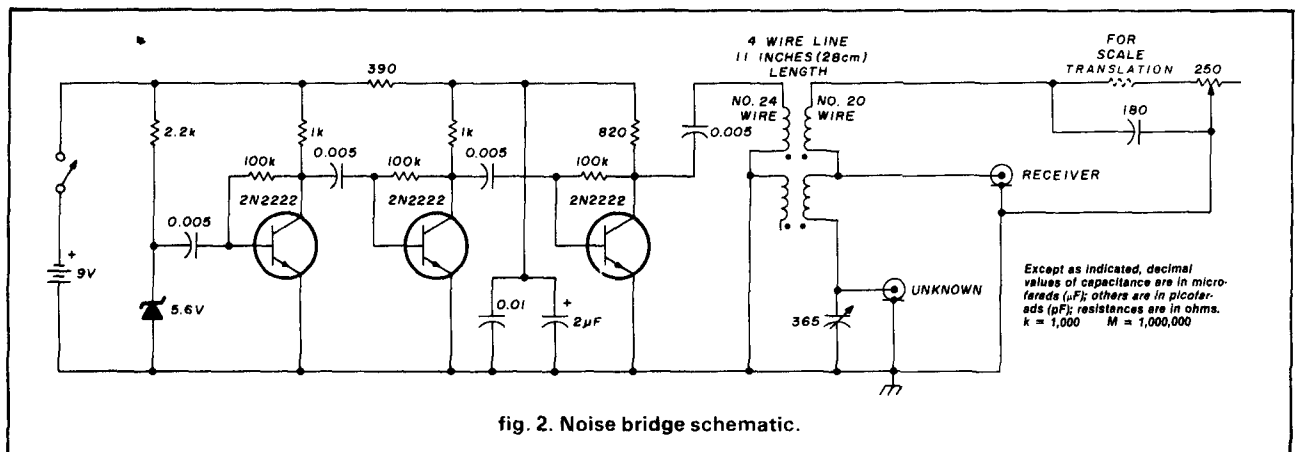


fig. 2. Noise bridge schematic.

watt carbon resistors connected by short leads are a second best alternative. These adjustments ensure that the calibration remains constant over the bandwidth and provides an accurate  $R_p$  readout. I have found multi-tester DWM's to be remarkably accurate ohmmeters and useful in measuring the calibration load resistances and the nulled potentiometer resistance.

The low-frequency adjustment is better described than explained. The adjustment cancels inductive and capacitive strays, and after building six bridge units, only two with identical layouts, I am confident of the procedure:

### low-frequency adjustment

1. Set the null-detecting receiver to a low frequency (I used 3.5 MHz) and null the bridge with a known resistive load of 50 to 100 ohms. The  $R_p$  and  $C_p$  controls are interdependent; repeat for best possible null.
2. Turn off the noise source and remove the load and null detector from their terminals. Measure the potentiometer resistance without disturbing its setting (measure the resistance seen at either the **Unknown** or **Detector** terminals). Measure the load resistance.
3. If there is agreement to within 0.1 ohm go to the high-frequency adjustment procedure. More likely, the agreement will be close but not good enough. With the bridge returned to operation as in step 1, gently bend the bridge line to a new position in any direction by a few degrees. The potentiometer null will change; renull and note whether the potentiometer resistance more closely agrees with the load. If it does not, bend the line in another direction, possibly opposite to the direction of the first trial. If after a few trials it is found that no direction of movement will cause good enough correspondence of readings, go on to the high-frequency compensation procedure. The two procedures are interdependent and

several iterations will be necessary.

### high-frequency adjustment

This adjustment equalizes any residual imbalance of inductance that may exist in the bridge. The adjustment is made at the high-frequency end of the range because this is where the influence of an imbalance is most noticeable. It consists of adding inductance (offset) to the secondary arm found to require compensation. Which arm this is depends upon the direction the  $R_p$  null must be shifted. I have found it always to be required in the reference arm connection if the noise signal is inserted to the primary connection at the same end of the line.

1. Using the same pure resistance load as in the low-frequency adjustment, null the bridge carefully at 30 MHz. Turn off the bridge, remove the load and detector, and compare the potentiometer resistance with that of the load. If this is the same you had at 3.5 MHz you're done. More likely, compensation will be necessary.
2. Add an inch (2.5 cm) of the same size wire used for the secondary to the secondary connection at the reference bridge arm. Returning the bridge to operation, if the  $R_p$  null resistance is higher than the resistance being measured and now goes higher still, remove the wire and add it in on the other secondary arm of the coupler. If it now goes lower, reduce the amount of added wire until the  $R_p$  null resistance matches the termination.
3. In making this adjustment be careful not to disturb the position of the line found for the low-frequency adjustment. In any case, recheck the low-frequency adjustment after completing this procedure. Several repetitions of the two adjustment procedures may be necessary. If more than an inch of secondary-size is required, you may encounter an unacceptable  $C_p$  calibration shift at 30 MHz, although  $R_p$  tracks over the whole range.

These two adjustment procedures force tracking of  $R_p$  at widely separate frequencies. But this does not guarantee flat tracking at frequencies in between, so after completing these adjustments check the calibration in the middle of the range, such as 14 MHz.  $R_p$  calibration should remain well aligned and  $C_p$  shift, compared with the 3.5-MHz position (the center, or "0" of this scale), minimal (less than 1 pF). If all is well, now check with other values of resistance. If you cannot obtain commercial terminations, quarter-watt carbon resistors of the deposited type are quite good, but connect them to the unknown

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*"An rf impedance bridge is an important tool . . . in the design of a phased-array antenna."*

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terminal with as short a lead length as possible. In spite of all your best efforts, you will see more  $C_p$  inductive shift at 30 MHz with these. If you used a 250-ohm potentiometer, expect to see more absolute  $R_p$  shift at 30 MHz than found at 3.5 MHz for resistance at the upper end of this range. My bridge reads 1 ohm low with a 237-ohm load at 30 MHz. At the low end of the range, using a 6-ohm resistor, the  $R_p$  reading is within 0.1 to 0.2 ohm over the whole frequency range, with one bridge reading on the high side at 30 MHz and the other on the low side. A DVM checked against a laboratory standard was used for a resistance calibration standard.

If the potentiometer is not tracking well with frequency, the circuit has layout problems. Here are three general areas to review:

1. Ground loops in the circuitry (multiple connections to ground for the same circuit path).
2. Too much inductance has been allowed to creep into the connections in known or unknown arms of the bridge. One clue is a continuous shift of  $C_p$  center as you step up in frequency bands from 3.5 to 30 MHz.
3. The noise source ground connection to the bridge has too much impedance. This is indicated by a pronounced shift in  $C_p$  center position at 30 MHz, with little or none at 21 MHz. The effect is very different

from the prior case, in which there is gradual shift. Note that this same effect also arises from a requirement for a large amount of compensation at high frequency. Still, this is related; the amount of compensation is an indicator of stray reactance in the circuit.

### circuit hints

Avoid creating ground loops. Have all ground returns going to a common point; this point is best located at some point midway between the circuit connections for the known and unknown bridge arms. Calibration accuracy to 30 MHz requires extreme care here. For example, in the last two units I built the detector coax terminal is isolated from the chassis. Coax (RG-174) is used to continue this line directly to the bridge connections, including ground. The bridge circuit and components are mounted on a printed-circuit board using copper foil on the circuit side. The chassis screw mountings for this board are insulated from the foil. The only connection with the chassis and the copper foil occurs via the SO-239 Unknown coax terminal.

Any stray reactance in this bridge creates calibration shift problems, particularly at 30 MHz, and if severe, even at 14 MHz. I found that liberal use of flat, braided shield (such as is used on RG-58/59 coax) for all bridge connections is indicated. In this regard, the miniature 365-pF variable capacitor recommended in the referenced article has too much internal inductance. The small Japanese air variable capacitors built on heavy aluminum frames have been found to be excellent and quite linear.

As a general rule disproportionate shifts in calibration with increasing frequency are indicative of stray capacitance problems, while linear shifts are due to unwanted inductance.

As for the potentiometers, Mil. Spec. linear-resistance carbon pots are recommended. I found Claro-stat Type 53C1, S-taper to be best. These are available in resistances as low as 50 ohms and are very linear.

As for the noise source, obtain several samples of the zener diode for the noise amplifier. Select the one with the highest noise output over the frequency range, emphasizing 30 MHz. Some diodes are so quiet there will be difficulty obtaining enough signal. Isolate the noise amplifier and its batteries from all grounds except the connection to the common ground at the bridge circuit. In one of my units, flat braid for this ground return solved a stubborn residual  $C_p$  shift problem at 30 MHz.

Unless you have a compelling need to carry a bridge in your hip pocket, don't try to crowd this device into a snuff box. My units are housed in a standard aluminum two-piece box 5 × 4 × 3 inches (13 × 10 × 8 cm). This is small enough, gives you space to work in, and keeps down strays. With the extra



space available another 9-volt transistor battery can be paralleled for extended operation.

## C<sub>p</sub> dial calibration

With R<sub>p</sub> calibration done, you are ready for calibration of the C<sub>p</sub> dial. Assuming you have found minimal or no C<sub>p</sub> shift of the "0" position with a 50-ohm termination at 30 MHz, I recommend that 3.5 MHz be used for this calibration, since it minimizes the effect of lead inductance errors.

In the referenced article<sup>1</sup> the authors suggest the following procedures: Calibrate the variable capacitor against a standard before connecting it to the bridge circuit. Or calibrate the capacitor, with bridge operating, using known values of capacitors in parallel with a resistance at the unknown terminal (50 ohms suggested). After completing the capacitive portion of the dial, temporarily disconnect the 180-pF bridge centering capacitor and work backward toward "0" for the inductive portion of the dial.

The first method is the preferred one. But since I had no capacitor standard available, I used a variation of the second method: After calibrating the capacitive portion of the dial, connect a 50-ohm quarter-wavelength coax cable cut for 3.8 MHz to the **Unknown** terminal. At the far end connect a 39 or 68 ohm resistance with minimum lead length. Find the frequency which nulls this resistance while the C<sub>p</sub> dial is set at "0." Note the frequency and remove the resistor. This is the exact frequency at which your cable is a quarter-wavelength. Now use the same known capacitors in parallel with 50 ohms at the end of this cable to calibrate the inductive side of the dial. The quarter-wavelength coax causes these capacitors to look "negative," that is, inductive by a nearly equal amount. Since these are parallel circuits in series with the coax, a small correction is necessary which increases with increasing capacitance. Assuming a 50-ohm coax with a 50-ohm termination at 3.8 MHz, use **table 1** correction points (interpolating for intermediate calibration points).

## upper frequency limitations

Frequencies higher than about 20 MHz begin to present problems for any circuit measurement device. This is why it is not easy to achieve a constant calibration with good accuracy as we go up in frequency. Here we enter a realm in which a resistor, a capacitor, or a coil can no longer be thought of as discrete elements; indeed, each can be a combination of all three. Since this also applies to the bridge circuit, calibration correction and circuit adjustment schemes become necessary. Reactance is involved, and so these compensations are frequency dependent. This means calibration is meaningful over relatively small frequency ranges. I have checked my noise bridge as high as 100 MHz and found that,

**table 1. Parallel-series-parallel correction for use in calibration of inductive side of C<sub>p</sub> dial.**

capacitance (pF)	
at end of coax	at terminal
50	49.82
75	69.51
100	98.60
120	117.59
130	126.94
140	136.20
150	145.34
160	154.37
170	163.28
180	172.06

while deep nulls can still be detected, the calibration shifts considerably.

## low resistance limitations

Measurement of very low R<sub>p</sub> circuits (below 5 ohms) poses two problems for this bridge: The low R<sub>p</sub>, coming quite close to a short circuit, reduces the C<sub>p</sub> null sensitivity, making it more difficult to determine balance. At the same time, any accompanying reactance, since we are measuring the parallel equivalent of the circuit, results in large excursions of the C<sub>p</sub> dial for relatively small reactances.

As R<sub>p</sub> approaches zero, bridge circuit strays become more significant. Do not depend upon any measurements made with this type of bridge with the R<sub>p</sub> dial setting near zero, since many potentiometers do not actually reach zero resistance at their mechanical stop. I have heard of attempting to determine a quarter-wavelength of coax this way. This bridge can do that and more, but not in this way. If measurement of very low impedances is an objective, use a series extender, or consider a series-type bridge.

## detector considerations

One of the reasons the noise bridge is a relatively simple circuit is that no null detector is included. For this an ordinary receiver is used. This can be the station receiver, and, in these days of accurate receiver frequency readout, it can be a frequency standard as well. For purposes of bridge calibration an Amateur-band-only receiver is adequate, but for most measurement needs a general-coverage receiver is better. The presence of an S-meter is helpful but not necessary.

When making measurements on antennas, a battery-operated receiver is convenient. I use an inexpensive all-band portable and a transistor crystal oscillator for marker frequencies. Since the noise signal is considerable, (S9 + 20dB, off null), receiver sensitivity is no great consideration; too much can lead to difficulty in finding a null because of receiver

AVC. The sharpness and depth of nulls provided by this bridge requires getting used to!

I have used as much as 30 feet of coax between the bridge detector terminal and the receiver with no effect upon calibration.

An fm receiver will not work well with this device, as the noise is almost entirely a-m. If the receiver has AFC it won't work at all, as AFC always shifts the receiver local oscillator off the null frequency.

## applications

Before discussing applications, I want to mention the difference between parallel and series circuits. Each type can be transformed into the other. This bridge measures parallel-circuit equivalent values for both real parallel and series circuits. If a real parallel circuit consisting of a 50-ohm resistor and a 180-pF capacitance is measured, then that is what is read out, regardless of the frequency. But a series circuit of 47.8 ohms and 4078 pF at 3.8 MHz also will be read by the bridge as 50 ohms  $R_p$  and 180 pF  $C_p$ . This point must always be considered when using this bridge.

An excellent tutorial on bridge applications and calculations using complex algebra exists.<sup>3</sup>

## coax cable measurements

If you used my method to calibrate the  $C_p$  dial, you have learned how to use this bridge to find very accurately the frequency for a quarter-wavelength of coax. If you have  $R_p$ , then taking the square root of the product of  $R_p$  and the resistive termination yields the characteristic impedance of this line. The ratio of the physical length of the line to the free-space quarter wavelength for the frequency measured is the velocity factor. If you were working with foam cable, you might be in for some surprises at this point. Neither of your calculations may agree very closely with the nominal values usually quoted. Characteristic impedance can vary as much as 10 percent, and velocity factor is seldom as high as the 0.82 usually given; 0.70 is more likely.

When using this method, choose a termination resistance different from the characteristic impedance, but one which will yield a transformed resistance within the range of the bridge. As a check on your work, try a resistance termination on the line equal to your calculation for characteristic impedance. You should find that the bridge, after being nulled for this resistance, maintains that null over a wide range of frequencies.

I have found that quarter-wavelength cables determined via grid-dip methods are 1 to 3 percent too long, when compared with my results. The grid-dip method introduces error because of the shorting link and the pickup coil.

## antenna measurements

An obvious application for this bridge is the measurement of self and mutual impedances of antenna elements and matching adjustments. Measurement of mutual impedances must be done indirectly and involves fairly complex calculations. (This will be discussed at greater length in forthcoming articles.)

Self impedance measurement is straight-forward: Simply connect the antenna leads to the bridge and adjust for null. Remember to keep lead length the same as in the actual installation, and remember too that the values being read are the parallel-circuit equivalents.

A very nice application for an accurate bridge is measuring the impedance of antennas right in the shack, having previously obtained an accurate measurement of the feedline length and its characteristic impedance. Using a Smith chart or programmable calculator to rotate the measurement back to the antenna saves a lot of legwork.

One point particularly applicable to antenna measurements is, make doubly sure of all connections and joints. This bridge operates with noise power measured in microwatts. Poor connections, which do not show up in normal operations, even when driven QRP, will become evident. A few watts may temporarily "weld" poor connections; the bridge hasn't enough power to do that.

## impedance transformers and networks

Another useful application for this bridge is measurement of the input and output impedances of transformers and networks. Remember that the readings are in parallel circuit form, and that the terminations may be in either form. It's sometimes easier to arrive at one than the other with available components.

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1. Robert Hubbs, W6BXI, and Frank Doting, W6NKU, "Improvements to the RX Noise Bridge," *ham radio*, February, 1977, page 10.
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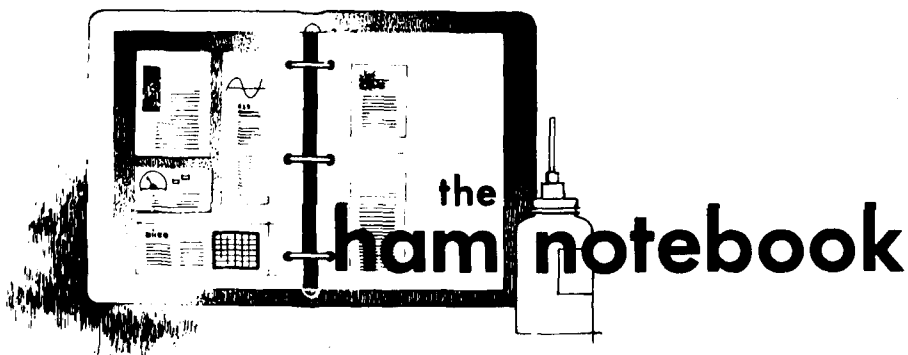
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## DC to 400-Hz ac converter

There are many inexpensive transceivers on the market that require 400 Hz. After I bought an ARC 38A, I was confronted with the problem of how to obtain 400 Hz for the servos.

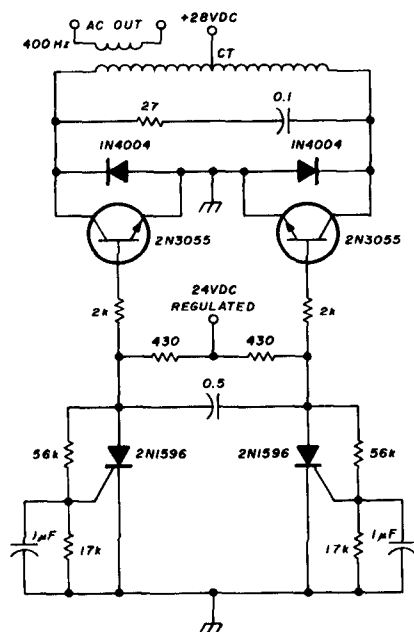


fig. 1. DC to 400-Hz ac converter.

Self-oscillating transformer circuits exist, however they need special transformer ratios to work and if load-

ed they like to change frequency or stop oscillating. A 555-oscillator chip with a driver and a push-pull output is another choice, but I wanted something simpler.

The circuit shown works well and is not too elaborate (fig. 1). It depends on inherent unbalance to start. Any small SCR will do.

In the oscillator<sup>1</sup> the voltage divider starts charging the capacitor until one SCR breaks down, initiating oscillation. The driver transistors (2N3055 or equivalent) drive the transformer with a square wave. Little power is dissipated in the transistors since they are mostly on or cut off. Diodes from collector to ground prevent a negative kickback voltage from appearing. 27 ohms and 0.1 µF are used to dampen spikes.

### reference

1. W.B. McCartney and E.O. Uhrig, "Astable High Power Multivibrator," *IEEE* 10;12, pages 30-31.

Kurt A. Bittmann, WB2YVY

## screen protection for the 5CX1500A

If anyone is considering building a linear amplifier around the 5CX1500A tube, he should not neglect to include screen protection circuitry. Because of the preponderance of grounded-grid triode amplifiers in the last few

years, this important point is now only superficially mentioned in the Amateur literature, and I have not seen it incorporated in published circuits. It is all too easy, however, to exceed the screen rating if there is instability during testing, or by resonating the tank with light loading.

During the construction of a 4CX-1000A amplifier I came across an article on tetrode screen current in an old *QST* (David D. Meacham, W6EMD, "Understanding Tetrode Screen Current," July, 1961). It convinced me of the importance of the screen overcurrent relay, and I designed an appropriate circuit based on the author's ideas (see fig. 2.)

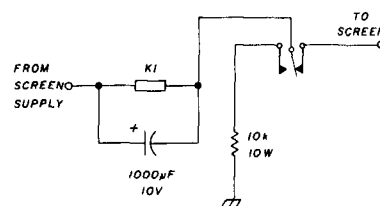


fig. 2. Screen current relay.

I selected a relay sensitive enough to pull in if the current exceeded 60 mA, and yet with resistance low enough not to impair screen regulation. In my case, this was a Japanese-made subminiature relay with a 6 volt, 80-ohm coil, but ITT also makes a line of similar relays. R1 must draw enough current to hold the relay on with the supply voltage in use. A relay requires far less current to hold it on than to pull it in.

During testing and tune-up, this relay has paid for itself many times over. Thanks to its quick action, my final has avoided the fate of irreversibly turning into a low- $\mu$  triode.

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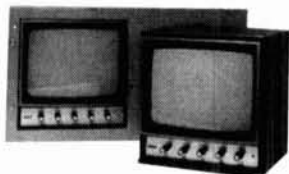
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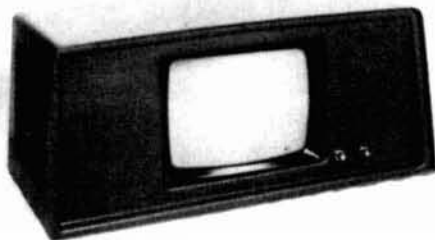
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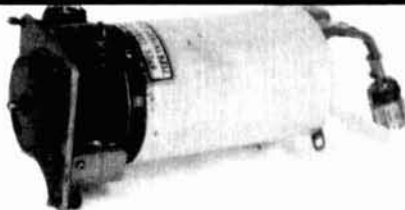
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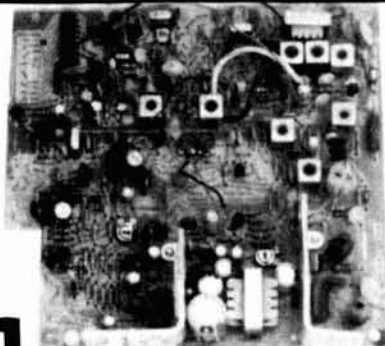
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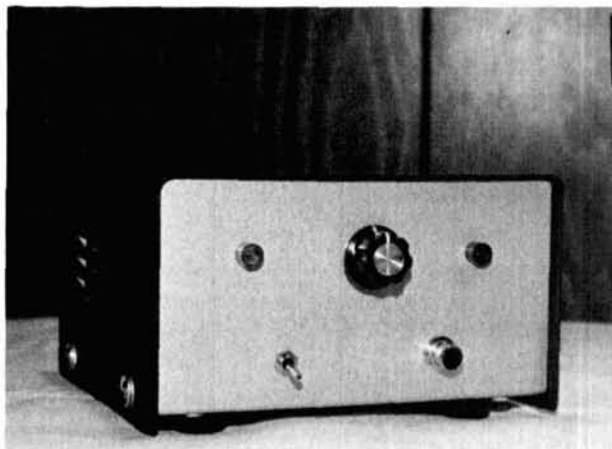
A changeover system  
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For the Novice operator and the CW fan, there are many older transmitters and receivers that will still perform with the best of the modern transceivers. And, for many of us, the price is right. They take up a bit more room, and we may have to work on them to hold down the drift and keep the performance up. But that is part of the challenge we accept when we decide to use such classics as the Heath DXs, the Globes, the Hammerlund HQs, or the National NCs.

Perhaps one of the greatest inconveniences inherent to most of these otherwise fine designs, however, is the lack of an automatic or semi-automatic transmit-receive system. In fact, most of these rigs lack an antenna change-over relay: switching is done manually. It's necessary to turn the transmitter from standby to operate and mute the receiver as well. In some cases, you must also turn on the VFO. All this switching can get in the way of smooth operating.

## CWX: tube style

Hams solved the basic problem of system switching long ago with a system called CWX, CW-operated relay. A CWX automatically switches all rig functions from receive to transmit with the first tap



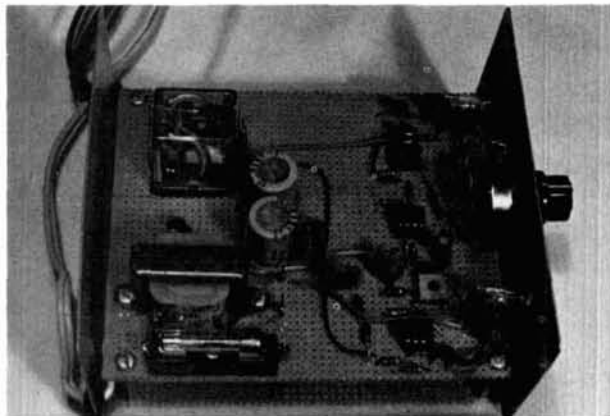
The front panel of SSCWX is simplicity itself. The top row consists of the ac power LED indicator, the timing potentiometer, and the transmitter-on LED indicator. In the bottom are the ac toggle switch and the key jack.

By **L.B. Cebik, W4RNL**, 5105 Holston Hills Road, Knoxville, Tennessee 37914



of the key. In most units, a delay circuit holds the rig in transmit mode between dots, letters, and words, but after a second or so the relay returns everything to receive.

Most of the good CWX circuits used tubes and plate relays. Fig. 1 shows a typical circuit, used at W4RNL for many years in conjunction with a DX-60A transmitter and a 2-B receiver. It is a modified version of a circuit that originally appeared in *QST* in February of 1966. The modification consists of eliminating one relay and tube section (which followed



The interior layout of SSCWX. In the left rear of the perf board are the power supply components and the function-changing relay. In the front right are (from bottom to top) the opto-isolator, the voltage comparator, and the switching transistors (mounted in an 8-pin DIP socket). Layout is non-critical if all leads are well bypassed as they enter the cabinet.

the keying) and replacing it with the pair of diodes near the key jack. With blocked-grid keying, this system is effective.

This old circuit will still work well with almost any miniature triode, and it has some advantages that are worth preserving. Note that the key line to the transmitter goes through spare contacts on the antenna relay (a Dow Key rf relay, in this case) so that the transmitter cannot generate rf until the antenna changes over. Spare contacts on the plate relay control the grid-block voltage to the VFO, holding the unit on during the entire transmit cycle. This feature eliminates chirp when the VFO frequency is multiplied to 20, 15, and 10 meters (a problem common to many inexpensive VFOs). The potentiometer permits adjustment of the delay time to suit varying code speeds. And finally, the unit is compact and self-powered, making it easy to fit either on the table or in a spare corner of transmitter or receiver.

The unit does use tubes, however, which have grown expensive in this silicon age. And plate relays are growing harder to find, although 12-volt relays are still plentiful. As a result, when one of the students in my license-upgrade class asked about an antenna change-over device for his DX-60, I wondered if I could preserve all the good qualities of the old CW in a solid-state device using easily obtainable components. SSCWX is the result.

### CWX: solid state

The solid-state version of CWX is designed for blocked-grid keying and a variety of auxiliary func-

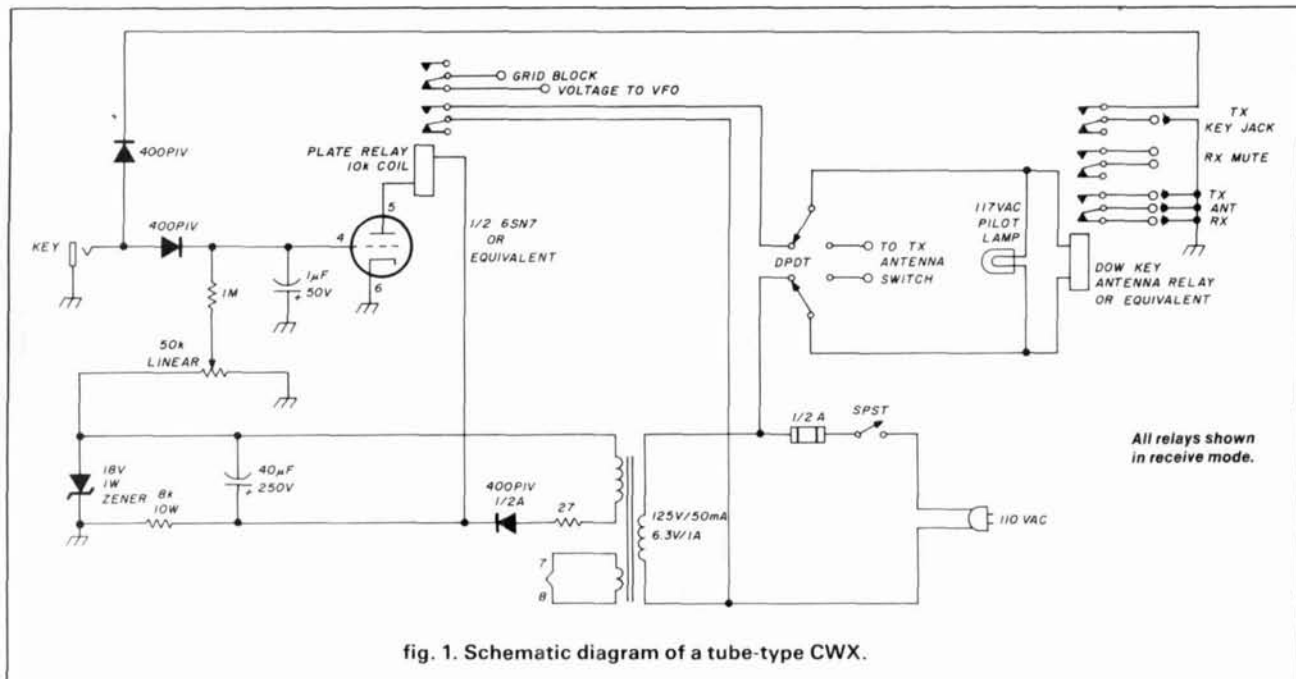


fig. 1. Schematic diagram of a tube-type CWX.

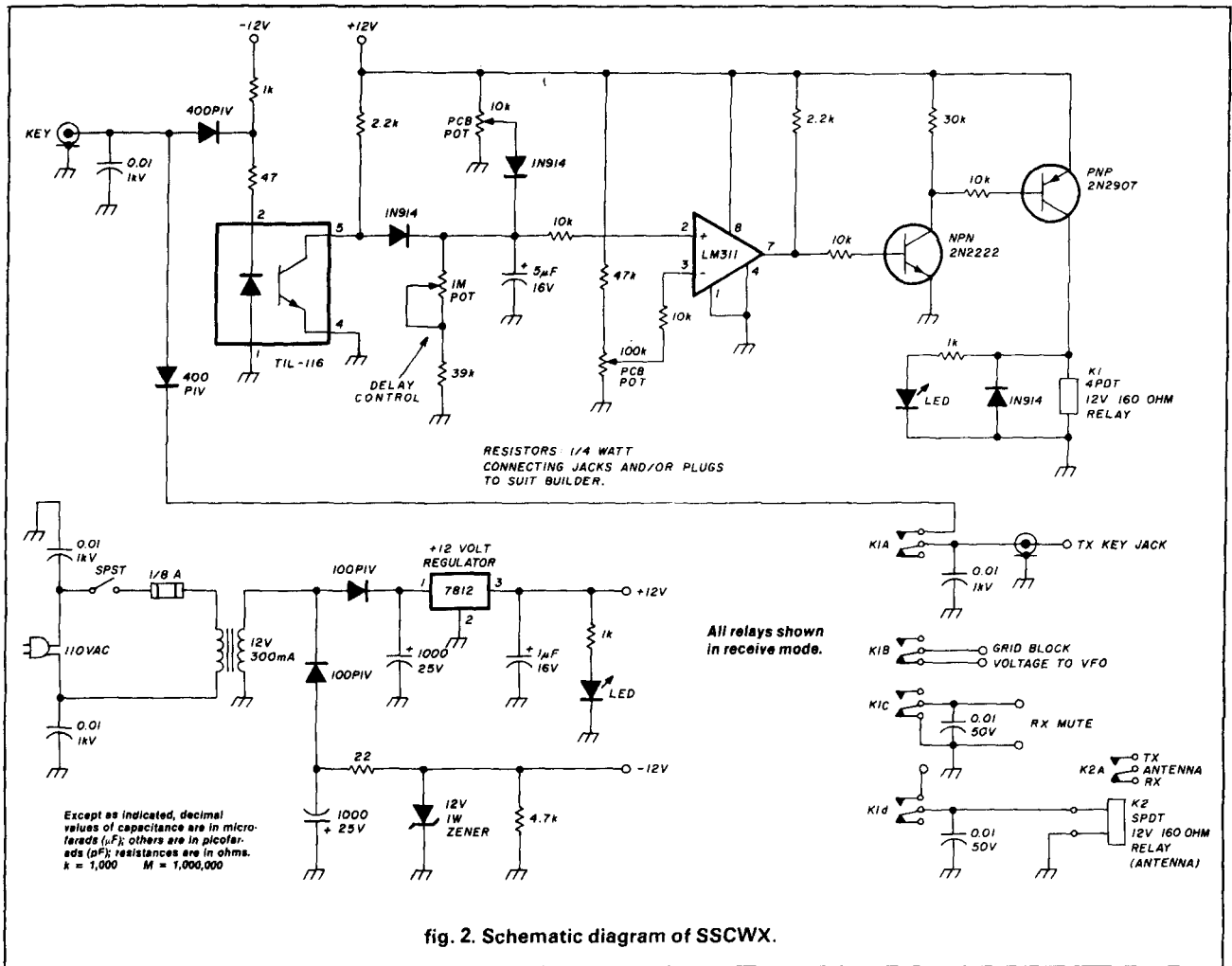


fig. 2. Schematic diagram of SSCWX.

tions, such as receiver muting, VFO control, and rf generation control. Given the variety of receivers available, each with a different muting system, control by relay offers more versatility than does solid-state switching. (Notes on customizing the design and expanding its use to other relay-switching applications appear at the end of this article.)

In older transmitters keying is usually either cathode (positive voltage, heavy current) or blocked grid (negative voltage, low current). In general, you should stay away from cathode keying, because the current will eventually wear key contacts or cause damage to the switching circuits of many keyers. Cathode-keyed transmitters can be modified to a blocked-grid system, but notes on altering SSCWX for positive keying also appear at the end of the article.

### circuit operation

The SSCWX circuit performs in the following sequence (see fig. 2). The TIL-116 is an optoisolator

used to separate the key line from the switching circuits. The diode is lit with negative voltage, corresponding to the grid block voltage on the key line. Key down turns off the diode and the transistor stops conducting, sending the collector voltage to the plus terminal of the LM311 voltage comparator and to the  $5\text{-}\mu F$  timing capacitor. The 1N914 diode in this line keeps the capacitor from discharging through the optoisolator when the key goes up, and the TIL-116 once more conducts.

Between the optoisolator and the voltage comparator is the timing circuit. The electrolytic capacitor charges quickly but discharges slowly through the 1-megohm timing potentiometer and the 39k fixed resistor. The fixed resistor controls the shortest time delay. The combination of the 47k resistor and 100k pot in the minus comparator input line controls the longest delay time, because the LM311 will change state whenever the voltages at its inputs pass each other. About 7 volts to pin 3 of the LM311 yields al-

most 3 seconds of delay — more than enough for most purposes.

The turn-on time is a function of the timing capacitor and the 2.2k collector resistor of the TIL-116. A single short dit will not yield the full delay time, without the supplementary charging circuit composed of the 10k pot and the second diode. By holding the timing capacitor at a level about 1 volt less than the comparator negative input line, the delay time is consistent for single key taps and actual CW transmissions.

The voltage comparator provides a full output swing whenever the capacitor voltage moves above or below the negative input level, going high when the plus line exceeds the minus and low otherwise. This swing triggers the two-transistor switch to the relay, keying it with definite action. The definiteness of the wave shape controlling the relay is the primary reason for using the comparator to isolate the timing circuit from the relay. Timing circuits tied directly to the 2N2222 base drive both transistors through the linear range, and the relay closes and opens as the voltage and current pass the threshold level for its coil. Neither of these phenomena is desirable. Since transistors draw base current (while the older tube circuit did not draw grid current), the devices are easily overloaded during long delay periods. In addition, many inexpensive low-voltage relays do not open cleanly as the voltage drops slowly, creating some sparking and consequent contact damage. Clean switching to the relay and to the controlling transistors is necessary, and the LM311 (or any similar voltage comparator having a fairly high input impedance and able to work from a single 12-volt line) performs admirably.

The 12-volt relay controls four circuits: receiver mute, VFO, antenna relay, and (possibly) the key line. Since Dow Key relays have grown very expensive, most hams have turned to ordinary heavy-duty contact relays for antenna switching. If the internal contact lines are fairly short, and if the relay is well shielded in a metal box with short leads to the three jacks (antenna, transmitter, and receiver), ordinary units perform well and do not increase SWR. If you have such a relay, however, do not use the extra contacts for key lines, muting or other such purposes, as they will be in a heavy rf field. Use a different relay. The circuit compromise in the SSCWX design does not guarantee that no rf will emerge from the transmitter until the antenna relay has changed over, but it does guarantee that malfunctions in the CWX unit itself will not permit you to pump rf into the receiver. Fig. 3 shows the relay circuit for use with Dow Key and similar rf-shielded relays.

Note that the VFO line shows the grid block voltage being controlled by the CWX relay. This repre-

sents a line from the VFO to the CWX to ensure that the grid blocking voltage is off for the entire transmit cycle. Modification for other schemes that help control VFO stability is simple. Some hams prefer to add voltage-controlled diodes to VFOs to shift them off frequency when the key is up, in which case the relay can handle the control voltage instead of the grid block voltage. Whatever the system, I do recommend that you use something to hold a tube-type VFO on for at least the entire transmit period, since the transition from cut-off to full plate current adversely affects short-term VFO stability, and signal quality as well.

Power for the SSCWX comes from a very small supply. The 7812 positive voltage regulator needs no heatsink in this application, but adding one cannot hurt, especially if the unit provides power to other accessories. If so, increase the transformer rating as well, since the entire unit, with two 75-mA relays, requires over two-thirds of the rated 300 mA from the secondary. The requirements for negative voltage are small, and the Zener may not even be required, although it does make the voltage predictable.

### construction and adjustment

The entire SSCWX unit fits on a small piece of perf board, cut to fit a standard 5-1/4 × 3 × 5-7/8 inch (13.3 × 7.6 × 14.9 cm) box, with room to spare. The relay and the power transformer are the largest components, as the photograph clearly shows. Layout is not critical. Only the timing potentiometer (1 megohm) goes to the front panel; the other two are PC-board types. IC sockets are handy, and I even mounted the switching transistors in a single eight-pin socket.

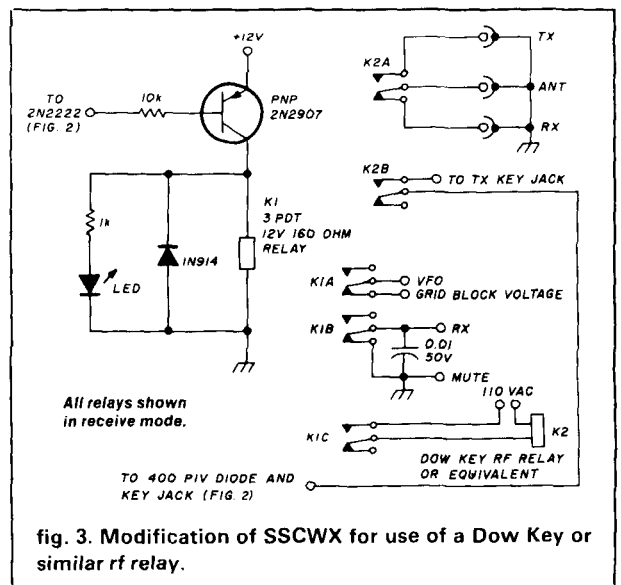


fig. 3. Modification of SSCWX for use of a Dow Key or similar rf relay.

The most critical components may be the 0.01- $\mu$ F bypass capacitors. Voltage ratings for these capacitors are indicated, because some lines may carry fairly high voltages. This is true of the ac line and the key line at both its input and output points. All control lines running from the CWX relay are shielded and bypassed. The object is to be as certain as possible that rf cannot enter the CWX and disrupt any of the circuits.

The power supply diodes are 1N4002 100-PIV units. 1N914s work well in the timing circuit. The key line requires diodes with a PIV rating twice that of the anticipated transmitter key line voltage. 1N4004 (400-PIV) is specified, but anything with a higher voltage rating will work as well. Finally, be sure not to omit the reverse diode across the relay coil; it prevents reverse voltages generated by the collapsing relay coil field from reaching the switching transistor.

Besides the timing potentiometer, the front panel contains only the ac switch, the key jack, and two LEDs, one to indicate that power is on, the other to show that the relay is closed. Although it is possible to use a timing pot with an ac switch, the separate power switch permits you to retain your timing setting from one day to the next. In addition, you might want to consider adding a SPST toggle switch from the +12 volt line to the relay in the collector circuit of the 2N2907; this would provide manual relay switching. The photo shows a phono jack used for the key, but any type will do.

The perf-board construction technique used for this project is good for one-of-a-kind projects in which lead length is no problem. As a group or club project, though, the circuit lends itself to printed-circuit-board fabrication. Build the circuit one step at a time, starting with the power supplies. If you build the remaining circuitry as a unit, at the very least test it progressively by plugging in only one solid-state device between checks. The 1k resistor in the TIL-116 diode line should work for most opto-isolators. The desired value is the highest that will just move the collector voltage (pin 5) from full to zero to full again under keying.

Next, add the LM311 and set the voltage comparator's minus input voltage to about 7 volts. Check the maximum time delay by measuring the output voltage swing. If the time is too long for your taste, increase the negative line reference voltage; if too short, decrease it. Perform these tests by holding the key down for at least one second to ensure the timing capacitor is fully charged, then release and time the delay. Be sure also that the supplementary charging voltage is at least a volt below the reference voltage. Once the long time delay is set, you can set the 10k pot so that the supplementary charging voltage

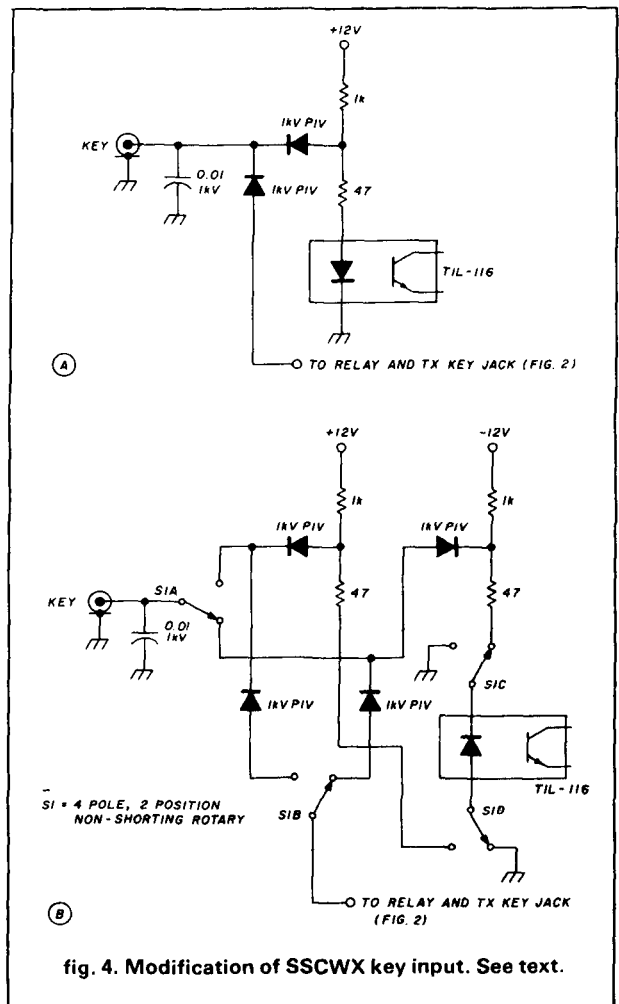


fig. 4. Modification of SSCWX key input. See text.

is between a half and a full volt less than the reference voltage.

Now check the delay at the minimum end of the line. If the timing pot allows the relay to open and close with every dit for more than 10 degrees of pot adjustment, then increase the series fixed resistor above the 39k value shown. If the relay does not come close enough to following your keying at minimum pot setting, then decrease the value of the fixed resistor. Finally, recheck the timing at the maximum end of the scale and do any final tweaking necessary.

Add the switching transistors, one at a time, and check their operation by measuring the collector voltages. When all is well, plug in the relay. If it operates well, mount the unit in the case. In the model shown in the photo, I set the perf board on 1-inch threaded pillars to keep the relay socket contacts well clear of the case. The only solder connection needed was for the key jack, since all other leads, indicators, and controls pass through the front panel from the rear.

The final test is with your transmitter and receiver.

Plug your key into the SSCWX unit. Be sure that the transmitter key jack is empty initially. Check the receiver mute line to be certain that the relay quiets the unit with the key down. Next, check the antenna relay for correct operation. The VFO control line is next. Finally, with a dummy load, plug the key line into the transmitter key jack. Only after you have tested the unit in this sequence should you put it on the air. In fact, you might want to unplug the key line from the transmitter (or shut the transmitter off) and get the feel of operating with the SSCWX, using only your receiver. It takes a little while to overcome the urge to grab a switch when going from transmit to receive and back again.

### modifications and other uses

Listed below are some optional modifications that make the basic circuit a more versatile tool around the shack, and applicable to many projects.

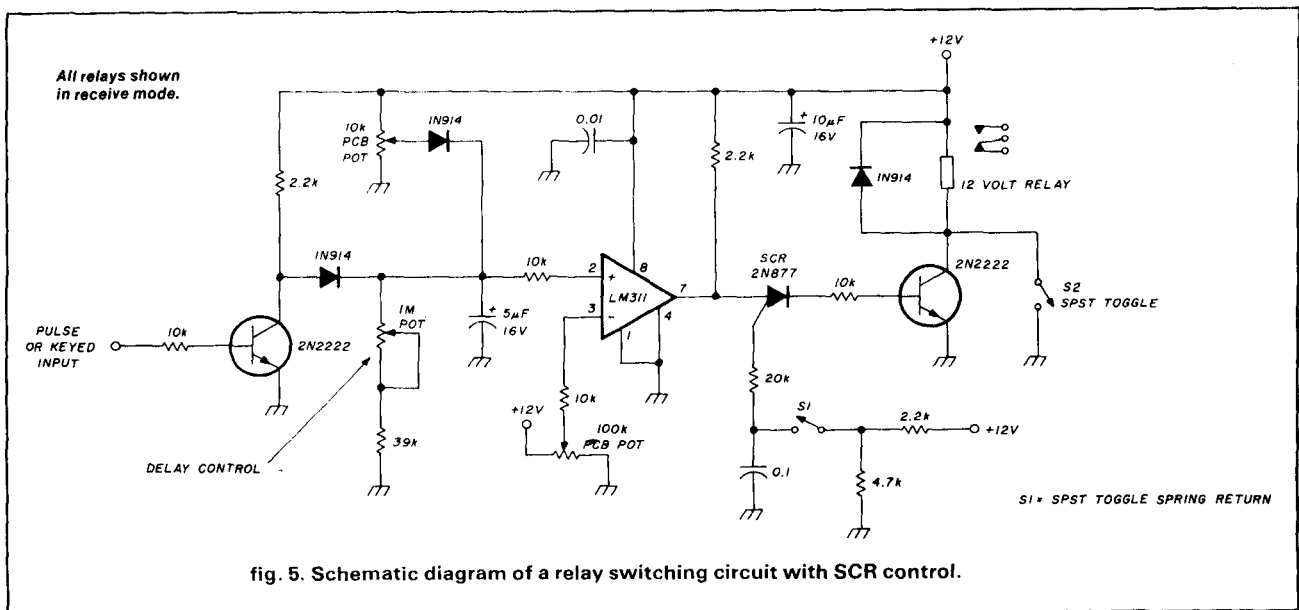
Fig. 4A shows a revision to the input circuit for positive keying voltages. Simply reverse the leads to the diode section of the TIL-116 and change the supply voltage to +12 volts. As shown in fig. 4B, a 4PDT non-shorting rotary switch permits easy conversion from negative to positive line keying. You may want to add indicator lights to this circuit.

The basic circuitry of SSCWX lends itself to almost any timed relay control application. Where isolation is unnecessary, a common NPN switching transistor can replace the opto-isolator. At the other end of the line, small relays with light coil-current demands do not require a double transistor switch. Placing the relay in the collector circuit of the 2N2222 provides reliable service for up to about 50 mA of coil

current. This arrangement provides excellent relay switching for CMOS controlled circuits, and the scheme is easily converted to TTL 5-volt levels.

To provide manual relay starting with automatic dropout, we can add a small SCR to the circuit, as shown in fig. 5. Any small plastic (TO-92) or metal (TO-5 to TO-18) case SCR, such as the 2N877 (rated at 30 volts and 0.5 amp) will work well in the circuit;

part	value or specification	Radio Shack No.
resistors	1 22-ohm, 1/4 watt	
	1 47-ohm, 1/4 watt	
	3 1-kilohm, 1/4 watt	
	2 2.2-kilohm, 1/4 watt	
	1 4.7-kilohm, 1/4 watt	
	4 10-kilohm, 1/4 watt	
	1 30-kilohm, 1/4 watt	
	1 39-kilohm, 1/4 watt	
	1 47-kilohm, 1/4 watt	
	1 1-megohm, panel mounting type	271-211
potentiometers	1 10-kilohm, pc board type	271-335
	1 100-kilohm, pc board type	271-338
	1 100-kilohm, pc board type	271-338
capacitors	2 1000- $\mu$ F, 25-volt electrolytic	272-1019
	1 1- $\mu$ F, 16-volt tantalum	
	1 5- $\mu$ F, 16-volt electrolytic	272-1024
	4 0.01- $\mu$ F, 1 kV disc	
	2 0.01- $\mu$ F, 50-volt disc	
diodes	2 400-PIV (1N4004)	276-1103
	2 100-PIV (1N4002)	276-1101
	3 1N914 switching diodes, silicon	276-1122
ICs	1 1N4742, 12-volt, 1-watt Zener	276-563
	1 7812 positive 12-volt regulator	276-1771
transistors	1 LM311	276-132
	1 TIL-116 or TIL-111	276-1617
relays	1 2N2222 NPN, 40-volt, 100 mA	276-1617
	1 2N2907 PNP, 40-volt, 500-mA	
transformer	1 4PDT 12-volt, 160-ohms	275-214
	1 socket for above	275-221
	1 DPDT 12-volt, 160-ohms, 10 amp contacts	275-218
switch	1 12.6-volt, 300-mA secondary	273-1385
	1 SPST miniature toggle	275-624
fuse	1 1/8 amp, slow blow	
	2 0.2-inch-diameter, with panel lens	
miscellaneous	3 8-pin DIP IC sockets	
	hardware perf board case (5-1/4 x 3 x 5-7/8 inches) key jack and output connectors or cables to suit builder	270-253



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• Toll Restrict	NO	YES
• LED Digital Display	NO	YES
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• Directly Interfaces with Repeater	NO	YES
• Rotary Dial System (incl. Last digit dial)	NO	YES—"Option"—\$49.95
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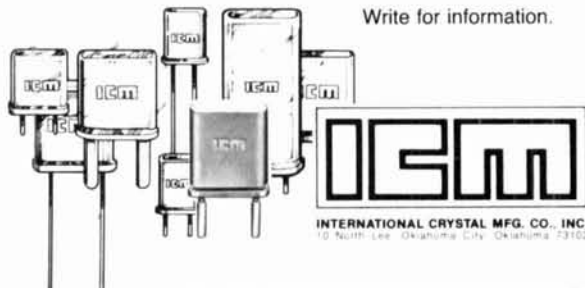
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avoid high-power SCRs. S1, a spring-return SPST toggle switch, provides triggering voltage to the SCR, which then conducts until the output of the LM311 goes low (even though the switch has been released). In short, the SCR blocks the comparator output from the relay transistor until fired by a manual touch of the switch, but then the SCR continues conducting until the input signal disappears and the time delay set by the LM311 input circuitry ends. A second SPST toggle, S2, provides manual relay control.

The SCR gate circuitry provides a low voltage and current to trigger the gate. For more rapid gate voltage decay after release of S1, parallel a 100 kilohm, 1/4-watt resistor across the 0.1- $\mu$ F capacitor. Although all circuit values have worked well for several unmarked SCRs in TO-92 cases, adjust the voltage divider and series gate resistor values to suit the devices you have on hand.

The circuit in **fig. 5** has numerous uses. Without the SCR, a train of digital pulses will key and hold the relay until a specified time after the last one. With the SCR, the relay pulls in upon manual starting with S1, a handy feature with very narrow passband CW audio filters. The use of an LED indicator in the filter, to indicate visually when the signal is locked, permits switch-over to the filter with no loss of copy while trying to tune the filter to the signal (or vice versa). A similar system can eliminate readout garbage in RTTY systems.

The SSCWX switching system provides a sound basis for designing time-controlled relay switching systems for many applications. Apart from the relay, which may range in size from DIP scale to one inch square and two high, the circuit requires a couple of square inches of perf or etched board space, plus the panel pot and any toggle switches. The solid state CWX unit is small enough (including power supply) to mount inside the cabinet of most older transmitters and receivers. There is room on most front panels for the few controls required. On the DX-60, for example, the audio gain control is convertible to CWX duty if the a-m feature is dispensable. A little rewiring of the key jack finishes the job, since the indicator lights are no longer needed. The only precaution is to be sure that the SSCWX is well shielded and bypassed from the DX-60's rf.

Hidden or in the open, the SSCWX makes a good addition to any shack using separate units to transmit and receive. The ease my old CWX gave to my CW operations back in the '60s made it deserving of an update. SSCWX is that update. The basic circuit and its variations can solve relay switching problems for most any ham or experimenter.

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# capacitively coupled hybrids

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Do you need a device that can split or combine power? How about a phase shifter or a circular polarizer? Power can be divided, combined, sampled, or phase-shifted in any increment by using a hybrid. A capacitively coupled hybrid is easily constructed and very compact.

The capacitively coupled hybrid is a four-port device consisting of two transmission lines. Output power at one port of the coupled transmission line is dependent on the direction of propagation in the main transmission line. In **fig. 1**, a wave traveling from port 1 (input) to port 2 is coupled to port 4. Ideally, no power appears at the isolation port 3. A wave traveling in the opposite direction, from port 2 to port 1, is coupled to port 3, indicating any mismatch at port 4. The coupling factor determines the amount of power coupled from the main line to the power output at the coupled-line output port, assuming matched loads present at all ports.

General expression

$$\text{Coupling} = 10 \log_{10} (P_1/P_4) \text{dB} \quad (1)$$

Coupling factor for the capacitively coupled hybrid

$$\text{Coupling (dB)} = 20 \log_{10} \cos \theta \quad (2)$$

where  $\theta$  is the electrical length of the transmission line. The capacitive reactance of the coupling capacitor is:

$$X_c = Z_o \cdot \tan \theta \quad (3)$$

where  $Z_o$  is the impedance of the transmission line. The line impedance is equal to the termination impedance at each port. The value of capacitance is:

$$C_c = \frac{1}{2\pi f X_c} = \frac{1}{2\pi f Z_o \tan \theta} \quad (4)$$

Because the device is reciprocal, a wave incident at port 2 is coupled to port 3 by the same coupling factor.

The hybrid described here is smaller than the Wilkinson hybrid (described in an earlier article<sup>1</sup>), easy to construct, and displays improved amplifier performance. The principal disadvantage of this hybrid is its limited bandwidth of 10 percent. This, however, is still adequate for most Amateur use.

## coupler operation

A wave traveling to the right (port 1 to port 2) on the main transmission line will have a portion of its energy coupled to the second transmission line at each end by capacitors. It is assumed that there is no inductive coupling. The two signals arriving from both paths at port 4 are in phase and combine. The signals at port 3 are also equal (**fig. 1B**) but 180 degrees out-of-phase and cancel. The longer path always has 180 degrees more phase shift because of the transmission lines. (The phase shifts that result

By Ernie Franke, WA2EWT, 63 Hunting Lane, Goode, Virginia 24556



from the coupling capacitors cancel out because they are present in both paths.)

In case of a wave traveling in the reverse direction (port 2 to port 1), the coupled energy combines at port 3 and cancels at port 4, with the capacitively coupled transmission lines acting as a directional coupler. A sample of the energy from a forward wave appears at port 4. A sample of the reflected wave appears at port 3. The ratio of power at port 3 to power at port 4 yields the return loss.

$$\begin{aligned} \text{return loss} &= -10 \log_{10} (P_3/P_4) \\ &= -10 \log_{10} (\text{reflected power}/\text{forward power}) \end{aligned}$$

### directional coupler

One's first thought would probably be to use the hybrid for measuring standing wave ratios. A detector placed at port 4 would indicate forward power, while a detector at port 3 would indicate reflected power. But how isolated are the two readings? How well is port 3 isolated from port 4?

Directivity is a measure of how well port 3 is isolated from port 4 with respect to power entering at terminal 1. Directivity is defined as:

$$\text{directivity} = 10 \log_{10} \frac{P_4}{P_3} \quad (6)$$

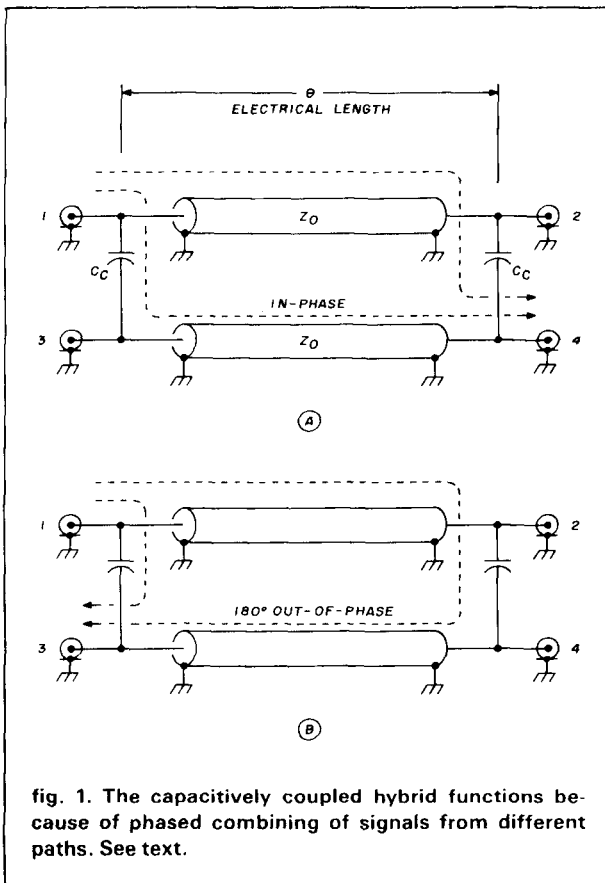


fig. 1. The capacitively coupled hybrid functions because of phased combining of signals from different paths. See text.

Ideally, the directivity should be infinite — but a directional coupler with a directivity of 20 to 30 dB is quite good. How high a directivity can be achieved is strongly dependent on the match at port 4, since reflected power is readily transmitted back to port 3.

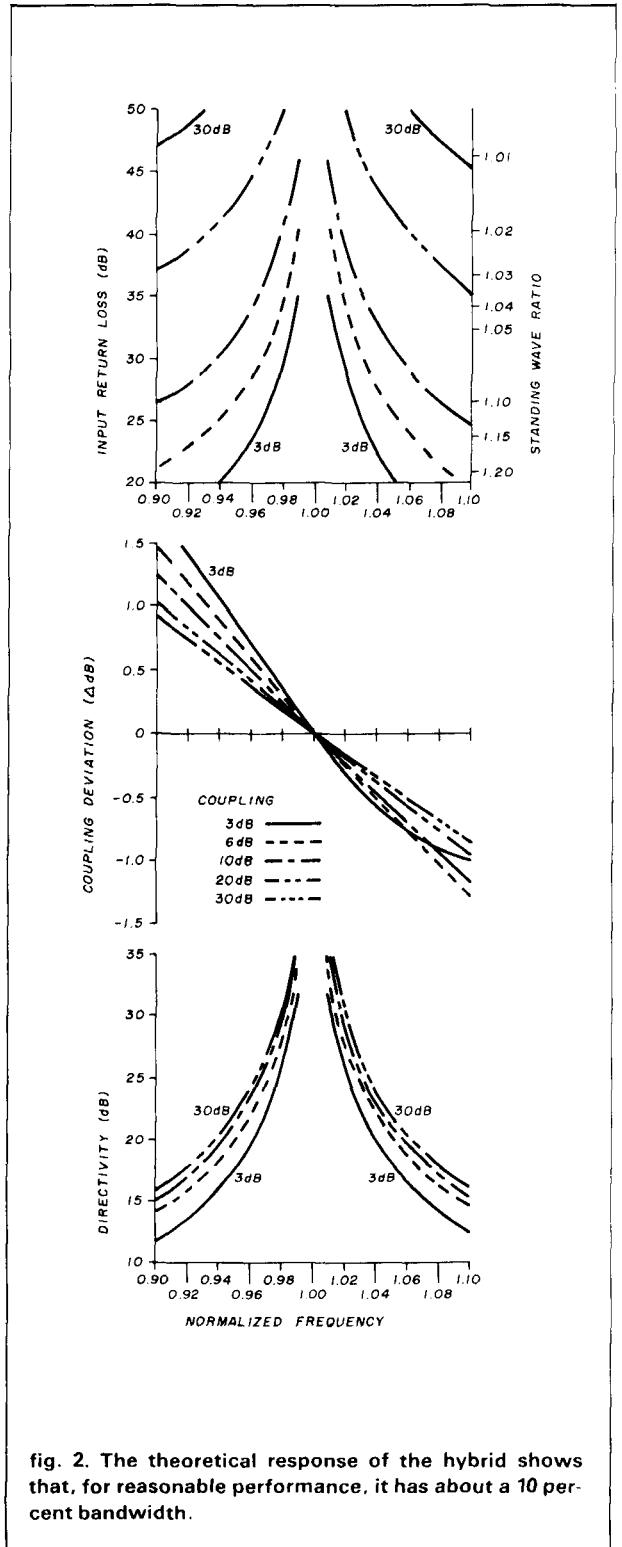


fig. 2. The theoretical response of the hybrid shows that, for reasonable performance, it has about a 10 per cent bandwidth.



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The theoretical response of the capacitively coupled hybrid is shown in **fig. 2**. The input return loss and directivity are better than 20 dB (VSWR  $\leq$  1.2:1) over a 10 percent bandwidth. The deviation from its nominal coupling is plotted for coupling from port 1

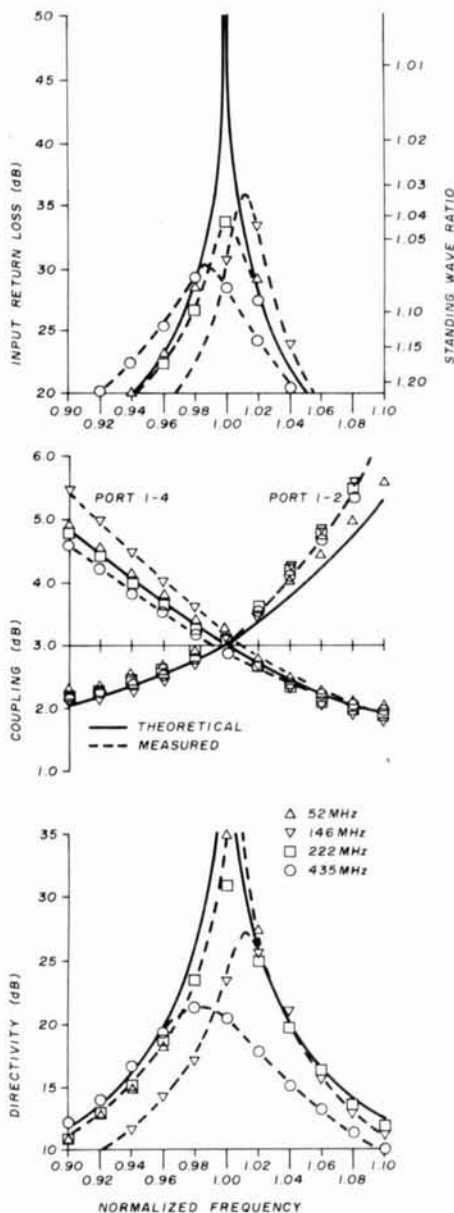
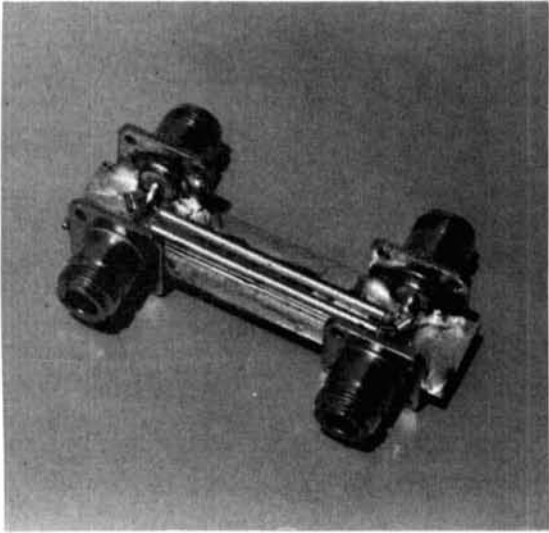
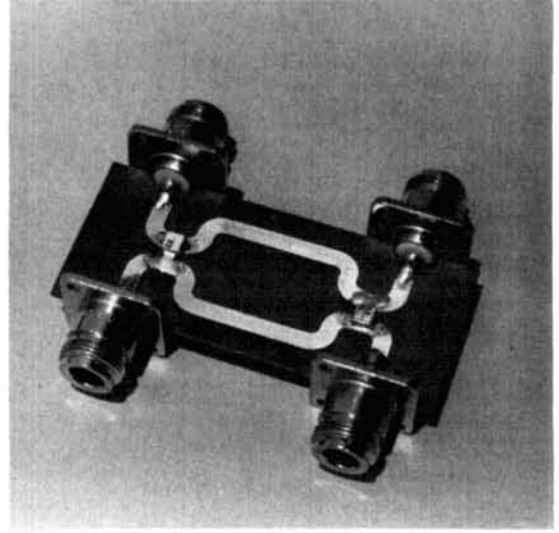


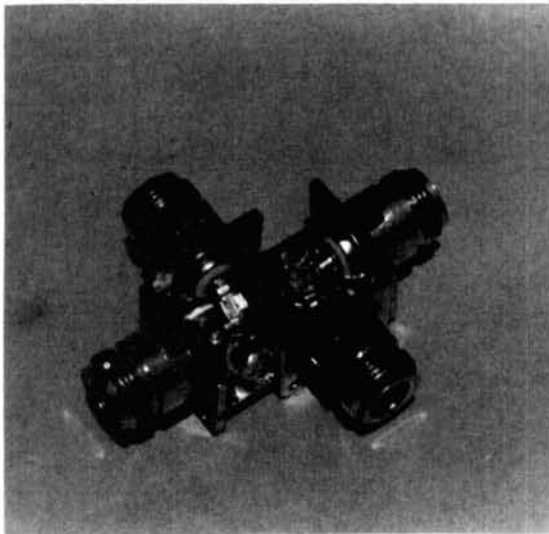
fig. 3. Models constructed using semi-rigid coax perform close to theoretical at VHF.



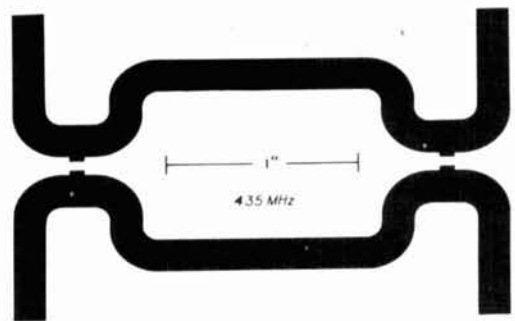
Micro strip hybrid.



Transmission line hybrid.



Lumped-constant hybrid.



Artwork for 435-MHz 3 dB coupler.

**table 1. Design equations for transmission line and pi lumped-constant hybrids.**

$$\theta = \cos^{-1} 10^{-K/20}$$

$$C_c = \frac{1}{2\pi f Z_o \tan \theta}$$

transmission-line hybrid

$$L = \left( \frac{Z_o}{2\pi f} \right) \sqrt{1 - 10^{-K/10}}$$

$$C_s = \left( \frac{1}{2\pi f Z_o} \right) \sqrt{\frac{10^{K/20} - 1}{10^{K/20} + 1}}$$

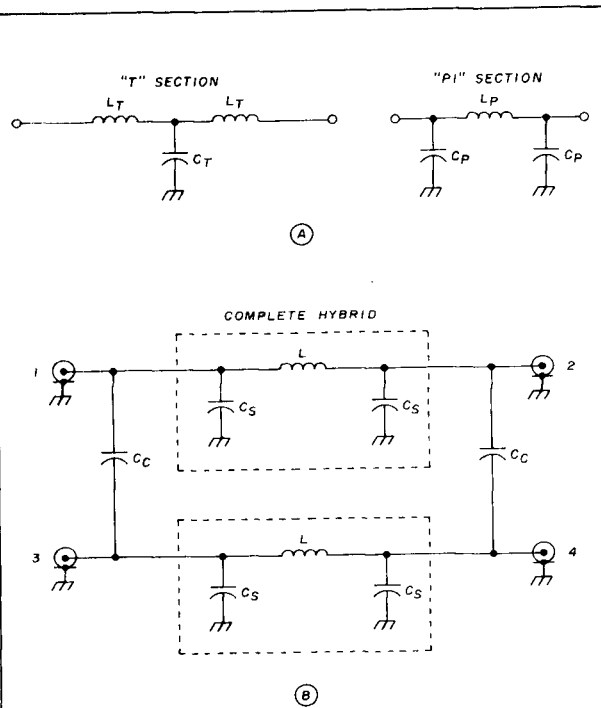
$$C_c = \left( \frac{1}{2\pi f Z_o} \right) \sqrt{\frac{1}{10^{K/10} - 1}}$$

pi lumped-constant hybrid

where  $K = 10 \log_{10} (P_1/P_4)$  absolute coupling

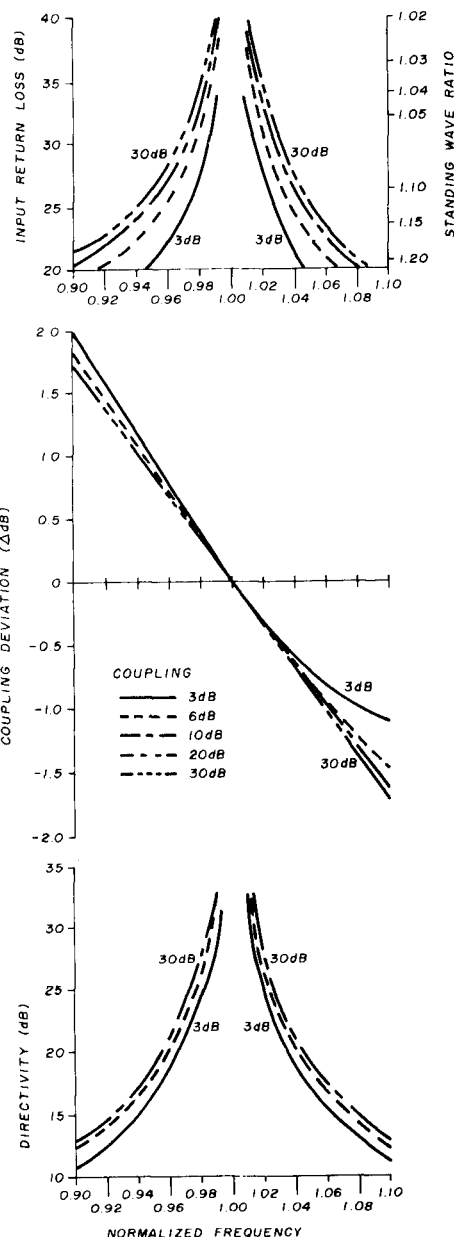
$Z_o$  = termination impedances

$\theta$  = electrical transmission line length



**fig. 4. (A) The hybrid may be formed by using the T or pi section artificial lines. (B) A complete hybrid.**

to port 4. The results from several experimental 3 dB models formed using eighth-wave semi-rigid transmission lines are shown in **fig. 3**. The theoretical response is shown as a solid line. Photographs are shown of transmission line, lumped-constant, and



**fig. 5. The theoretical response of the lumped-constant version of the hybrid is not as good as that of the transmission-line model.**

microstrip hybrids. The artwork for a 435-MHz, 3-dB coupler is also shown for use on 1/16-inch, one-ounce Teflon-glass printed-wire board.

### lumped constant coupler

The capacitively coupled hybrid may be formed with lumped constants in place of the coaxial cables. An artificial transmission line may be formed using lumped constants configured as a T section or a pi section, as shown in fig. 4A. The pi section is usually chosen to simulate the transmission line because it involves only one inductor and acts as a lowpass filter. The lowpass filter decreases harmonic energy, and inductors are more difficult to measure exactly

than capacitors.

The values for the  $T$  artificial line are given as:

$$L_T = Z(1 - \cos \theta) / 2\pi f \sin \theta \quad (7)$$

$$C_T = \sin \theta / Z 2\pi f \quad (8)$$

where  $Z$  refers to the input and output impedances and  $\theta$  is the phase delay through the network.

For a pi network the needed inductance and capacitance are calculated from:

$$L_p = Z \sin \theta / 2\pi f \quad (9)$$

$$C_p = (1 - \cos \theta) / Z 2\pi f \sin \theta \quad (10)$$

table 2. Component values for transmission-line and pi lumped-constant hybrids.

coupling (dB)	line length $\theta$ (deg.)	fractional wavelength $(\lambda/\lambda_0)$	coupling capacitance ( $\pm f$ in MHz) (pF)	series inductance ( $\pm f$ in MHz) (nH)	shunt capacitance ( $\pm f$ in MHz) (pF)
3	45.0	0.125	3183	5,627	1,318
4	50.9	0.141	2589	6,174	1,514
5	55.8	0.155	2165	6,580	1,685
6	60.0	0.167	1838	6,886	1,835
10	71.6	0.199	1061	7,549	2,294
15	79.8	0.222	575.2	7,831	2,659
20	84.3	0.234	320	7,918	2,879
25	86.8	0.241	179.3	7,945	3,007
30	88.2	0.245	100.7	7,954	3,084

table 3. Component values for Amateur band transmission-line and pi lumped-constant hybrids.

frequency (MHz)	coupling (dB)	coupling capacitance (pF)	series inductance (nH)	shunt capacitance (pF)
3.750	3	849	1,500	352
	20	85.3	2,112	768
7.150	3	445	787	184
	20	44.7	1,107	403
14.175	3	225	397	930
	20	22.6	559	203
21.225	3	150	265	62.1
	20	15.1	373	136
28.850	3	110	195	45.7
	20	11.1	274	99.8
52	3	61.2	108	25.4
	20	6.15	152	55.4
146	3	21.80	38.50	9.03
	20	2.19	54.20	19.7
222	3	14.3	25.30	5.94
	20	1.44	35.70	13
435	3	7.37	12.90	3.03
	20	0.74	18.20	6.62
1296	3	2.46	4.34	1.02
	20	0.25	6.11	2.22

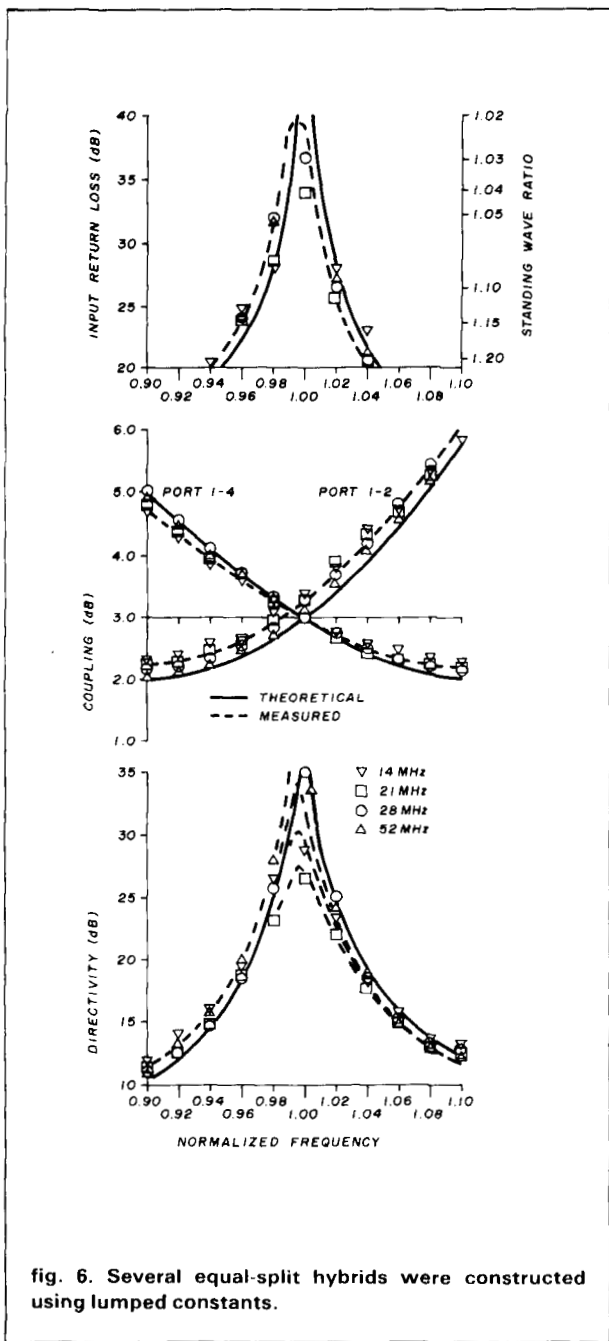


fig. 6. Several equal-split hybrids were constructed using lumped constants.

By substituting the coupling equation,

$$\cos \theta = 10^{-K/20} \quad (11)$$

where  $K$  is the coupling between port 1 and port 4 in decibels, into the above equations for the pi network and using a few trigonometry identities, we arrive at the design equations for the lumped-constant, capacitively coupled hybrid given in table 1.

A capacitively coupled hybrid is shown in fig. 4B. The predicted response of the lumped constant ver-

sion is shown in fig. 5. The graph of the input return loss shows that the lumped constant version has slightly less bandwidth than the lightly coupled transmission line version ( $\geq 10$  dB). At frequencies below 100 MHz the length of transmission line is too great for compact implementation. Values for the construction of transmission-line or lumped-constant, capacitively coupled hybrids are catalogued in table 2. Values for each of the Amateur bands are given in table 3.

Lumped-constant models were constructed for several Amateur bands. The results for 3-dB and 20-dB couplers are shown in figs. 6 and 7. The 3-dB hybrid is commonly used for power splitting/combining, while the 20-dB hybrid is normally used as a

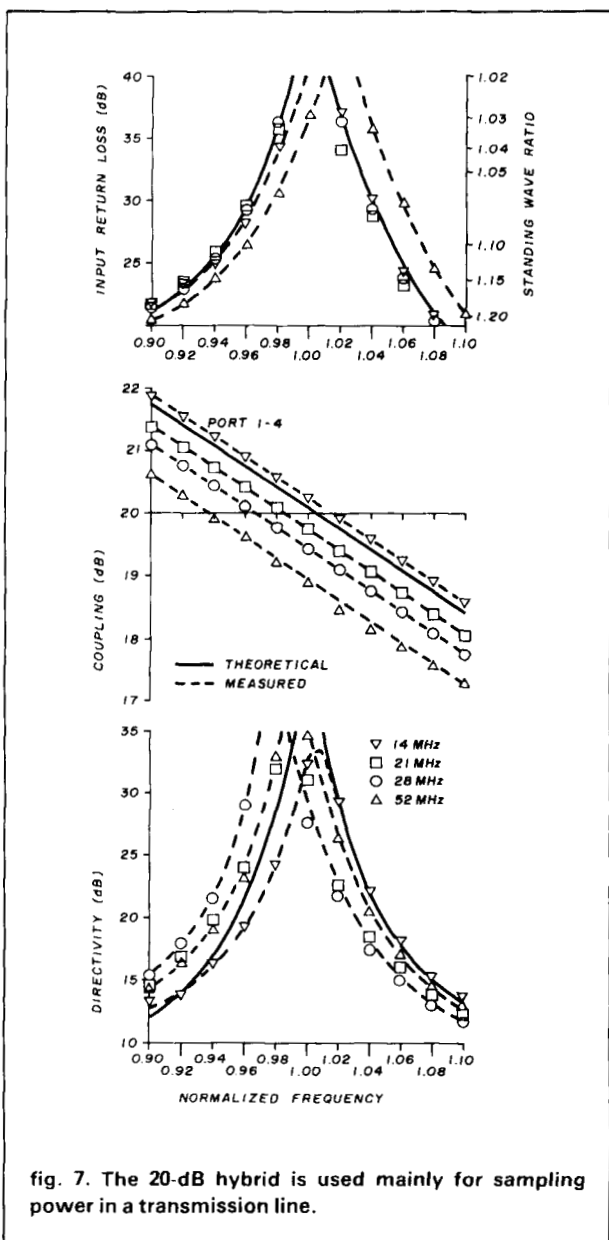


fig. 7. The 20-dB hybrid is used mainly for sampling power in a transmission line.

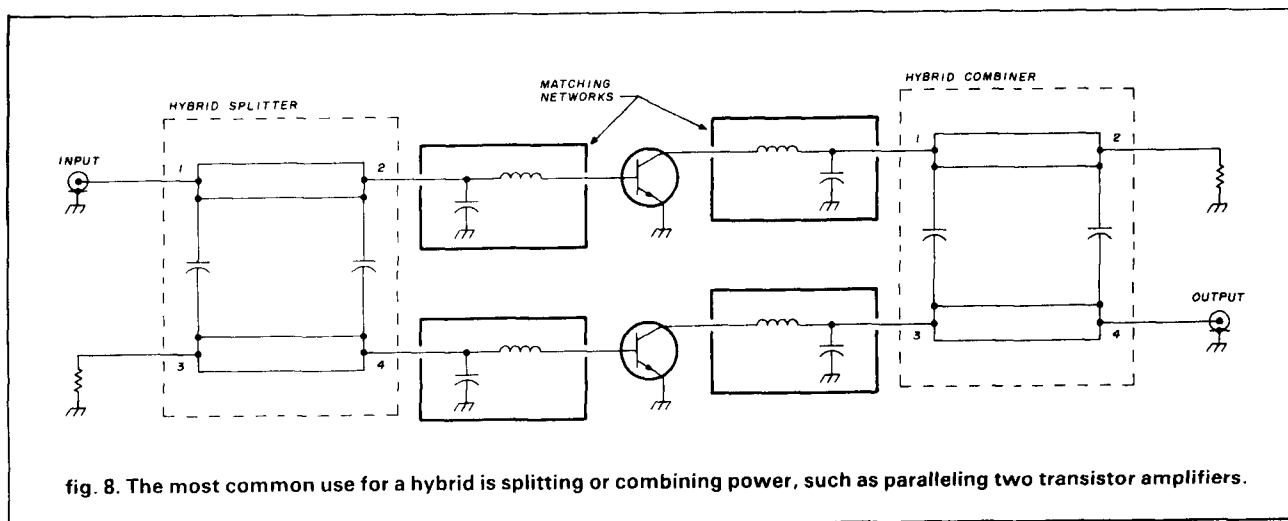


fig. 8. The most common use for a hybrid is splitting or combining power, such as paralleling two transistor amplifiers.

directional coupler. The results agree quite well with the theoretical values shown as a solid line. The responses could be adjusted for perfect agreement if one used greater precision in component selection. Capacitors with marked values close to the calculated values were used in these hybrids. With matched terminations on each port, the inductor closest to the input port is adjusted for best return loss. The input was then applied to port 4 with all the other ports terminated and the match adjusted using the other inductor. Spreading or squeezing of the inductor is determined using a "diddle" stick. This stick is constructed by gluing a piece of brass to one end of a plastic stick and a piece of ferrite to the other end. If the performance improves with the proximity of brass, the coil needs to be spread for less inductance. The coupling between transmission lines is mainly a factor of the coupling capacitance. The capacitance is easily measured using a digital capacitance meter.

The most popular use for the capacitively coupled hybrid is splitting the input and combining the output power from two transistors in a power amplifier, see fig. 8. The transistors remain isolated from each other. Isolation is a measure of how much of the power incident at port 1 appears at the isolation port 3. It is also the same as the reflected power at port 4 appearing at port 2, because the device is reciprocal. Isolation is equal to the coupling (port 1 to port 4) plus the directivity. For this hybrid the isolation between ports 1 and 3 or between ports 2 and 4 is the same as the input return loss. Thus one can expect the mismatched power at the input to one transistor to be reduced by 20 dB (X100) before appearing at the input to the other transistor. This isolation improves stability by decreasing interaction. Any mismatch difference at the input to the two transistors appears at the isolation port. If the mismatch is iden-

tical the hybrid still provides a good input match, and no power appears at the isolator port. Power at the isolation port of the output hybrid indicates gain differences of the transistors.

### intermodulation performance

With respect to a signal applied to input port 1, the output signals at ports 2 and 4 on a 3-dB hybrid are 90 degrees out of phase. One advantage that the 90-degree hybrid has over the Wilkinson combiner is the improvement in amplifier output intermodulation performance. A signal ( $F_1$ ) from a nearby interfering transmitter coupled through the transmitting antennas mixes with the second harmonic of the amplifier ( $2F_0$ ) at the collector in a power amplifier to form an intermodulation product ( $2F_0 \pm F_1$ ). This mixing product is typically close to the desired carrier frequency and thus very difficult to filter. Examination of the 90-degree hybrid shows that the intermodulation products ( $F_{IM}$ ) from the amplifier shown

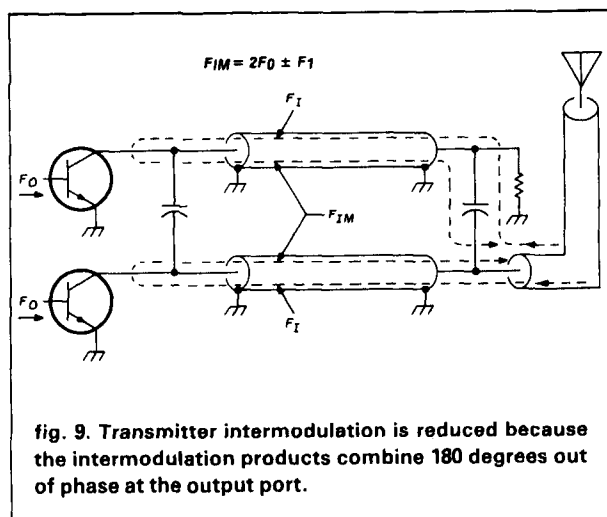
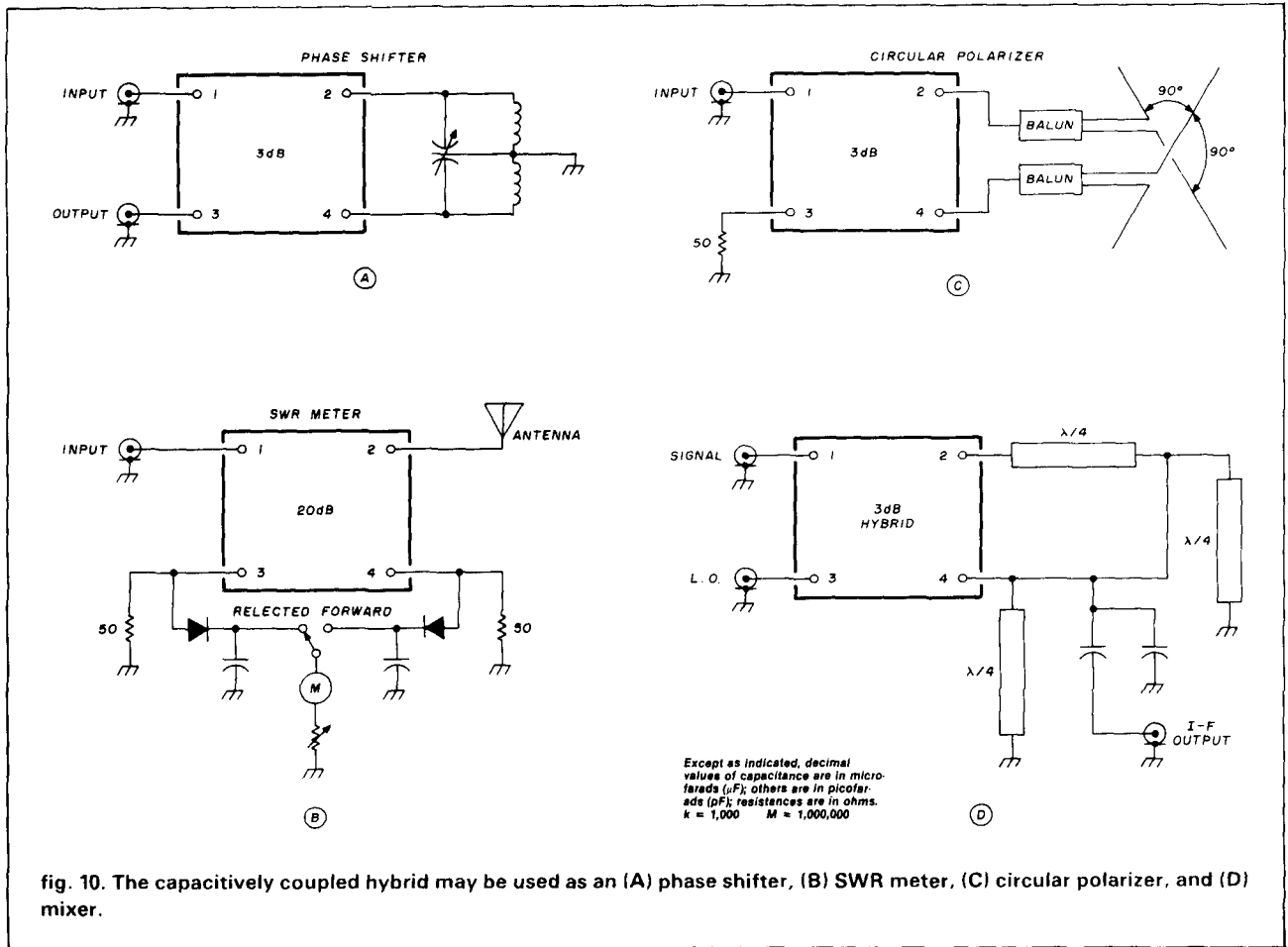


fig. 9. Transmitter intermodulation is reduced because the intermodulation products combine 180 degrees out of phase at the output port.

in fig. 8 combine at the output port 180 degrees out of phase to cancel out, fig. 9. The round trip path of one interference-intermodulation signal is 180 degrees longer than the round trip path to the other transistor.

if peak detector diodes are used. A 3-dB quadrature hybrid may be used to induce circular polarization<sup>3</sup> in a pair of crossed dipoles, fig. 10C. By using a 3-dB hybrid to establish a 90-degree phase shift, a receiver mixer<sup>4</sup> can be formed, fig. 10D. The input signal and



As long as each collector provides identical terminations to the combiner hybrid, the level of the intermodulation signal at the output port should be reduced by the value of the input return loss, graphed for each hybrid.

### other uses

The 3-dB coupler may be used as a phase shifter for varying the phase in one leg of a phased array.<sup>2</sup> See fig. 10A. These antenna systems typically cover a single Amateur band. By varying the purely reactive load at ports 2 and 4 the phase may be adjusted over 180 degrees. The standing wave ratio coupler shown in fig. 10B uses a 20-dB hybrid. The measured return loss is given as

$$\text{return loss (dB)} = -20 \log_{10} \frac{V_3}{V_4} \quad (12)$$

local oscillator are isolated typically by 30 dB over a 4 percent bandwidth. The additional quarter-wave line at port 2 is used to establish an additional 90-degree phase shift with respect to port 4. Thus the mixer diodes are driven 180 degrees out of phase with respect to each other and the local oscillator signal is balanced at the i-f port.

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1. Ernie Franke, WA2EWT, "Wilkinson Hybrids," *ham radio*, January, 1982, pages 12-18.
2. Henry S. Keen, W5TRS, "Electrically Controlled Phased Array," *ham radio*, May, 1975, pages 52-55.
3. Reed E. Fisher, W2CQH, and Richard H. Turrin, W2IMU, "UHF Directional Couplers," *QST*, September, 1970, pages 26-31.
4. Kurt Bittmann, WB2YVY, "Easy-to-Make 1296-MHz Mixer," *73*, July, 1972, pages 33-35.

ham radio





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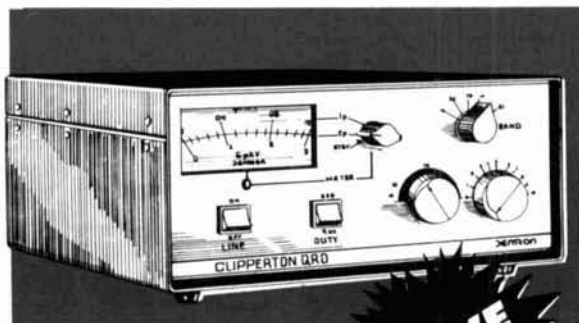


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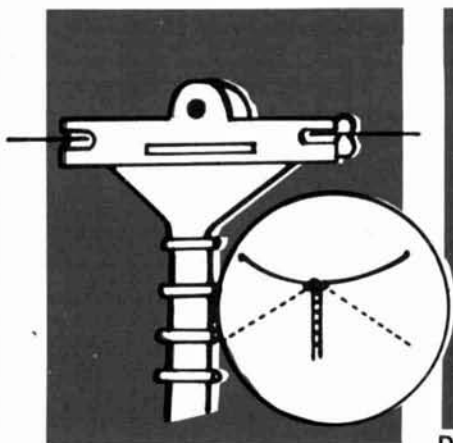
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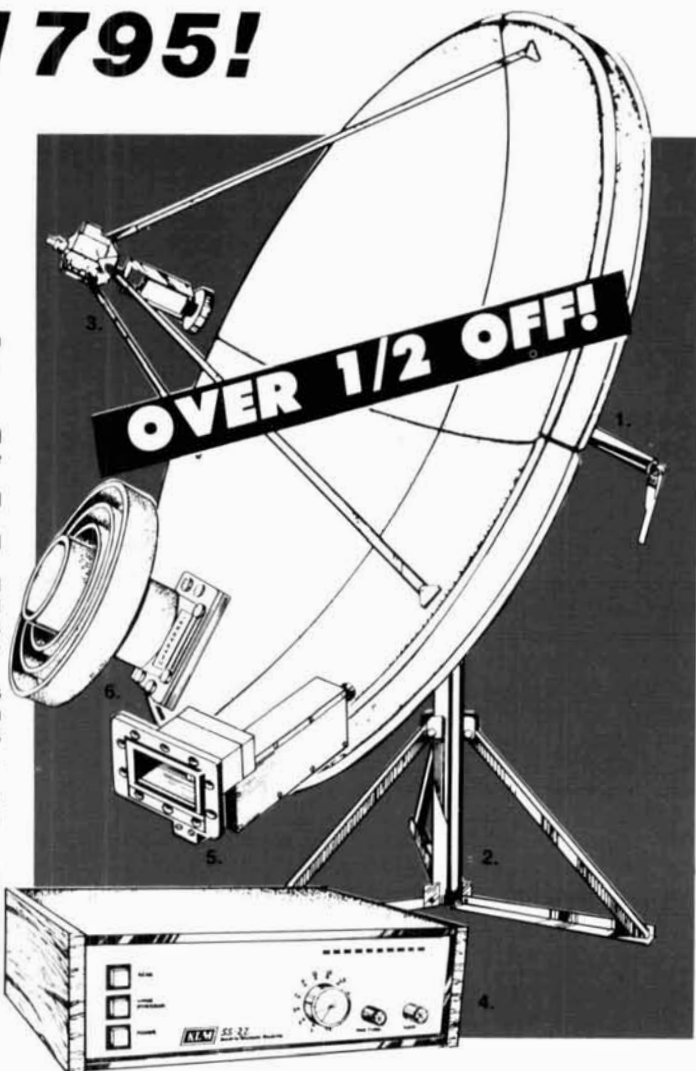
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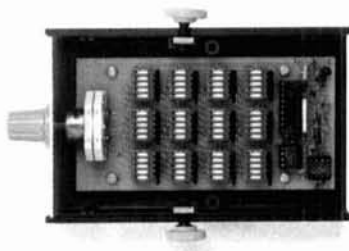
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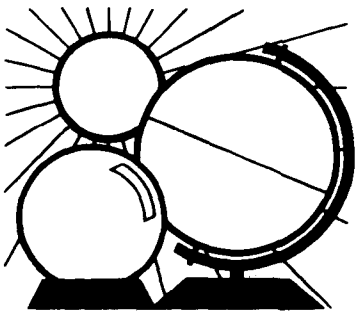
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## last minute forecast

DX conditions on 10, 15, and 20 meters are expected to peak during the second and third weeks of this month because of high MUFs resulting from increased solar activity. Disturbed conditions, because of solar flare emissions, are expected on the 10th and 18th. Other periods of geomagnetic and ionospheric (recurrent) disturbances are likely the 1st and 27th because of enhanced solar wind from coronal holes. Nighttime DX on the lower bands should continue to be good during the entire month, with best performance during the first and last weeks.

With the advent of the vernal equinox this month (21st), gray-line DX-ing should occur during both local sunrise and sunset. North/south paths over the polar regions should

be useful, during lulls in geomagnetic storm activity. Geomagnetic disturbances, which are more evident as the equinox approaches, cause considerable signal attenuation and fading on polar paths. (Gray-line operation is explained in *ham radio*, September, 1982, page 56, and *CQ Magazine*, September, 1975, page 27).

## band-by-band summary

*Ten, fifteen, and twenty meters* will be open from morning to early evening almost every day, and to most areas of the world. The openings will be shorter on the higher bands and occur more frequently at local noon. Trans-equatorial propagation will be more likely on these bands during conditions of high solar flux and a dis-

turbed geomagnetic field.

*Thirty meters* will be useful almost twenty-four hours a day. Daytime conditions will resemble those on 20 meters, except that signal strengths may decrease during midday on some days, those days coinciding with high solar flux values. Nighttime use will be good except following days of very high MUF conditions. Generally, the usable distance is expected to be greater than achieved on 80 at night but less than that on 20 meters during the day.

*Forty, eighty, and one-sixty meters* are the night DXer's bands. The bands are open just before sunset and last until sunrise, local time. Except for daytime short-skip signal strengths, high solar flux values won't greatly affect these bands.

### WESTERN USA

GMT	PST	N	NE	E	SE	S	SW	W	NW
0000	4:00	10	20	15	10	15	10	10	10
0100	5:00	10	20	15	10	15	10	10	10
0200	6:00	15	40	20	10	15	10	10	10
0300	7:00	15	40	20	15	15	10	10	15
0400	8:00	15	40	20	15	15	10	10	15
0500	8:00	20	40	20	20	15	10	15	20
0600	10:00	—	40	—	20	15	15	15	20
0700	11:00	—	—	—	20	20	15	20	40
0800	12:00	—	—	—	20	20	20	20	40
0900	1:00	—	—	20	20	20	20	20	40
1000	2:00	—	—	—	—	40	20	20	40
1100	3:00	—	—	—	—	40	20	20	—
1200	4:00	—	—	—	—	40	40	20	—
1300	5:00	—	—	—	—	—	40	20	—
1400	6:00	—	—	20	20	15	—	—	—
1500	7:00	—	15	15	10	—	—	—	—
1600	8:00	20	15	10	10	—	—	15	—
1700	9:00	20	15	10	10	—	15	15	—
1800	10:00	20	15	10	10	—	10	15	—
1900	11:00	15	15	10	10	—	10	15	—
2000	12:00	15	20	10	10	—	10	15	20
2100	1:00	15	20	15	10	15	10	10	15
2200	2:00	10	20	15	10	15	10	10	10
2300	3:00	10	20	15	10	15	10	10	10

**MARCH**

### MID USA

GMT	MST	N	NE	E	SE	S	SW	W	NW
0000	5:00	15	20	15	10	15*	10	10	10
0100	6:00	15	20	15	10	15	10	10	10
0200	7:00	15	40	20*	15	15	10	10	15*
0300	8:00	20	40	20	15	15	10	15	15
0400	9:00	20	40	20	20*	15	15	15	15
0500	10:00	—	40	20	20	15	15	15	20
0600	11:00	—	40	20	20	20	20	20	20
0700	12:00	—	20	20	20	20	20	20	40
0800	1:00	—	—	—	20	20	20	20	40
0900	2:00	—	—	—	—	40	40	—	—
1000	3:00	—	—	—	—	40	40	—	—
1100	4:00	—	—	—	20	—	—	—	—
1200	5:00	—	—	15	15	—	—	—	—
1300	6:00	20	—	15	15	—	—	—	—
1400	7:00	20	20	10	10	—	—	—	—
1500	8:00	20	15	10	10	—	—	—	—
1600	9:00	20	15	10	10	—	—	15	—
1700	10:00	20	15	10	10	—	20	15	—
1800	11:00	20	15	10	10	—	15	15	—
1900	12:00	20	15	10	10	—	15	15	—
2000	1:00	—	15	10	10	—	10	10	20
2100	2:00	—	20	10	10	15	10	10	15
2200	3:00	—	20	10	10	15	10	10	10
2300	4:00	—	20	10	10	15*	10	10	10

### EASTERN USA

GMT	EST	N	NE	E	SE	S	SW	W	NW
0000	7:00	15	20	15	10	15	10	10	15
0100	8:00	15	20	15	10	15	10	10	15
0200	9:00	15	40	20	15	15	10	15	15
0300	10:00	15	40	20	15	15	15	15	20
0400	11:00	20	40	15	15	20	15	20	20
0500	12:00	—	40	15	20	20	20	20	20
0600	1:00	—	20	20	20	20	20	20	40
0700	2:00	—	20	—	20	20	40	—	40
0800	3:00	—	—	—	20	20	40	—	—
0900	4:00	—	—	—	20	—	—	—	—
1000	5:00	—	—	—	—	—	—	—	—
1100	6:00	—	—	15	15	—	—	—	—
1200	7:00	—	—	10	15*	—	—	20	—
1300	8:00	20	15	10	10	—	—	15	—
1400	9:00	20	10	10	10	—	—	15	—
1500	10:00	20	10	10	10	—	—	15	—
1600	11:00	15	10	10	10	—	—	15	—
1700	12:00	15	10	10	10	—	20	15	—
1800	1:00	15	15	10	10	—	20	15	—
1900	2:00	—	15	10	10	15*	10	10	—
2000	3:00	—	15	10	10	15*	10	10	20
2100	4:00	—	15	10	10	10	10	10	15
2200	5:00	—	20	10	10	10	10	10	10
2300	6:00	15	20	10	10	10	10	10	10

Look at next higher band for possible openings.

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|---|---|
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| 7.5 wpm code for 25 minutes               | 25 wpm code for 20 minutes                |
| 10 wpm code for 25 minutes                | 30 wpm code for 20 minutes                |
| 15 wpm code for 25 minutes                | 35 wpm code for 20 minutes                |
| <input type="checkbox"/> HR-STC2 — \$4.95 | 40 wpm code for 20 minutes                |
| 15 wpm code for 50 minutes                |   |
| 22.5 wpm code for 35 minutes              |   |

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consists of characters sent at a 15 wpm rate, but with 2.5 wpm spac-  
ing between each character. These tapes are excellent for the beginner  
who wants to practice copying higher speed code without the frustra-  
tion of constantly getting behind.

- |   |   |
|---|---|
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| 22.5/2.5 wpm code for 80 minutes          | 15/5 wpm code for 28 minutes              |
| <input type="checkbox"/> HR-HLC2 — \$4.95 | 15/7.5 wpm code for 28 minutes            |
| 22.5/5 wpm code for 20 minutes            | 15/10 wpm code for 28 minutes             |
| 22.5/7.5 wpm code for 20 minutes          | <input type="checkbox"/> HR-HLC4 — \$4.95 |
| 22.5/10 wpm code for 20 minutes           | 15/2.5 wpm code for 80 minutes            |
| 22.5/13 wpm code for 20 minutes           |   |

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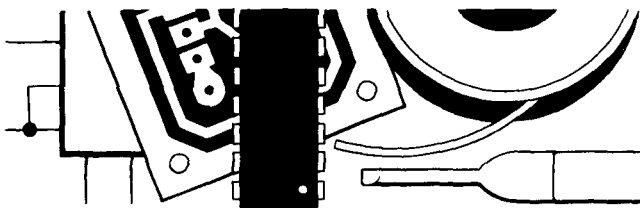
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## the **weekender**



### **CW zero-beat indicator for transceivers**

For many CW operators who use transceivers, zero beating the station they are working is very difficult. I, too, had this problem, and after several years of searching for an answer I devised the circuit illustrated in **fig. 1**. It puts me within 50 Hz of the frequency of the station I am working.

Terminal A (see **fig. 1**) is connected to the audio output line of my transceiver. When I am zero beat with the CW station I'm working, the LED lights when the other operator keys his transmitter. If the LED does not light, my rig is not zero beat.

The LM567 (U1) contains an oscillator, with output frequency controlled by the value of R1 and C1. The value of R1 is adjusted so the LM567 oscillator frequency is equal to the offset frequency of your transceiver. The LM567 output is brought out at pin 8 and is an open collector. As you examine **fig. 1**, you will see that the collector load for the output transistor is the LED and R3, which are connected to +5 volts. When the input frequency to terminal A is equal to the LM567 oscillator frequency, the transistor is turned on and the LED lights up to provide an indica-

tion that the audio frequency entering terminal A is equal to the oscillator frequency.

The amplitude of the audio signal fed to terminal A will effect the bandwidth (the band of frequencies that will cause the LED to light). If the LM567 oscillator is one kHz, and the input signal at pin 3 is 300 mV<sub>p-p</sub>, the bandwidth of this circuit will be approximately 350 Hz. Under this condition, you will get a zero beat indication when the audio input signal is at any frequency between 825 and 1175 Hz. So, you may be off-frequency by 175 Hz and still get a zero beat indication. We can do much better than this, however. A 25 mV<sub>p-p</sub> signal on pin 3 will provide a 100 Hz bandwidth and the most that you will be off is 50 Hz.

The lower the amplitude of the audio signal arriving at pin 3 the more accurate you will be in zero beating the station you are working. However, you will still want to hear the audio signal, so R2, CR1 and CR2 function to limit the maximum amplitude of the audio signal at pin 3 of the LM567 to a usable level while maintaining adequate volume from the speaker or headphones. My HW-101 has an 8-ohm audio output, and I find the 100-ohm resistor value for R2 adequate. However, 100 ohms may be too high for use with a 4-ohm audio output stage. Therefore, it may be desirable to install a 500-ohm potentiometer, and set the resistance of R2 to suit your own taste.

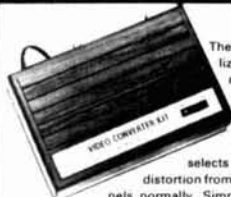
The capacitance of C1 must be temperature stable. I had the sad experience of using a capacitor for C1 with capacitance that changed with temperature, and the oscillator frequency of the LM567 changed

---

By Glen Carlson, W6KVD, 2588 Hermosa Street, Pinole, California 94564

## SWD-1 VIDEO CONVERTER

FOR CABLE TV



The SWD-1 Video Converter is utilized on cable TV systems to remove the KHz's signal from a distorted video (channel 3 in/out) and also pass thru the normal undistorted/detected audio signal. Rocker switch selects operating mode to remove KHz's distortion from the video or pass all other channels normally. Simple to assemble—less than 30 minutes. Pre-tuned. Input/output Channel 3. Impedance 75 ohms. 117VAC.

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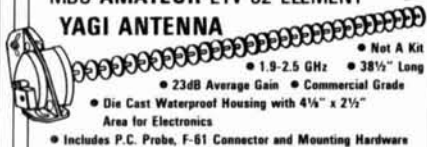
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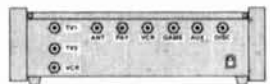
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3	3TP11-PWD	PCB Potentiometers 4-20K, 1-5K, 2-10K, 2-5K, 1-1K, and 1-50K. (11 pieces)	8.95
4	4FR-31-PWD	Resistor Kit, 1/4W, 5% 28-pcs, 1/2 W 2-pcs	4.95
5	5PT1-PWD	Power Transformer, PRI-117VAC, SEC-24VAC at 500ma	9.95
6	6PP2-PWD	Panel Mount Potentiometers and Knobs, 1-1KBT and 1-5KAT with switch	5.95
7	7SS17-PWD	IC's 7-pcs, Diodes 4-pcs, Regulators 2-pcs Transistors 2-pcs, Heat Sinks 2-pcs	29.95
8	8CE14-PWD	Electrolytic Capacitor Kit, 14-pieces	6.95
9	9CC20-PWD	Ceramic Disk Capacitor Kit, 50 WV, 20-pcs	7.95
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with the temperature of my rig. The capacitor type listed for C1 in the parts list works well for me.

Owners of solid-state rigs having a +5 Vdc supply will need only the circuit illustrated in fig. 1. Owners of tube type rigs may need to build a 5 volt supply, illustrated in fig. 2.

## construction

I built the circuits illustrated in figs. 1 and 2 on a Radio Shack™ 276-024 printed circuit board. These boards are inexpensive and are made to accept a 16-pin IC. They provide pads at the edge of the board connected to the 16 IC pins. I recommend installing an 8-pin IC socket for IC1. Tin the traces on the board before drilling small holes in the pads.

I drilled a hole in the front panel of my HW-101 and installed a grommet to hold the LED in place. I used single-conductor shielded audio cable for running power to the LED. The shield must be isolated from the chassis as it will be used as one of the conductors to the LED. \* Find a place for installing the PC-board, and then construct a small U-bracket for mounting the board while maintaining isolation from the chassis.

Tie terminal A to a point in your transceiver where this circuit will operate when using either the speaker or headphones. If you tie terminal A to a point where there is a dc voltage, CR1 or CR2 will be biased into conduction. In this case, use a suitable blocking capacitor between terminal A and the point you pick up the audio signal. A capacitor value between 0.01  $\mu$ F and 0.1  $\mu$ F should suffice. Adjust the value of R3 for suitable brilliance of the LED.

## alignment

Alignment consists of setting the LM567 oscillator frequency equal to the offset frequency of your transceiver. There may be several ways to accom-

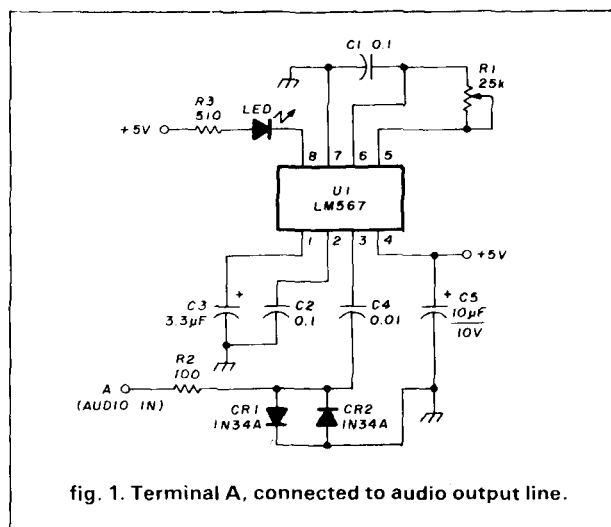


fig. 1. Terminal A, connected to audio output line.

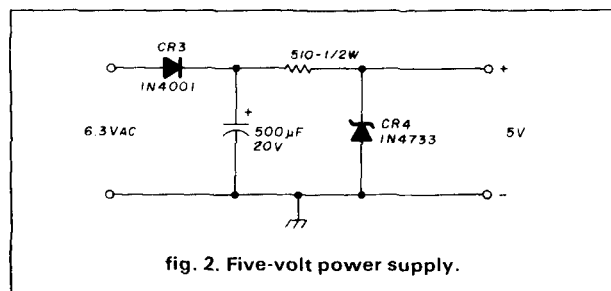


fig. 2. Five-volt power supply.

C1	0.1 $\mu$ F 200 V (see text) RS 272-1053
C2	0.1 $\mu$ F 10 V
C3	3.3 $\mu$ F 10 V tantalum
C4	0.01 $\mu$ F 10 V
C5	10 $\mu$ F 10 V (tantalum)
C6	500 $\mu$ F 20 V
CR1 & CR2	small signal germanium diodes
CR3	rectifier diode 1N4001 RS 276-1101
CR4	5 V 1/2 watt zener diode 1N4733 RS 276-565
R1	25 K 1/4 watt potentiometer (RS 271-336)
R2	100 ohm 1/4 watt (see text)
R3	510 ohm 1/4 watt
R4	510 ohm 1/2 watt
LED	RS 276-032
	printed circuit board (RS 276-024)
LM567	phase-locked loop (RS 176-1721) (XR-567 is identical and usable)

plish this adjustment. You will set the oscillator frequency by adjusting R1.

If you have a ham friend who can zero beat your transmit frequency, adjust R1 while listening to his signal. As you adjust R1, keep reducing the volume to narrow the bandpass for the best adjustment of the oscillator frequency. Then practice zero beating your friend's frequency, and have him check the accuracy of your adjustment.

A more accurate method uses a frequency counter. Measure the frequencies of the oscillators which determine the frequency difference between your transmit and receive frequencies. Subtract the smaller frequency from the larger frequency, and you will have your offset frequency. Using a low capacity probe, connect the frequency counter to pin 5 of the LM567, and adjust R1 until the counter reads your offset frequency.

When using the frequency counter method, you may be surprised to find that your offset frequency is not exactly as listed by the transceiver manufacturer. In my case, the offset frequency was off by 70 Hz.

If enough Amateurs build this circuit, we may find room in the CW bands for more QSOs.

\*A twisted pair made from No. 24 insulated hookup wire would work just as well and is easier to handle. Editor

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115	#1-840 CT @ 230 Ma #2-620 CT @ 25 Ma #3-5 @ 2 A #4-6.3 @ 5.5 A #5-6.3 @ 5 A #6-6.3 @ 5 A	5x6x7	10	\$19.95
115	#1-45 @ 50 Ma #2-68 @ 3 A #3-34 CT @ 25 A #4-20 CT @ 2 A	7x6x6	43	\$24.50
115	#1-105 @ 75 Ma #2-475 @ 200 Ma #3-6.5 @ 9 A #4-6.3 @ 3.3 A #5-6.3 @ 2A	4x5x3	7	\$ 9.95

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2	10,000	13¼x4x17½	40	\$ 29.00
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New low-noise microwave transistors make preamps in the 0.9 to 1.0 dB noise figure range possible without the fragility and power supply problems of gas-fet's. Units furnished wired and tuned to ham band. Can be easily retuned to nearby freq.



Models LNA( ), P30, and P432 shown

Model	Tunable		Noise Figure	Gain	Price
	Freq Range	Gain			
LNA 28	20-40	0.9 dB	20 dB	\$39.95	
LNA 50	40-70	0.9 dB	20 dB	\$39.95	
LNA 144	120-180	1.0 dB	18 dB	\$39.95	
LNA 220	180-250	1.0 dB	17 dB	\$39.95	
LNA 432	380-470	1.0 dB	18 dB	\$44.95	

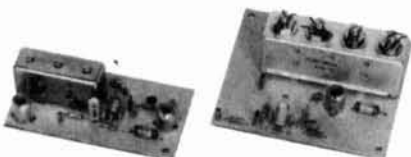
## ECONOMY PREAMPS

Our traditional preamps, proven in years of service. Over 20,000 in use throughout the world. Tuneable over narrow range. Specify exact freq. band needed. Gain 16-20 dB. NF = 2 dB or less. VHF units available 27 to 300 MHz. UHF units available 300 to 650 MHz.

- P30K, VHF Kit less case \$14.95
- P30C, VHF Kit with case \$20.95
- P30W, VHF Wired/Tested \$29.95
- P432K, UHF Kit less case \$18.95
- P432C, UHF Kit with case \$24.95
- P432W, UHF Wired/Tested \$33.95

P432 also available in broadband version to cover 20-650 MHz without tuning. Same price as P432; add "B" to model #.

## HELICAL RESONATOR PREAMPS



Our lab has developed a new line of low-noise receiver preamps with helical resonator filters built in. The combination of a low noise amplifier similar to the LNA series and the sharp selectivity of a 3 or 4 section helical resonator provides increased sensitivity while reducing intermod and cross-band interference in critical applications. See selectivity curves at right. Noise figure = 1 to 1.2 dB. Gain = 12 to 15 dB.

Model	Tuning Range	Price
HRA-144	143-150 MHz	\$49.95
HRA-220	213-233 MHz	\$49.95
HRA-432	420-450 MHz	\$59.95

## RECEIVING CONVERTERS



Models to cover every practical rf & if range to listen to SSB, FM, ATV, etc. NF = 2 dB or less.

	Antenna Input Range	Receiver Output
<b>VHF MODELS</b>	28-32	144-148
	50-52	28-30
Kit \$44.95	50-54	144-148
Less Case \$39.95	144-146	28-30
Wired \$59.95	145-147	28-30
	144-144.4	27-27.4
	146-148	28-30
	144-148	50-54
	220-222	28-30
	220-224	144-148
	222-226	144-148
	220-224	50-54
	222-224	28-30
<b>UHF MODELS</b>	432-434	28-30
	435-437	28-30
Kit \$54.95	432-436	144-148
Less Case \$49.95	432-436	50-54
Wired \$74.95	439.25	61.25

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## TRANSMIT CONVERTERS

For SSB, CW, ATV, FM, etc. Why pay big bucks for a multi mode rig for each band? Can be linked with receive converters for transceive. 2 watts output.

	Exciter Input Range	Antenna Output
For VHF, Model XV2 Kit \$79.95 Wired \$119.95 (Specify band)	28-30	144-146
	28-29	145-146
	28-30	50-52
	27-27.4	144-144.4
	28-30	220-222
	50-54	220-224
	144-146	50-52
	50-54	144-148
	144-146	28-30

For UHF, Model XV4 Kit \$99.95 Wired \$149.95	28-30	432-434
	28-30	435-437
	50-54	432-436
	61.25	439.25
	144-148	432-436*

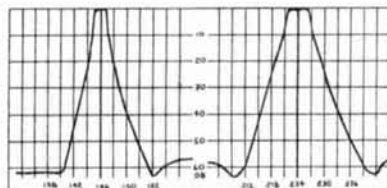
\*Add \$35 for 2M input

**FREE OFFER**

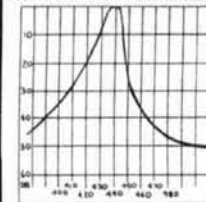
For limited time, buy a transmit converter above with 40-45W PA (\$129.95) and get \$39.95 cabinet FREE.



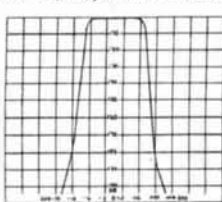
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R144 & R220 Front Ends. HRA 144/220, & HRF-144/220

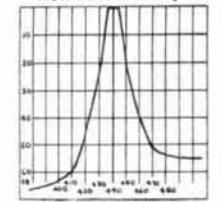


R451 Receiver Front End



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HRA-432, HRF-432

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Band	Kit	Wired/Tested
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440	\$645	\$795

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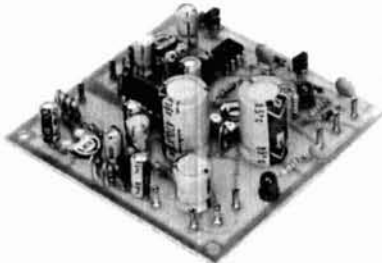


R144 Shown

- **R144/R220 FM RCVRs** for 2M or 220 MHz. 0.15uV sens.; 8 pole xtal filter & ceramic filter in i-f, helical resonator front end for exceptional selectivity (curves at left). AFC incl., xtal oven avail. Kit only \$119.95
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- **R76 FM RCVR** for 10M, 6M, 2M, 220, or commercial bands. As above, but w/o AFC or hel. res. Kits only \$109.95. Also avail w/4 pole filter, only \$94.95/ kit.
- **R110 VHF AM RECEIVER** kit for VHF aircraft band or ham bands. Only \$84.95.
- **R110 UHF AM RECEIVER** for UHF uses, including special 296 MHz model to hear SPACE SHUTTLE. Kit \$94.95.



- **HELICAL RESONATOR FILTERS** available separately on pcb w/connectors.  
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 HRF-432 for 420-450 MHz \$44.95  
 (See selectivity curves at left.)



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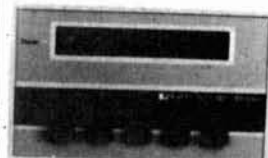
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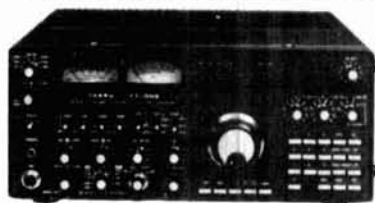
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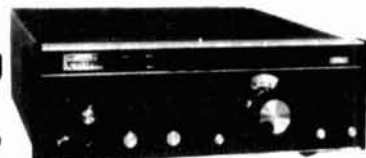
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switch for use with various optional filter combinations. The TS-430S carries a factory suggested retail price of \$899.95.

For additional information, contact your local Kenwood Amateur Radio dealer, or write to Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220. Reader Service Number 302.

### C22 and B23 HT power amplifiers upgrade

Mirage Communications' pocket-size B23 (2 meters) and C22 (220 MHz) power amplifiers have been upgraded for greater versatility. A new power switch permits selection of full amplifier power or a non-energized bypass mode when only HT power is desired. The fm/SSB switch controls choice of rapid or delayed relay action. Power amplification is linear in either mode. The B23 produces 30 watts (minimum)



for 2 watts in, 15 watts for 1 watt, etc. The C22 produces 20 watts (minimum) for 2 watts in, 10 watts for 1 watt, etc. The duty cycle of both amplifiers is continuous. The B23 and C22 are made in the U.S., and carry a five-year general warranty with one year for rf power transistors.

For more information, contact Everett L. Gracey, Director of Marketing, KLM Electronics, Inc., P.O. Box 816, Morgan Hill, California 95037. Reader Service Number 303.

### automatic antenna coupler

A new automatic antenna coupler, the Model H-402CU, has been introduced by Hull Electronics Company, Inc., of San Diego, California. The coupler employs a microcomputer to exactly match the ship's antenna system to any SSB radiotelephone in the frequency range of 2-22 MHz. The H-402CU initially tunes itself each time a new channel is selected and also fine tunes the antenna as necessary whenever the transmitter is on the air.

Rated at 250 watts power, the coupler combines an L network with high-efficiency toroid inductors to provide maximum transfer of the transmitter energy to the antenna. Initial tune-up time is typically 1/2 to 2 seconds. A unique test panel is included to allow the technician to observe operation of the coupler and to test the various digital circuits.

Installation is simple and requires no preliminary adjustment to the coupler. The H-402CU operates with antennas from 8 to 80 feet in length. Two antennas may be used with the system as an optional feature.

Model H-402CU is housed in a rugged, weather-tight enclosure 16 1/2 inches high x 11 1/4 inches wide x 5 1/2 inches deep and weighs only 9 1/2 pounds. Operating voltage of 12 Vdc is supplied via the SSB transceiver.

For more information, contact Hull Electronics Company, Inc., 7563 Convoy Court, San Diego, California 92111. Reader Service Number 304.

### noise bridge with built-in range extender

MFJ Enterprises, Inc.'s MFJ-202B noise bridge allows quick adjustment for maximum performance of any antenna — single, multi-band, dipole, inverted vee, beam, vertical whip, or random systems. You can measure resonant frequency, radiation resistance and reactance. It tells you whether to lengthen or shorten your antenna for minimum SWR over any portion of the band.

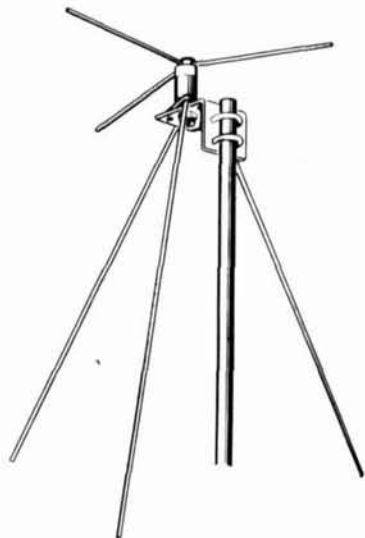
The MFJ-202B will measure resistance to 250 ohms and has a wide capacitance range of ± 150 pF. It includes a built-in range extender that shunts large unknown impedances down to its measuring range. You can tune transmatches, adjust tuned circuits, measure inductance, rf impedance of amplifiers, baluns, transformers and other rf circuits. It can also be used to determine electrical length, velocity factor and impedance of coax cable. With a transmatch and dummy load, it can synthesize rf impedances.

The MFJ-202B measures 4-1/2 x 2 x 4-1/2 inches and is housed in a rugged black aluminum cabinet with eggshell white front. It is available from MFJ Enterprises for \$59.95, plus

### monitor antenna increases scanner performance

The Antenna Specialists Co. has introduced the Mon-64 Discan™, a lightweight monitor antenna that can increase base station reception by as much as 100 percent.

The Mon-64 Discan provides excellent reception on the popular low band, VHF, UHF, and T-band frequencies, from 25-512 MHz. The Mon-64 significantly improves reception of distant stations or low-powered portables. Weighing less than two pounds, the Discan antenna is easy to install, and comes complete with SO-239 connector and double U-clamp bracket; cable is not provided. The antenna mounts easily on any pipe or tubing up to 1 1/4 inches in diameter.



Suggested list price is \$19.95. For complete specifications, contact The Antenna Specialists Co., 12435 Euclid Avenue, Cleveland, Ohio 44106. Reader Service Number 301.

### TS-430S hf transceiver

The TS-430S, a recent addition to Kenwood's line of high-frequency transceivers, is an all-solid-state SSB, CW, and a-m transceiver, with fm optional. Designed to cover the 160-10 meter Amateur bands, including the new WARC bands, it also incorporates a 150

\$4.00 shipping and handling. For more information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762. Reader Service Number 305.

## 5/8-wave UHF antenna for handhelds

RF Products announces the addition of UHF to its existing line of 5/8-wavelength VHF telescopic gain antennas for handheld transceivers. The new models are available with a BNC type connector in 10-MHz frequency segments for the 440-512 MHz band; the most popular are now in production along with the 144-174 and 220-255 MHz versions. Typical gain is 6 dB (ref. 1/4-wave helical) or 3 dB (ref. 1/4 wave). Maximum gain and minimum VSWR is achieved by a tunable LC network.

The antennas include a base spring to prevent whip damage to the telescopic radiator. Minimum bandwidth for 1.5:1 VSWR is 10 MHz with a maximum rf power rating of 5 watts. The maximum extended length with connector is 17-3/16 inches (435 mm) and the collapsed length is 6-5/16 inches (160 mm). The operating frequency range for each model is identified by the color of the base spring cover. The model/frequency ranges available are as follows: 191-914 (440-450 MHz), 191-954



(450-460 MHz), 191-964 (460-470 MHz), 191-974 (470-480 MHz), 191-984 (480-490 MHz), 191-994 (490-500 MHz), and 191-904 (500-512 MHz). Suggested list price for all models is \$19.95 with dealer and OEM discounts available.

For more information, contact RF Products, P.O. Box 33, Rockledge, Florida 32955. Reader Service Number 306.

## handheld counter-timer

The 5000 Counter-Timer combines all the important features and performance capabilities of a benchtop unit with the convenience of a fully portable, battery operated instrument. It is priced at \$349.95, and measures 7.6 x 3.75 x 1.7 inches, and weighs 14 ounces (without batteries).



The 5000 is designed to measure frequency, period and pulse width with extreme accuracy and exceptional reliability. It features full signal conditioning, including attenuator (X1, X10, X100); slope selection (+ or - edge for pulse-width measurement); ac or dc coupling and variable-trigger level.

A high contrast 0.43-inch LCD display offers eight-digit precision for fast and accurate readings. LCD annunciators indicate overflow, gate open, and low battery conditions. A switch allows the display storage mode to maintain the last reading in the display indefinitely.

The 5000 has automatic master reset logic, which instantly clears the display and initiates a new measurement cycle, eliminating erroneous partial measurement. A self-diagnostic function performs analysis of internal logic and provides instant assurance of accurate operation.

The 5000 has three modes of operation: frequency, period and pulse width. Signal input is via BNC connector — input impedance is 1 megohm at 25 pF for all modes. In the frequency mode, the 5000 can handle inputs from 0.1 Hz to 50 MHz. Gate times of 0.01, 0.1, 1.0, or 10 seconds can be selected. Frequency will be displayed in kilohertz on the LCD screen. The 5000 will measure any periods from 25 ns to 10 seconds and deliver a single cycle measurement or an average of 10, 100, or 1000 cycles. Time will be displayed in mS. Pulse width measurement from 25 ns to 10 seconds can be made. Either the high or low portion of the input signal can be selected.

The 5000 is powered by six AA NiCd or alkaline batteries or an optional ac adapter/charger. Optional accessories are available for the 5000.

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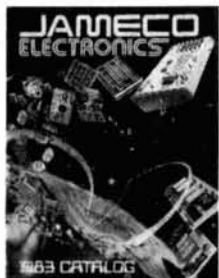
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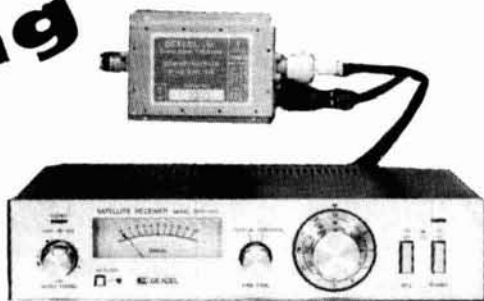
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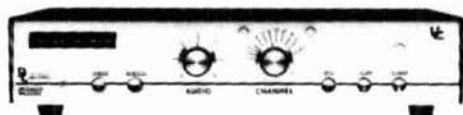
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code and voice channels lets the student practice the code with or without voice help. Turning down the code channel lets students practice code letters and words in unison with the tape with their own code oscillators. Playing both channels simultaneously in a car allows students to practice the code without having to write down any letters.

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## happy 30th

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The Antenna Specialists Company got its start in 1953 by producing land mobile antennas for the Motorola and Zenith companies. Today, the Cleveland-based company is a member of the Allen Group, and it is the world's largest and most diversified manufacturer of communications antennas and accessories for professional land mobile services. It is also an antenna supplier for Amateur Radio, aviation, marine, personal communications, and monitor receivers.

Examples of A/S leadership are the recent introduction of the first completely concealed

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In 1979, the company's Professional Division embarked on its first major departure from the antenna business with its RESCU<sup>®</sup> division, which manufactures emergency electronic devices for the public safety and industrial markets. Two years later, the firm introduced the LIFEGARD<sup>™</sup> personal distress-alerting device for firefighters and industrial personnel working in hazardous conditions.

In 1981, A/S acquired Avanti Communications, a Chicago-based manufacturer of spe-

cialty antennas for land mobile, Amateur, auto-sound, and CB communications. Under the Avanti brand name, A/S has continued marketing a variety of products including CB base station antennas and an on-glass, no-ground-plane CB and land mobile antenna. The unique Avanti on-glass concept also led A/S to the development of a new mobile pager antenna system in 1982, the "Beeper Booster<sup>™</sup>."

Throughout its thirty-year history, A/S has enjoyed a reputation for product quality and reliability. Its Gold Seal Warranty on base station antennas, for example, is the only one in the industry providing reimbursement for re-installation as well as replacement of a failed antenna.

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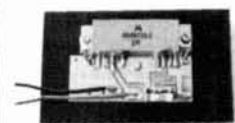
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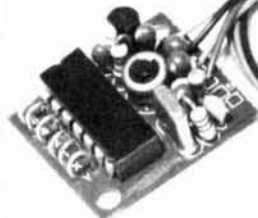
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## Coming Events ACTIVITIES "Places to go..."

**ILLINOIS:** The 1983 LAMARSFEST, Sunday, March 27, Lake County Fairgrounds, Routes 45 and 120, Grayslake. Public admitted 8 AM. Tickets \$2.00 advance, \$3.00 door. 9' tables \$5.00 each. Free parking. Refreshments available. Talk in on 147.63-03 or 146.94 simplex. For tickets, table reservations, exhibitor information SASE to LAMARS, PO Box 751, Libertyville, IL 60048.

**INDIANA:** The Martinsville Hamfest, March 13, sponsored by the Morgan County Amateur Radio Club, Morgan County 4-H Building and Fairgrounds. Admission: \$4.00 at door, \$3.00 advance, children 11 and under free. Flea market with table \$5.00; without table \$3.00; premium table \$20.00. Free parking. Doors open to public 8 AM. Vendor setup starts 5 AM. Talk in on 147.66/06. For tickets, table reservations and information SASE to Aileen Scales, KA9MBK, 3142 Market Place, Bloomington, Indiana 47401.

**MARYLAND:** The Baltimore Amateur Radio Club's 1983 Greater Baltimore Hamboree and Computertest, March 27, Maryland State Fairgrounds Exhibition Complex, Timonium. Gates open 8 AM. Admission \$3.00, children under 12 free. Large indoor dealer and flea market area. Large outdoor tailgate area. Refreshments, free parking. Guest speakers include Vic Clark, President ARRL. For information and table reservations: G B H & C, PO Box 95, Timonium, MD 21093-0095. (301) 561-1282.

**MINNESOTA:** The 6th annual Rochester Area Hamfest, Saturday, April 9, 8:30 AM, John Adams Junior High School, 1525 NW 31 Street, Rochester. Large indoor flea market, refreshments, free parking. Talk in on 146.22/82. For further information: RARC, c/o WB0YEE, 2253 Nordic Ct. N.W., Rochester, MN 55901.

**MISSOURI:** The J.B.A.R.C.'s Amateur Radio Auction, March 11, Carondelet Sunday Morning Athletic Club, 1012 Loughborough, St. Louis. Open 6 PM; auction 7:30 PM. Call in on 146.94/34.

**NEW HAMPSHIRE:** The 3rd annual Hamfest/Flea Market sponsored by the Great Bay Radio Association, Saturday, April 9, Somersworth Armory, Somersworth, 9 AM to 3 PM. Entrance fee \$1.00 per person. Refreshments available. Free parking. For advance registrations and further information: Great Bay Radio Association, PO Box 911, Dover, NH 03820.

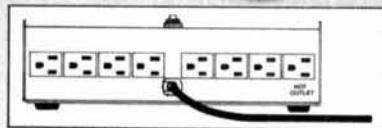
**NEW JERSEY:** The Delaware Valley Radio Association's 11th annual flea market, Sunday, March 13, 8 AM to 4 PM, New Jersey National Guard 112th Field Artillery Armory, Eggerts Crossing Road, Lawrence Township. Advance registration \$2.50, \$3.00 door. Indoor/outdoor flea market area. Refreshments. Sellers bring own tables. Talk in on 146.52 and 146.07-67. For information: D.V.R.A., PO Box 7024, West Trenton, NJ 08628. (SASE please.)

**NEW JERSEY:** The Chestnut Ridge Radio Club's Ham Radio Flea Market, Saturday, March 19, Education Building, Saddle River Reformed Church, East Saddle River Road at Weiss Road, Upper Saddle River. Tables: \$10.00 for first, \$5.00 each additional. Tailgating \$5.00. Refresh-

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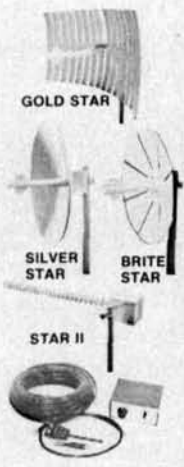
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ments. Free admission. Contact: Jack Meagher, W2EHD (201) 768-8360 or Roger Soderman, KW2U, (201) 666-2430.

**OHIO:** The TEAYS ARC will hold its sixth annual "King of the Pumpkin Hamfest," Sunday, March 20, 8 AM to 4 PM, Pickaway County Fairgrounds Coliseum. Tickets \$2.00 advance, \$3.00 door. 8' tables \$4.00 advance, \$5.00 door. Talk in .52-.52 and 147.78-18. Open for setup Saturday 4 PM. Overnight security provided. For information SASE to Dan Grant, W8UCF, 22150 Hulse Road, Circleville, Ohio 43113. (614) 474-6305.

**OHIO:** The Lake County Amateur Radio Association's fifth annual Hamfest and Computer Fest, Sunday, March 27, Madison High School, Madison. All indoors. 8 AM to 4 PM. Admission: \$2.50 advance and \$3.50 door. Table/display space \$5 per 6' table; \$6.50 per 8' table. Talk in on 147.81/21. For information and reservations SASE to Lake County Hamfest Committee, 37778 Lake Shore Blvd., Eastlake, Ohio 44094. (216) 953-9784.

**OHIO:** The 14th annual B\*A\*S\*H Friday night of Dayton Hamvention, April 29, Convention Center, Main and Fifth Streets. Adjacent parking. Free admission. Refreshments and entertainment. Two exciting top awards and more. For further information contact the Miami Valley FM Association, PO Box 263, Dayton, Ohio 45401.

**PENNSYLVANIA:** The Penn Wireless Association's Tradefest '83, Sunday, March 27, National Guard Armory, Southampton Road and Roosevelt Blvd., Philadelphia. Sellers space (6X8) \$5.00. Bring tables, limited number of power connections, \$3.00. General admission \$3.00. Refreshments, displays and surprises. Talk-in on 146.115/715 and .52. Contact: Mark Pierson, KB3NE, 12517 Nanton Drive, Philadelphia, PA 19154.

**PENNSYLVANIA:** The Conemaugh Valley Amateur Radio Club's sixth annual Hamfest, Sunday, March 27, East Taylor Fire Hall, Route 271, 4 miles north of Johnstown. 8 AM to 4 PM. Plenty of food and refreshments available. Check in on the 146.34/94 repeater.

**ROCHESTER HAMFEST:** Atlantic Division/New York State Convention. Saturday, May 21, Monroe County Fairgrounds. Hotel headquarters, Rochester Marriott Thruway. More info? Write or call Rochester Hamfest, 300 White Spruce Blvd., Rochester, NY 14623 (716) 424-7184.

**WISCONSIN:** The Tri-County Amateur Radio Club's annual Hamfest, March 20, 8 AM to 3 PM, Jefferson County Fairgrounds, Jefferson. No price increase! Tickets \$2.50 advance, \$3.00 door. Tables \$2.50 advance, \$3.50 door. Free parking and plenty of food and refreshments. Talk in on 146.52, 146.22/82 and 144.89/145.49. For information, tickets and tables SASE to Horace Hilker, K9LJM, PO Box 204, 261 E. High Street, Milton, WI 53563.

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**OPERATING EVENTS**  
 "Things to do..."

**MARCH 12:** The Green Mountain Wireless Society will operate N1VT from 1400Z to 2100Z from the Paul P. Harris Memorial Building, the site where the founder of Rotary International attended school. Frequencies: 7.235 and 21.360. Certificate available for QSO number, QSL and large SASE to Wallingford Rotary Club, PO Box 456, Wallingford, Vermont 05773, att: Ted Lidstone.

**MARCH 19:** B.A.R.T.G. Spring RTTY Contest. 0200 GMT Saturday, March 19 until 0200 GMT Monday, March 21. Total contest period is 48 hours but not more than 30 hours of operation is permitted. Bands: 3.5, 7.0, 14.0, 21.0 and 28 MHz. Stations: May not be contacted more than once on any one band but additional contacts may be made with the same station if a different band is used. Messages: Time GMT, RST and contact number. All logs must be received by May 31, 1983 to qualify. Summary and log sheets available from contest manager for two IRC's: Ted Double GBODW, 89, Linden Gardens, Enfield, Middlesex, England EN1 4DX.

**MARCH 19:** The 13th annual Tennessee QSO party, sponsored by the Tennessee Council of Amateur Radio Clubs from 2100Z March 19 to 0500Z March 20 and 1400-2200Z March 20. Exchange signal report and county. Out-of-state stations send signal report and state, province or country. Suggested frequencies: CW - 1815 kHz and approx. 50 kHz from bottom of HF bands. Phone: 1860, 3980, 7280, 14280, 21380, 28580 kHz. Novice: 3725, 7125, 21125, 28125 kHz. Logs must be postmarked no later than May 1, 1983. Send business SASE with your log for complete results and any certificates earned to Oak Ridge Amateur Radio Club, Att: Contest Coordinator, PO Box 291, Oak Ridge, Tennessee 37830.

**MARCH 26:** The 4th annual Spring VHF/UHF QSO Party, sponsored by the Ramapo Mountain Amateur Radio Club, from 2100 UTC Saturday, March 26, to 0400 UTC Sunday, March 27. The grid square and range scoring system is being used. SASE to RMARC, PO Box 364, Oakland, NJ 07436 for logentry forms and other information.

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DC 800	120V/60Hz	8VDC 800mA	\$1.95
DC8112	120V/60Hz	6.9V 12VDC 300mA	\$8.95
DC8490	117V/60Hz	6.5VDC 275mA	\$2.95
DC860	120V/60Hz	9VDC 500mA	\$3.95
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30008 National Memory Data Book (1980) (464 pages) RAMs, ROMs, PROMs, EPROMs Series	\$6.95
30009 Interchip Data Book (1983) (1306 pages) Complete line	\$9.95
30010 National Audio/Radio Handbook (1980) (240 pages) Pre-Amps, AM, FM & FM Stereo, Power Amps	\$5.95
30011 National Linear Application Handbook (1980) (726 pages) Application Notes, Linear Briefs, etc.	\$15.95
30012 National PAL Data Book (1982) (176 pages) Data Sheet, PAL Design	\$5.95
30013 Zilog Data Book (1983) (84 pages) Microprocessors and Support Chips	\$7.95
010400 Intel Component Data Book (1982) (1405 pg) Full data sheets for Intel's products incl. mem. devices, microproc, periph. & ind./mil. products	\$14.95
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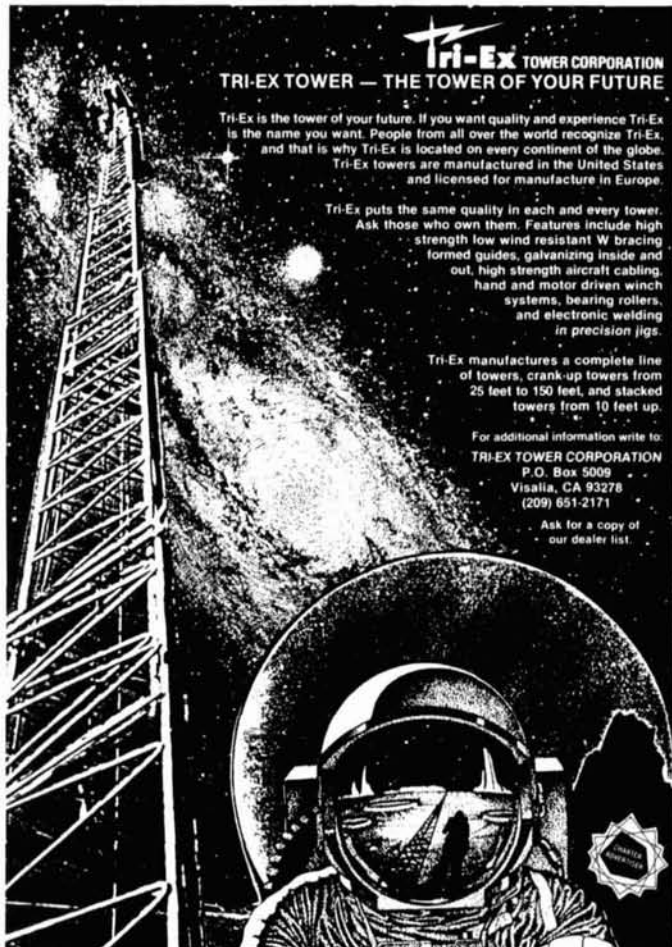
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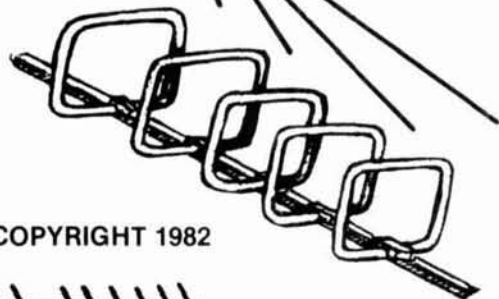
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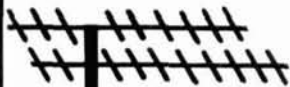
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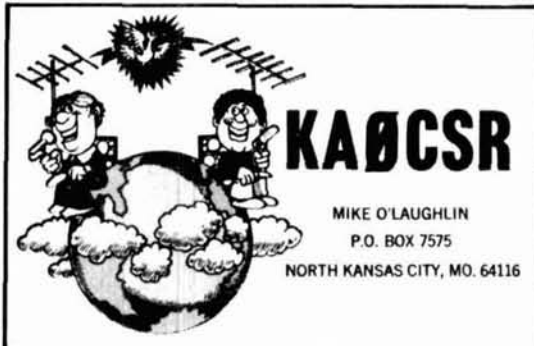
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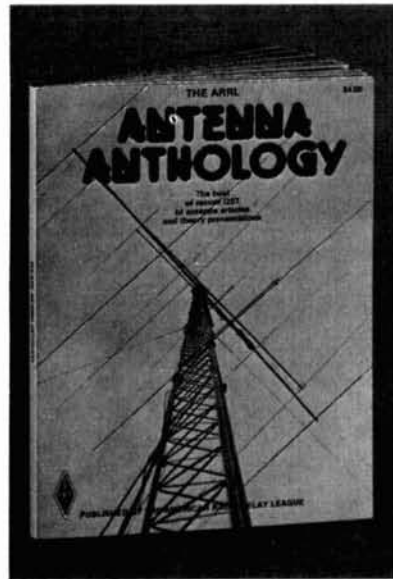
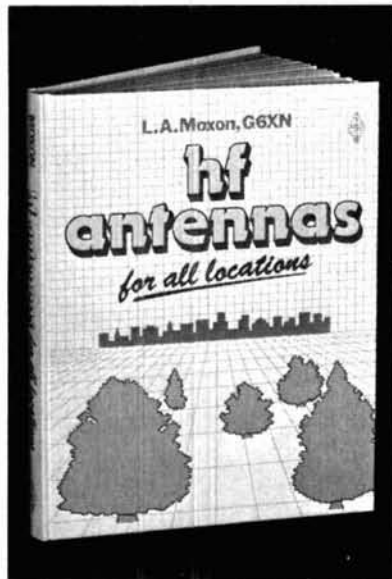
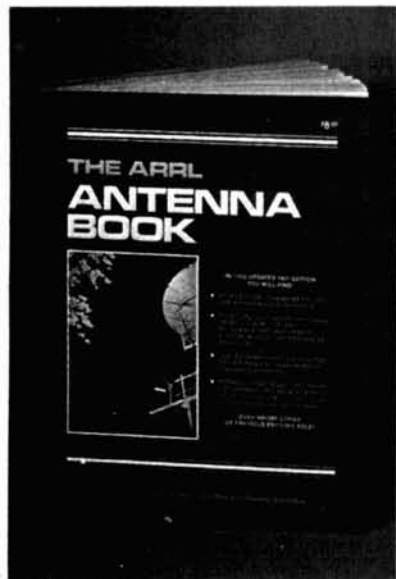
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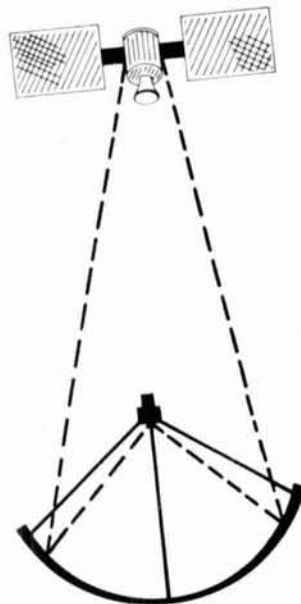
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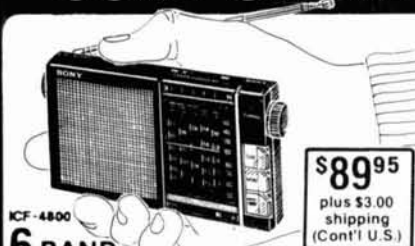
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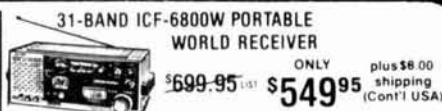
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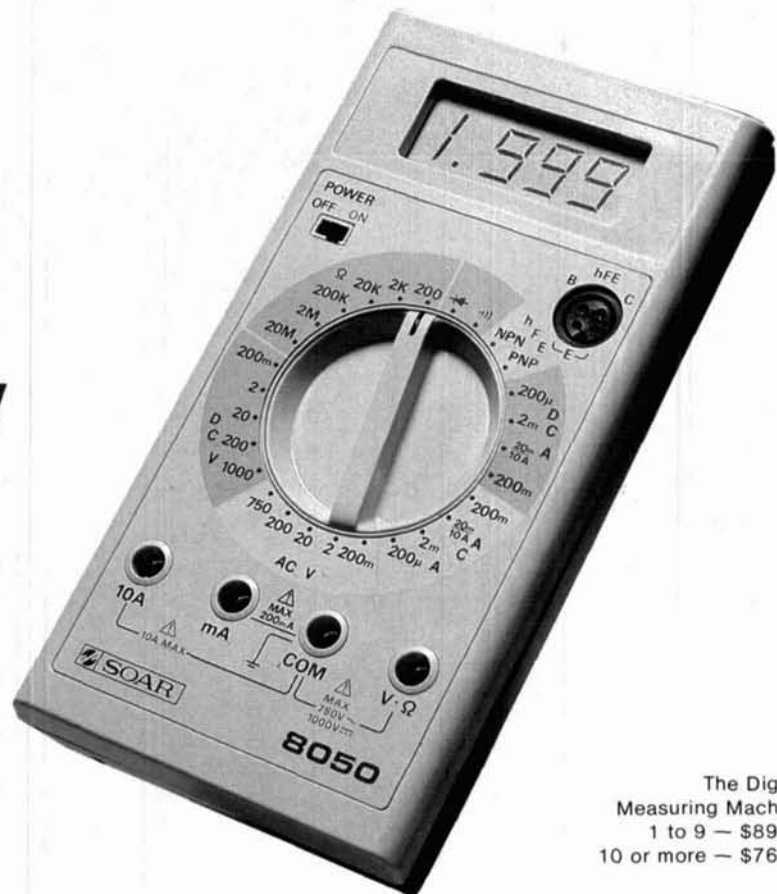
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# 10

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## SPECIFICATIONS

### TRANSMITTER

Power Input: (1.8-25 MHz) (28-29.9 MHz)	
SSB, CW	240W DC 160W DC
AM	80W DC 80W DC
FM	160W DC

### RECEIVER

Image Rejection:  
 Better than 70dB from 1.8-21.5 MHz  
 Better than 50dB from 24.5-29.9 MHz

IF rejection:  
 Better than 70 dB

Selectivity (-6 dB/ -60 dB):  
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1082



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# Scan the World.

# NEW



## SSB, CW, AM, FM, digital VFO's, 10 memories, memory and band scan, dual 24-hour clocks...

# R-2000

The R-2000 is an all mode SSB, CW, AM, FM receiver that covers 150 kHz-30 MHz in 30 bands. New microprocessor controlled operating features and an UP conversion PLL circuit provide maximum flexibility and ease of operation to enhance the excitement of listening to stations around the world. Key features include digital VFO's, ten memories that store frequency, band, and mode information, memory scan, programmable band scan, fluorescent tube digital display, and dual 24-hour clock with timer.

### R-2000 FEATURES:

- **Covers 150 kHz-30 MHz in 30 bands.**  
Uses innovative UP-conversion digitally controlled PLL circuit. UP/DOWN band switches (1-MHz step). VFO's continuously tuneable across 150 kHz-30 MHz.
- **All mode: USB, LSB, CW, AM, FM.**  
Provides expanded flexibility in receiving various signal types. Front panel mode selector keys, with LED indicators.
- **Digital VFO's for best stability.**  
50-Hz step, switchable to 500-Hz or 5-kHz, using front panel pushbutton switches. F. LOCK switch provided.
- **Ten memories store frequency, band, and mode data.**  
Complete information on frequency, band, and mode is stored in memory, assuring maximum ease of operation. Each memory may be tuned as a VFO. Original memory frequency may be recalled. AUTO. M switch for automatic storage of current operating data, or when off, selective storage of data using M. IN switch.
- **Lithium battery memory back-up. (Est. 5 yr. life.)**
- **Memory scan.**  
Scans all memories, or may be programmed to scan specific memories. HOLD switch interrupts scanning. Frequency, band, and mode are automatically selected in accordance with the memory channel being scanned. The scanning time is approximately 2 seconds per channel.
- **Programmable band scan.**  
Scans automatically within the programmed bandwidth. Memory channels 9 and 0 establish upper and lower scan limits. HOLD switch interrupts scanning. Frequency may be adjusted, using the tuning control, during scan HOLD.
- **Fluorescent tube digital display (100-Hz resolution).**  
Built-in 7 digit fluorescent tube digital display indicates frequency or time, plus memory channel number. DIM switch provided. The display may be switched to indicate CLOCK-2, FREQUENCY, CLOCK-1, and timer ON or OFF by the front panel FUNCTION switch.
- **Dual 24-hour quartz clocks, with timer.**  
Permits programming two different time zones. Timer for ON and OFF programming. Timer REMOTE output on rear panel (not for AC power).
- **Three built-in IF filters with NARROW/WIDE selector switch. (CW filter optional.)**  
6 kHz wide or 2.7 kHz narrow on AM. 2.7 kHz automatic on SSB. 2.7 kHz wide on CW, or, with optional YG-455C filter installed, 500 Hz narrow. 15 kHz automatic on FM.
- **Squelch circuit, all mode, built-in, with BUSY indicator.**
- **Noise blanker built-in.**  
Eliminates pulse-type noise on SSB, CW, and AM.
- **Large front mounted speaker.**
- **Tone control.**
- **RF step attenuator. (0-10-20-30 dB.)**  
Four step attenuator, plus antenna fuse.
- **AGC switch. (Slow-Fast.)**
- **"S" meter, with SINPO "S" scale.**
- **High and low impedance antenna terminals.**  
A high impedance (500 ohm) terminal, and a low impedance (50 ohm) co-axial connector are provided.
- **100/120/220/240 VAC, or 13.8 VDC operation.** (Optional DCK-1 cable kit required for 13.8 VDC.)

### Other features.

- RECORD output jack.
- Audible "beeper" (through speaker).
- Carrying handle.
- Headphone jack.
- External speaker jack.

### Optional accessories:

- HS-4, HS-5, HS-6 headphones.
- DCK-1 DC cable kit.
- YG-455C 500-Hz CW filter.
- HC-10 World digital quartz clock.

More information on the R-2000 is available from all authorized dealers of Trio-Kenwood Communications  
1111 West Walnut Street  
Compton, California 90220.

# KENWOOD

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