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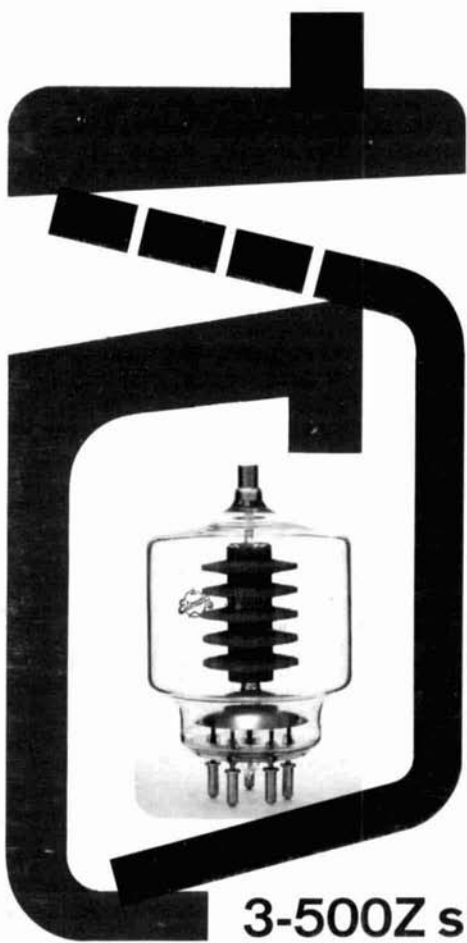
ham radio

magazine

FEBRUARY 1969

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3-500Z s
in a 2-kw PEP amplifier



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... or write for details ...

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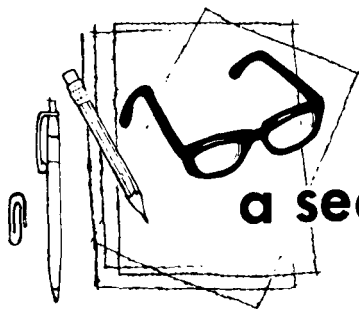
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a second look

by jim
fisk

If you were asked to generate a generous amount of rf power you'd probably ask, "at what frequency?" For the high frequencies, it would be transistors or vacuum tubes; at vhf you could add varactors. For microwave you would tend to think in terms of large klystrons, magnetrons . . . or semiconductors that use drift time in their operation such as klystrons use transit time.

The first of the semiconductor devices to actually work was built by Ian Gunn at IBM in 1963. In the basic Gunn-effect device, a simple wafer of n-type gallium arsenide—without junctions—generates microwaves directly when a voltage is placed across it. When a constant voltage is applied to the material, the current through it fluctuates wildly at an extremely rapid rate, although somewhat randomly; however, if the semiconductor wafer is thin (about 0.005 inch or less) the current no longer fluctuates in a random fashion, but rises and falls in a cyclic way. Presto: microwave energy.

The fundamental frequency range of Gunn-effect oscillators is 5 to 35 GHz, although the Gunn device is not a narrow band gadget. On the contrary, the frequency is governed by the circuit. The active semiconductor material has a natural resonant frequency, so if it is imbedded in a completely resistive load, it will oscillate at its natural frequency. If the material is connected into a tuned circuit, it is possible to change the frequency and provide essentially a widely tunable oscillator.

What about power output? Not too bad; there are continuous-wave Gunn oscillators available which provide nearly a half a watt at 10,000 MHz. If you are willing to go to pulse, you can obtain about 50 watts at the same frequency with increasing amounts of power as you slide down the spectrum, going up to a kilowatt or so at 5000 MHz.

If you think the Gunn-effect device sounds interesting, consider one of the other microwave-generating diodes. The avalanche type or IMPATT diode (for IMPact Avalanche Transit Time) is just one; IMPATT diode oscillators operate because of a negative resistance that results from a combination of internal secondary emission and bunched current carriers (electrons and holes) that drift through the solid-state material and deliver rf power by causing an external circuit current 180° out of phase with the applied voltage. The main disadvantages of IMPATT devices are high noise and low efficiency—on the order of 5%. The CW output powers of IMPATT diodes are slightly higher than Gunn devices and pulse powers are about the same.

Although these diodes are bound to whet your appetite for more information, there is one device that is even more interesting; silicon avalanche diodes have been built that have several hundred watts output in the 400 to 1250 MHz range! Results are so good they do not agree with theory for these devices so they are called—most appropriately—**anomalous** diodes. As an example of a device of this type, at a frequency of 425 MHz, the output power was 435 watts with a 164-volt supply at 12 amps. This represents an efficiency of 22%, low by vacuum-tube standards, but the same diodes oscillated at 10,000 MHz (in the IMPATT mode) with their ordinary efficiencies of less than 5%.

So far I haven't mentioned cost—for good reason—it's high; so was the varactor when Sam Harris described the first practical parametric amplifier ten years ago, but that didn't stop interested amateurs from getting devices for their own experiments.

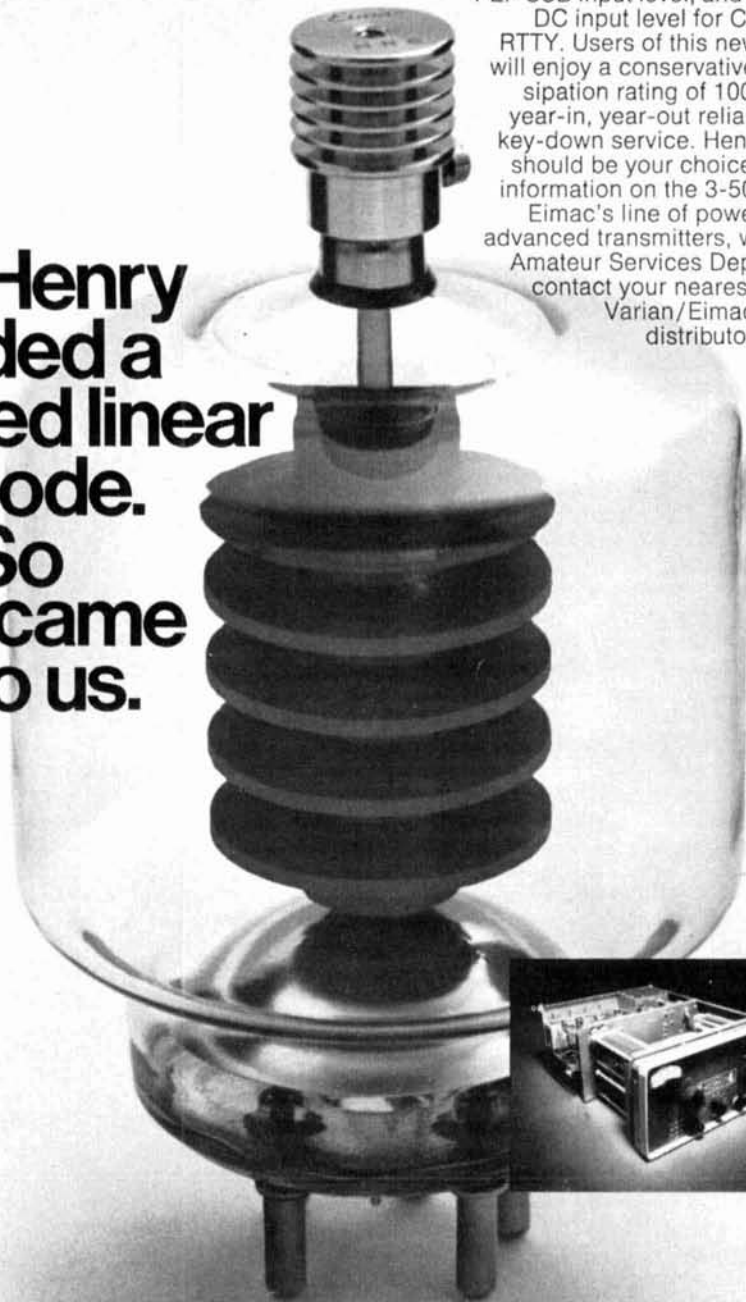
Jim Fisk, W1DTY
Editor

**Ted Henry
needed a
rugged linear
triode.
So
he came
to us.**

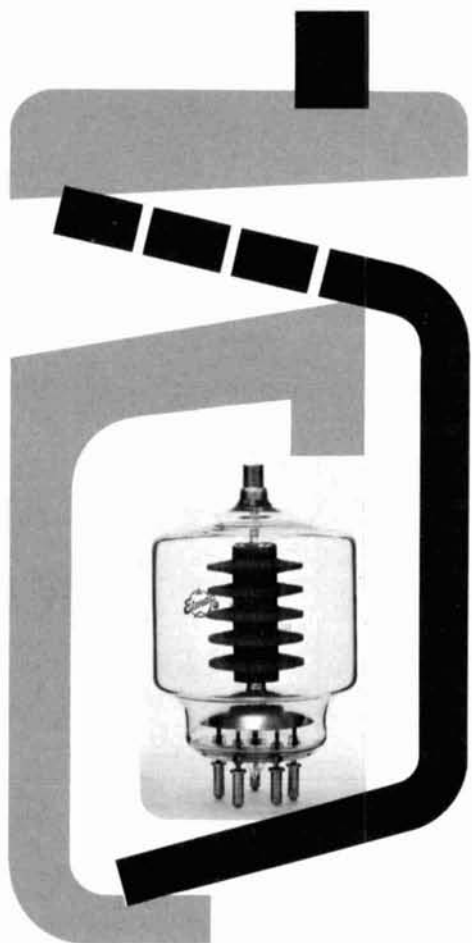
Two rugged Eimac 3-500Z high-mu triodes are featured in Henry Radio's new 2K-3 linear amplifier. Henry designed the amplifier around versatile Eimac power tubes because these popular triodes are ideal for grounded-grid operation at the 2 kW PEP SSB input level, and at the 1 kW

DC input level for CW, AM and RTTY. Users of this new Henry rig will enjoy a conservative plate dissipation rating of 1000 watts for year-in, year-out reliability under key-down service. Henry's choice should be your choice. For more information on the 3-500Z and on

Eimac's line of power tubes for advanced transmitters, write Eimac Amateur Services Department or contact your nearest Varian/Eimac distributor.



a grounded-grid two-kilowatt pep amplifier



This "full-house" linear
uses a pair
of 3-500Z's
in a highly
efficient circuit—
W6SAI
designs it for you
and
explains its operation

The popularity of the grounded-grid amplifier is well established. This interesting circuit combines high efficiency, good linearity, and minimum expense—three attributes not usually found together in one amplifier. A quick on-the-air check reveals that an amazingly large number of amateurs are using grounded-grid amplifiers and, judging from the robust signals, are having much success with them.

the grounded grid

To the uninitiated the grounded-grid or cathode-driven circuit is puzzling. A simplified sketch of this amplifier is shown in fig. 1. The driving signal is applied to the filament, or cathode, and the grid is at ground potential, thus acting as a shield between input and output. When the cathode is driven negative, the grid is positive with respect to the cathode, and plate current flows. The driving voltage is thus added to the plate voltage, and the resultant increase in power input to the stage appears as feed-through power in the plate circuit. The plate circuit is conventional, being either parallel-tuned or the more popular pi or pi-L network.

At first glance it might appear that the grid receives no exciting signal. The driving voltage is developed **between the grid and cathode** of the tube, and it matters little which of these two elements is at ground potential. That is to say, the grounded-grid amplifier may be thought of as a grid-driven amplifier "standing on its head." This is not strictly correct, as other factors such as feed-through power and grid-lead inductance enter the picture. However, if you understand the conventional grid-driven amplifier, you can understand the grounded-grid amplifier. The fact that the grid is at ground potential should cause you no distress.

If a well-shielded triode having a reasonably high amplification factor is used in the grounded-grid amplifier, neutralizing circuits and bulky bias supplies can be eliminated, reducing the amplifier stage to the basic es-

entials needed to make it function properly. The new Eimac 3-500Z, and its older companion the 3-400Z, are ideally suited to this class of service. They combine the characteristics mentioned above with modest cost, providing more watts output per dollar.

Of interest to the user is that the drive requirements of a grounded-grid, 2-kilowatt PEP amplifier using a pair of these tubes is of the order of the output power level of typical amateur exciters and transceivers; that is, in the 100-watt PEP region. Bulky and expensive power absorbers between amplifier and exciter aren't required. Best of all, a portion of the drive power appears as output in the grounded-grid amplifier. This almost repeals the ancient law that says "You can't get something for nothing." In this case, it is nearly true!

A pair of 3-500Z or 3-400Z power triodes will run comfortably at 2 kilowatts input

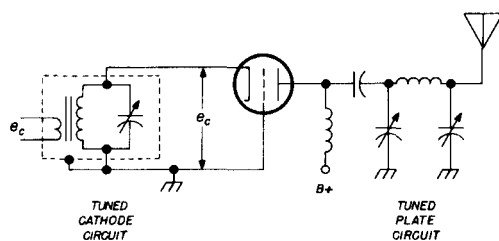


fig. 1. The basic grounded-grid amplifier. The driving voltage, e_c , is impressed between cathode and grid. The grounded-grid circuit isolates input from output and affords good shielding up to 100 MHz or so in tubes of the 3-500Z class.

PEP (1000 watts average power) under continuous ratings in CW, ssb or RTTY service without distress. "Fireman-quick" tune-up style, wherein the equipment must be tuned up in 30 seconds or less else the whole works goes up in a puff of smoke, is not necessary. This fire drill may have an element of excitement about it, but it can result in tube casualties when done in a

hurry, or under pressure, as in the closing moments of a DX contest. Adequate plate dissipation, on the other hand, is insurance against catastrophe from operator panic or error and provides a comfortable margin of safety that leads to equipment reliability and lower operator blood pressure.

two-kw pep amplifier

The Eimac 3-400Z and 3-500Z tubes are air-cooled power triodes designed for zero-bias, class-B rf amplifier service. They are particularly well-suited for ssb linear amplifier service. Intermodulation distortion products are 30 dB or better below one tone of a two-tone test signal for either tube type. A comparison of the 3-400Z and 3-500Z types is given in Eimac Amateur Service Bulletin number 35*.

A circuit for either the 3-400Z or 3-500Z is shown in **fig. 2**. The tubes are cathode driven with grids at both rf and dc ground potential. A tuned-cathode network achieves an optimum impedance match to the exciter and reduces intermodulation distortion to a minimum. The plate pi-network circuit is suitable for operation on amateur bands between 80 and 10 meters. Separate meters monitor grid and plate current. A VOX-operated circuit reduces plate dissipation during standby. Power output is better than 1200 watts PEP on all bands, and driving power is about 75 watts PEP. Plate potentials between 2500 and 3000 may be used, with reduced output for plate voltages as low as 2000.

the plate circuit

The amplifier plate circuit may be of several forms. The easy and more expensive assembly is to use a bandswitching pi-network inductor. The less-expensive but time-consuming procedure is to make your own bandswitching network from commercially available coil stock, copper tubing and a heavy-duty switch. Either approach will work well, and your choice will probably be made on an economic basis. Data for either type of circuit is given in the parts list, **table 2**.

One idea that hasn't been popularized is designing the amplifier for single-band oper-

fig. 2 (right). Schematic of the grounded-grid linear amplifier. RFC1 should be mounted reasonably in the clear so that the series-resonant frequency does not drop below 23 MHz. Capacitors marked C are disc ceramic, Centralab DD-103. Power-line bypass capacitors are Centralab DD16-103.

blower 13 cubic feet per minute at 0.2-inch back pressure; use number-3 impeller at 3100 rpm (Ripley 8472, Dayton 1C180 or Redmond AK-2H-01AX)

sockets Eimac SK-510 socket and SK-416 chimney for 3-400Z; Eimac SK-510 socket and SK0406 chimney for 3-500Z

table 1. Cathode circuit components.

T1, 10 meters 0.15 μ H. 4 turns number-14 AWG enamelled wire on $\frac{1}{2}$ " form, $\frac{1}{2}$ " long (National XR-50 form); parallel capacitance: 200 pF, 1 kV silver mica; resonate to 28.7 MHz

T2, 15 meters 0.15 μ H. Same as T1, resonate to 21.3 MHz with 470-pF, 1-kV silver-mica capacitor

T3, 20 meters 0.31 μ H. 6 turns number-14 AWG enamelled wire on National XR-50 form; parallel capacitance: 470 pF, 1 kV silver mica; resonate to 14.3 MHz

T4, 40 meters 0.31 μ H. Same as T3, resonate to 7.2 MHz with 940 pF (two 470 pF, 1 kV silver micas in parallel)

T5, 80 meters 1.3 μ H. 13 turns number-18 AWG enamelled wire on National XR-50 form; parallel capacitance: 940 pF, as in T4

table 2. Plate circuit components.

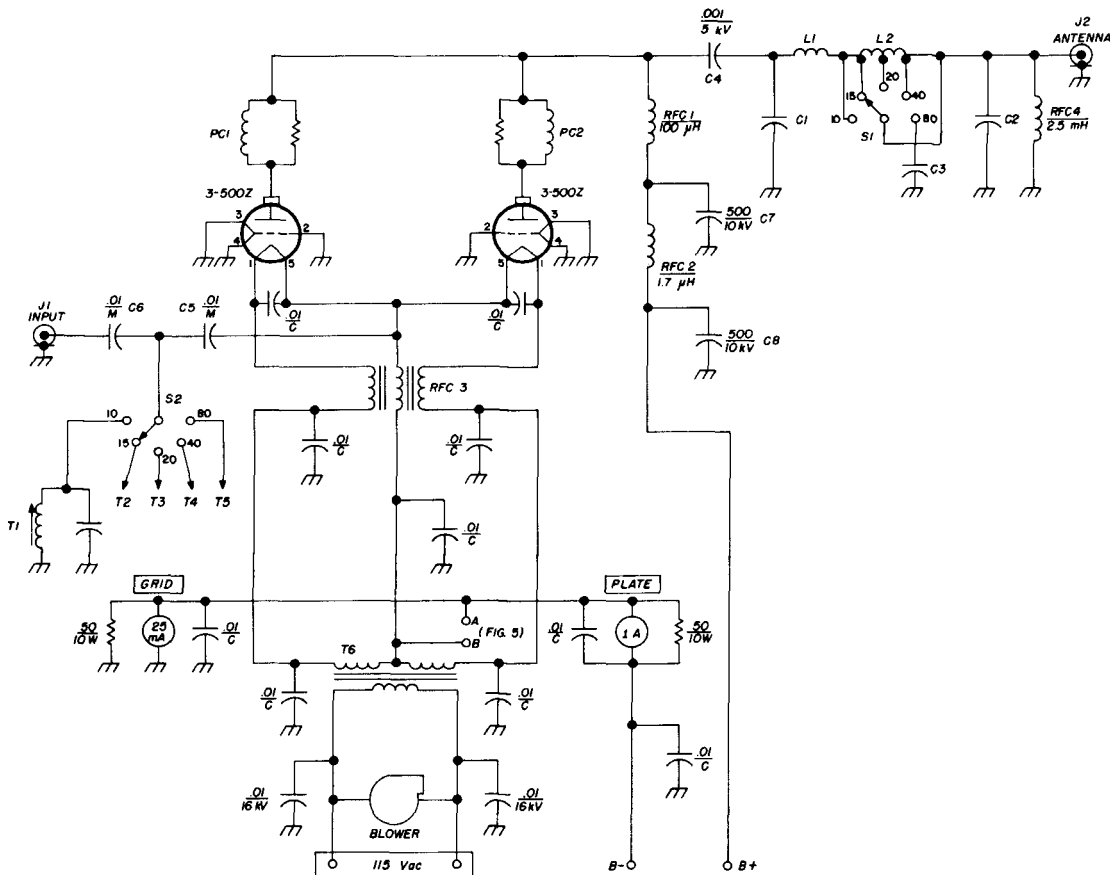
combination 1 B&W 850A bandswitching inductor

combination 2 Air-Dux 195-2 pi-inductor

combination 3 homemade tapped inductor; 80 meters, 13.6 μ H; 40 meters, 6.5 μ H; 20 meters, 1.75 μ H; 15 meters, 1.0 μ H; 10 meters, 0.8 μ H

Combinations 2 and 3 should be tapped and trimmed to resonate as follows: 80 meters, 210 pF; 40 meters, 105 pF; 20 meters, 52 pF; 15 meters, 30 pF; 10 meters, 30 pF. Inductor should be parallel resonated and grid dipped with these capacitance values before placement in circuit. For combination 3, the 10-, 15- and 20-meter coil should be wound with 3/16-inch diameter copper tubing, and the 80- and 40-meter coil wound with number-12 AWG wire.

* Available from the author upon request.



- C1 250 pF, 4.5 kV (Johnson 154-16)
 C2 1500 pF (LaPointe Industries, Rockville, Connecticut, type J-1500-S-30)
 C3 500 pF, 2500-volt mica (surplus)
 C4 .001 μ F, 5 kV (Centralab 858S-1000)
 C5, C6 .01 μ F, 500-V mica (two EIMenco CM-30 in parallel)
 C7, C8 .001 μ F, 6 to 10 kV (Centralab DD60-12)
 L1, L2 see table 2
 PC1, PC2 three 100-ohm, 2-watt carbon resistors in parallel (Ohmite "Little Devil"); three turns number 14 spacewound around one resistor
 RFC1 100 μ H, 1A (B&W 800)
 homemade versions: A. 112 turns number-26 AWG, spacewound (wire diameter) on 1" diameter, 6" long ceramic form (Centralab X-3022H insulator) series resonant at 24.5 MHz; grid dip with terminals shorted
 B. 136 turns number-20 AWG, 1" diameter, 4 1/4" long (B&W 3016 or Air-Dux 832T), slipped over varnished wood dowel for

- mounting, series resonant at 25 MHz; grid dip with terminals shorted
 RFC2 1.7 μ H (J. W. Miller RFC-144 or Ohmite Z-144) homemade: approximately 20 turns number-26 AWG enamelled wire wound on 100k, 2-watt resistor
 RFC3 for series filament operation: 12 1/2 bifilar turns number-12 AWG enamelled wire wound on ferrite core 3" long, 1/2" diameter (Indiana General CF-503, Newark electronics catalog number 59F-1521); notch core with file and snap to break
 for parallel operation: overwind third winding with number-18 AWG enamelled wire (see photo)
 S1 single-pole, 12-position ceramic switch, 7 kV test (Radio Switch Corporation, Marlboro, New Jersey, type 86A)
 S2 single-pole, six-position ceramic switch (Centralab PA-2001)
 T1 - T5 see table 1
 T6 series filament operation: 10 volts at 15 A (Thordarson 21F-146); parallel operation: 5 volts at 30 A (Chicago-Stancor P-6468)

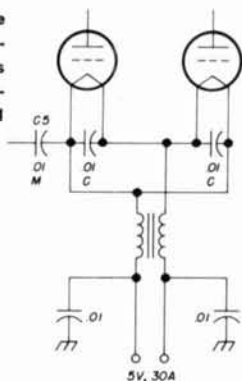
ation. If your station operation (or the greater portion of it) is confined to a single hf band, a worthwhile reduction in cost may be achieved by constructing the amplifier for single-band operation. In this case, only one tuned input circuit is used, and the other transformers and switch S-2 may be omitted. The plate circuit may be simplified to a single tank coil, and the expensive and rather bulky plate switch may be omitted, along with the problem of placing the proper taps on the bandswitching coil. No hard and fast rules apply, and the choice is yours.

the input circuit

This is the feature that really distinguishes the grounded-grid amplifier from its conventional grounded cathode counterpart. A fixed-tuned cathode circuit is used, with each input circuit (T1 through T5) tuned to the center of each band (table 1). A grid dip meter is used for resonating the cathode circuits. Each circuit should be peaked before it's installed.

Because the cathode circuit replaces the grid circuit as the input system to the amplifier, the cathode circuit must be isolated from ground. This means that cathode voltage has to be applied through an rf choke.

fig. 3. Filament choke for parallel filament connection. Core material is Q-1 type having a permeability of 125 at 1 MHz.



These chokes are available commercially, but are rather expensive. Why not make your own?

The filaments may be operated either in series, as in fig. 2, or in parallel (fig. 3). Series operation requires a current balancing re-

turn to the junction of the series filaments. This means a third winding is needed on the isolation choke. Here's how to make your choke.

the cathode choke

Use a ferrite core, but beware of the transistor-radio ferrites used for loopstick antennas. They **won't** work in this application and are inferior at higher frequencies. I used Q-1 ferrite material with a permeability ra-



Filament choke for grounded-grid amplifier is wound on ferrite core, one-half inch in diameter and three inches long. The text explains how to build your own.

tio of 125 at 1 MHz. My core is one-half inch in diameter and three inches long. (The choke in the photo was wound on a two-inch-long core, which makes winding a bit tight; a slightly longer core makes the winding job easier.)

I used 12½ turns of number-12 AWG enameled wire, bifilar wound, and held in place with epoxy cement. If you want to use series-connected filaments, wind another layer using number-18 AWG enameled wire. Wind it over the primary. Epoxy cement will hold it in place. Tin the end of each winding and apply tape as shown in the photo. The assembled choke is extremely rugged when the epoxy dries, and you

will save at least \$5.00 by making your own.

If the filaments are series-connected for 10-volt operation, the tertiary winding on the filament choke must make a connection between the center tap of the filament transformer and the junction between the series-connected tube filaments. If this connection is omitted, the filament voltages for the two tubes will not necessarily be balanced, as one tube may draw slightly more current than the other. Manufacturing tolerances

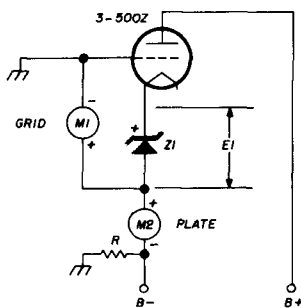


fig. 4. Bias and metering circuit for 3-500Z tubes. Meter M1 reads grid current between cathode and grid; meter 2 reads plate current between cathode and plate. The bias voltage, E1, is provided by the zener diode, Z1.

permit a current variation of about five percent between two tubes of this size. It's therefore possible for one tube to absorb the filament voltage at the expense of the other. Parallel filament operation eliminates this problem. However, 10-volt filament transformers are sometimes easier to obtain than 5-volt units.

the grid circuit

This circuit, by careful inspection of the schematic, is almost nonexistent. All three grid pins of each tube socket are grounded to the amplifier chassis. The proper Eimac air-system socket and chimney should be used, regardless of your choice of tubes. The less-expensive ceramic socket often used with tubes of this type applies too much lateral force to the base pins and glass header. The Eimac SK-410 socket has special, flexible connectors that allow the tube to

move about. This equalizes any stress when plugging the tube into the socket. Also, this socket has small slots in the shell through which grounding straps may be passed to permit a short, low-inductance ground connection to grid pins 2, 3 and 4.

the metering circuit

You may wonder, "How is grid current measured when the grid is directly grounded?" The answer is that grid current may be measured in the grid return circuit to the tube filaments. Grid meter M1 handles this chore. The grid circuit electron flow, then, is from filament to grid within the tube, from grid to ground, from ground to the negative terminal of meter M1, through the meter, then from the positive terminal of meter M1 to the filament center tap.

The plate-current meter may be handled similarly. Instead of placing it in series with the B-plus lead, which may be highly dangerous, the meter is in the B-minus return lead to the power supply. No grid current flows through plate meter and no plate current flows through grid meter. This circuit is shown in fig. 4.

standby and vox control

The amplifier should be turned off to conserve power during listening periods and to reduce diode noise in the receiver. Under certain conditions, it's possible for large transmitting tubes in standby to generate a powerful hiss when they pass resting plate current. Normally, you don't hear this unless a T-R switch is used, but it can be a nuisance and is often noticed even when coaxial antenna changeover relays are used.

Common practice calls for the amplifier plate supply to run during standby, with the transmitter energized by the voice-controlled (VOX) system. A number of alternatives, then, present themselves to solve the diode noise problem and allow effective VOX control.

The schematic of fig. 2 shows the filament circuit interrupted by two terminals (A-B). When these are open, the amplifier is in a cut-off mode, and no amplifier plate current flows. I don't recommend the amplifier be used under these conditions, but fig. 5

illustrates several acceptable circuits that may be used across terminals A-B to do the job.

Fig. 5A shows A-B shorted, in which the amplifier operates normally, with no VOX control and no provision for reducing noise. **Fig. 5B** shows the VOX relay contacts in parallel with a 10-k ohm, 10-W bias resistor. When the VOX relay is in standby, its contacts are open, and the resistor is in the filament-return circuit. A small plate-current flow through this resistor creates a voltage drop across the resistor. This applies a negative blocking bias to the amplifier grids. The standby plate current is thus reduced to near zero, and the noise is eliminated. When the VOX circuit is energized, the relay contacts short circuit the bias resistor. Am-

If asked for my circuit preference, I'd vote for the circuit in **fig. 5B**. The addition of the bias, in my opinion, is "gilding the lily," unless the power supply is of marginal rating and every extra watt must be carefully conserved, or unless you want to reduce cabinet temperature during extended periods.

the air cooling circuit

Until some genius miniaturizes the watt, a means will be necessary to remove the heat generated in an amplifier of this size. When you realize that the filaments of the two tubes have a power capacity of 145 watts, and the total plate dissipation may reach **1000** watts during the tune-up process, you can appreciate that the amplifier cabinet temperature may reach astounding propor-

A. No bias and no vox; short circuit terminals A and B of fig. 2.

B. VOX-only circuit; relay contacts are part of external vox circuit.

C. Zener diode bias only; use 1N4551 zener diode for one or two 3-500Z's.

D. Vox circuit and zener diode bias.

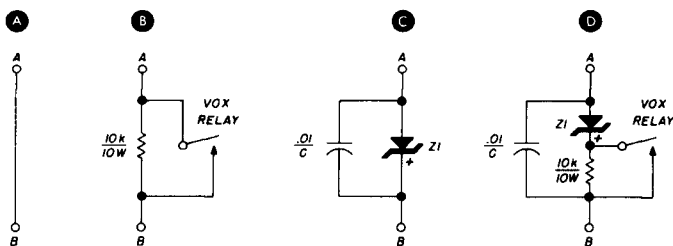


fig. 5. Bias and vox circuits for the grounded-grid amplifier.

plifier operation then returns to normal.

Fig. 5C shows how a zener reduces amplifier standby plate current. At a plate potential of 3000 volts, for example, the 3-500Z and 3-400Z resting plate currents are 160 mA and 100 mA respectively. This corresponds to a plate dissipation of 480 watts and 300 watts. Unless VOX bias is used, this amount of dissipation per tube can warm the operating room and run up the power bill in short order. Addition of a zener provides a few volts of regulated bias, which reduces plate current to a moderate value. Standby plate dissipation is reduced accordingly.

Fig. 5D shows the use of both circuits. The VOX relay and resistor bias provide near zero standby plate current, and the zener provides low static resting plate current when the amplifier is activated.

tions unless the heat is conducted elsewhere. Large commercial transmitters that must deal with thousands, or hundreds of thousands, of watts of dissipation often forsake air cooling for more efficient water cooling or vapor-phase (steam) cooling. At the amateur power level, however, simple convection or forced-air cooling is still the most popular method of removing heat.

Either the 3-400Z or the 3-500Z requires cooling air supplied at a rate of 13 cubic feet per minute capable of overcoming a back pressure created by tube, socket and chimney expressed as a pressure drop of 0.13 inches of water. Blowers that will meet this specification are specified in the parts list. A word of warning: don't cheat on air supply. Heat is the enemy of vacuum tubes, and a war-surplus blower or cheap fan may result in decreased tube life due to filament

or plate seals overheating. A few pennies saved in the purchase of the proper blower is false economy, indeed, and may lead to premature tube failure after a few hundred hours.

circuit stability

The grounded-grid amplifier is inherently a *degenerative circuit*, and the *grid shields* the input from the output circuit very effectively. The builder, however, is responsible for feedback occurring **around** the tube caused by coupling between input and output circuits. If sufficient coupling exists, it's possible for a fundamental frequency parasitic oscillation to occur. Proper amplifier shielding and correct isolation of power leads will eliminate this problem.

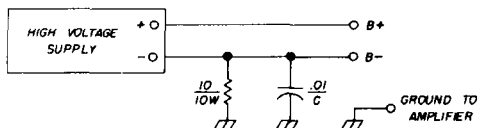


fig. 6. Power supply connections for proper "floating B-minus" circuit. Negative of supply is lifted above ground a few volts by the 10-ohm resistor.

The circuit shown in **fig. 2** has each power lead bypassed at the terminal board on the amplifier. Coaxial input and output rf connectors are used. If the cable run to the power supply is rather long, it's a good idea to run the leads in shielded cable. The B-plus lead can be a length of RG-8/U coaxial line, with the shield grounded at both ends.

The power supply B-minus circuit is above ground because plate meter M2 is in the return lead. A protective resistor should be in the power supply, as shown in **fig. 6**, to keep the supply from getting too far above ground.

The tuned-cathode transformers should be in a shielded enclosure to separate them from the rf currents and fields in the output circuit. There are several ways of doing this. One is to enclose the amplifier underchassis completely making it airtight. The blower is mounted on the rear wall of the enclosure,

transforming it into a plenum chamber, with the air exhausted through the tube sockets. The cathode circuit, in turn, is automatically shielded by the enclosure.

On the other hand, if the underchassis area is not pressurized, and the air is conducted separately to each tube socket by hoses, the cathode circuit should be in a *shielded enclosure* of its own. The plate circuit components, usually mounted above chassis, should be in a shielded enclosure that affords proper ventilation. This can be done with a cover shield made of perforated aluminum sheet. With both grid and plate circuits properly shielded, minimum intrastage coupling will exist.

A good way to determine the amount and degree of intrastage coupling is to observe the action of the grid and plate meters after the amplifier has been tuned and loaded. If the stage is operating properly, maximum grid current and minimum loaded plate current will occur at the same setting of the plate tuning capacitor. If, on the other hand, grid current reaches maximum when the plate circuit is slightly detuned from resonance, it's an indication that unwanted feedback exists from plate to grid. A small amount of unwanted coupling probably will do no harm, but instability may often be found in such an instance, particularly on the 10- and 15-meter bands. If the degree of coupling is great enough, the amplifier will break into oscillation at the operating frequency. Judicious use of proper shielding can easily eliminate this problem.

parasitic oscillation

Most modern tubes perform well into the vhf region and demonstrate ample gain and efficiency, even though they may be designed primarily for high-frequency operation. The 3-400Z and 3-500Z are no exceptions. High-frequency circuit and construction techniques, however, usually create a condition wherein input and output circuits are not especially well shielded or otherwise isolated from each other at frequencies above 100 MHz or so.

This isn't surprising, as the amplifier is not designed to work in this portion of the spectrum. The upshot of this combination

of forces is that the tube is capable of amplifying (and oscillating) in the vhf region. Unless steps are taken to prevent it, this very oscillation will occur. Auxiliary neutralizing circuits may be added to suppress the tendency to oscillate, or, more simply, the parasitic circuit may be loaded to suppress the oscillation. This is most easily done by

citator at random. If parasitics are absent, the meter readings will remain unchanged. If, however, the amplifier exhibits a tendency toward parasitic instability, a grid-current reading will be indicated, and a rise in resting plate current will be noted.

When a new item of equipment is being tested for the first time in this fashion, it



New Swan Mark II linear amplifier, designed to use either Eimac 3-400Z (foreground) or 3-500Z (installed in amplifier). The Swan incorporates zener bias to reduce zero-signal plate current of the 3-500Z tubes; bias may be switched out for the 3-400Z's.

the use of plate circuit parasitic suppressors, a nearly universal adjunct to most high-frequency amplifiers. The parasitic choke described in the parts list should do the job for either tube type. One choke is placed in each plate lead near the plate connector.

The check for vhf parasitics is simple. The amplifier plate and filament voltages are applied, but excitation is not. Observe the resting plate current and grid-current, then tune the plate-tuning capacitor and loading capa-

may be wise to drop the plate voltage in half with the aid of a variable voltage transformer in the power supply primary. If all looks well, the voltage may be raised a bit and the test repeated. The most drastic parasitic test of all is to boost plate voltage about 30 percent above normal while the test is being run. Higher-than-normal plate voltage raises the power gain of a tube and enhances the tendency towards parasitic oscillation, if such tendency is present.

If a parasitic oscillation occurs, the plate circuit parasitic suppressors must be altered in unison until the amplifier is "cold." Addition of a turn or two to each suppressor coil is suggested as the first step. Too many turns on the coil will cause the suppressor resistor to overheat, particularly on the 10-meter band. A proper compromise between suppressor heating and parasitic suppression can be worked out without too much effort.

amplifier tuning and adjustment

Once the amplifier has been tested and is free of parasitic oscillation, it is ready for final adjustment and tuning. Using a grid-dip oscillator, the cathode circuits may be tuned to frequency before they are installed and will need no further tuning. A check of these should show each to be resonant at about the mid-frequency point of each amateur band.

Adjustment of the pi-network output circuit may also be approximated with the aid of the grid-dip oscillator. The output (loading) capacitor is set to the approximate value of capacitance for a particular band, the plate leads are attached to the tubes, and the plate tuning capacitor varied to provide resonance on a grid-dip oscillator coupled to the network coil. Settings of tuning and loading capacitors should be logged. A dummy antenna and power output device (an swr meter, for example) are attached to the amplifier, as is the exciter. Plate voltage is applied, and a **small** amount of drive (carrier) is introduced into the amplifier. The plate tuning capacitor is adjusted for maximum indication on the power output meter.

The idea is to achieve maximum power output with proper grid and plate currents, combined with a minimum of grid drive. This is how you do it:

As an example, assume the amplifier has a pair of 3-500Z tubes operating at 2500 volts. To achieve 2000 watts PEP input (usually assumed to equal 1000 watts under voice conditions) the amplifier **must be tuned and loaded at the 2000-watt level**, unless some rather sophisticated test equipment is at hand. Accordingly, 800 mA of plate current must be run to the two tubes,

and the data sheet for the 3-500Z shows a grid current of 120 mA per tube, or a total of 240 mA for two tubes at this power level.

Here we go! Carrier is gradually inserted, and the amplifier is loaded toward the target plate current of 800 mA. Grid current is deliberately held on the low side as a safety measure. Plate loading is increased (pi-network output capacitance decreased), plate tuning is resonated and grid drive is slowly raised. At resonance we approach 700 mA plate current and 200 mA grid current. Tuning is "on the beam." Loading is juggled a bit to peak power output, and cathode drive is boosted a bit. The plate current is now 800 mA, grid current is 220 mA, and **power output drops if loading is either increased or decreased**. Power output also drops sharply when excitation is reduced. Under these conditions, the amplifier is loaded to 2000 watts dc input with the proper plate loading and with the correct ratio of cathode drive to plate loading.

Excitation is now removed, grid current drops to zero, and plate current returns to zero-signal value. Now, instead of a steady-carrier driving signal, a voice signal is impressed onto the amplifier. Voice gain is raised until voice peaks **occasionally** hit 400 mA. This indicates a "dc meter reading" of 1000 watts peak input (2500 volts at 400 mA). The PEP input, if viewed on an oscilloscope, would be in the neighborhood of 2000 watts or so. Grid current, under voice conditions, will peak about 110 mA.

Note that to establish a 2000-watt PEP input condition, the amplifier must be loaded and adjusted at the 2000-watt PEP level.

This is extremely important; it must be capable of handling 2000-watt peak signals without distortion. It is wise to use an oscilloscope to monitor voice peaks and ensure that over-excitation or "flat-topping" of the signal does not occur.

The experienced operator will find that the combination of grid current and power output indications are important adjuncts to proper tuning. Once the proper tuning and loading technique is mastered, adjustment of a linear amplifier will be simple and uncomplicated.

ham radio

signal detection and communication

in the presence
of white noise

An informative treatment
of the
radio noise problem,
with some answers
based on
detection theory

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Many years ago, when radio started, the detection of signals was receiver-sensitivity limited, since rf amplifying devices were not known. When these devices came about, it was found that receiving sensitivity was limited by natural and man-made interference rather than by the ever-present thermal noise of the antenna and its environment. This is still true for all but the higher amateur bands.

During the past several years new communications modes such as moonbounce, troposcatter and meteorscatter have become popular with pioneering radio amateurs. These modes have in common tremendous path losses ranging up to 280 dB, and all operate against a background of white (or thermal) noise. With receiver or antenna noise a limiting factor on one side and transmitter power and antenna size on the other, much brain power has been expended by amateur and professional alike to develop improved signal processing techniques and coding systems to achieve maximum performance.

Many claims have been made by amateurs for particular "circuits" that will pull signals out of the noise, "cancel" the noise, etc. Some of these claims are true, while others are oversimplifications and not ap-

plicable. The object of this article is to bring some of the findings of detection theory to bear on the problem and show what is not possible. I'll try to do this in simple language, with a minimum of math, but without yielding to the temptation of oversimplification.

the problem

Let's assume that we have reduced the incoming thermal noise to the lowest pos-

cw transmission

For ease of understanding we'll start with the simple proposition of detecting the presence of a transmitted rf carrier. Let's assume the carrier is transmitted at constant amplitude and constant frequency. At first glance it might appear that a bandpass filter of arbitrarily narrow bandwidth in the receiver would be the optimum choice for highest sensitivity, since the amount of noise in the filter bandwidth can be reduced to

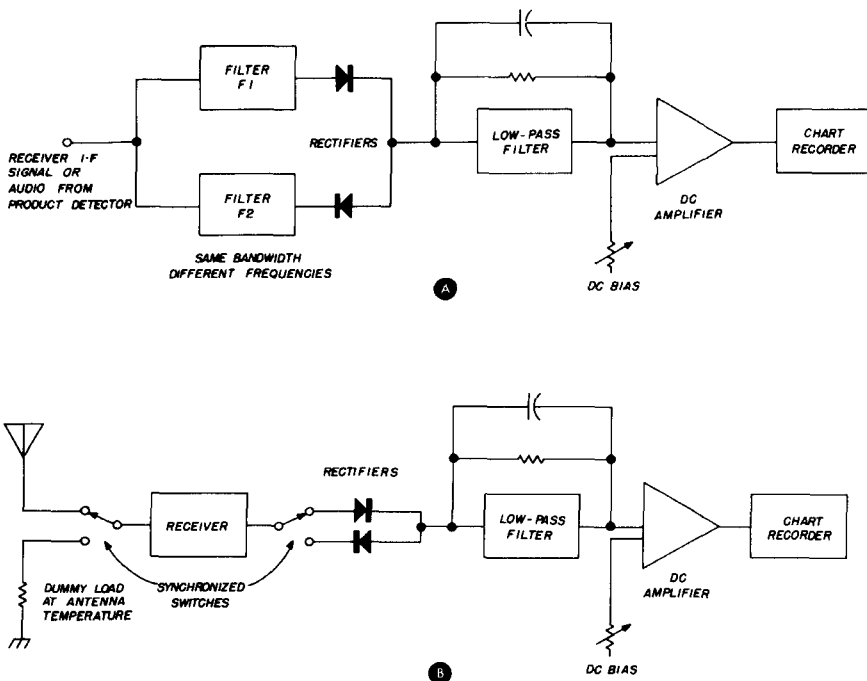


fig. 1. Weak signal cw receiving systems. The two-frequency method is shown in A; the Dicke radiometer system is shown in B.

sible amount by low-noise antennas, feed-lines and receiver front ends. What we've done is reduce the noise power spectral density in the frequency band of interest. This quantity is measured in watts per Hz of bandwidth. Let the problem now be: given this background of thermal noise density, how can we detect signals and communicate by modulating the signal in some way?

as low a value as necessary for detection of the carrier. Unfortunately this is not true.

Between transmitter and receiver exists a transmission path that always modifies the transmitted signal. Such phenomena as fading, scintillation, Doppler shifts, and multi-path modify the signal in the direction of increased bandwidth. The best receiving system, then, is that which maximizes the signal-to-noise ratio (snr) for the received

signal; not the one being transmitted. A filter that maximizes the ratio of signal power to noise power is commonly called a matched filter. Such a filter for the above case would be one with a response approximately as wide as the signal spectrum; however, the exact frequency response of the filter is similar to the power spectrum curve of the signal being received. I won't go into the exact mathematics here, because we don't ordinarily know enough about a specific path at a given time to really shape the filter curve for optimum snr. Besides, the snr is not too sensitive to changes in the shape of the filter curve.

noise. Typically, the output of the filter is then displayed on a chart recorder or scope. The signal-to-noise ratio at the output of the low-pass filter can be computed by the following relationship:

$$\frac{S}{N_{out}} = \left(\frac{S}{N} \right)_{IF}^2 \sqrt{\frac{B}{2\Delta f}}$$

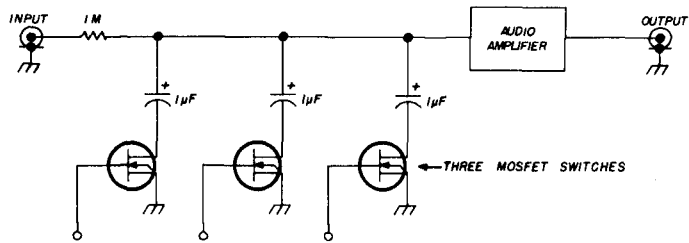
(B >> Δf)

where

B = i-f bandwidth.

Δf = lowpass filter bandwidth.

fig. 2. Electronic digital filter. A three-phase square wave at the frequency of interest is used to drive the three mosfet switches.



pre- and postdetection

Let's assume we've put into the receiver i-f this approximately matched filter, yet the signal-to-noise ratio is not high enough to detect the cw carrier. Can anything be done to improve snr using additional processing? Yes, but the price paid for improved detection capability is a drastic increase in *time-to-detection*. The additional processing techniques are called postdetection integration.

Basically, they all involve reading the long-term value of the noise and thereby detecting small increases of power caused by the presence of extremely low-level signals. They are nonlinear systems, and the output signal-to-noise ratio is not proportional to the input signal amplitude.

The implementation of postdetection systems is always to rectify the i-f, rf, or audio (in a heterodyne receiver) and smooth the resulting dc in a low-pass filter. It's intuitively apparent that the longer the time constant of the filter, the easier it is to detect a small change in average signal plus

Note that the i-f signal-to-noise ratio appears as a square term in the equation, so that even very large values of integration times produce only a small reduction in permissible i-f snr. The time-to-detection can go clear out of sight before there is a significant output signal-to-noise ratio or improvement in sensitivity.

While the equation is exact, it has to be applied to a system with care. If the signal amplitude is not constant you can't put in a fixed number for the i-f snr. To be exact you must integrate over a period of time. The low-pass filter bandwidth in a system involving a recorder and a human observer is really the combination of the electrical filter and the brain of the observer looking at the chart.

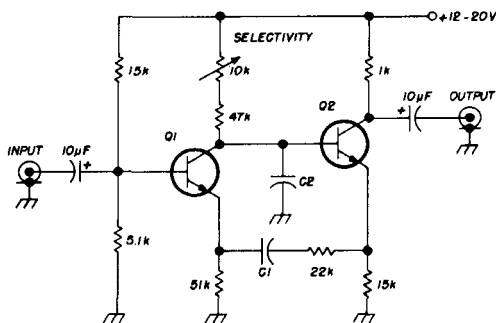
In a practical system the observer may be contributing most of the filtering.* I've found with systems of this type that there is little point in using integration times over 2 seconds on a cw carrier. With rapidly fad-

* And, depending on the observer, some of the noise.

ing moon echoes I prefer an integration time of 0.5 second, doing mental filtering on the chart output.

As a point of reference I estimate that the signal path contributes about 10 to 20 Hz of additional bandwidth on amplitude modulation. As an example, a cw system designed for highest sensitivity on moon echoes would use a predetection filter with bandwidth appropriate to the band used (20 Hz on 1296), followed by a postdetection in-

Regenerative audio filter. Transistors Q1 and Q2 are high-gain silicon npn units. C2—about 0.05 μ F—is one-half the value of C1 and depends on the desired audio frequency.



tegrator of between 0.1 and 0.5 second, followed by a recording device and the operator's mental filter. This is the optimum technique, and no fancy gadgets or trick circuits can do better than that.

practical considerations

In order to keep small drifts in receiver gain from affecting the recording, most people use two different audio filters and two rectifiers in a bucking arrangement working into an integrator, so that the output is normally zero in either filter with no carrier. Since the noise output from the two filters adds, while the dc component cancels, the snr of this system is 3 dB worse than that with the single i-f filter. This is well worthwhile, though, to obtain system stability.

If frequency-shift keying is used, performance is identical to a keyed carrier in a single filter. A less-popular technique is to

have a reference noise source at the receiver input and synchronously switch it and a pair of rectifiers at the integrator input to cancel receiver gain drifts. This is the technique used by radio astronomers. For cw use it's more difficult to implement and no better in sensitivity. (See fig. 1).

A disadvantage of the Dicke system is that it can't discriminate against radar interference. The two-frequency technique does discriminate, because the radar pulse adds to both channels equally if they have the same bandwidth.

predetection filter schemes

The two-meter moonbounce enthusiast might wonder how to implement a two-to-five-Hz bandwidth filter for the predetection (i-f) circuit. There are several possible methods:

1. Use a circuit similar to a Q multiplier, but at audio frequencies. This requires well-regulated supplies and a temperature-stable transistor circuit. Any transistor oscillator circuit will do if adjusted for class-A operation and if the feedback is adjusted to a point just before oscillation.
2. Use a passive RC circuit; with good components and regulated supplies, this circuit is capable of about 1-Hz bandwidth. Its operation is similar to that of the Q multiplier concept in method number 1.
3. Use a digital filter (shown in fig. 2). This is the ac equivalent of a synchronous detector. It's extremely stable and can generate an arbitrary narrow response. I've built some of these with 0.1-Hz bandwidth at 1 kHz. Their resonant frequency is that of the ac source driving them. Some care is required in layout to avoid coupling the oscillator output into the amplifier input. Their Q depends on the basic cutoff frequency; the lower the cutoff frequency, the higher the Q. This network can't oscillate by itself.
4. Use a synchronous detector followed by a low-pass filter. This is equivalent to another heterodyne process, translating the high audio frequencies to near dc, where they can be filtered without high-Q devices. This is a

very useful device if phase coherence exists between the drive source and the carrier. It is more difficult to use than method number 3.

Other considerations in choosing your predetection bandwidth are transmitter and receiver stability and the search problem with a too-narrow system. At 1296 MHz, 50-Hz bandwidth is probably as low as you can go and be sure the carrier is centered in the passband. At two meters a system of a few Hz bandwidth should be practical.

Fig. 3 shows a system that I am using on 1296 MHz. On moon echoes it is about 3 to 6 dB more sensitive than the human ear, with a 0.5-second integration time. **Fig. 4** shows typical moon echoes with a 0.5-second integration time.

pulsed systems

Let's now consider pulsed systems. Can anything be gained by transmitting a high ratio of peak-to-average power? The answer is a qualified "yes" for the relatively low power in amateur transmissions.

In general, pulsed and cw systems can transmit the same information rate if both are limited by white noise, and both use the same average power, provided each system is not limited by path a-m or fm. Both pulsed and cw radars perform identically with respect to maximum range and time-to-detection if both use the same average power.

However, for very high average power, transmission lines and antennas can't support the energy without breakdown. It's much easier to handle high average power if the peak-to-average power ratio is small. For these reasons new, long-range, high-performance radars use the cw mode. Range and range rate data are obtained by pulse compression at the receiver.

These radars are sometimes called "chirp radars." They transmit a "long" pulse of linear frequency. At the receiver the long pulse goes into a special filter that has a delay which also changes linearly with frequency to generate a short, high-peak power pulse. This overcomes the peak power limitations of the antenna and feedlines. These filters are very critical and are not yet within the realm of amateur work.

To show the equivalence of the cw and pulsed modes, let me go through an example. Assume two transmitters of equal average power output, no path modulation, and both transmitters operating for a 1-second interval. Allowable time-to-detection is one second.

Since we have an allowable "waiting time" of one second, corresponding to an information rate of one bit per second, the best pulse repetition rate is one pulse per second, because signal power is maximized in each pulse.

The optimum cw bandwidth for a 1-second "long" pulse is 1 Hz. (This is again the bandwidth of a matched filter.) It can be shown¹ that the shortest pulse length an amplifier can produce as useful output is equal to $1/B$ seconds, where B is the bandwidth in Hz. Conversely, the bandwidth for a pulse of any length, L , that will maximize the snr is equal to $1/L$.

For the pulsed transmitter, the bandwidth required for maximum i-f snr varies proportionally with the peak power and inversely with the pulse length. As the noise power increases in the i-f bandwidth, the i-f snr is constant for any pulse length between zero and one second. On the surface one might think that pulse transmission can offer a tremendous increase in snr for a given average power level. Short-duration, high-power pulses could be used in conjunction with a gate to remove the noise during quiescent periods, resulting in a high snr during gate-on periods.

In the real world, with path distortion due to a-m or fm, it's a different matter. If the receiver bandwidth is chosen on the basis of the path rather than the desired information rate (as in the above example) there is much to be gained.

For the 1296-MHz system, we arrived at a bandwidth of 50 Hz for optimum detection. On the basis of information rate, however, we could live with a system transmitting one bit per second. For these conditions, then, a system transmitting one pulse per second with a pulse width of 20 milliseconds would be about optimum. This system would have a 17-dB improvement over the cw threshold.

It may be easy to generate such long pulses, but there is also a question of legality for that amateur frequency. However, this does show what is possible. The mode would be legal on 2300 MHz and higher bands. The bandwidth of this system is no larger than that required for high-speed cw.

further enhancement of snr

The comments made earlier about post-detection systems still apply to pulse trans-

There is one other consideration with pulse transmission. The moon is of considerable size, and if a short pulse is sent up, the pulse will arrive at different points on the moon at different times. The reflected pulse is therefore stretched out in time. It turns out that the reflected pulse width is about 100 μ s, which corresponds to a reflection from the near parts of the lunar disc only. Also, there's a very much longer tail that corresponds to the full radius of the moon;

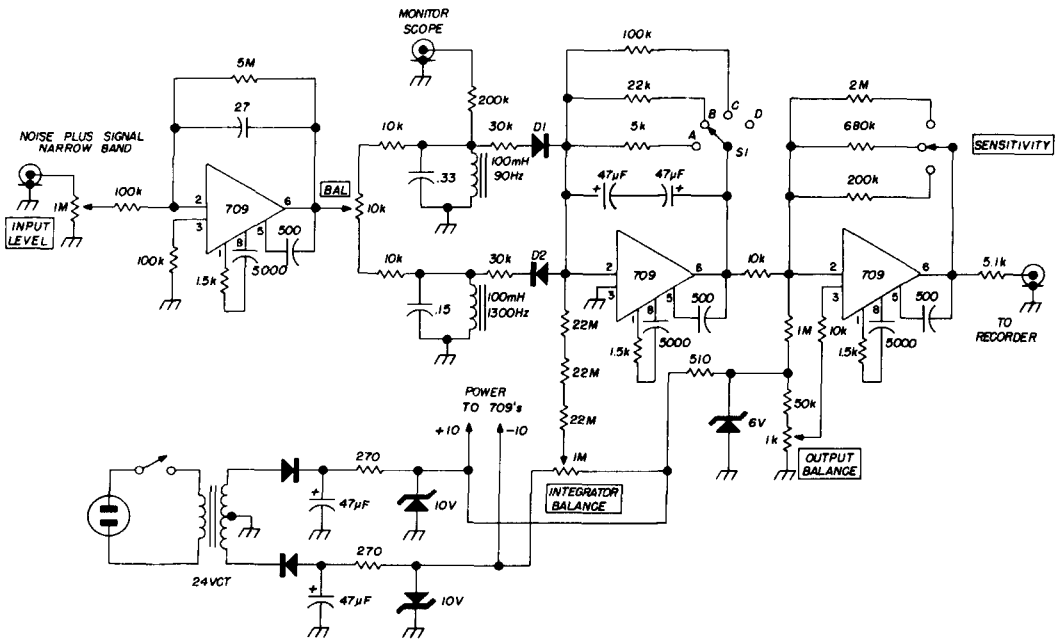


fig. 3. Versatile postdetection system used at WB6IOM. The positions of S1 are A, low sensitivity, fast response (0.1 second); B, high sensitivity, slow response (0.5 second); C, maximum sensitivity, slow response (2 seconds). D is the echo accumulator position.

mission. If the i-f snr is not high enough with the 17-dB improvement due to pulse transmission, the pulse can be integrated to enhance the output signal-to-noise ratio. It's desirable to gate out the noise between pulses, as those intervals don't contribute to the signal. The calculation of output snr is:

$$S/N_{out} = (S/N_{i-f})^2 \sqrt{N}$$

where N = number of pulses integrated.

however, the tail doesn't contain much power.

bandwidth reduction on moonbounce

It would seem, then, that a bandpass filter of about 10 kHz bandwidth at the receiver should result in optimum snr, but this isn't necessarily so. If the transmitted pulse has a width of 1 μ s, corresponding to a bandwidth of 1 MHz, the returned pulse may be ex-

panded in time without a corresponding reduction in bandwidth due to the incoherent nature of the moon reflection.

It's not clear, from any of the papers I've read on the subject, as to how much of a bandwidth reduction does occur, if any. If no significant reduction in bandwidth occurs, pulses less than 100 μ s wide aren't desirable since they actually force the receiver bandwidth up without the corresponding benefit of a short, high-peak power pulse to separate signal and noise.

If a considerable compression of bandwidth occurs due to the physical size of the moon, systems using short pulse widths may be as efficient as systems using longer pulses. There are a variety of planar triodes around that can generate power at good efficiency when pulsed as 2300-MHz amplifiers, so this appears to be a useful approach to low-cost power generation. It would even be possible to send short pulses spaced less than 100 μ s and get essentially cw returns. However, in the absence of hard facts relating to the phase coherence of the moon at these frequencies, this must await expert opinion.

voice transmission

The optimum transmission mode for voice signals against a background of white noise is still surrounded by a lot of confusing statements, even in the professional literature. However, properly defined and stated, it is a closed subject. For amateur purposes we are interested in a system that requires the lowest amount of signal power at the receiver for good voice intelligibility; not high

output signal-to-noise ratio. Some possible transmission modes to consider are single sideband suppressed carrier, a-m, narrowband fm, wideband fm, pulse code modulation, pulse amplitude modulation, pulse position modulation and others.²

If voice transmission systems are compared on the basis of peak and average power, ssb is the obvious choice. However, a-m is superior to wideband fm and equal to narrowband fm, on the basis of the same carrier power.

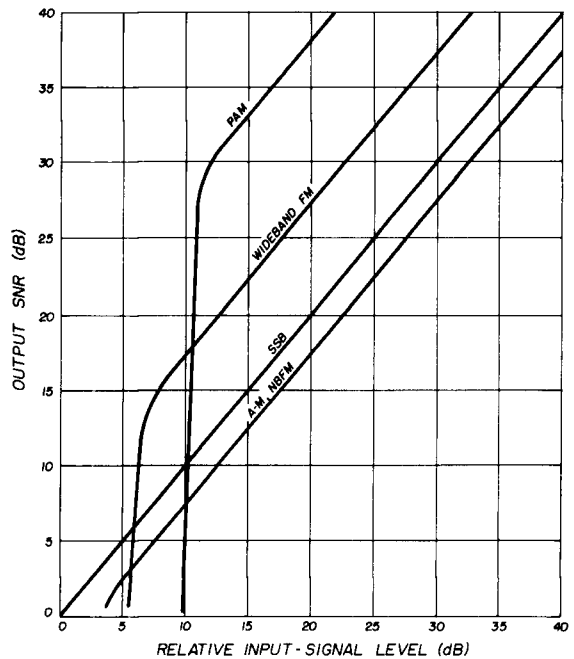
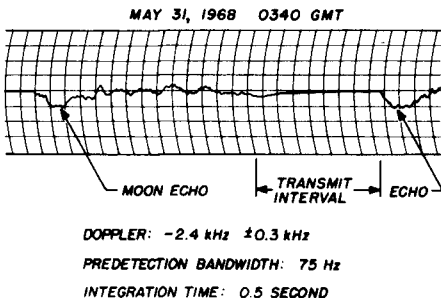


fig. 5. Signal-to-noise ratio characteristics of various modes of communication.

fig. 4. Chart recording of moon echoes received by WB6IOM on 31 May 1968.



The advantage of fm, PCM, PAM, etc., is that these systems have noise improvement properties. In all these systems, the output snr may be higher than that of the carrier. The price paid for this is a higher signal threshold below which there is no intelligibility. Fig. 5 compares the different systems qualitatively on the basis of equal input power to the receiver.

While the noise improvement property of

fm, etc., is important for data transmission and high fidelity entertainment, it doesn't help the voice intelligibility very much. On the graph of **fig. 5** the voice intelligibility is already near 100 percent before the fm or pulse curves result in a larger snr than ssb.

in summary

Whether you listen to weak cw signals by ear or use a postdetection system and detect the signal by eye, it's extremely impor-

black box contains several sophisticated computers, but it's not in the realm of anything as simple as, say, a couple of transistors.

By the way, it's quite difficult to build any system that's better than the human ear listening to an audio note in white noise. The ear acts like a tracking filter with a tracking range of about 3 kHz and an instantaneous bandwidth of 25 to 50 Hz. From about 200 Hz to 1 kHz the ear's sensitivity is constant and will detect signals with unity signal-to-

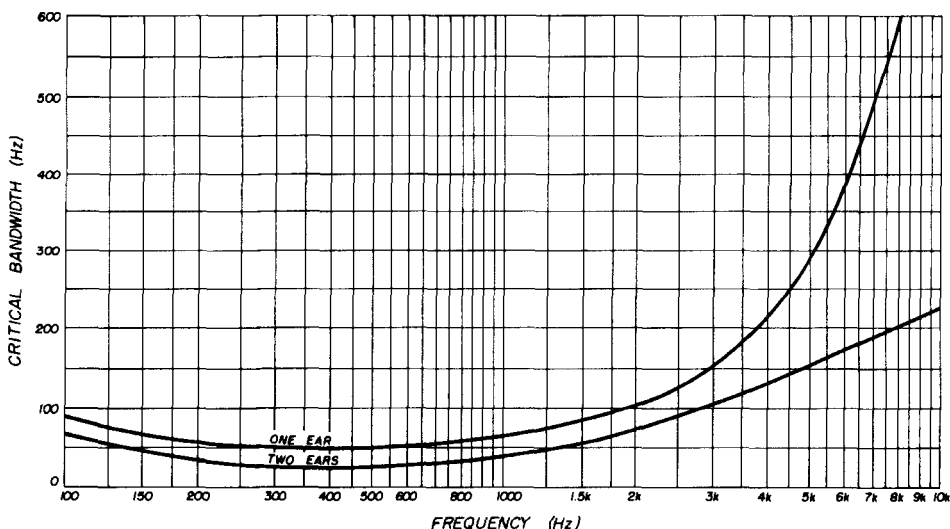


fig. 6. Critical bandwidth of the human ear as a function of frequency.

tant to keep the system linear. The human brain is a very complex signal processor that can adapt itself optimally to a variety of signals, but only if you don't make decisions for it with relatively unsophisticated devices like level thresholds, keyed oscillators, or clipping circuits. Avoid all chances of clipping in the system, and particularly avoid level thresholds as in some popular "black box" devices.

The philosophy here is simply this: if you put a circuit in your black box that makes decisions about the presence or absence of signals, then you have presented your brain with a problem about which it has no choice other than to reject or accept the black box decision. This is probably not too bad if the

noise ratio in a 25-50 Hz bandwidth (**fig. 6**). For signal detection of a carrier in white noise, about the best one can do is use a maximally flat (less than 1-dB ripple) 100-Hz to 1- or 2-kHz bandpass filter, with earphones or speaker that are also flat with frequency. Small resonances or peaks are quite distracting, as are some of the shenanigans now going on in the sub-bands reserved for extra class amateurs who are trying to work DX on these frequencies.

It's a good idea to check your ear for its sensitivity peak, because this peak can vary from about 200 to 600 Hz, depending on your age. Narrowband audio filters can be very tiring on the ear, because signal and noise begin to sound alike after an extended

period of listening (again depending on age of the listener). These filters are really no more sensitive than flat, wideband filters.

If you are an experienced weak-signal operator, you'll also have noticed that the brain can make a better decision about the presence of a signal if the signal's frequency is changing very slowly. (Any DX'er will certainly recognize this as a subtle phase shift between signal and noise; it requires extreme concentration, however.)

Audible detection of a signal, and copying it in the form of code, are two different things. The signal has to be at least 3 dB above threshold and peaked at your particular aural resonance point before you can decipher intelligence. This is because the minimum change in signal strength detectable by the human ear is 3 dB. The advantage of the postdetection scheme is that you can get positive copy, whereas the ear can just barely detect the presence of a carrier.

Another point to remember, when using ear filtering, is that the bandwidth of the ear is very dependent on volume; maximum sensitivity is usually at a relatively low volume when copying ssb and cw signals.

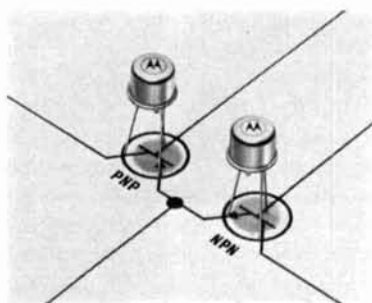
Carrier and cw detection by means of the ear are almost as sensitive as the best post-detection system known. They have the advantage of being much less sensitive to frequency instability and tuning errors. However, full sensitivity can only be realized if the system is treated with about the same care as a sophisticated high-fidelity receiving system. A good system should have separate volume adjustment for each ear, some indication of noise level for each ear, and at least two filters of 400 Hz and 2 kHz bandwidth. Optional would be adjustable frequency shaping to compensate for the different response of the two ears.

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ham radio

large signal vhf/uhf transistor



Motorola has just made the first silicon PNP large signal vhf/uhf power transistor available to the electronics industry. Designated the 2N5160, it is the forerunner of several PNP transistors intended to complement existing NPN devices. The 2N5160 was designed specifically as the PNP complement

of the NPN 2N3866 for use in complimentary circuit configurations such as vhf and uhf amplifiers.

This new transistor is very helpful where a positive ground must be maintained and eliminates many problems with bypass capacitors. Other uses for these power transistors are in unsaturated switch applications at higher current levels than are presently available with PNP switches. This transistor features a minimum power gain of 8 dB with no emitter tuning and 1-watt minimum power output at 400 MHz. It is suitable for use in class A, B, or C output, driver or oscillator applications at frequencies up to 800 MHz. The 2N5160 uses the multiple emitter overlay geometry and is available in the TO-39 package. For further information and complete specifications, write to Motorola Semiconductor Products, Inc., P. O. Box 13408, Phoenix, Arizona 85002.

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converting a vacuum-tube receiver to solid state

You
can reduce drift
in older receivers
by an order
of magnitude—
a modification scheme
for the BC348
using MOSFET's
and JFET's

The BC-348 receiver is well known for its excellent selectivity and mechanical stability. At the same time, this old iron box can be an instrument of torture for the user. Some of the earlier BC-348's were bedeviled with the "scratchy dial" syndrome. This is a phenomenon of unknown origin and is virtually impossible to cure. It appears when the receiver is tuned and sounds like a thousand demons rubbing a tin washboard. If you're unfortunate enough to have a BC-348 with this ailment, and the receiver has a drift problem, it can be a real chore indeed to keep a signal in the i-f passband.

In resuming an interrupted conversion of the BC-348 begun a dozen years ago, I decided the only way to reduce frequency drift was to reduce internal temperature rise. There are several ways to do this. Reducing B plus takes off seven watts; a transistorized audio circuit eliminates another nine. To reduce drift by ten times, however, the power consumption must be cut by 90 percent. Answer: **all** transistors.

If the conversion includes a class-B audio amplifier, drain under normal conditions would be well under ten watts, and two watts of this would be in the pilot lights. Fine. Now the question is, how to do this without sacrificing any of the features this receiver is noted for?

Hank Cross, W100P, 11 Birds Hill Avenue, Needham, Massachusetts 02192

choosing replacements

The BC-348, -312, and -342 receivers use rf coils with low-impedance plate windings (or taps) and high-impedance grid windings. Bipolar transistors might be used by turning things around (interchanging grid and plate), or junction FET triodes might be used in a cascode circuit, as in the H. H. Scott fm tuners—two JFET's replace one 6K7.

However, the RCA dual-gate MOSFET's,

tiometer was ganged to the tuning capacitor. I felt that this was an added complication not needed for amateur work, so I cut the wires going to the pot and those from the pot to the bandswitch inside the coil box.

Tubes I replaced with FET's were the 6K7 rf amplifiers, a 6J7 mixer, and 6C5 local oscillator. Characteristics of these tubes at 18 MHz are given in **table 1** with those of some replacement semiconductors. Problems oc-

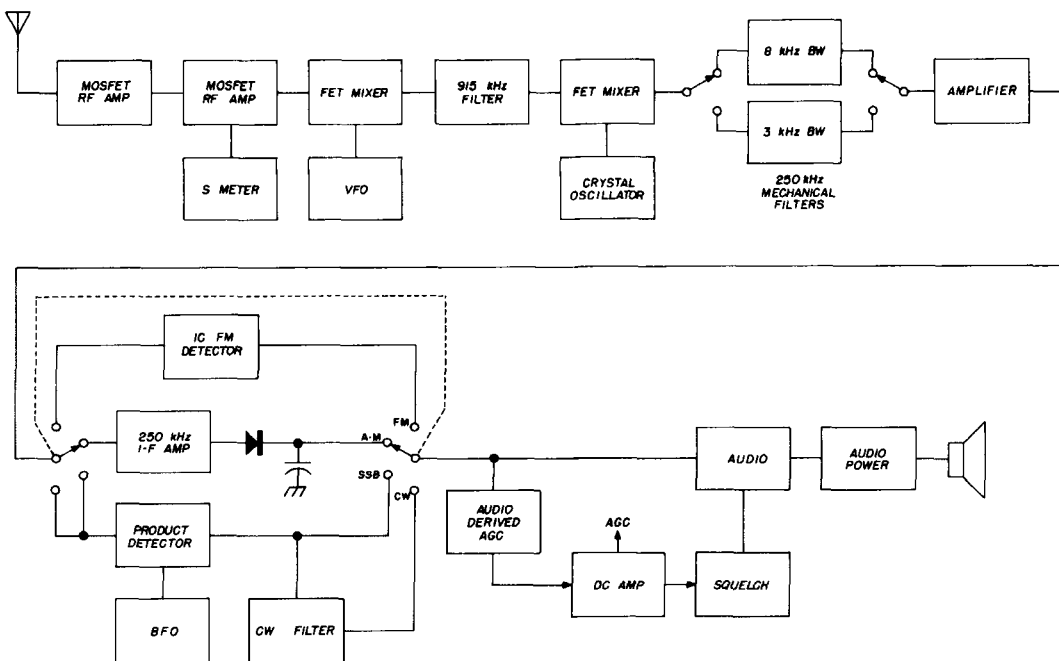


fig. 1. Simplified block diagram of the solid state BC-348 receiver. In the actual receiver, the audio-derived agc circuit is switched out of the system in the a-m mode.

recently put on the market at a reasonable price, have the correct input and output impedances for this use and afford a gain control with lower distortion products. It turns out that the 3N140 MOSFET, at rated current, has about five times the transconductance of the 6K7. The relatively low output impedance of the 3N140 (**table 1**) is not objectionable when used with low-impedance inputs.

Because of the interstage coupling, stage gain in the BC-348 receivers increases from the low to the high end of each band. In some receiver models a gain control poten-

tiometer was ganged to the tuning capacitor. I felt that this was an added complication not needed for amateur work, so I cut the wires going to the pot and those from the pot to the bandswitch inside the coil box.

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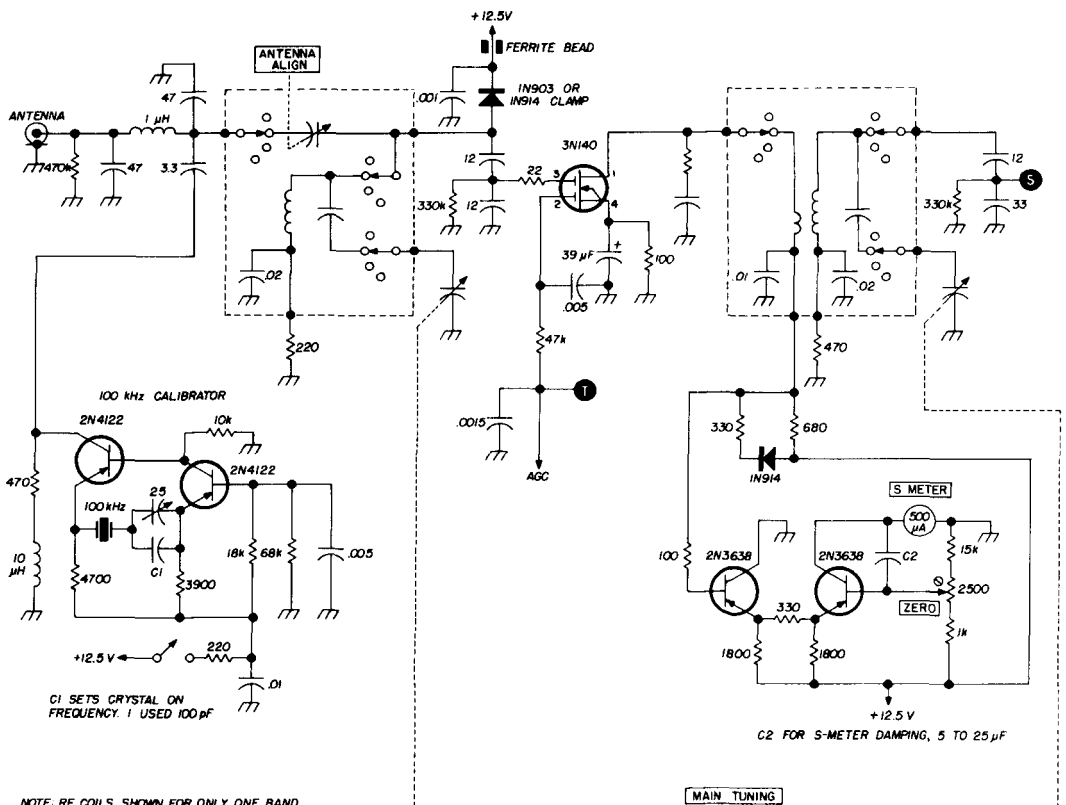


fig. 2. Schematic diagram of the solid-state BC-348. Coils in the rf section are shown for one band only; band-switch shaft end should be grounded with a spring to remove 144-MHz birdies. Resistors are $\frac{1}{4}$ -watt composition unless otherwise marked; 1% resistors are $\frac{1}{8}$ -watt metal-film types (RN55 or RN60). Polarized capacitors less than $25\text{-}\mu\text{F}$ are tantalum types (150D or CS13); aluminum electrolytics are ok if not used for agc or audio coupling.

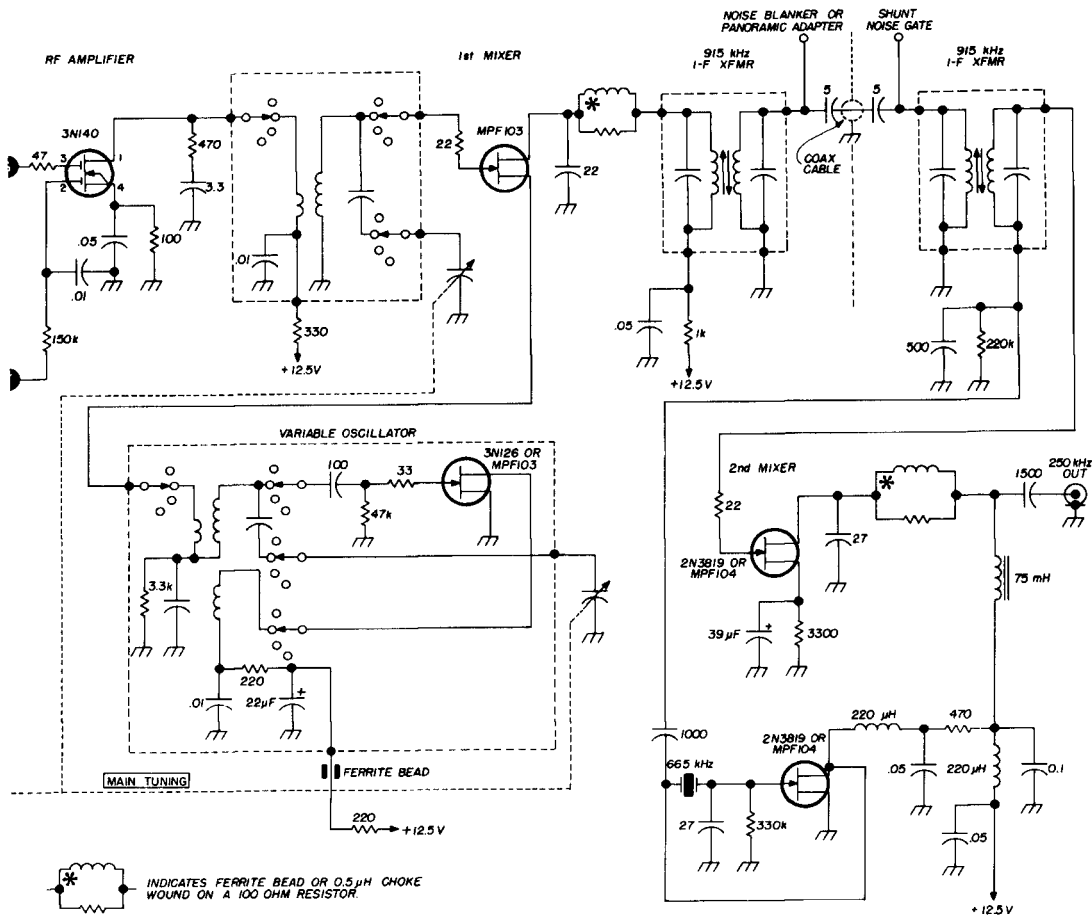
at the input of each 3N140. Because the semiconductor triode noise is lower than that of the 6K7, the measured noise figure of the modified receiver was somewhat better than the original, even though there wasn't always a peak in noise as the antenna trimmer was turned.

Gain control is provided by changing the bias on the second gate. Insulated gates don't draw current, but the control voltage should be +4 V for maximum gain, and complete cutoff is obtained at -3 V. The first gate is returned to ground. There's no chance of rectified current causing the whole receiver to block when the rf stage conducts (a common cause of receiver overload).

A triode FET can replace the 6J7 mixer. The high feedback capacitance is a minor problem so long as the input and output circuits

are tuned to widely different frequencies. The conversion gain is a function of the available oscillator drive power. With enough injection the mixer gain could be ten times that of the original. To replace the 6C5 oscillator a low-transconductance triode FET is needed, as discussed below.

I went to double conversion to obtain better selectivity. If 455-kHz filters were to have been used, I would have moved the first i-f to 930 kHz or higher in order to avoid two intermediate frequencies in harmonic relationship. Of course the second oscillator would have been put on the high side. (Think where the image would be if it were on the low side.) However, I had some 250-kHz mechanical filters on hand. The point of this conversion was, after all, to use the six-band, twenty-four-coil tuning appara-



tus of the original receiver with as little modifications as possible.

Fig. 1 is a block diagram of the complete receiver, including auxiliary features (to be covered in another article). Everything fits inside easily. Fig. 2 is a schematic of the main rf and i-f circuits.

input circuit

The antenna input starts out with a low-pass filter, with nominal cutoff at about 30 MHz. Aside from obvious advantages if you are near strong TV stations (I am), there are also less "birdies" when the receiver is used with a vhf converter. Because of the input coupling method in this model receiver, the filter is especially important.

There is a protective diode clamp on the antenna coil. 6BA6 rf amplifiers will stand

several hundred volts swing without damage. If the grid return resistance is high, however, the diode is essential because the insulated gate FET is rated for only a few tens of volts peak. A type 1N914 of any manufacture will have low loss at 12 volts of reverse bias. The added capacitance is about 1 pF.

The change of drain (plate) current in the first rf stage is used to indicate signal strength. Because of the slope of the 3N140 control characteristic, the meter scale is non-linear in dB, even with the diode in the circuit. Other approaches, including a separate signal indicator channel, were less satisfactory.*

* The use of an s-meter as a quantitative measuring device for received signals is open to question, but an s-meter is fine as a tuning indicator. Editor

local oscillator

The local oscillator FET socket is supported by its leads, which pass through the keyway from the octal socket pins. The 6C5 ran at around 90 volts, with a starting g_m of less than 2 millimhos. I tried various adaptations, and found that "hot" FET's (those having high I_{dss} * and a high zero-bias g_m) were noisy and unstable, especially at about 15 MHz.

Good results as a local oscillator were obtained with the Motorola 3N126 and some MPF103's. The 3N126 was used with the second gate (substrate and case) hooked to source. If a MOSFET is used, something like a 1N34 would be needed to generate "grid-leak" bias.¹

In the original oscillator the grid leak was 100k ohms. In the course of chasing troubles I changed the value to 47k, but 100k probably would be satisfactory. The 15k mixer cathode resistor, also inside that box, probably would be satisfactory without change, although mixer FET's having very high I_{dss} would be more suitable in that case. The 3.3k resistor I used was based on experience with the breadboard model of the second converter. The 33- (or 47-ohm) resistor in series with the gate lead of the oscillator can be put in between the octal socket (pin 5) and the transistor socket.

mixer

The mixer proved to be rather critical as to the FET that was used. For type MPF103 the I_{dss} range is 1 to 5 mA; for the 2N3819 it is 2 to 20 mA. I found that anything be-

tween 1 and 4 mA worked in my receiver, but the higher-current units gave a bit more gain. Still higher I_{dss} units had low gain and were noisy in this socket. Other places in the receiver are not so finicky in this respect. I'm not worrying about spares: if the FET's last as long as 6SK7's, I'll need another one in 1990.

In a double-conversion receiver, there should be as much selectivity at the first i-f as other considerations permit, and as little gain between the first and second mixer as possible without compromising noise figure (but in no case less than unity gain, or the first stage will overload before the second). A few calculations show that three very high-Q tuned circuits will reject the nearest spurious response (789 kHz), but four tuned circuits of moderate Q will do even better.

I could have used two of the original cans, but I rewound a couple of miniature J-trans to save space. (For details on how to adjust the coupling using a Q-meter, see reference 2.) The "tee" of capacitors between the cans is used in setting up the gain, as the coupling factor between pairs is much less than critical. The shunt capacitance is that of the coax running from the rf board to the second converter module.

second conversion and detector stages

The second mixer and oscillator are built on a small plate that fits in place of the original output transformer. The values were juggled to get best operation with the FET's and crystal I had. With my setup, the oscillator drain swing was about 9 V p-p, and the mixer developed no rectified bias across the 220k resistor.

Strong second oscillator harmonics were found at the mixer output and on the B-supply lead beyond the feed-through capacitor. The additional rf choke and capacitor fixed the power-lead leakage, while the natural attenuation of the mechanical filter is enough protection for the output harmonics,

* I_{dss} is a common classification parameter, measured by shorting the gate to source and applying 5 to 10 V to the drain. Thus by definition, I_{dss} is the drain-to-source current when the gate-to-source voltage is zero.

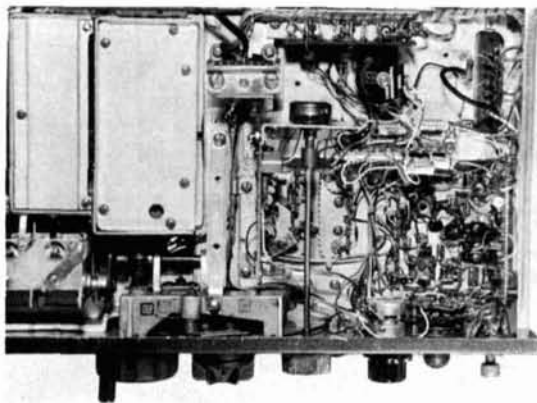


table 1. comparison of tube and semiconductor characteristics.

Type		V_B (volts)	I_B (mA)	g_M (millimho)	C_{FBK} (pF)	R_{in} at 18 MHz	R_{out} (ohms)	R_{eq} (noise) (ohms)
6C5	tube	80	6	2	2	50k	10k	—
6J7	tube	100	2	1	0.005	60k	800k	10k
6K7	tube	100	9.5	1.6	0.005	50k	150k	15k
2N2084	padt	12	1	34	2.0	800	40k	200
3N126	jfet	12	2	1.4	0.5	25k	30k	—
3N128	mosfet	15	2	4	0.2	100k	80k	—
3N140	mosfet	12	10	8	0.02	100k	12k	<1k
MPF103	jfet	12	2	3	2	100k	50k	<2k
40481	npn	7.5	2	64	0.2	1500	50k	200

if the filter input terminals, etc., are well shielded.

Second mixer output is parallel-fed via a choke and blocking capacitor to the filter switch. Either the 3-kHz or the 8-kHz band-pass filter may be selected, the other one having its terminals shorted.

The output section of the switch (8 pole, 5 position) goes to a feedback amplifier that has a high-impedance input and a gain of ten, followed by a potentiometer that permits the background noise level to be adjusted without modifying the agc action. The range of the control is 23 dB.

The stage following the gain control is the third 3N140. It is coupled by an auto radio i-f can, padded down from 262 to 250 kHz with additional 22-pF capacitors, to the product detector and to an i-f power amplifier that drives the diode detector. The two i-f cans shape the top of the passband in the 8-kHz bandwidth a-m position to get the best weak-signal a-m reception. (This seems to want a rounded top, symmetrical, down about 6 to 10 dB at 3-kHz off center.)

The front-panel switch selects five combinations of bandwidth and detection, as

per table 2. The agc source for a-m is the diode detector, but for sideband and CW reception agc is derived from rectified audio, so that the effect of audio selectivity is included.

mode-selection circuit

The mode switch is shown for 8-kHz bandwidth a-m reception, with agc in operation. The second-gate control voltage on the three 3N140's varies from plus three or four volts with no signal to negative two or three on very strong signals. The loop gain is high, giving a dynamic range of 40 dB. The first volt of bias change makes very little change in receiver gain, but the squelch operates reliably at that point.

agc/mvc

In the gain control circuit the two dc amplifiers, one to handle agc and one for manual control, are hooked to a transistor that supplies -4 V to the emitter returns and which, incidentally, provides temperature compensation—a sort of three-transistor differential amplifier.

A negative voltage supply makes the am-

table 2. Mode switch setup.

function	FM-8	AM-8	mode AM-3	SSB	CW	REMARKS
filter (kHz)	8.5	8.5	3	3	3	4 - poles
200-Hz audio filter	no	no	no	no	yes	1 - pole
agc	car	car	car	af	af	1 - pole
B+ to SSB and bfo	off	off	off	on	on	
audio from	fm	a-m	a-m	prod	prod	1 - pole
B+ to fm det	on	off	off	off	off	1 - pole for B+

notes:

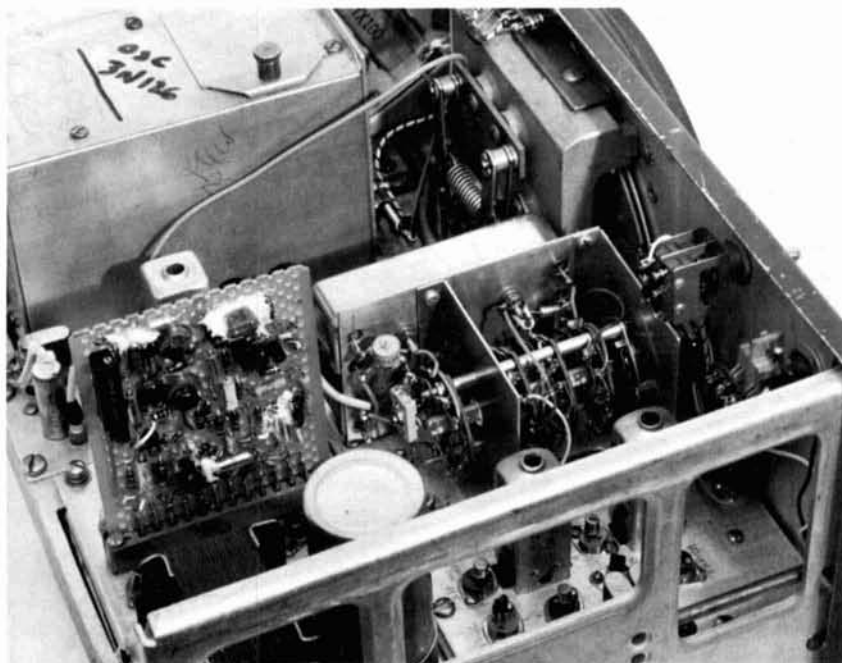
1. SB1-SB2 (SB1 is lower 40-75, upper, 20).
2. Agc-off-mvc functions similar to original.

plifier practical and is a handy way to make dc level changes without extra stages. Also the audio stage operating point is centered automatically.

The audio-derived agc voltage is from a little three-transistor feed-back amplifier (with automatic temperature compensation; again, this is from the pnp follower driving the npn stage) and a voltage-doubling rectifier. The attack time is fast for small changes, but the maximum effect from a single pulse

there would be worse thumping at the beginning of a sideband transmission. If the action is too fast, there may be a stability problem or there may be a tendency for the receiver to block on a burst of QRM or ignition noise.

For fm reception, a Sprague integrated circuit is used, type ULN-2111A. This has a bunch of emitter-coupled clipping amplifiers, a rather fancy six-transistor phase detector, and various diode bias networks. All that



Chassis view of the revamped BC-348 shows the new mode switch and two mechanical filters. The power supply and audio circuits occupy the old dynamotor space; the plug-in board contains voltage regulating and low-level audio circuits.

is limited by a 0.1-microfarad capacitor that couples the amplifier to the rectifier, while the rectifier has fifteen microfarads to charge up.

For 1-kHz CW tones, this means that 1 Hz makes a maximum difference in charge of the 15- μ F capacitor of 10 V times 0.1 divided by 15, or 0.067 V. However, the gain of the dc amplifier after that is considerable, and gain changes of 20 to 40 dB/millisecond can be obtained. If action were much slower,

is required outside is a single-tuned circuit. (A typical transistor radio i-f can is suitable; just use the high impedance winding, padded to frequency with external capacitors and some by-pass capacitors.) The output in my receiver swung three volts peak-to-peak for a signal moving across the 8-kHz filter. When the signal was injected after the filter, an s-shaped curve about 18-kHz wide was obtained, with an amplitude of more than five volts peak-to-peak.

For manual gain control (I happen to like to dive for the same volume control knob whether agc is on or not), the agc circuits are still operative to prevent blocking on unexpectedly strong signals. The audio gain is high enough so that normal operation is in a region where the agc threshold is not exceeded. Because of the highly nonlinear control function (dB versus control volts), tapered sections are used for both audio and manual rf gain.

diodes. However, 1N914's could be used in place of the 1N457's as well.

The 2N2925's and 2N2924's were actually the green- or yellow-coded 2N2926's, while the brown- and red-coded 2926's were used for bias compensation diodes, labeled "BD." You get 25 assorted for \$7.25.** You'll be able to use 20 if you're crazy enough to copy this receiver.

There are also two 2N3638's, three 2N4122's (two in the calibrator, which is

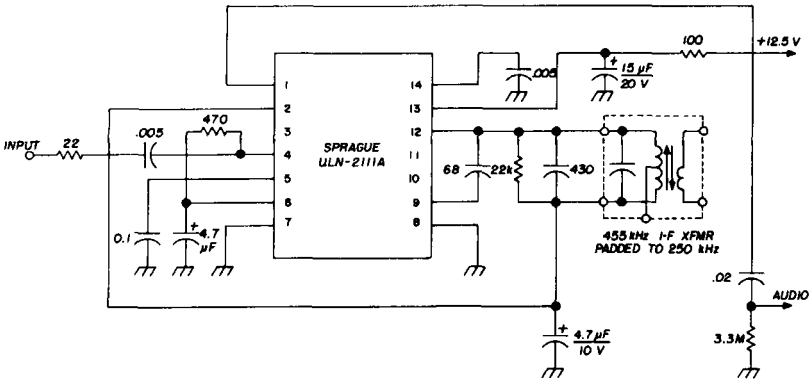


fig. 3. Integrated-circuit fm detector using a Sprague ULN-2111A. This circuit provides 3 volts p-p output with 1 millivolt input.

I found that a dual 100k audio taper unit was satisfactory; the value could be as high as 250k and as low as 50k with a few minor adjustments. A stereo amplifier control should be adequate. The **reserve gain** pot in the 250-kHz i-f amplifier could have either a linear or audio taper such as used in transistor radios.

The rest of the controls (for agc thresholds, product detector bias, manual volume control "off" bias,* and the power supply voltage setting) can be any handy trim pot, either surplus from computer cards, or small trimmers such as the IRC 2CI or Mallory MTC.

bfo circuit

High-speed computer diodes such as the 1N903 or 1N914 are used for the input voltage clamp and the bfo crystal switch; I used 1N457's or 1N629's for other spots, as I had them, and I felt that there was less chance of generating harmonics with the "slow"

mounted on the rf panel near the input) and six pnp's, which could be either type. The first type costs fifty cents; the second eighty. (The whole count is 50 transistors and FET's, 16 various diodes, and one integrated circuit.)

audio section

After the mode of reception is selected (with or without narrowband CW filter), the signal goes, via the audio gain pot in the agc position, to a source-follower first audio stage, an active low-pass filter, then to a complementary class-B output stage. The squelch is in the active filter section. The circuit is

* Minimum gain setting must be at minimum gain bias.

** Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order stock number 49F3-2N2926. Kit of 25 is \$7.25 plus postage.

similar to one previously described.³

The power regulator, power transformer (Triad F-40X 26 V, 1 A center-tapped), squelch control, and all except the first audio are mounted in the space vacated by the dynamotor. The high-power transistors are on the subchassis (insulated by mica washers), and the low-power circuits are on a plug-in card (very easy to pull out and change).

selectivity filter

The CW filter is switched into the sideband channel for more selectivity when needed. The values seem reasonably satisfactory in practice, although the filter performance has not been measured in place in the receiver. Any other narrowband audio filter of ten or twenty thousand ohms impedance should do as well. Obviously, an extra amplifier stage could be put in the chain if the filter on hand was of lower impedance, as the method of switching makes this practical. The audio agc is based on what passes through the filter, so it can be used even under quite rough conditions. Because the squelch can also be used in this mode, I have been able to find a fairly weak (though **stable**) signal and park on him with the receiver on squelch, waiting to get a chance to work him.

crystal calibrator

The crystal calibrator is almost a necessity in a wide-range unit such as the BC-348. It appears that there would be some benefit in making a *more complex calibrator*, such as a 200-50-10 kHz type. Higher output than I have is desirable when running some converters, but I haven't done anything about it as yet.

in retrospect

The main problems with transistors almost always stem from overloading and temperature variation. Resolving the overload problem takes effort: care to see that the last i-f amplifier, for instance, is capable of driving the second detector hard enough, so that the second detector can generate enough agc voltage, so that it won't be driven too hard—like a tight servo system. (The more I read that the more I wonder if the English language is up to today's problems.)

What all this means is that the last i-f stage should be designed for best power output, not best gain. Also, it means that agc can't be used on the last i-f stage. (Auto radios with **one** i-f stage have some rather fancy solutions to this problem.)

Another device to reduce overload is feedback. At "low" frequencies (under a couple MHz for today's transistors), feedback will hold gain and bias constant despite temperature variation. Distortion will also be kept low.

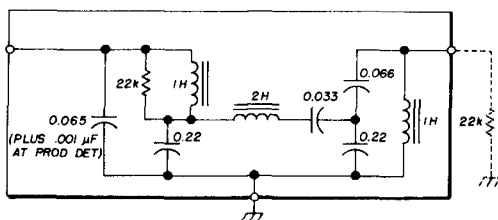


fig. 4. Audio CW filter with nominal 700-Hz center frequency and 160-Hz bandwidth. Inductors are Triad epoxy-molded toroids with Q of 50 to 80 at 1000 Hz. Capacitors are mylar insulated.

In many places where dc levels are critical, differential amplifiers are used. The drift with temperature of one transistor is balanced almost exactly by another of the same type. In other circuits, a diode-connected transistor (or any diode) is used as part of the dc bias circuit, so that temperature effects of the transistor are pretty well cancelled by the diode as it conducts a few mA in the forward direction. In class B or AB circuits this is particularly important. For best results, the bias diodes must be placed physically so that they are at about the same temperature as the power transistors.

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for

single sideband

A power supply's a power supply. Or, so it seems. But not all of them suit all purposes. Sometimes extra demands are made on power supplies for single-sideband equipment—such as wider regulation, multiple voltages, cleaner filtering, and so on. That's why these dc supplies deserve special attention.

First, though, it's just as well to review the characteristics common to all power supplies. What are they supposed to do, and how do they go about it?

Mainly, the power supplies in ham equipment furnish dc voltages to operate tubes and transistors in the stages. Circuits may need positive voltages or negative voltages, high voltages or low voltages, high currents or low currents. The power supply generally converts 117-volt ac power to the needed dc voltages.

A power supply also furnishes certain specific ac voltages. The power transformer usually develops them with an extra winding or two; you don't even notice them. Without them, though, tubes wouldn't get proper heater voltages—and sometimes relays wouldn't work.

getting the voltage you want

High-power ssb transmitters naturally need much higher voltages than those

needed by, say, a receiver. Transistor equipment needs lower voltages. There are three chief methods for getting the exact voltage needed by each stage of a transmitter or receiver.

The first way is by transformer. By specifying transformer windings with certain turns ratios, the ssb-equipment designer gets a range of voltages. After the ac voltages are at the right values, they can be changed to dc.

A step-down turns ratio of 117:6.3 supplies the ac voltage for most parallel-wired tube heaters. If a special transmitter tube needs 5 volts or 10 volts for its filament or heater, a winding ratio of 117:5 or 117:10 is used. For higher voltages, set-up ratios are used. For 350 volts, a turns ratio of 117:350. For 1000 volts, a ratio of 117:1000.

As the transformer schematic in **fig. 1A** suggests, any reasonable number of windings can be included in a single housing. Also, as **fig. 1B** illustrates, one winding can be tapped to provide several voltages. The transformer method is expensive, but it's a practical way to step the power-line voltage either up or down.

The second way to get specific dc voltages from a power supply is by **resistive divider**. This method is used after the volt-

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age has been changed from ac to dc. A drawback is that you can only step voltages down—not up.

The power transformer steps the power-line ac up to the highest voltage desired. After the ac is changed to dc, the dc voltage is fed to a series of resistors; the voltage divides across the resistors, in proportion to the resistance values. An example of this appears in **fig. 1C**.

The third way to get the voltage you want is by **voltage doubling**. This is a trick that builds voltages up in value, but can't step them down. Voltage doubling is done

before the dc is fit to use, the ac must be removed; electrolytic capacitors and filter chokes do this. The output of a rectifier-filter combination is relatively pure dc. A simple example appears in **fig. 2**.

The dc output voltage depends mostly on how much ac is applied to the rectifier. The line voltage can be applied directly, as it is in **fig. 2**, or, ac can be applied from a stepup winding of the power transformer to make the dc output voltage greater. Ac can also be applied from a stepdown winding, as it is to develop low-voltage dc for powering transistors.

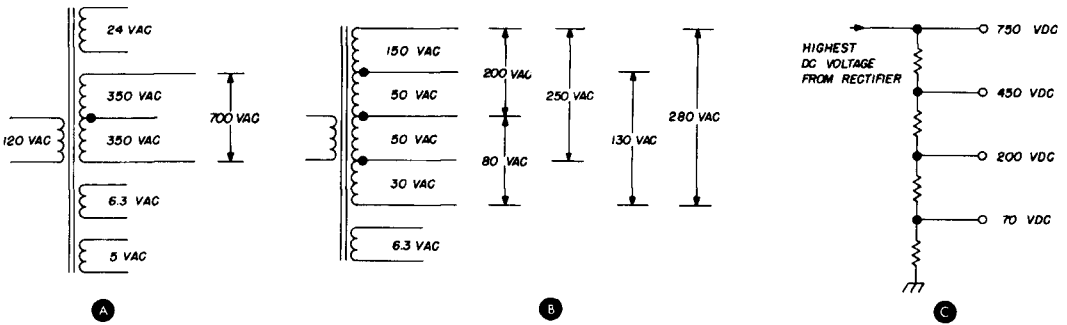


fig. 1. There are several ways to get the voltage you need; by selecting the power transformer, by using a tapped secondary or with a resistive voltage divider.

in the rectifier part of the power supply. The result is a dc voltage that's twice the value an ordinary rectifier would produce. Where especially high voltages are needed, tripling and quadrupling can be arranged.

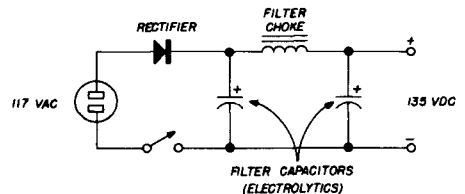
In commercial ssb equipment, you'll find combinations of these three ways. Each has certain advantages, and in combinations they are versatile. You should consider all of them for homebrew ssb gear.

from ac to dc

Ac power is changed to dc power by rectifying and filtering. A tube or semiconductor is the rectifier. Nowadays, semiconductors are more popular. The rectifier conducts on half of each ac cycle and blocks on the other half. The output is pulsating dc—which is dc voltage with a rather large ac component still riding along. Be-

fore the dc output is usually slightly higher than the measured ac voltage applied to the rectifier. The output of **fig. 2**, for example, is about 135 volts dc, although the input is only 117 volts dc. That's because the output depends on the peak value of each cycle, whereas the measured ac voltage is rms or effective value.

fig. 2. Simple half-wave rectifier circuit with capacitive input filter.



Rectifiers are broadly categorized as half-wave and full-wave. A half-wave circuit is shown in **fig. 3A**. Half-wave rectifiers waste one half-cycle of the input waveform. Both halves are applied, but only one can pass. In **fig. 3A** it is the positive half, because of the direction the rectifier diode is connected. The filter capacitors and choke smooth out the deep ripples, leaving a positive dc voltage with almost no ac in it.

A full-wave rectifier is shown in **fig. 3B**. It requires a center-tapped transformer winding; the center tap is grounded in this version.

Considering the whole secondary winding, each positive half-cycle makes the top rectifier conduct and each negative half-cycle makes the bottom one conduct. With respect to ground, however, both rectifiers receive a positive half-cycle. The combined output of the two rectifiers is a series of positive half-cycles. The full input wave is used, instead of just half of it. The filters make the output nearly pure dc.

A bridge rectifier circuit is shown in **fig. 4**. This one is a full-wave system because it uses both half-cycles. Compare this transformer with the one in **fig. 3B**, and you can see why the bridge circuit is popular. The 280 volts dc output is obtained with a simple 250-volt transformer instead of a center-tapped 500-volt unit.

One more circuit that converts ac to dc is the doubler. There are two common kinds, both shown in **fig. 5**. The one in **fig. 5A** is a half-wave doubler. Consider the bottom of the secondary as a reference point. The first negative-going half-cycle charges C1 through D1; D2 can't conduct because the voltage is in wrong polarity. Then, the positive half-cycle that follows charges C2 through D2. The previously stored charge on C1 is in series, and it adds to the positive half-cycle. The charge

that develops across C2 is therefore double what it would otherwise be. This doubled charge is applied to C2 with each positive half-cycle; the negative half-cycle doesn't affect C2 at all—only C1. The output is therefore a half-wave pulsating dc voltage almost double the peak input ac voltage.

The full-wave doubler in **fig. 5B** uses both half-cycles. Each positive half-cycle charges C2, because D2 can conduct. The negative half-cycle that follows then charges C1, because D1 can conduct. Each half-cycle of the input wave therefore

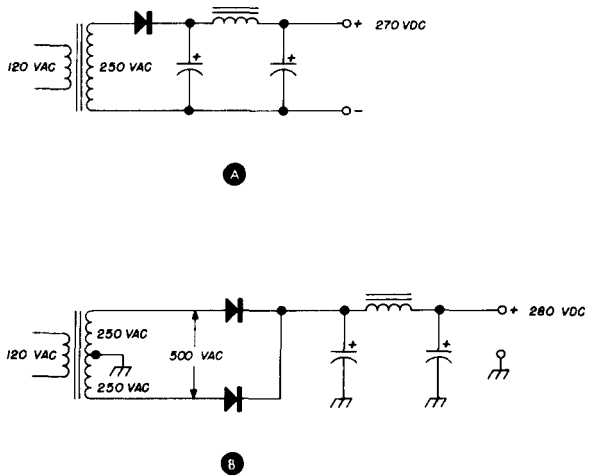


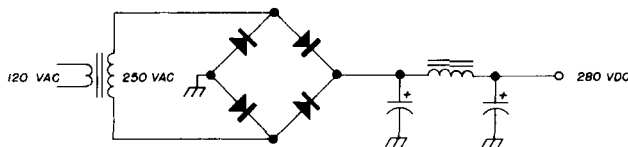
fig. 3. Two types of rectifier circuits—half wave in **A** and full wave in **B**.

charges one capacitor or the other. The output is taken across the two capacitors in series, so their voltages are added. The output is a full-wave pulsating dc voltage that's about double the peak input ac voltage.

the current you need

Transistors don't require much dc current in most uses. Tubes used as small-

fig. 4. The full-wave bridge rectifier.



signal amplifiers need more; tubes for receiver power amplifiers need still more; tubes in transmitters need even more; and transmitter power tubes need plenty. Single-sideband linear power amplifiers don't use as much average current as their a-m or cw counterparts, but their demands during modulation peaks are even greater.

The range of current that single-sideband power supplies must furnish is wide, then. Three factors about a dc power supply determine its ability to deliver current without overheating or damage: the trans-

be drawn from the power supply as current. If the filters are small and the current drain is heavy, the pulsations of dc don't have time to "fill them up;" the current keeps being drawn out too rapidly. Larger capacitors store more electrons, and current demands don't deplete them between half-cycle pulsations.

Full-wave rectifier circuits supply more current than half-wave circuits. Their pulsations are twice as frequent as those of half-wave rectifiers, so they keep the filters charged up better.

Voltage doublers reduce the current a rectifier can supply. Double the voltage, and you halve the current that can be drawn safely. From a tripler, only a third as much current is available as from the same set of components in a simple rectifier and filter.

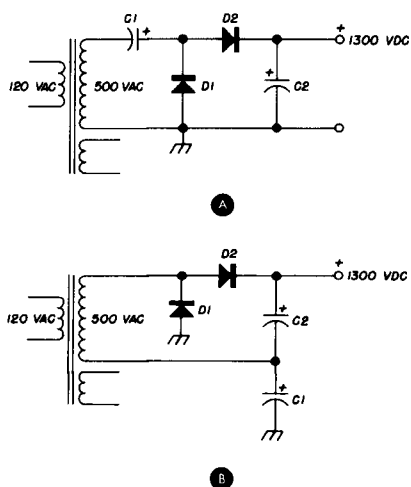


fig. 5. Here are two circuits which rectify the ac voltage and double it at the same time.

former-winding current rating, the rectifier current rating and the size of the filter capacitors.

When a designer knows how much current all the tube plates and screens draw, he picks a rectifier (or group of them) that can handle the current. The rectifier must also be able to withstand the voltage that is to be applied. (Rectifier diodes can be wired in series to increase voltage ratings.) Then the designer picks a transformer that supplies enough ac voltage to develop the maximum dc voltage needed and makes sure its windings are rated to carry the needed current.

Filter capacitors store electrons that will

smoothing it out

Getting rid of whatever ac is still a part of pulsating dc is the job of the filter capacitors and choke. The larger the capacitance is, the less ac gets through. Also, the larger the choke inductance, the smoother the dc that results.

Some typical pi-network power-supply filter circuits are shown in fig. 6. The input filter (C1) has a lot to do with the output voltage. The larger it is, the higher the dc — up to nearly the peak value of the input ac. The output filter controls how much current the supply can furnish without leaving a lot of ac ripple.

Aside from their effects on voltage and current, the two capacitors cut down ripple. They are low impedances for power-supply ac. The input capacitor "shorts" a lot of the leftover ac (from the rectifier) to ground. The choke has a high impedance to ac but passes the dc easily if it's rated to carry enough current. The output capacitor then "shorts out" any remaining ripple.

When not too much current is to be drawn, a wirewound resistor can take the place of the choke. Fig. 6B shows the circuit. The resistor isn't as effective as a choke, but it works.

The diagram in fig. 6C shows the elec-

trolytic capacitors "upside down." That's because the input is negative pulsating dc. Electrolytics are polarized and must be connected correctly. The output is smooth negative dc.

dc plus or minus

The last power-supply characteristic we'll review is polarity. In tube-type ssb receivers, the voltages needed most are for plates and screens; the supplies therefore supply positive voltages. Transmitters and linear amps sometimes need high negative bias voltages, so the power supply must develop a negative output voltage. Transistors, too, may demand negative supply voltages—usually much lower than those for biasing tubes.

Polarity of the dc output from a power supply is determined by the direction in which the rectifier is hooked up. If the cathode (bar) is on the side toward the output or the filter network, the output is positive. If the anode is in that direction, the output is negative. The sketches in **fig. 7A** show both.

There's another way to get negative voltages, without using a separate rectifier. It's diagramed in **fig. 7B**. This is an ordinary positive-output dc supply, furnishing 105 volts dc across the output filter. The bottom of the transformer secondary is not grounded; instead, the ground point is between the two resistors that form a bleeder across the output.

Dc voltages for stages in modern equipment are measured with respect to ground. A voltmeter connected to the bottom of the bleeder in **fig. 7B** measures -5 volts dc. Thus, by the simple trick of moving the ground point up the bleeder chain, a negative voltage becomes available. In ssb equipment that requires both positive and

negative supply voltages, this is a favorite way to get them.

ssb power supplies at work

The power supply in **fig. 8**, chosen from an ssb receiver, demonstrates many of the principles just discussed. The main supply

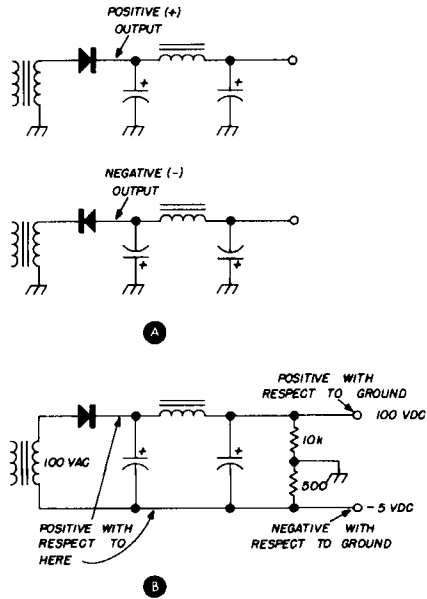
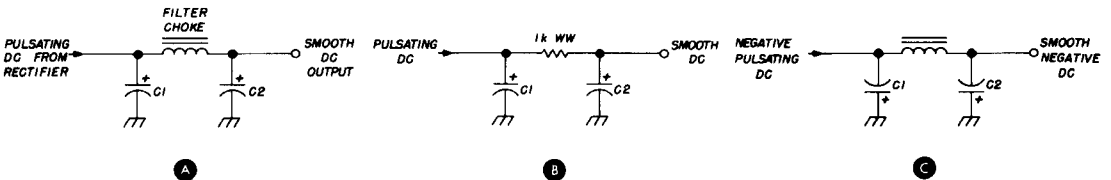


fig. 7. Positive and negative voltages with a single rectifier.

is a full-wave rectifier. D1 and D2 supply positive-going pulsations to a large input filter. With 180 volts ac from each transformer winding, dc across the output filter is 195 volts.

Some circuits in this receiver need closely controlled voltage. So, a regulator tube holds the output at the end of the 4k re-

fig. 6. Pi-network power-supply filters.



sistor at exactly 150 volts—no matter how much current is drawn from that branch of the supply (up to 35 mA).

Also, a negative voltage is needed. For it, the lower half of the secondary winding applies 180 volts ac to a resistive voltage divider, lowering the ac voltage applied to

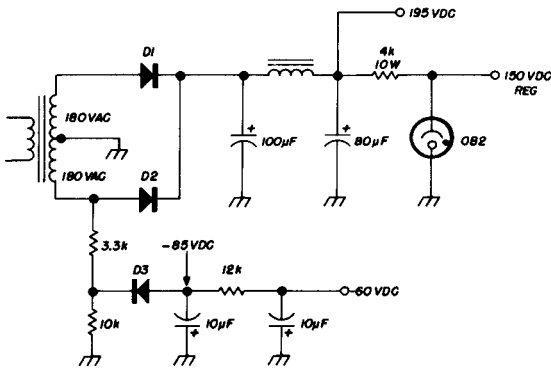


fig. 8. Power supply for an ssb receiver demonstrates many of the power-supply principles discussed in the text.

D3. The rectifier is connected for anode-output, so it supplies negative pulsations to its filter network. The capacitors, connected for negative output polarity, work with the 12k resistor as a pi-network filter to smooth out any ripple. The output is -60 volts dc.

One transmitter supply uses some interesting variations. Take a look at fig. 9. The arrangement is part full-wave doubler and part ordinary half-wave rectifier. The rectifier diodes are in series merely to increase their voltage rating, so less-expensive diodes can be used.

First, the half-wave circuit. It consists of D3, D4, the full transformer winding, and filter C2. When the top of the transformer winding is negative with respect to the bottom, D3 and D4 can conduct, charging C2. On the next half-cycle, when the top is positive-going, they can't conduct. No more voltage is added to C2, nor can it discharge through the rectifiers. Thus, a train of positive half-cycles applied to C2 develops a positive pulsating dc voltage

there. The capacitor is large enough that the pulsations are mostly smoothed out. The output is about 260 volts of fairly smooth dc.

The 200-ohm resistor and C3 form a pi-network with C2. The network smooths most remaining ripple out of the dc. Also, the 200-ohm resistance, with current flowing through it, drops the output voltage for this branch to 190 volts dc.

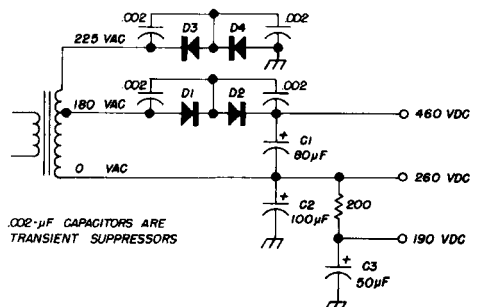
During the half-cycle when the top of the transformer winding is positive, the 180-volt center tap is also positive with respect to the bottom end. Rectifiers D1 and D2 can conduct with this polarity of voltage applied; they do, and charge C1. The peaks of these positive cycles reach about 250 volts. Capacitor C1 smooths out the pulsations and leaves an average dc voltage of about 200.

With respect to ground, C1 and C2 are in series, so the dc voltages across them add. The dc output is 260 volts (C2) plus 200 volts (C1), or 460 volts. The C1 voltage is developed during one half-cycle, and the C2 voltage is developed during the other—which makes the 460-volt output a result of full-wave rectification.

ssb on wheels

Single-sideband is a natural for mobile operation. Transmitting efficiency is high, which helps overcome the limitations of available power for the rig. In fig. 10 is a supply that converts 12 volts dc to high-voltage ac and then back to even higher-voltage dc.

fig. 9. Interesting transceiver power supply provides three positive voltages.



The part of this circuit that turns dc into ac is called the inverter. The two pnp transistors are switches that alternate in conducting current, each through its own part of the primary winding. To begin the action, when 12 volts dc is first applied to point A, one transistor can conduct slightly better than the other, so it dominates. Suppose Q2 is that transistor.

Current flows through the winding from A to B to C and to ground through the 195-ohm resistor. A magnetic field builds up around those portions of the windings. The

cyclic rise and fall of the magnetic field around the primary winding. That's ac.

The secondary of this transformer is a stepup winding for developing a high ac voltage. Its efficiency is improved tremendously by "tuning" it with a capacitor—the 470 pF. The output of this winding is about 1000 volts ac.

The rectifier diodes are series-connected in two banks of six each, called stacks. The series connection divides up the voltage across each rectifier. The two stacks make a full-wave doubler, producing an

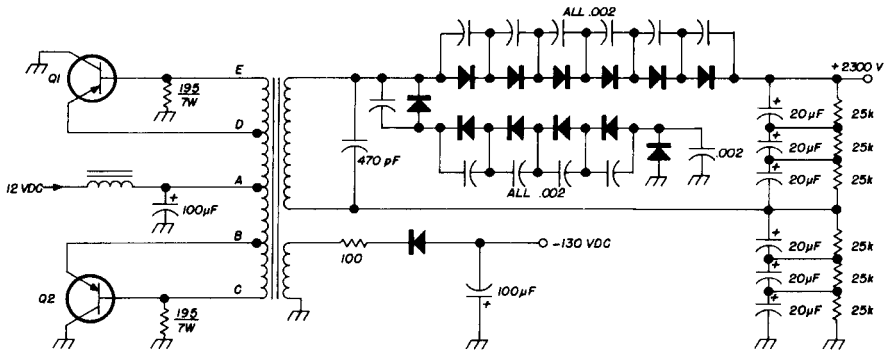


fig. 10. Mobile dc-to-dc converter that provides 2300 volts output from a 12-volt battery.

base of Q2 is a little less positive than the emitter, forward-biasing the pnp transistor. The collector, being grounded, is the most negative. Q2 conducts more and more heavily.

Whenever a magnetic field is expanding, it creates a counter-force that's opposing. When current in Q2 reaches saturation, the magnetic field stops expanding. The counter-force takes over and starts cutting down current in Q2 and in its winding sections. This reduction continues until Q2 is cut off by the reverse-bias that's self-induced by the winding.

Meanwhile, Q1 was kept cut off by the current rise that made Q2 conduct, because the two transistors are wired in opposite phase from each other. Once current in Q2 starts diminishing, however, current in Q1 starts rising. That makes current in the upper half of the winding increase, which raises the magnetic field around that half.

The result of this push-pull switching is a

output of 2300 volts dc. As you probably guessed, the capacitors are wired in series—three to a stack—to permit using lower-voltage types. (High-voltage capacitors are very costly.)

The bottom secondary winding is connected to a negative-output rectifier, and supplies —130 volts ac. This is a bias voltage for the high-power ssb transmitter this power supply works with.

dc power for transistors

For operating transistors, input ac is first stepped down, then rectified. For critical transistor circuits, the low dc output may also be regulated. The power supply in fig. 11 combines all three features.

There's something odd in fig. 11: transistors are used as rectifiers! This is the ac-input version of a supply that can also be used on 12 volts dc. The dc mode uses transistors Q1 and Q2 for inverting dc to ac. They're big, husky transistors, so why

not use them as rectifiers, too, and take advantage of diode action between their bases and collectors?

After full-wave rectification, the dc is smoothed by filter capacitors and a choke. The 12 volts dc is fed through R1 and R3 to the output terminal. The zener diode holds the base of Q3 at a steady voltage with respect to point A. Because of voltage drop across R1, current drawn through R3 makes the voltage vary at point A—and therefore at the base. The variation at the base makes shunt regulator transistor Q3

exactly the same amount—becoming that much more positive than normal. Voltage at the base becomes more positive, too, but not as much, because the voltage (and any change) is divided between R1 and R2. The net effect is to reduce the forward bias of Q2, lowering conduction. That reduces conduction in the base-emitter junction of Q1; that junction is in series with the collector of Q2. Less base current in Q1 raises its emitter-collector resistance. And that lowers the output voltage to its normal value.

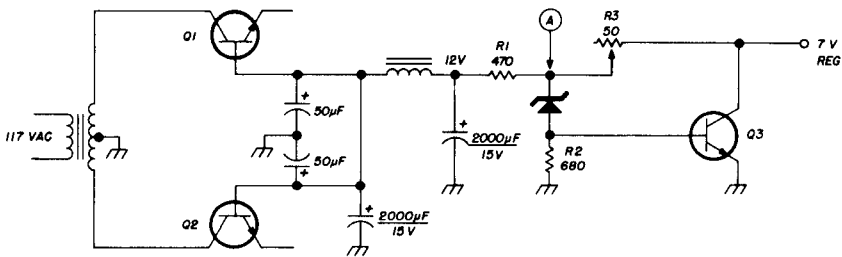


fig. 11. Regulated low-voltage power supply using transistors as rectifiers.

draw more current or less through R3 and thus hold the output voltage constant.

Consider what happens if the output voltage (and the voltage at A) rises. The base of Q3 goes more positive, and the transistor draws more current, reducing voltage at the collector (tied to the output). If the output voltage drops, the base goes less positive, and Q3 draws less current through the series resistors. R3 is adjustable to allow for aging of the zener or the transistor.

Transistors are used more often as series regulators. The example in fig. 12 has an extra transistor, often known as an error amplifier. Operation is simple. Transistor Q1 is a series resistance, variable by the voltage at its base. This variable resistance parallels R2 in series with the output. The resistance of Q1 depends on conduction of Q2. An example illustrates how output voltage variations are corrected.

If the 12-volt output goes up, voltage at the emitter of Q2 goes up, through D2, by

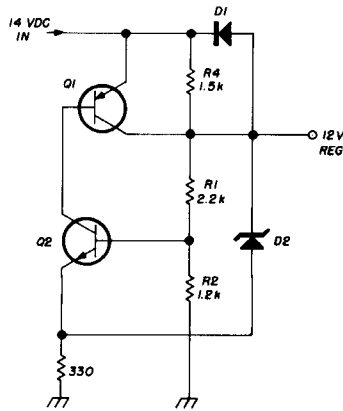
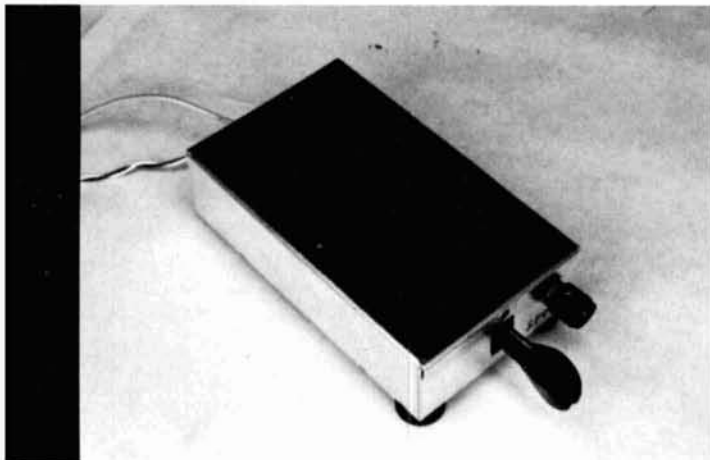


fig. 12. Solid-state series-type voltage regulator.

If the output goes down, forward bias is increased for Q2, and thus for Q1. The lowered emitter-collector resistance of Q1 lets the voltage go back up to normal. The result is a steady 12 volts, no matter what tries to lower it.

ham radio



Mini paddle installed in a homebrew keyer.

the mini paddle

Here's a simple
low-cost keying paddle
for the new
electronic key
you are
putting together

Del Crowell, K6RIL, 1674 Morgan Street, Mountain View, California 94040

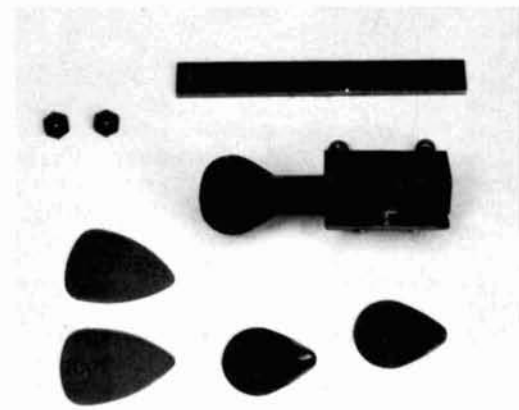
Many **electronic-keyer** circuits have appeared in the past several years, and they're getting so simple that almost anyone can build one. However, the paddles are a different story. The paddle described here requires a minimum of tools and is inexpensive to build.

construction

The paddle should be built with good quality materials. The arm is built from a section of 1/8-inch double-sided epoxy board; the sides must be flat and smooth. You can use aluminum or other material, but it will require much more work.

A pair of guitar picks are epoxied to the

Disassembled mini paddle.



paddle to provide a good surface for the fingers. By carefully using a drill, the bearing holes can be recessed in the edges of the epoxy material. For accurate alignment, these holes must line up with the holes in the bracket. The bearings can be salvaged from an old broadcast variable capacitor. Silver contacts removed from an old relay are soldered to the copper side of the paddle.

The bracket should be built from .040 brass, although tin or steel can also be used.

side of the paddle. By adjusting the thickness of the spring you can adjust the tension to your own desires. The spring is soldered in place after adjusting. The paddle is held in place by bending the bracket to provide proper spring tension on the pivot bearings.

adjustment

Adjustment of the paddle is quite simple once the spring is adjusted for centering and soldered to the bracket. The contacts can be

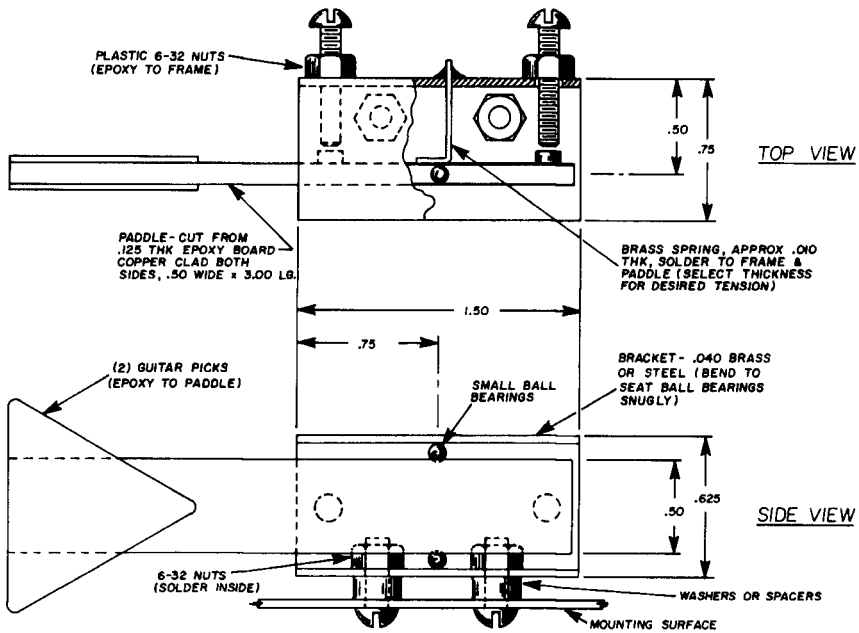


fig. 1. Constructional details of the mini paddle.

Mounting holes are located as shown in fig. 1 and 6-32 nuts are soldered to the inside surface of the bracket. Guide nuts for the contact screws are provided by epoxying two plastic or nylon 6-32 nuts to the sides of the bracket. The clearance holes must be larger than the screw you use and the nuts must be centered to prevent shorting. Contact screws should be brass with the ends filed round and polished for good contact.

Spring tension is provided by installing a strip of brass material as shown in fig. 1. The size of strap is approximately 0.13 x 1 inch with one end bent and soldered to the flat

adjusted to provide for proper travel and centering. I installed my paddle in an LMB 139 box along with the Micro TO Keyer described in the 1968 ARRL Handbook.

This keyer was originally built for mobile operation but it does a nice job as a second unit for the home station. It operates from a pair of flashlight batteries and has a built-in zener diode and dropping resistor so I can plug it into the cigarette lighter in the car. The monitor oscillator was eliminated for simplicity since my equipment has built-in sidetone.

ham radio

a sloping inverted-vee dipole

Here is
an interesting antenna
for small antenna farms
that bears
further investigation

Alf Wilson, W6NIF, 3928 Alameda Drive, San Diego, California 92103

Most of us at one time or another have dreamed of the ultimate antenna. If your ambition, for example, is to accumulate new country DX contacts, then this ultimate antenna might be a family of rhombics covering 360 degrees, or maybe a 100-foot telephone pole with full-sized Yagis for each band. The majority of us never realize the ultimate antenna, unfortunately. As a matter of fact, there are hams who, for various reasons, will never be able to own and operate **any** kind of antenna but a dipole or one of its relatives.

It is for these hams that this piece is written. I hope they may benefit by my experience with a restricted-space antenna that evolved over several months of testing to achieve the best possible compromise. In general, those who are restricted to low-gain antennas fall into one or more of the following groups:

- a. Thin pocketbook city dwellers on small lots
- b. Not-so-thin pocketbook city dwellers on small lots who have zoning restrictions, unreasonable neighbors, unco-operative spouses or all three.
- c. Timid city dwellers who are reluctant to live under a potential catastrophe such as a ton of tubular steel crashing onto the house

Many of these hams, knowing it is futile to participate seriously in the big contests, nevertheless would like to have a chance at working some DX at least once in awhile. Being antenna dreamers, these people have read extensively and have learned the pros and cons of horizontally- and vertically-polarized low-gain antennas. They have resigned themselves to using one of the dipole family; what, then, is the best possible arrangement they can build that will provide at least a fair chance in the DX bands?

Let's assume a typical city dweller has room to erect an inconspicuous mast at

riod on 14 MHz. It is a variation of the well-known inverted vee. The conventional inverted vee has been covered fairly extensively in the literature^{1,2,3,4}. Very little has been published, however, on any but the standard configuration.

the inverted vee

The standard inverted vee is a half-wavelength dipole with the elements in a vertical plane, oriented in an acute angle, α , and fed at the apex (fig. 1). The size of the apex angle apparently has little effect on the vertical radiation pattern but significantly changes the horizontal (earth

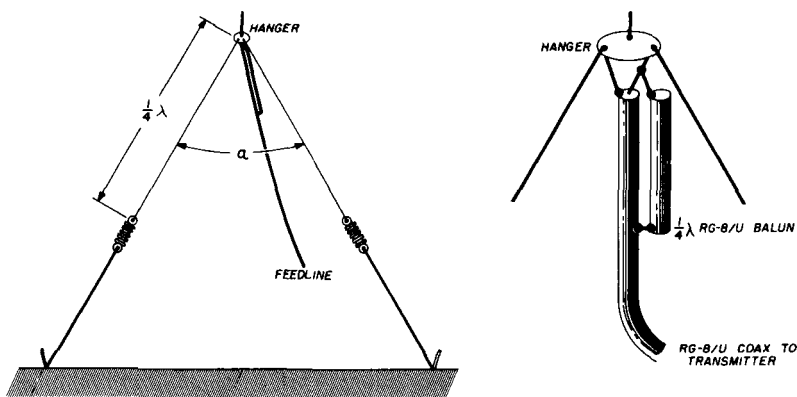


fig. 1. Basic inverted-vee antenna. Resonant frequency depends upon height above ground and angle. The length of the quarter-wave balun is equal to $246v/f$, where $v = 0.659$ if balun is taped to feedline, and f is frequency in MHz. Center conductor of the "bazooka" section is floating.

least one-half wavelength high on 14 MHz, but doesn't have enough room laterally to erect a second mast one-half wavelength from the first. Full-sized horizontal dipoles are therefore out. He could put up a shortened (trap) dipole or one of the verticals, say a ground plane. Both have advantages and drawbacks, all of which are well known and will not be discussed here. Which is most effective, considering the restrictions involved?

Based on these restrictions the antenna to be described, while not in the competition class by any means, is a modified version of the dipole family that has given a good account of itself over a six-month pe-

plane) pattern such that the nulls off the ends normal to the wire axis tend to fill in as the apex angle decreases. (Reference 5 discusses this at length and provides typical patterns in both planes for several apex angles.) Thus as the antenna approaches a vertical configuration, its horizontal pattern approaches the omnidirectional pattern of the conventional vertical, which is to be expected. Reference 5 shows that the vertical pattern is virtually unaffected for antenna apex angles from 120 to 60 degrees. Reference 6 shows the vertical radiation pattern of any antenna should be lower than 30 degrees for effective DX work.

The problem is to somehow get the vertical pattern down into the useful angles without sacrificing energy in high-angle lobes that penetrate the ionosphere. Increasing the antenna height above one-half wavelength does lower the vertical angle but introduces high-angle lobes that are useless for DX propagation. Also, recall that we are restricted to one-half wavelength in height on the 14 MHz band.

the sloping inverted vee

After using a conventional inverted vee at a height of one-half wavelength on 14 MHz for a month or so, I wanted to find out what would happen to antenna performance if the elements were rotated out of the x, y plane. The elements were kept broadside to the x, y plane but were elevated at a slope angle, θ (fig. 2).

At this point I must emphasize that no amount of wishful thinking will alter a basic physical fact; namely that the directional characteristics of a true vee antenna cannot be realized with anything less than one wavelength on the legs. A vee-shaped dipole is exactly that: its shape is that of a vee, but there its similarity to a vee antenna ends. The inverted vee dipole behaves like a dipole regardless of the apex angle, α , or the slope angle, θ . However, it appears that if the slope angle is

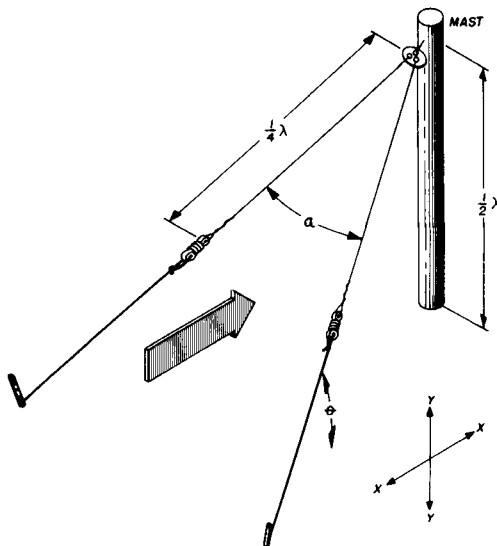


fig. 2. Geometry of the sloping inverted-vee antenna. The angle of the vertical lobe decrease in the direction of the arrow.

table 1. Characteristics of several simple antennas; all assumed one-half wavelength high and fed with the same current.

antenna	nominal radiation resistance (ohms)	vertical pattern useful lobe angles (degrees)	first reflection zone* (miles)
horizontal dipole	73	15 - 45	200 - 1100
ground plane	53**	10 - 55	100 - 1500
inverted-vee dipole	50	15 - 45	200 - 1100
sloping inverted-vee dipole	50	15 - 25 (estimated)	600 - 1500 (estimated)
yagi	8	15 - 45	200 - 100

*for F-layer heights between 125 and 250 miles

**with radials sloped approximately 50 degrees

tests

My sloping inverted vee dipole was oriented with the elements broadside to the long-path direction (about 230 degrees azimuth for my location). Over a six-month period, the antenna was used in tests with South African, Australian, Southeast Asian, and Asian stations on 14 MHz. Approximately 100 contacts were made. During these tests the sloping inverted vee antenna was compared with a shortened dipole and a quarter-wave ground plane. The driving points of all three antennas were within about 5 feet of being at the same height above ground. The inverted vee was rotated out of the x, y plane to slope angles as great as 70 degrees. This was accomplished by rigging a lightweight spreader between the antenna elements and raising

them to various angles with a rope and pulley arrangement. An interesting effect was that the optimum slope angle seemed to depend on the time of day. During band openings on 14 MHz (1300-1400 GMT) best results were obtained with a slope angle of about 40 degrees. As the sun rose, reports improved with the slope angle approaching 70 degrees.

results

No quantitative measurements were made. Pattern and field strength measurements are difficult enough even under laboratory conditions; consequently any data taken with the limited facilities available would certainly be open to question. However, after testing the antenna under all band conditions, the following observations are offered:

- a. The sloping inverted vee antenna apparently radiates better in the vertical plane at lower angles than a dipole or ground plane.
- b. The first reflection zone of the sloping inverted vee appears to be at a greater distance than the comparison antennas.
- c. As a receiving antenna, the sloping inverted vee is noisier than the dipole, but not as noisy as the ground plane.
- d. Band openings occur earlier by about a half hour with the sloping inverted vee.
- e. The sloping inverted vee requires less space than the dipole, but more than the ground plane.

Based on the data in reference 6 and the empirical results of the tests described above, the characteristics of the sloping inverted vee are summarized in **table 1**.

conclusions

In conclusion I would like to point out that the sloping inverted vee is not in the same league as even a two-element Yagi, which has both directional and power gain. While the sloping inverted vee does seem to have more power at the lower angles than a conventional horizontal dipole, it simply cannot compete with a Yagi or any other directional array. The estimated first reflection zone for the sloping inverted vee is based on the curves of reference 6, which in turn are based on various vertical pattern lobe angles. Experience with this antenna seems to indicate performance close to the data shown in the table.

I hope that others will try this antenna with different slope angles. It will be very interesting to see how it performs at other locations and on different bands.

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ham radio

vtvm modification

Most garden-variety vtvm's use a 1.5-volt flashlight cell for the ohmmeter supply. This works fine if you replace the flashlight cell every few months. However, the ohmmeter becomes very inaccurate on the low-ohms scale as the cell's internal resistance increases with age.

By adding a few parts and making a few circuit modifications, it is possible to do

away with the flashlight cell altogether. The diagrams show the modifications I made to my Heathkit V-7A. This is typical of all inexpensive dc vtvm's, whether made by Heath, Allied, Eico or RCA.

As shown in **fig. 1**, the 6.3-volt filament voltage is rectified by a conventional bridge rectifier; the resulting dc is dropped by the 18-ohm resistor in conjunction with the regu-

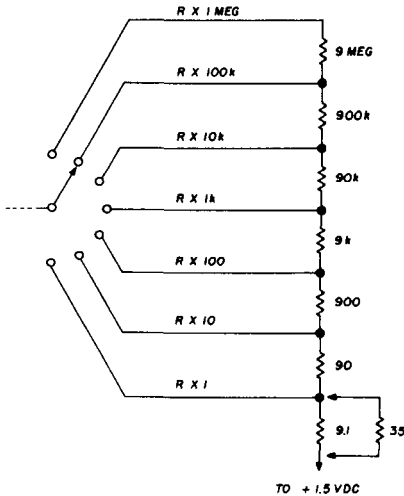
lating diodes, giving the required 1.5 volts dc.

It's necessary to lift one side of the heaters from ground before you can use the 6.3-volt winding in a bridge rectifier. Heathkit uses a printed-circuit board, but the conductors that must be removed can be easily scraped off the board with a sharp knife.

Two 1N538's in series act as a zener to regulate the output voltage. The barrier potential for a silicon diode is about 0.75 volts; two in series regulate the output nicely at 1.5 volts.

The original flashlight cell was connected to a resistor string tapped by the ohmmeter range switch as shown in **fig. 2**. Notice that the total resistance switched into the circuit is always 10 times the range switch position; this assumes cell internal resistance of 0.9 ohms. The internal resistance of this power supply is higher than 0.9 ohms, so it's necessary to lower the 9.1-ohm resistance by shunting it with 35 ohms as shown. The two in parallel then make 6.75 ohms. When this is

fig. 2. Modifying the ohmmeter circuitry for the 1.5 Vdc supply.



resistance, 600-PIV silicon diodes. This saves about as much heater current as consumed by the 1.5-volt power supply, so the transformer is still running cool.

The diodes that replace the 6AL5 should be forward biased slightly to overcome their barrier potential. This is done with a voltage divider consisting of a 24k- and 150-ohm

fig. 1. Deriving a 1.5-volt dc supply for the vtvm from the 6.3 Vac filament transformer. Added circuitry is enclosed by the dotted line.

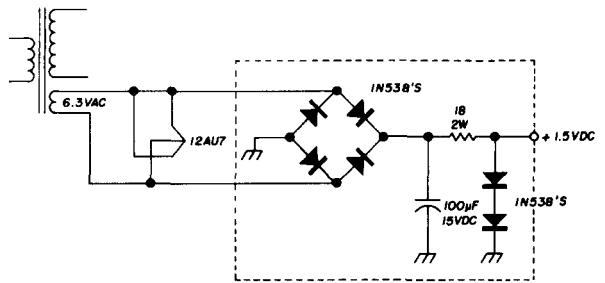
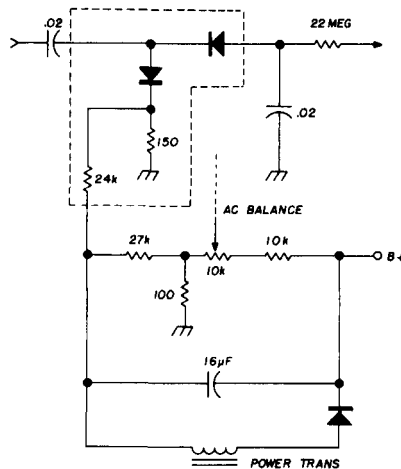


fig. 3. How to replace the 6AL5 with forward-biased silicon diodes.



added to the power supply internal resistance of 3.25 ohms, it makes up the required 10.0 ohms.

In my vtvm I also replaced the 6AL5 "ac volts" rectifier with a pair of high back

resistor connected to the negative side of the original plate supply (**fig. 3**). Linearity is rather poor below levels of 1 volt ac, but is still superior to the original 6AL5.

Fred Brown, W6HPH



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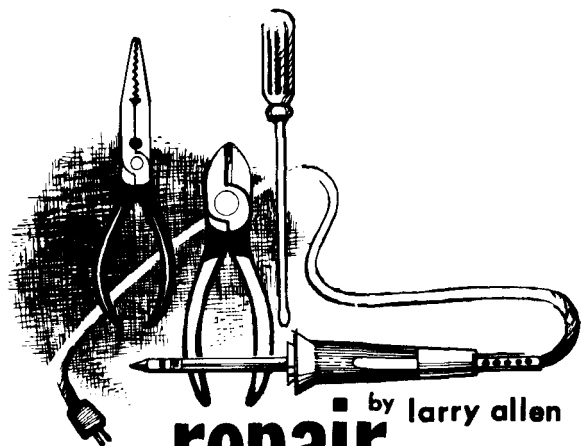
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by Larry Allen

the repair bench

analyzing wrong dc voltages

There isn't really a "best" way to troubleshoot electronic equipment. The most successful radio and electronic technicians combine several methods. I've already told you about some of them in past columns, but this month's method is probably the most popular; at least it's one you have to understand if you expect to repair every trouble in all sets. It's called **voltage troubleshooting**.

Just because this method is the best known doesn't mean it's always best. Most experienced guys save this one for last. It's a quick way to find the faulty part in most circuits, once you know which circuit to look in.

You can find which circuit this way too, but other ways are usually quicker. I'll use this month's column to tell you how to pin down faulty parts within the circuits; that's the best way to use voltage troubleshooting.

how the method works

In theory, it's simple. You start by measuring each dc voltage in the suspected circuit. Then compare yours with correct voltages. Most instruction books and schemat-

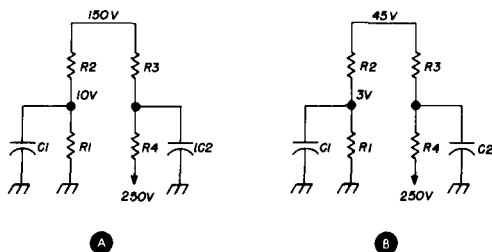
ics nowadays include voltages—at least at each tube or transistor terminal. When your circuit diagram doesn't show the working voltages, use past experience and your knowledge of how the circuit works.

When you find a voltage that isn't right, figure out what could cause it. If you know Ohm's law for voltage, current, and resistance, it's not too hard to decide what's causing the voltage to change. In fact, it's really cut-and-dried.

Here's an example. In **fig. 1** are two versions of a simple series resistance circuit connected across a dc voltage. The first, **fig. 1A**, shows normal dc voltages (which I'll call operating voltages). The other, **fig. 1B**, shows voltages you might measure in this circuit when you start troubleshooting. (Voltages are always measured with respect to ground.)

There's 150 volts at the R2-R3 junction in **fig. 1A** because current flows through R3 and R4, causing (says Ohm's law) a drop of 100 volts. There's only 10 volts at the

fig. 1. Simple series resistance circuits show how voltages divide according to the various resistance ratios.



R1-R2 junction because another 140 volts is dropped across R2. The same amount of current flows in all four resistances, since they're in series, so the voltages divide up exactly according to the resistance ratios.

You analyze voltage to determine why the two voltages in **fig. 1B** have changed from normal. What part has gone bad? For some reason, the resistance ratios must have changed. Your first step, then, is to figure out what the new ratios are.

If you consider the 45 volts, you can quickly see two possibilities. The first is that R1 or R2 may have lowered in resistance. In that case, the ratio of their re-

the ratio of voltages, and then figure out what's causing the foulup.

voltage dividers and ohm's law

You should already know about series and parallel circuits.* For voltage troubleshooting, you must know how voltages spread themselves around both kinds of circuits.

Start with the simple series circuit in **fig. 2A**. The resistors divide the voltage in proportion to their values. The ratio is 2:2:1. The R1-R2 junction is 80 volts below the supply voltage, and the R2-R3 junction is 80 volts further below supply voltage. Look-

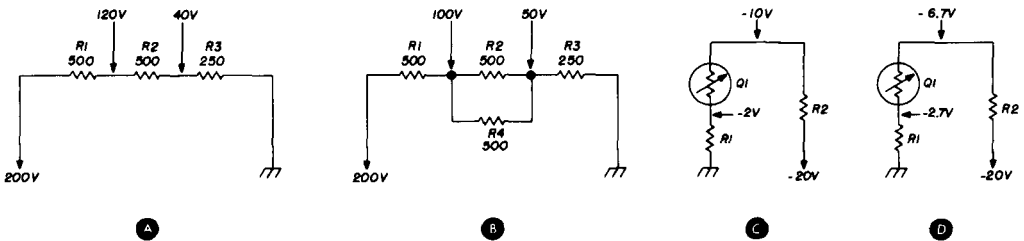


fig. 2. How voltages change when one of the resistance values changes.

sistance to that of R3-R4 lowers, and less of the total voltage appears across them. On the other hand, the combined resistance of R3-R4 may have increased. That causes more voltage to drop across them and lowers the voltage reaching the top of R2.

So . . . which is it? The clue lies in the fact that **both** the R1-R2 voltage **and** the R2-R3 voltage have kept the same ratio between them, even though both are lower than normal. The 3 is to 45 as the 10 is to 150; the ratio is still 1 to 15. The trouble is more likely in R3 or R4; the value of one of them must have increased. To find out which one, you could disconnect them and measure with your ohmmeter. However, further voltage measurements will show one of them still retains its ratio to R1-R2; the other is the culprit.

This simple reasoning is the basis for analyzing all dc voltages in series circuits. You study the ratio of resistances, compare

ing at the divider string from the standpoint of the voltmeter: R1 drops 80 volts, R2 drops 80 volts, and R3 drops 40 volts. The 2:2:1 ratio is maintained.

Look what happens when you alter a resistance value, as in **fig. 2B**. Some additional resistance, in the form of R4, has been added in parallel with R2. It lowers the effective resistance; R2 and R4 are both 500 ohms, so their parallel value is 250 ohms.

That goofy up the resistance ratio of the divider chain. It has become 2:1:1, and the voltage divides accordingly. R1 drops 100 volts, R2 (with R4) drops 50 volts, and R3 drops 50 volts. So you see, among the voltages, the 2:1:1 ratio holds.

Now, imagine that **fig. 2C** is a supply circuit in a solid-state receiver. The variable resistance labeled Q1 is actually the tran-

* For a rundown on series and parallel circuits, check the **repair bench** in the November, 1968 and January, 1969 issues of **ham radio**.

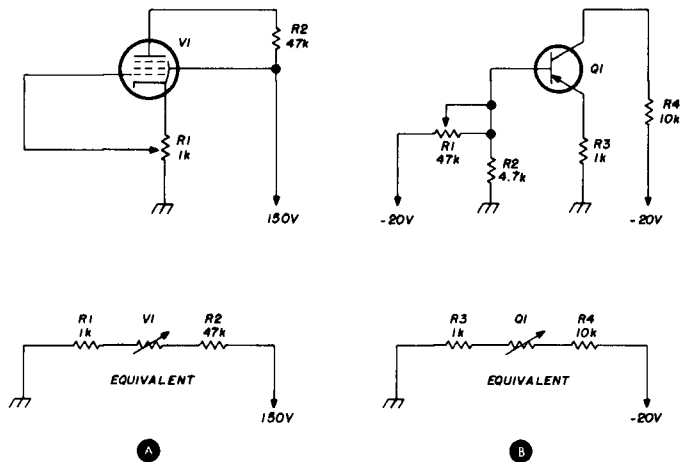
sistor, but imagine it here as a variable part of the divider chain. Voltages are shown at each point in the circuit, so you can figure out the voltage (and therefore resistance) ratio. Across R1 is dropped 2 volts, across Q1 is dropped 8 volts, and across R2 is dropped 10 volts. The ratio is 1:4:5.

Suppose something goes wrong with the

sistance in the dc supply circuit. In **fig. 3A**, the tube bias can be changed by moving the slider on cathode-bias resistor R1. The equivalent circuit shows the tube's dc plate resistance as variable; it actually changes as bias is varied.

Simplicity is lost because R1 is also in series with the dc supply circuit. When the resistance of V1 changes, current changes

fig. 3. Tube and transistor circuits with their equivalent resistive divider circuits. Drawing the equivalent circuit makes it easier to see the effect when a resistance changes.



circuit. You measure with your voltmeter and find the voltages changed to those in **fig. 2D**. Now you have a new ratio of voltages. Across R1 is 2.7 volts, across Q1 is 4 volts, and across R2 is 13.3 volts. If you write out the new ratio, however, it's easy to see where the voltage changed. The new ratio is 2:3:10. The old ratio was 1:4:5, which is the same as 2:8:10; only the center factor has changed. That means the resistance of Q1 has changed, and that's what upset the voltages.

tubes and transistors as dividers

Thinking in ratios is easy as long as the resistances are simple. Experience will teach you to estimate voltage ratios close enough to give you a clue to which resistance has changed.

Tube and transistor circuits are different. Changing the bias on either one alters its dc operation. **Fig. 3** suggests how to consider a tube or transistor as a variable re-

all through the series circuit. A change in current through R1 affects bias, which changes the plate resistance of V1, which changes the bias, which changes plate resistance, and so on. The interdependency makes it tough to treat tube circuits as simple ratio dividers.

This problem also applies to transistor circuits; to see how, check **fig. 3B**. Bias for the transistor base is controlled by voltage divider R1-R2; since R1 is a potentiometer, bias is variable. As bias is altered, more or less current can flow in the emitter-collector circuit. The effect is to alter the emitter-collector resistance, represented by resistance Q1 in the equivalent circuit.

To see how complex these relationships can get, take a look at **fig. 4**—which is from a ham rig. Equivalent diagrams are shown at the right. The one at **fig. 4B** is for the collector circuit; it's the supply path through which dc must flow. **Fig. 4C** is the dc supply path for the base circuit; it in-

cludes the biasing voltage divider and the path through the base-emitter junction (shown as Q1B). Notice that R3 is in this path, too, so a current change in either circuit changes voltage across that resistor.

Finally, notice the combined paths in **fig. 4D**. Figuring out ratios in this circuit would be hard even if the resistances were simple. But they're complicated by the fact

plate voltage is low—only 180 volts instead of the 235 called for by the schematic. All the other voltages are about normal.

The plate-supply circuit (all the resistances plate current flows through) includes R5, one winding of T1, the tube itself, and R3. One at a time, consider the effect each part **might** have on voltage at the plate. Redrawing the equivalent plate-supply cir-

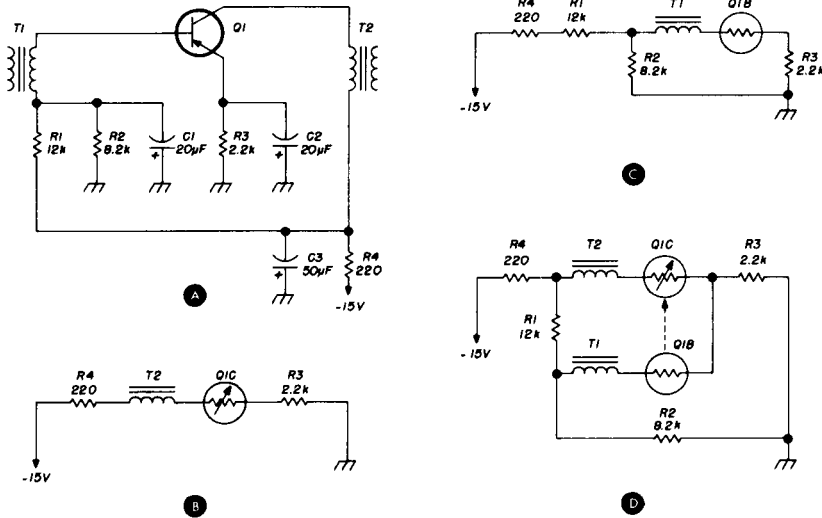


fig. 4. Transistor audio amplifier; the equivalent circuits in B, C and D show how difficult it is to use the "ratio analysis" method of troubleshooting a stage like this.

that the value of Q1C is controlled by Q1B. It's just too much to be simple.

a step at a time

However—and this is important—even though you can't use voltage ratios directly in tube and transistor circuits, you can apply the principles. Concentrate on one supply path at a time. At each step, ignore the effects of other circuits. When you decide what happened in one circuit, then study whether another could possibly be causing the voltage upset.

The transmitter tube stage in **fig. 5** is an easy example. Suppose you've traced trouble to this stage and have decided to measure voltages to help you find the faulty part. Your voltmeter tells you the

circuit, as in **fig. 5B**, can help. Here's the way your thinking might go.

If R3 increases in value, the dc voltage at the plate increases instead of decreases. If R3 decreases, the voltage lowers—but only a little, since R3 is such a small portion of the over-all circuit. Conclusion: forget about R3 for the moment.

If the V1 resistance increases for some reason, the dc plate voltage goes up. However, if the resistance of V1 decreases, the plate voltage goes down—exactly the symptom you have. Conclusion: the trouble might be V1. But don't rush; check other possibilities, too.

The winding of T1 can be eliminated as a suspect. If it opens, there's no voltage at the plate of V1. If it shorts, it changes the

voltage too little to notice. If the value of R5 increases, the voltage at the plate of V1 lowers. So, there's another suspect! If R5 decreases in value, voltage goes up instead of down.

Now you have two suspects: R5 may be too high in value, and the dc plate resistance of V1 may be too low. Which is it? To find out, you measure other voltages in the stage.

For example, what about screen voltage? It's okay. If the dc plate resistance of V1 is low, it also lowers the voltage at the screen. But it isn't, so the tube and its bias must be okay. (You can make sure bias is okay by measuring it.) That leaves R5 as the likely culprit. You can just measure R5 with your ohmmeter to be sure.

Suppose you do find the screen voltage low, too. What then? The cause is probably the tube plate resistance, but what's making it low? It could be a faulty tube or incorrect bias. Either one could allow too much current through the tube, and thus lower the plate resistance. So—measure the bias. If it's okay, the tube must be bad. If the bias is wrong, go into that circuit to find out why.

The secret to voltage troubleshooting is to take it step-by-step, through each portion of each dc supply circuit. Decide how each part could cause the symptom you observe. If you trace the trouble to the tube or transistor, find out if it's bad or if some other operating voltage is making it work wrong.

voltages in transistor stages

The emitter-collector resistance of a transistor is set by bias current in its base circuit. Just as plate voltage of a tube may change if bias isn't right, the collector voltage of a transistor can, too.

The transistor stage in **fig. 6** is a mixer for a communications receiver. The transistor is a pnp type, forward-biased (base more negative than emitter). Dc operating voltages are listed alongside the emitter, base, and collector leads.

Suppose you trace trouble to this stage and want to use voltages to pin down the faulty part. The voltages you measure are

listed in **fig. 6B**. Collector voltage is high, suggesting there's either too little current through the transistor or R4 is too low in value. If the trouble is in R4, however, the voltage at the emitter must also be high—and it isn't.

The best clue is the positive voltage on the base. It is reverse-biasing the base-emitter junction, cutting off collector cur-

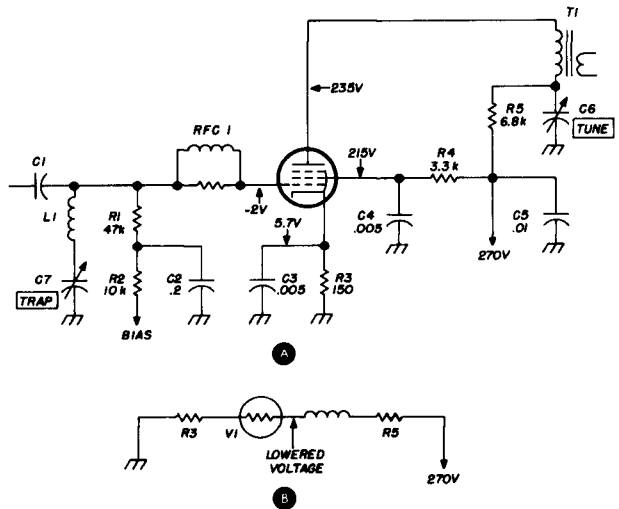


fig. 5. Driver circuit, with normal voltages listed at tube pins; B show equivalent plate circuit for troubleshooting.

rent through the transistor. You have to trace the cause of the positive voltage. As it turns out, the other side of C1 goes to a positive voltage in the preceding stage; C1 is leaky, and is coupling the wrong voltage to the base of this stage.

Suppose, instead, you measure the voltages listed in **fig. 6C**. Collector voltage is very low, meaning there's excess current through the transistor or R4 is too high in value. Base voltage is a little low, but it hasn't changed nearly as much as emitter voltage has. Study the circuit. You'll quickly decide that C2 has shorted. With 1.9 volts of forward bias (instead of the normal 0.5), the transistor conducts heavily, lowering collector voltage drastically. The base voltage is slightly lower because of

the effect of heavy current through R4; the base also gets its voltage from R4, through R2.

There are countless transistor-supply circuits you can analyze this way. Always pick out the voltage that is most wrong and find the cause of that first. Then check possible causes of other wrong voltages. The part that's common to all symptoms is usually the culprit. Each symptom leads to another, and ultimately to the bad part.

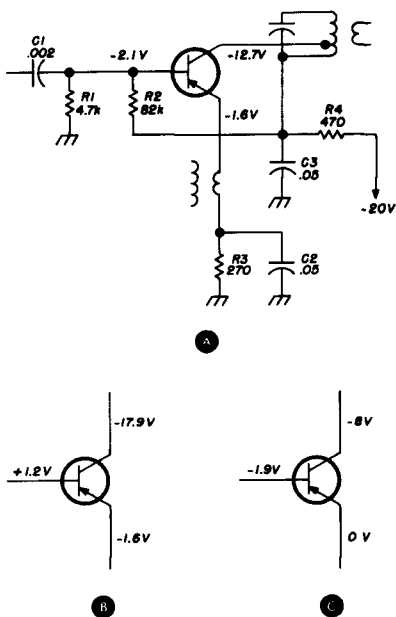


fig. 6. Transistor receiver mixer circuit; B and C show voltage changes in different troubleshooting situations.

testing individual components

This is the end of the line in troubleshooting—testing individual parts to make sure they're really bad. Resistors you can test most easily with your ohmmeter. Capacitors and other parts can be evaluated by voltage tests, combined with simple logic.

For example, C1 in fig. 6 should couple absolutely no dc voltage from the preceding stage. If the preceding voltage is positive and the capacitor is leaky, the result is what you get in fig. 6B. If the preceding voltage is negative, it could drive the base

of the transistor more negative and wind up reducing collector voltage. Leakage through C1 might also reduce voltage on the preceding stage—another helpful clue.

There is a reasonably sure way of testing a capacitor for leakage. If one end is connected to a dc voltage, disconnect the other end. With a vtvm, measure there for dc voltage. If voltage is leaking through, the voltmeter will detect it.

Don't be misled by electrolytic capacitors. They have a certain amount of natural leakage. Best way to check them is with an ohmmeter; leakage resistance should be no lower than 30k-50k.

You can tell if a coil or transformer winding is open, provided the winding carries dc. Just check at both ends with your voltmeter. An open winding blocks voltage.

You can also check for leakage between windings. Make sure one winding is connected to a voltage. With the other winding completely disconnected from any circuit, touch the voltmeter probe to one of its wires. If you measure any dc voltage, the transformer is leaky.

Any short-circuit can be pinned down quickly. Just find where the voltage has dropped completely to zero. This technique is most helpful in supply circuits where a filter, a bypass, or a high-power tube has shorted.

coming up

Awhile back, I promised to devote a column to using an oscilloscope at the repair bench. Apparently more amateurs own scopes than I suspected, but not many know how to use them or for what. Of course, everyone (just about) knows how to hook a scope up as a modulation monitor for a-m. But there are dozens of uses for this versatile instrument.

In my next column, I'll tell you exactly what a scope is and how to go about using it most easily. I'll give as many uses for it as I have room for. Then, after that, I'll explain a bunch of troubleshooting shortcuts with the scope. If you've got one, you should be using it on the repair bench. It's doggone handy.

ham radio

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propagation

predictions for february

Special
tabular predictions
for the ARRL
DX contest

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

In the past, I have stayed clear of forecasting propagation for specific paths. Forecasts have been as general as possible so they would be applicable to any of the paths that might interest a radio amateur in the 48 contiguous United States. The format I have used makes you “work,” but you will have obtained more accurate forecasts than otherwise possible.*

Still, the idea of doing a special propagation forecast for the CQ World-Wide DX contest or ARRL DX contest with the opening times tabulated for many paths **was** appealing. Now this copy is way past deadline, a \$46 computer bill has just arrived and I am swearing off! No more worldwide specific path predictions—at least until the next contest.

If you live within 300 miles of San Francisco, Dallas or New York City, and are interested in knowing when ten meters will open to Outer Mongolia or 18 other rare or not-so-rare DX spots, the tables are made for you. If you live between San Francisco and Dallas or between Dallas and New York you can interpolate the path opening times on many paths (not over the pole). Otherwise, the usual propagation charts, **fig. 1** through **6** will suffice.

* If you're looking for propagation forecasts for specific paths, there's always George Jacob's propagation column in CQ magazine.

Upon investigating the tabular path predictions, you will find, perhaps to your amusement, that opening times have been predicted to the nearest third or half hour on the lower bands and to the nearest 10 minutes on the higher bands.

Lest I be called to task for missing an opening by 10 minutes, or even a half hour, let me point out that these are median short-path predictions, and that variations of 10 percent in path muf and 10 dB in absorption are not uncommon. In addition, station parameters may vary considerably from the 100 watts CW or 800 watts ssb assumed on 80 through 10 meters (12 watts CW or 100 watts ssb on 160 meters). Antenna gains (over an isotropic radiator) are assumed to be -6 dB for 160 and 80 meters, zero dB for 40 meters and +6 dB for 20 meters and higher frequencies.

The ten-meter paths with no openings forecast **may** be open during the middle of the fifteen-meter opening—because the muf may be higher than forecast or the long path may be “open.” There are some marginal openings forecast that may catch the average contester unaware, such as the reopening of 20 meters after midnight or the short over-the-pole openings near dawn and dusk.

The serious contest operator will be prepared with at least one extra receiver, perhaps a panadapter, as well as a receiving antenna for checking “dead” bands. However, a serious weekend of bandscanning and **listening** just before the big contest can give more useful propagation information than any column.

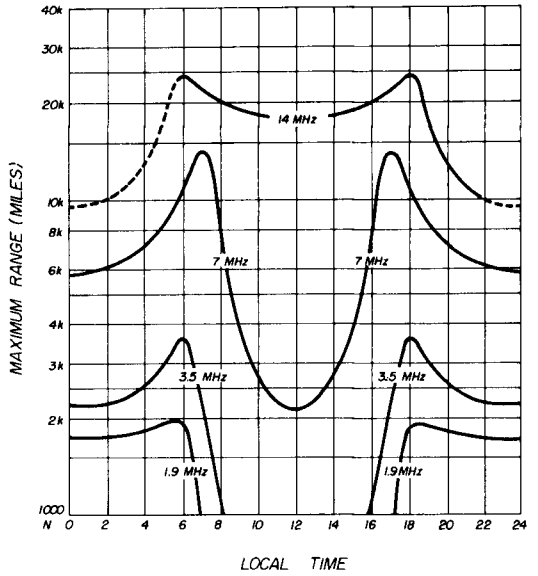


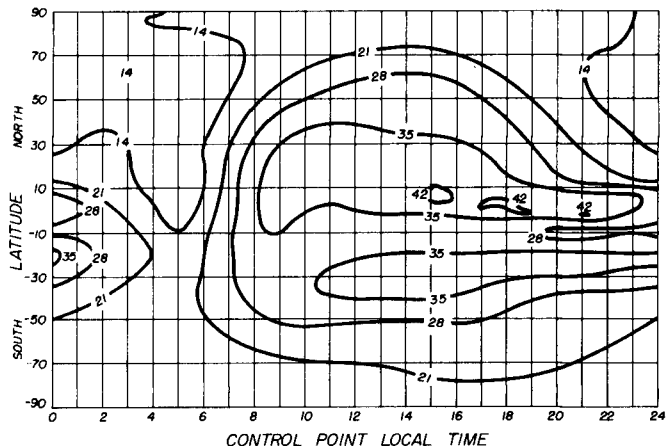
fig. 2. Maximum range to the north from 38° N latitude due to absorption.

tenna for checking “dead” bands. However, a serious weekend of bandscanning and **listening** just before the big contest can give more useful propagation information than any column.

six meters

Judging by last fall’s openings, six meters is expected to open infrequently (about 10

fig. 1. Maximum usable frequency curves for February 1969 based on 90° W longitude.



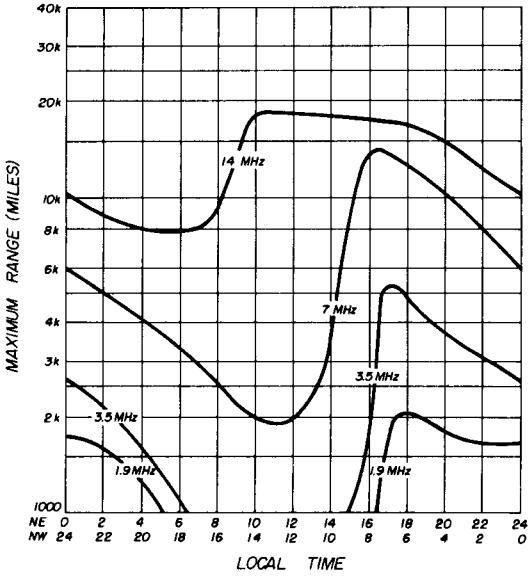


fig. 3. Maximum range to the northeast (top time scale) and to the northwest (bottom time scale) from 38° N latitude due to absorption.

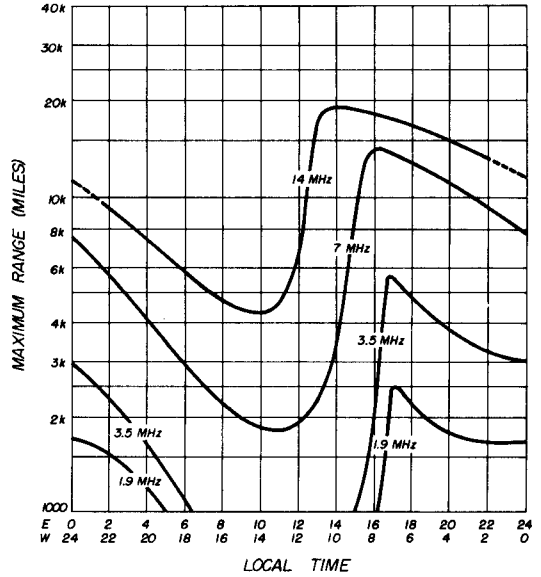


fig. 4. Maximum range to the east (top time scale) and to the west (bottom time scale) from 38° N latitude due to absorption.

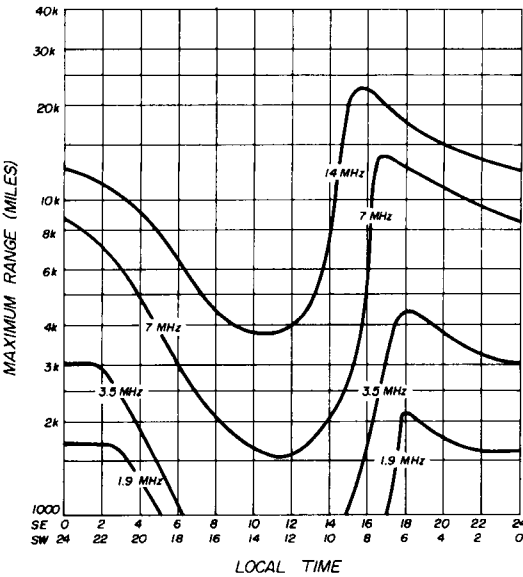


fig. 5. Maximum range to the southeast (top time scale) and to the southwest (lower time scale) from 38° N latitude due to absorption.

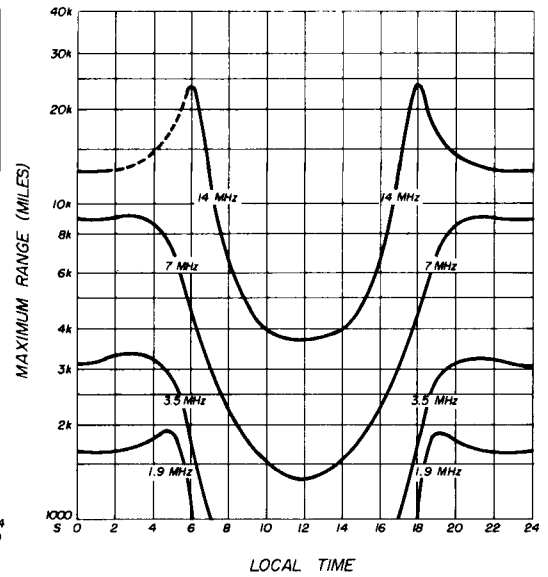


fig. 6. Maximum range to the south from 38° N latitude due to absorption.

table 1. Predicted 160-meter openings during the ARRL DX contest (time in GMT).

to from	San		
	Francisco	Dallas	New York
XE, Mexico City	0200-1200	0030-1240	0100-1100
KP4, Puerto Rico	0200-1000*	0100-1000	2300-1000
KH6, Oahu, Hawaii	0500-1400	0500-1300*	0500-1100*
KL7, Anchorage	0400-1400	0300-1300*	0300-1100*
KZ5, Canal Zone	0200-1100*	0100-1100	0000-1000
OX, Thule, Greenland	0200-1230*	0100-1130*	2300-1000
PY, Belem, Brazil	0200-0900*	0100-0900*	2300-0900*
ZK1, Cook Islands	0600-1400*	0600-1200*	0900-1100*

* Difficult path, 10 dB additional circuit gain assumed.

table 2. Predicted 80-meter openings during the DX contest.

to from	San		
	Francisco	Dallas	New York
XE, Mexico City	0120-1320	0000-1330	0000-1200
KP4, Puerto Rico	0100-1100	2340-1120	2220-1040
KH6, Oahu, Hawaii	0400-1500	0400-1320	0500-1200
KL7, Anchorage	0230-1530	0200-1320	0200-1200
KZ5, Canal Zone	0100-1140	0000-1140	2320-1100
OX, Thule, Greenland	0140-1330	0000-1220	2200-1100
PY, Belem, Brazil	0100-0930	2340-0900	2200-0920
ZK1, Cook Islands	0530-1440	0500-1300	0800-1200

table 3. Predicted 15-meter openings.

to from	San		
	Francisco	Dallas	New York
UAØ, Sakhalin Island	2040-0400	2120-0300	2200-0010
JT1, Ulan Bator	2300-0340	0000-0220	None
UA, Astrokhan	None	1440-1600	1220-1640
UA, Leningrad	None	1400-1600	1220-1740
DL, Berlin	1600-1800	1330-1900	1140-1910
VU2, New Delhi	0130-0320	1430	1240-1440
VS1, Singapore	2240-0440	2320-0300	1420-2000
	1730-1900	1620-1930	
DU, Manila	2200-0440	2200-0300	2240-2320
KG6, Guam	2000-0500	2010-0310	2020-0120
VK6, Perth	2100-0530	2130-0350	2130-0140
		1530-1830	
VK3, Melbourne	1940-0550	1930-0420	1940-0220
			1410-1540
VR2, Fiji Islands	1720-0550	1700-0420	1630-0220
		1500-1520	1410-1520
FO8, Tahiti	1600-0630	1530-0500	1530-0220
CE2, Valparaiso	1400-0340	1240-0440	1130-0120
PY1, Rio De Janeiro	1340-0300	1200-0200	1050-0040
ZD8, Ascension Island	1330-0220	1140-0100	1020-2340
CR6, Angola	1400-0200	1140-0040	1010-2300
ZS1, Capetown	1400-2200	1140-2220	1010-2320
5R8, Tananarive, Malagasy	1600-0240	1210-0020	1020-2300

table 4. Predicted 20-meter openings.

to from	San		
	Francisco	Dallas	New York
UAØ, Sakhalin Island	1930-0540	2000-0500	2020-0300
JT1, Ulan Bator	2140-0510	2240-0440	2320-0300
			1200-1240
UA, Astrokhan	1530-1700	1240-1730	1100-1800
UA, Leningrad	1500-1900	1240-2030	1100-2010
DL, Berlin	1430-2030	1220-1230	1030-2100
VU2, New Delhi	0130-0320	1430	1240-1440
	1540	1330-1600	1120-1620
VS1, Singapore	2150-0630	1500-0440	1200-0220
	1630-2030		
DU, Manila	2200-0440	2200-0300	2240-2320
	1630-1830	1500-1800	1330-1800
KG6, Guam	1900-0650	1910-0520	1940-0350
	1610-1620	1450-1610	1400-1620
VK6, Perth	1600-2140*	1440-2000*	1400-0400
	0200-0730	0200-0610	
VK3, Melbourne	1530-2000*	1400-1900*	1250-1830*
	0120-0900	0000-1050	0030-0430
VR2, Fiji Islands	1530-0900	1400-1050	1250-1840*
			2200-0430
FO8, Tahiti	1500-1140	1340-1100	1240-0440
			0700-0900
CE2, Valparaiso	1300-1030	1200-0900	1020-0800
PY1, Rio De Janeiro	1240-0600	1050-0830	0940-0720
ZD8, Ascension Island	1240-0430	1030-0800	0920-0200
			0440-0620
CR6, Angola	1300-0430	1030-0600	0920-0130
			0300-0530
ZS1, Capetown	1830-0430	1030-1240*	0920-1140*
		1740-0230	1620-0130
		0520-0800	0300-0600
5R8, Tananarive, Malagasy	1420-0420	1400-0500	0920-0120
			0300-0400

* Between this and next time period, path may be open for 10-dB additional circuit gain or via long path.

• With 10-dB additional circuit gain, San Francisco-Cape-town path open as early as 1240 GMT; Dallas-Tananarive open as early as 1110 GMT.

percent of the days of the month) from mid-latitudes (38° N.) to the south of east and west, with more frequent openings to latitudes below 33° N. Transequatorial forward scatter during evening hours may furnish your best chance to work 50 MHz DX. The TE season should open on February 10 at 8 PM local time.

Late word on 50 MHz beacon frequencies and liaison schedules may be obtained by sending me a self-addressed stamped post-card.

ham radio

table 5. Predicted 40-meter path openings during the ARRL DX contest.

to from	San		
	Francisco	Dallas	New York
UAØ, Sakhalin Island	0500-1740	0500-1520	0500-1420
JT1, Ulan Bator	0720-1740	0800-1500	0800-1330 2020-0100
UA, Leningrad	2300-1530	2200-1020	1900-0920
DL, Berlin	2240-0900	2140-0900	1900-0900
VU2, New Delhi	1120-1700	1200-1400 2300-0220	1920-0240
VS1, Singapore	1000-1740	1000-1540	1930-2330
DU, Manila	0820-1740	0820-1540	0900-1500
KG6, Guam	0640-1740	0700-1540	0700-1430
VK6, Perth	0930-1620	0930-1500	1000-1340
VK3, Melbourne	0800-1600	0800-1420	0800-1320
VR2, Fiji Islands	0500-1600	0500-1440	0520-1330
FO8, Tahiti	0320-1520	0300-1420	0300-1320
CE2, Valparaiso	0040-1120	2340-1100	2240-1030
PY1, Rio De Janeiro	2340-0940	2300-1000	2130-0930
ZD8, Ascension Island	2330-0800	2200-0800	2040-0800
CR8, Angola	2300-0600	2200-0600	2000-0620
ZS1, Capetown	0000-0520	2230-0520	2030-0520
5R8, Tananarive	2300-0320	2200-0340	1940-0340

table 6. Predicted 10-meter openings.

to from	San		
	Francisco	Dallas	New York
UAØ, Sakhalin Island	2200-0300	2250-0100	None
JT1, Ulan Bator	0030-0215	0030	None
UA, Astrokhan	None	1430	1350-1440
UA, Leningrad	None	1430	1350-1440
DL, Berlin	None	1430-1630	1320-1700
VU2, New Delhi	0200	None	1340
VS1, Singapore	2320-0300	0000-0040	None
DU, Manila	2230-0300	2250-0110	None
KG6, Guam	2040-0320	2050-0200	2100-2300
VK6, Perth	2240-0400	2250-0230	2250-2330
VK3, Melbourne	2200-0420	2150-0250	2310-0100
VR2, Fiji Islands	1840-0420	1810-0250	1820-0100
FO8, Tahiti	1710-0430	1630-0300	1620-0100
CE2, Valparaiso	1450-0200	1320-0140	1210-2340
PY1, Rio De Janeiro	1430-0100	1240-0030	1140-2300
ZD8, Ascension Island	1440-0050	1220-2330	1110-2200
CR8, Angola	1510-0020	1230-2300	1120-2120
ZS1, Capetown	1430-2100	1220-2100	1110-2130
5R8, Tananarive, Malagasy	1800-2300	1310-2230	1130-2100

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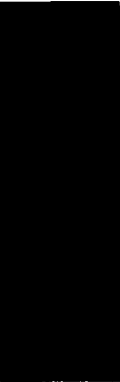
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from
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featuring
the 2N2360 transistor

“Easy to construct” and “low cost” could very well be added to the title of this article. The preamplifier described is a solid-state device that will give excellent performance on any single band from 80 through 6 meters. Its size, construction, and cost allow it to be incorporated easily into any fixed station, mobile or portable receiver or transceiver, or to be used as a mast-mounted antenna preamplifier.

A good preamplifier can perform wonders in improving reception and reducing fading on bands where a receiver lacks gain, sensitivity, or has an image problem. This is true even with expensive receivers in single bands, because compromises must be made with receiver r-f stage design due to band-switching requirements and other constraints.

A good preamplifier should have several important characteristics, the main ones probably being a low noise figure and reasonable gain. Unless the preamplifier noise figure is less than that of the receiver's r-f stage, the preamplifier self-generated noise will mask signals that the receiver itself is capable of detecting. Even in the case where the preamplifier noise figure is not less than that of the receiver, the gain provided by the amplifier will still reduce the effects of fading and raise the level of all signals.

For best weak-signal reception, low noise

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

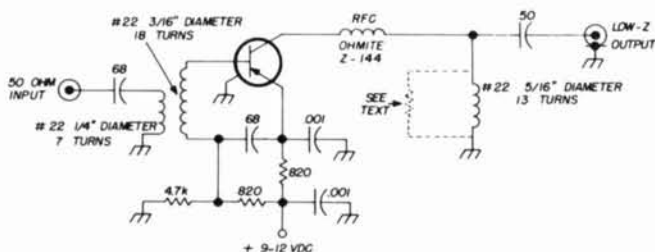
figure in a preamplifier is more important than gain, and most preamplifier designs sacrifice some gain to achieve low-noise performance. Reasonable selectivity to reject out-of-band signals and good overload response are secondary characteristics. If the receiver front end has good selectivity and overload response, then it's not necessary to incorporate these features in the preamplifier, thus keeping it simple.

The preamplifier can be easily wired to be

vhf-TV circuits, and although it may not seem as striking as an FET, it performs very well, allows very simple circuitry, and has recently become reasonably priced. (It formerly cost between \$1.40 and \$2.40 from most Sprague distributors.) If the 2N2360 is not available locally, a "substitute" type transistor should **not** be used, because its high-frequency performance will rarely match that of the 2N2360 as far as gain and noise figure are concerned.

The 2N2360 will provide a 4- to 7-dB noise

fig. 1. Broadband preamplifier. Although coil values shown here are for 10 meters, basic circuit may be used on any band from 80 to 6 meters.



switched out of the antenna circuit if the preamplifier becomes saturated by strong signals. Achieving good overload performance using a simple circuit and a transistor is particularly difficult, although rarely does any problem result with the signal levels encountered in most amateur bands. Trying to use a sensitive preamplifier with a kilowatt transmitter operating a few block away, however, is quite another problem.

the 2N2360 transistor

The heart of the preamplifier is the Sprague 2N2360. This transistor has been proved in

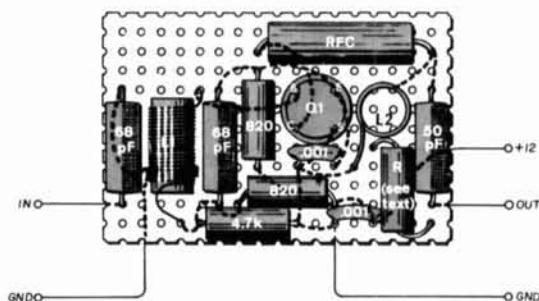
figure on any band from 80 through 6 meters. The gain will vary from about 15 to 25 dB or higher if tuned instead of broadband input and output circuits are used. The circuit used here requires a minimum of parts, and the entire preamplifier can be constructed for the cost of the 2N2360 alone if a "junk-box" selection is available.

circuit

The complete preamplifier circuit is shown in fig. 1. The basic circuit uses a grounded-emitter configuration. The input is double series tuned. A series capacitor matches the low impedance of a coaxial line. The secondary side is similarly tuned to match the 2N2360 input. As shown in fig. 1, the input circuit tunes to the high-frequency end of the band, and the secondary circuit tunes to the low-frequency end (the opposite condition could just as well be used).

Circuit resonance is easily checked with a grid-dip meter or a signal generator and receiver. Variable capacitors may be used initially to determine the necessary values and later replaced by fixed capacitors. These capacitors should have a good Q. Dipped silver micas (CDE type CD6 or similar) are recom-

fig. 2. Layout of the broadband preamp.



mended. Disc ceramics designed for bypassing should not be used, because their low Q will result in a loss of gain and selectivity.

The bias conditions, determined by the two 820-ohm and 4.7k-ohm resistors, were chosen as the best compromise between gain and noise figure. The 4.7k-ohm resistor can be varied if you want to experiment with the circuit or provide a manual gain control.

The parasitic choke in the 2N2360 collector lead prevents amplifier instability. The output circuit impedance is about 200 ohms and works quite well with simple capacitive coupling to the output coaxial line. If a more exact match is desired, a link of 2 or 3 turns can be used over the output coil. The output coil is resonated in the center of a band by using a grid-dip meter and adjusting the coil dimensions.

No loading resistor is used for operation on 80 through 15 meters and the phone portion of 10 meters. A loading resistor of about 1000 ohms is needed for complete 10- or 6-meter coverage. The only power required is from 9 to 12 volts. A battery, a simple rectifier circuit off the heater line in a tube-type receiver or any other well-filtered source may be used. Required current is less than 10 mA.

construction

A small Vector board is the handiest base for construction. With care, the preamplifier can be built to occupy about 1 inch by 1½ inch, which is small enough to fit into any receiver or transceiver. A suggested component arrangement is shown in **fig. 2**. This layout should be followed fairly closely, with particular care given to the separation and orientation of the coils to prevent intercoupling and oscillation. Normally, no instability or oscillation should exist. If this does occur, however, a shield across the middle of the 2N2360, connected to the case lead of the transistor, will solve the problem.

No special precautions are necessary if the preamplifier is installed inside an equipment enclosure, except that the ground connection should be as short and as direct as possible. If the preamplifier is used in a separate box (such as a minibox) as a mast-mounted unit, a 1-mH rf choke should be installed from the input coaxial line center conductor to

ground. This will prevent any high static voltage on an antenna from damaging the 2N2360.

The component layout of the completed universal high-gain transistor preamplifier is shown in **fig. 2**. (Note particularly that the output coil is mounted **vertically** to prevent coupling to the other inductors). The construction shown may not deserve any kudos for neatness, but it does illustrate the ease with which the preamplifier components can be assembled.

in summary

Considering ease of construction and cost versus performance, the 2N2360 preamplifier circuit is certainly one of the best. It should be possible for the average amateur to build this preamplifier in one evening for any selected band with the aid of only a grid-dip oscillator. The improvement in performance that this preamplifier provides when used with most receivers and transceivers, particularly on 15 through 6 meters, is truly amazing.

I have presented only the simplest constructional use of the 2N2360 transistor preamplifier. Many variations are possible for those who would like to construct a more elaborate unit. Discrete tuning of the amplifier instead of the broadband operation, for instance, will increase receiver gain somewhat and improve image rejection. When used as a tuned amplifier, the input circuit series capacitor is omitted, and the series capacitor in the input circuit secondary is variable. The primary circuit serves, then, as a simple coupling link. The output coil can be either broadbanded or tuned by a variable capacitor.

If you want to construct a multiband preamplifier, there is no reason why the input and output circuits cannot be bandswitched. Provision should be made for disabling the preamplifier, if possible. Although it would take an unusually strong signal to overload the preamplifier, it is worthwhile to be able to determine positively, when signal distortion does occur, whether the preamplifier or another part of the receiving system is at fault.

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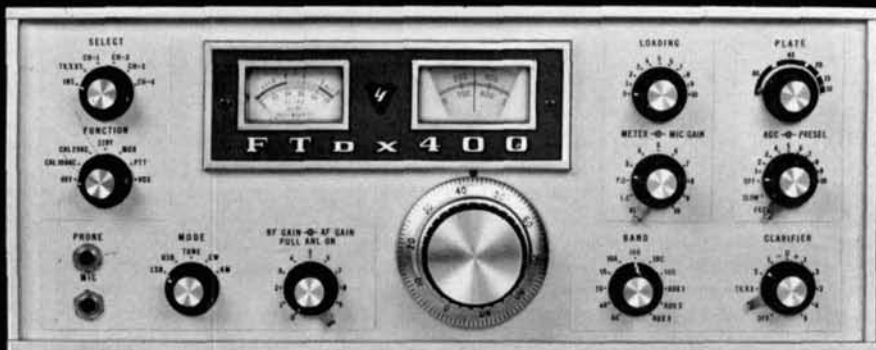
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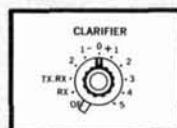
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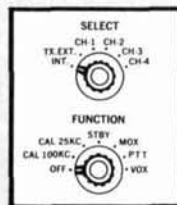
Design features include double conversion system for both transmit and receive functions resulting in, drift free operation, high sensitivity and image rejection • Switch selected metering • The FT dx 400 utilizes 18 tubes and 42 silicon semi-conductors in hybrid circuits designed to optimize the natural advantages of both tubes and transistors • Planetary gear tuning dial cover 500 KHz in 1 KHz increments • Glass-epoxy circuit boards • Final amplifier uses the popular 6KD6 tubes.

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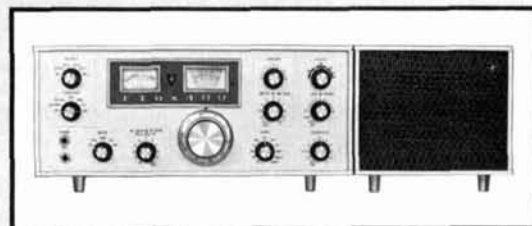


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Other projects include QRP transmitters, absorption wavemeters and field-strength indicators. All projects include complete parts lists and use parts available from almost any distributor; the experimenter won't be stymied by non-availability of needed parts.

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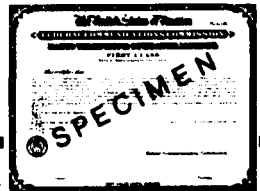
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NOISE BLANKER FOR THE SWAN 250

Westcom Engineering is now offering the TNB Noise Blanker in a version specifically designed for use with the Swan 250 transceiver.

The TNB-250 Noise Blanker effectively suppresses noise generated by auto ignitions, appliances, power lines, etc., permitting the recovery of weak DX and scatter signals normally lost in noise.

Features include modern solid state design techniques utilizing dual-gate MOS FET transistors and two stages of IF noise clipping for the efficient removal of impulse noise at the transceiver IF frequency. The use of MOS FETs and a special gain controlled amplifier circuit provide excellent cross-modulation characteristics in strong signal locations.



TNB-250 shown installed on Swan 250 by means of the pre-punched accessory holes.

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Refer to the New Products column of the August '68 issue of Ham Radio Magazine for additional information on the TNB Noise Blanker or write for technical brochure.

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motorola power transistors

Motorola has just introduced the first six members of a new line of intermediate-power transistors. These new devices, in plastic packages, will dissipate up to 8 watts with suitable heat sinking. Although the package is only about the size of a TO-5 transistor, it can dissipate approximately 1 watt at an ambient temperature of 25° C.

The npn MPSU01 and pnp MPSU51 are complementary transistors designed for use in audio amplifiers delivering up to 5 watts output. They have high gain and excellent current-gain linearity from 1 to 1000 mA. The maximum V_{CE0} rating for these devices is 30 volts, maximum continuous collector current is 1.5 amperes, and device dissipation is 8 watts.

The npn MPSU02 and pnp MPSU52 are intended for general purpose amplifier and driver applications. These silicon transistors have excellent gain linearity, low saturation voltage, minimum f_t of 150 MHz, maximum V_{CE0} rating of 40 volts and 6 watts dissipation. The npn MPSU03 and MPSU04 are 1-ampere devices for video output circuits and horizontal driver applications in TV sets. The collector-to-emitter breakdown voltage rating is 120 V for the MPSU03 and 180V for the MPSU04; total dissipation at 25° C case temperature is 8 W.

For more information on these new transistors, write to the Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

in-circuit transistor tester

Amphenol has recently introduced a new compact transistor analyzer for checking transistors—in or out of a circuit—that also doubles as a very sensitive dc voltmeter. In addition, the unit functions as an accurate diode analyzer capable of measuring both forward and reverse current. This instrument was developed by Amphenol to minimize the time and effort required to pinpoint defective semiconductors. It can be used for troubleshooting all solid-state equipment, including receivers, hi-fi systems, TV sets, fm

tuners and tape recorders.

The new **Transistor Commander** can check high- and low-power npn and pnp power transistors for in-circuit dc-current-gain characteristics, check out-of-circuit devices for current gain, I_{CBO} and I_{CEO} leakage, check diodes and rectifiers for in-circuit opens and shorts and check out-of-circuit diodes for forward and reverse currents.

The new analyzer contains a current-limiting circuit for protection against accidental shorts; this prevents shorted transistors and diodes from damaging the instrument. In addition, simplified meter scales facilitate rapid testing, output test jacks are arranged so a quick-test socket can be easily adapted and the cabinet is constructed with a tilt leg for adjusting the meter to eye level.

The Transistor Commander provides a dc-current-gain range of 1 to 1000 ($\pm 5\%$), adjustable collector current range of 0 to 10 mA and collector-base leakage and collector-emitter leakage measurements from 0 to 5 mA. In diode testing, a 100-mA limit circuit protects the diode from damage. The built-in 100-Vdc meter permits accurate measurements with the same probes that are used for the other checks.

For more information on the Model 830 Transistor Commander, write to the Amphenol Distributor Division, Amphenol Corporation, 2875 S. 25th Avenue, Broadview, Illinois 60153.

10 watts at 1000 mhz

Looking for high power solid state on the uhf bands? If so, here's a transistor that may interest you: the new S1050 from the Electronic Components Division of United Aircraft. The S1050 is a silicon npn epitaxial planar transistor designed specifically for high-power output in class-C uhf amplifier service. It uses an overlay emitter design and delivers 10 watts at 1000 MHz with 5-dB gain. The stud-mounted package is modified for optimum performance in high-frequency, high-power applications. Price, \$200. For more informations, write to Electronic Components Division of United Aircraft, Trevose, Pennsylvania 19047.



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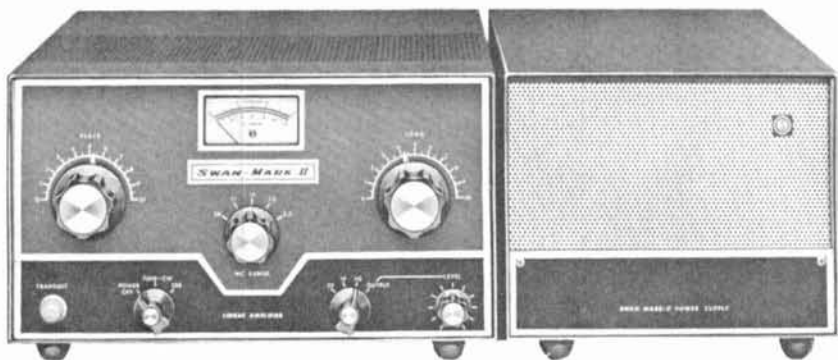
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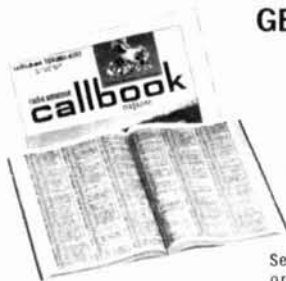
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Oakland, California 94607
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COLE ELECTRONICS SUPPLY COMPANY

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Tyler, Texas 75701
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ELECTRONIC CENTER, INC.

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Tel: 612-338-5881

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Evansville, Indiana 47710

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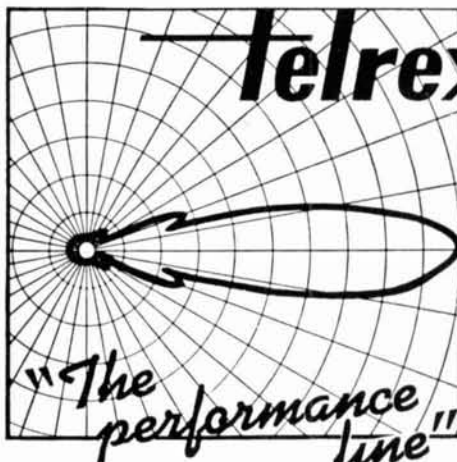
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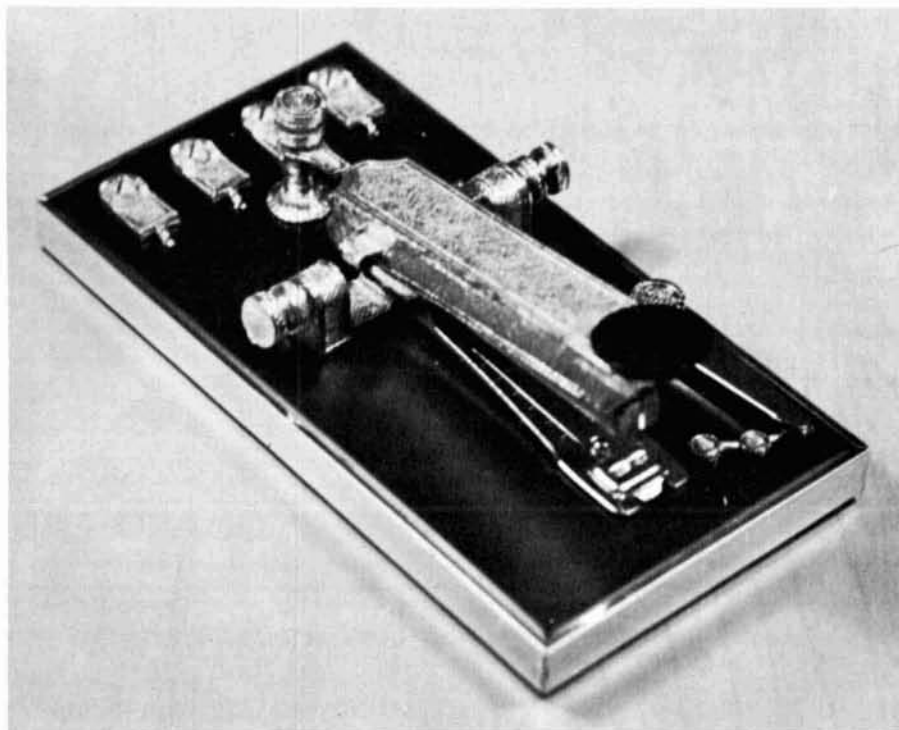
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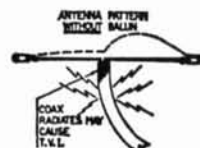
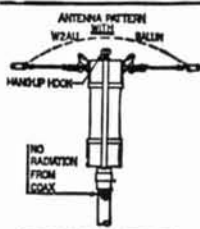


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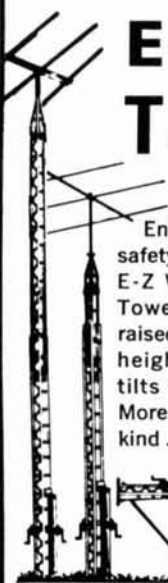
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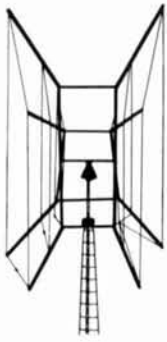


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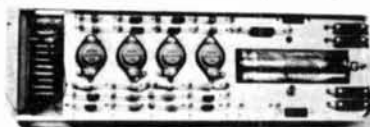
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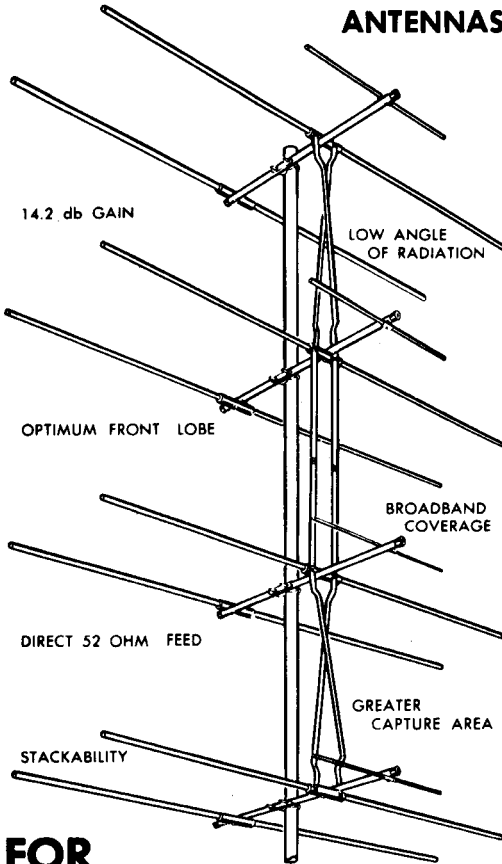
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