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2-kw
strip-line amplifier
for
432 mhz

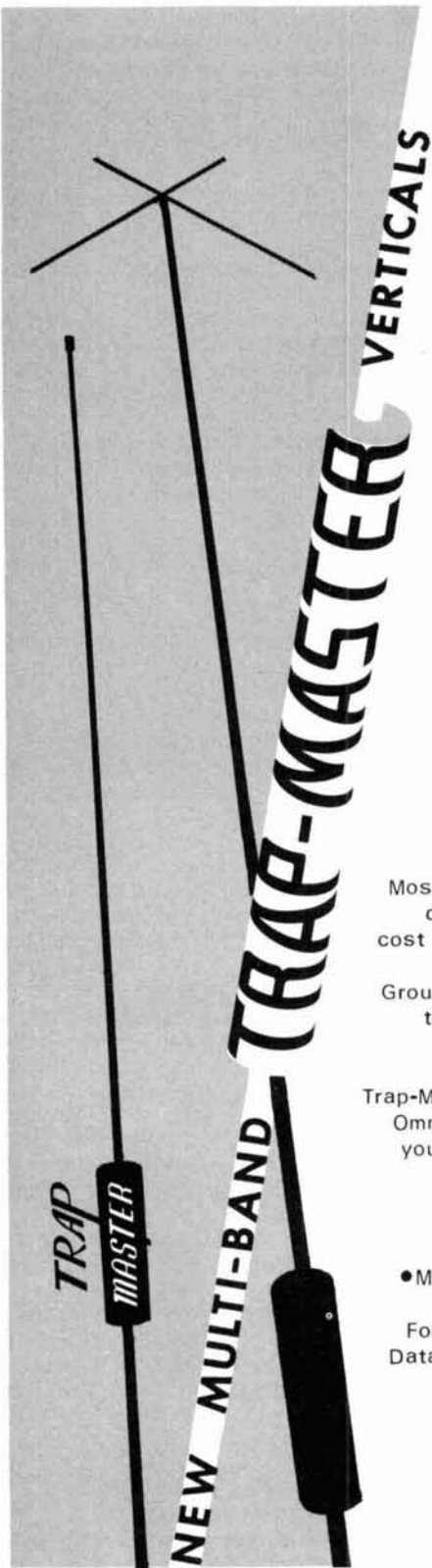


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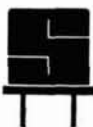
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staff

editor

James R. Fisk,
 W1DTY

roving editor

Forest H. Belt

vhf editor

Robert M. Brown,
 K2ZSQ/W9HBF

associate editors

A. Norman Into, Jr.,
 W1CCZ/K2IEO
 Nicholas D. Skeer,
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art director

Jean Frey

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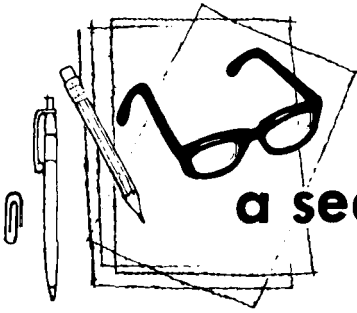
offices

Greenville, New Hampshire 03048
 Telephone: 603-878-1441

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a second look

by Jim
Fisk

In case you don't know it, the frequency spectrum above 30 MHz is in serious trouble. Communications technology has outstripped our methods for controlling it and many harried users are clamoring for more space. The FCC has traditionally allocated blocks of frequencies to various users—in the vhf and uhf range they have taken advantage of line-of-sight propagation to allocate the same frequencies to various sections of the country. However, with satellite communications, a new problem arises: since a single satellite signal blankets a large region, the same frequency can't be allocated in other areas.

In the past, when a particular band of frequencies became too crowded, we simply improved our technology and moved higher in the spectrum. However, now the only higher bands lie in the region above 10,000 MHz. These frequencies are not too attractive, even if we develop equipment to use them, because of high atmospheric attenuation.

A four-year study sponsored by the IEEE and the EIA has recommended the allocation of space on the basis of need rather than the existing "block" approach. This is where the amateur frequencies are liable to be looked at very closely. Although our bands represent a small portion of the total spectrum between 30 and 960 MHz (about 5%), a complete overhaul may reduce this—particularly if the overhaul is based on use and need!

Some of the other recommendations proposed for putting more transmitters into the same amount of space include: more "splitting" to reduce the originally-assigned bandwidth to make room for other users, narrowing the 6-MHz now allowed for TV stations so that adjacent and in-between frequencies can be used for other purposes, and tightening of standards for both receivers and trans-

mitters to conserve bandwidth. This last proposal could have some interesting effects on amateur operation—for one thing, improved standards governing receiver susceptibility could reduce TVI problems. On the other hand, improved standards might carry a "type-acceptance" clause that could eliminate our traditional privilege of building our own equipment.

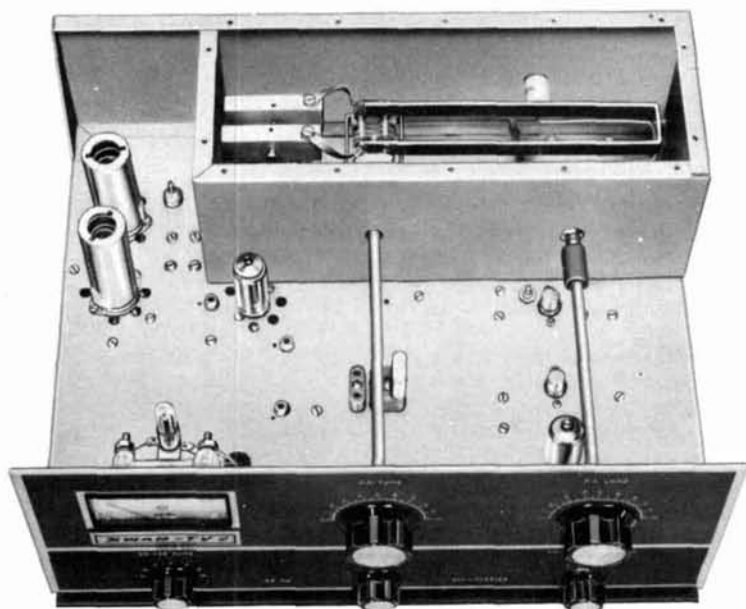
In addition to the four-year study sponsored by the IEEE and the EIA, the President's Task Force on Communications Policy is working on the problem. Although their report was originally due in the summer, it's not expected to be released until this month. Among other things, it's expected that they will recommend a new Department of Telecommunications with a Cabinet-level administrator or the creation of a telecommunications agency along the lines of NASA or the FAA.

Although no action is expected this year, things are apt to happen when the new President takes office. However, before the government makes any drastic changes, they'll probably set up a pilot project in a representative region to test the recommended frequency allocation concepts. **Now** is the time to think about vhf—not after the serious action starts. Conduct your close-in communications and rag chews on the bands above 100 MHz, resolve to get on 220 and 432, and if you're already on 432, consider going to 1296 or 2304 MHz. If you're a member of a local traffic net, consider moving from 75 meters to vhf—you'll get the same coverage with less QRM. We may not be able to save everything from other frequency-hungry services, but we should give it a jolly good try.

Jim Fisk, W1DTY
Editor

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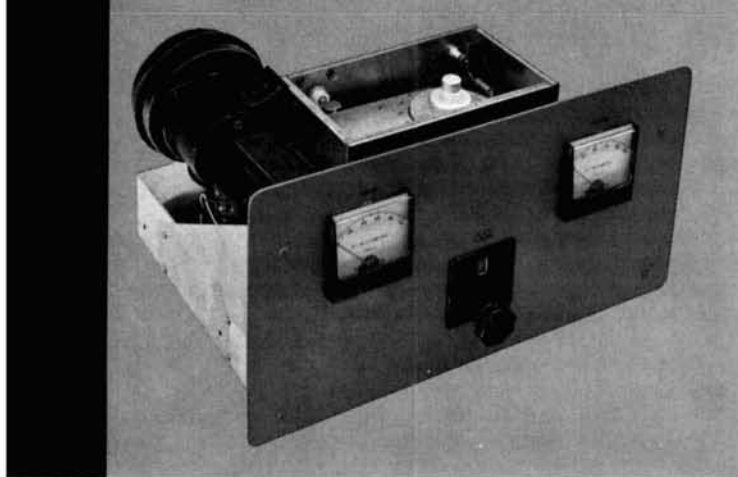
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William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California
John T. Chambers, W6NLZ, 2228 Via La Brea, Palos Verdes Estates, California

The design of a high-power linear amplifier for 432 MHz poses some interesting problems. Most important is choosing the tube to be used. Most medium-cost transmitting tubes tend to "run out of gas" in this portion of the radio frequency spectrum while others require extensive and expensive cavity enclosures to make them play. The vhf enthusiast, then, is severely limited in his approach to high power on 432 MHz, and the goal of a so-called "two-kilowatt PEP" amplifier has been outside practical radio amateur capabilities until now.

The new Eimac 3CX1000A7 ceramic zero-bias triode shows promise of superior performance in the vhf region. This interesting "bottle" resembles the well-known 4CX1000A in general shape and outward appearance, although electrically it is a cousin to the 3-1000Z. Although the 3CX1000A7 is rated for operation to 220 MHz, it has been run on an experimental basis at higher frequencies.

Suggested maximum peak-power input in the high-frequency region is 2000 watts (2500 volts at 800 mA) with maximum peak power output of 1200 watts. Peak drive power, according to the data sheet, is 67 watts. The all-important question, of course, is, "How closely can these low-frequency operating conditions be duplicated at 432 MHz where circuit losses are higher, tube efficiency lower and driving power dearer?"

Since no operational data for the 3CX-

1000A7 was available for 432 MHz, the only answer was to build an amplifier, try the tube at this frequency and obtain the results on a first-hand basis. A grounded-grid cathode-driven configuration was chosen, as shown in fig. 1.

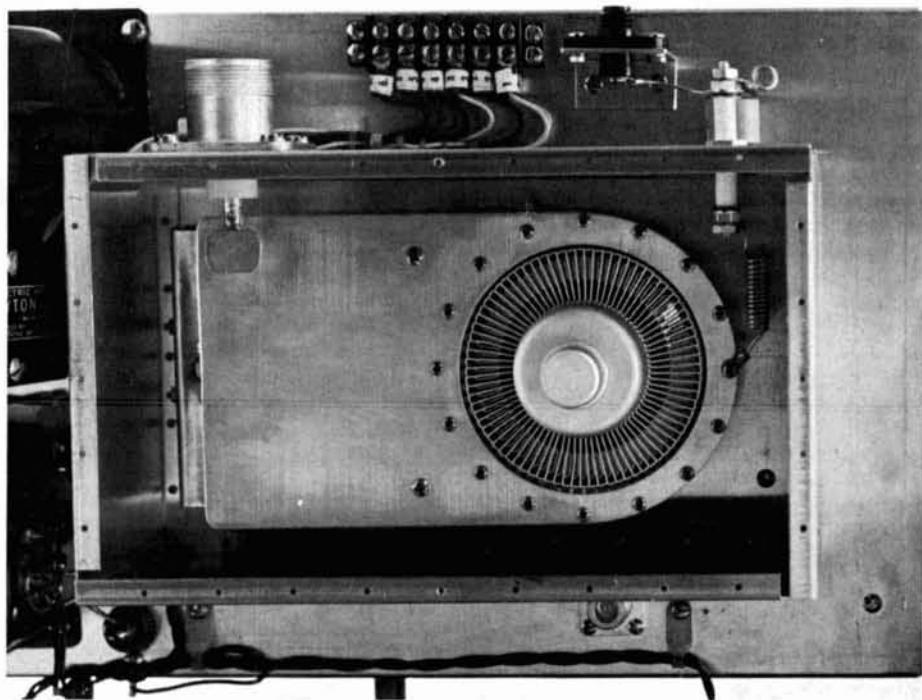
strip-line plate circuit

In the interest of simplicity, it was decided to try the simple strip-line plate tank circuit shown in fig. 2. Since the output capacitance

seemingly large value would pose problems at 432 MHz. We hoped that excessive driving power wouldn't be expended in the charging currents flowing through the input capacitance and that the power gain of the stage wouldn't be excessively low. Both of these difficulties were overcome by properly designed tank circuits.

The plate circuit strip line consists of a $\frac{1}{8}$ -inch-thick copper plate, rounded at one end and placed in an aluminum cavity box. The

The half-wavelength strip line is supported at anode height by two ceramic insulators. Anode of 3CX1000A7 is encircled by copper collar bolted to strip line; the inner circumference is lined with flexible finger stock. Plate rf choke is at right and antenna capacitor plate is mounted to coaxial plug. Edge of the plate-tuning capacitor is visible below end of strip line.

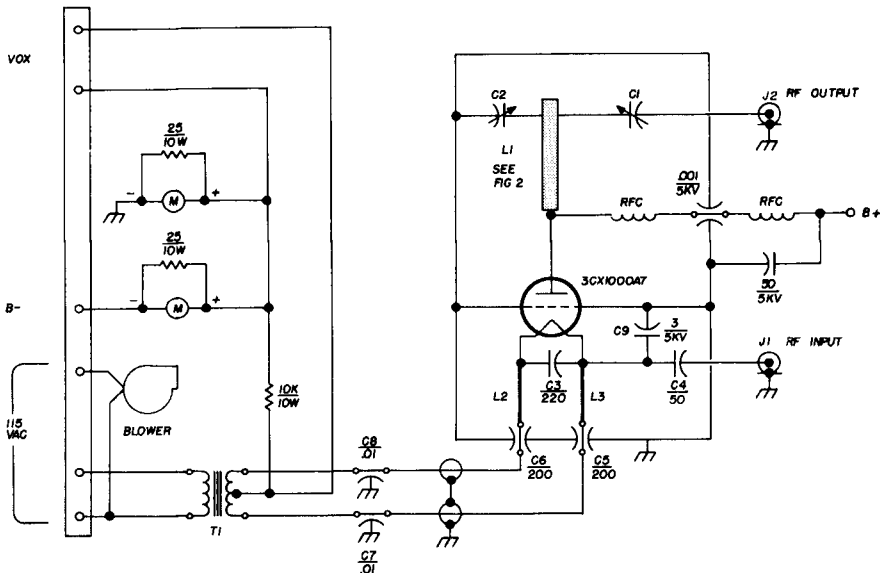


of the 3CX1000A7 is only about 15 pF in the grounded-grid configuration, a half-wavelength plate line was chosen. Even so, a large portion of the line is swallowed up inside the tube because of its internal capacitance and rather large physical size. The half-wavelength line eased coupling problems and promised better operating efficiency than a shorter line.

The input capacitance of the 3CX1000A7 is approximately 30 pF. We thought this

rounded end of the strip line encircles the anode of the 3CX1000A7 and is connected to it with a matching copper collar, fig. 3, with flexible finger stock soldered to its inner circumference.

The collar is bolted to the strip line which, in turn, is supported at the center by a pair of 2-inch high ceramic insulators. The free end of the strip line is capacitance tuned by means of a copper flipper that is hinged to the chassis and moved to and fro by the arm



- C1** Coupling capacitor. Copper tab 1" x 5/8" spaced approximately 1/2" from plate line. Tab is supported by copper rod 0.188" diameter, soldered to center pin of coaxial connector J2. Rod may be made from center conductor taken from RG-17/U coaxial cable
- C2** Tuning capacitor. Aluminum tab 1" x 4" spaced about 1/4" to 3/8" from plate line. Tab is portion of longer strip bent in an inverted L with brass hinge at bottom. Hinge is jumpered with copper shim stock to provide low-impedance ground path. Tab is moved by an eccentric arm and 3/8" diameter teflon drive rod driven by worm gear
- C3** 220-pF dipped silver mica capacitors (4 required) mounted from heater terminals to socket ring
- C4** 50 pF (Centralab 850S-50Z)
- C5, C6** 200 pF, 30 A capacity (Erie 482-463-10)
- C7, C8** 0.01 μF, 30 A capacity (Sprague 80-P3)
- C9** 3 pF. Grounded to two through-bolts of the socket assembly. Connect bolts in parallel and to one side of capacitor (Centralab 855-3Z)
- J1** UG-58A/U
- J2** UG-352/U
- RFC** 15 turns number 16, 1/4" diameter, 1-1/8" long
- T1** 5 V, 30 A filament transformer (Stencor P-6468)
- Blower** 80 cubic feet per minute (Dayton 1C-180)
- Socket** Eimac SK-870

fig. 1. Schematic diagram of the kilowatt amplifier for 432 MHz.

and worm gear arrangement shown in the photo. The antenna circuit is capacitively coupled at this end of the line.

grid circuit

The grid of the 3C1000A7 is at nominal ground potential since the Eimac SK-870 air-system socket grounds the multiple grid terminals to the chassis. Even so, the grid structure within the tube is above ground at 432 MHz by virtue of the small but discrete

cumulative inductance of the socket, grid terminals and grid cone assembly within the tube. A portion of the driving voltage appears across this cumulative grid inductance, and makes the circuit degenerative and reduces the input-to-output isolation of the tube.

Improved circuit stability and increased stage gain is achieved by adding a small capacitance between the input circuit and the grid circuit at the tube socket terminals. This partially compensates for the effects of the

inductance of the socket and internal grid structure of the tube.

filament circuit

The broadly resonant tuned filament circuit is composed of a segment of parallel transmission line run from the socket terminals to feedthrough capacitors mounted on a nearby aluminum bracket. From this point, shielded filament leads run to a second pair of feedthrough capacitors mounted on the chassis deck. The filament transformer is mounted above the deck in a corner of the main chassis out of the field of the amplifier plate-tank assembly.

amplifier construction

Amplifier construction is relatively easy and simple. The complete assembly is mounted on a 10 x 17 x 3-inch aluminum chassis, supported behind a 19-inch relay-rack panel. A space is left between the chassis and panel to accommodate the gear mechanism of the plate-circuit counter dial.

The plate-circuit assembly box is a standard aluminum enclosure measuring 6 x 4 x 10

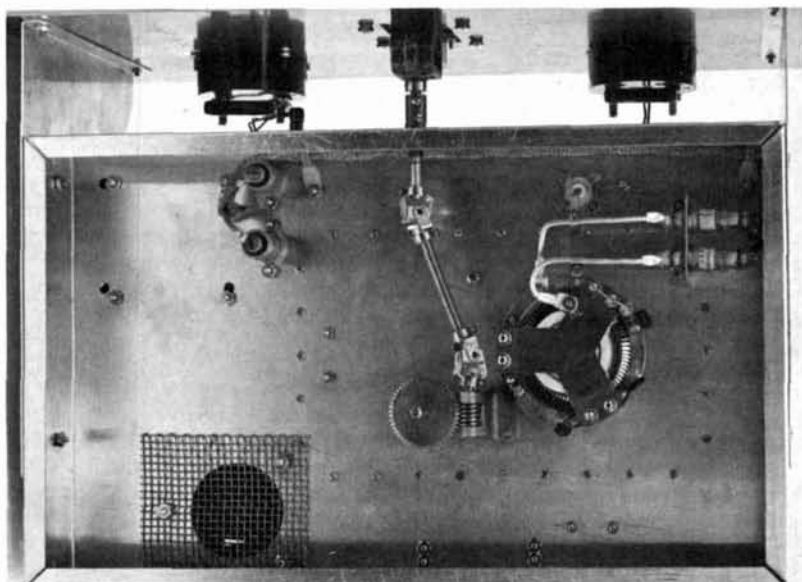
inches, with removable 6 x 10-inch sides. The box is firmly bolted to the chassis by the lips of one of the open sides and the other side serves as the top of the enclosure.

In order to cool the 3CX1000A7 properly, a blower is used, and the under-chassis area is pressurized by a bottom plate. The air is fed through the tube socket, past the anode of the tube, and exhausted through the perforated top plate of the amplifier box.

The half-wavelength anode strip line is shown in **fig. 2**. A copper ring is cut on a lathe, and the anode finger stock* is soldered to the inner diameter of the ring. The ring is drilled for sixteen 4-40 screws and firmly bolted to the strip line. The whole assembly is supported over the tube socket opening by two ceramic insulators. The electrical center of the plate line is very nearly at the outer anode diameter of the tube, so the plate rf choke is attached at this point as shown in the photograph.

* Eimac CF-300. Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog number 47E2087. \$5.80 plus postage; shipping weight 12 ounces.

Underneath the amplifier. Socket and filament lines are at right; input connector is tapped on one line at approximately mid-point. The surplus gear drive and flexible couplers connect the plate-tuning capacitor to the counter dial on the front panel.



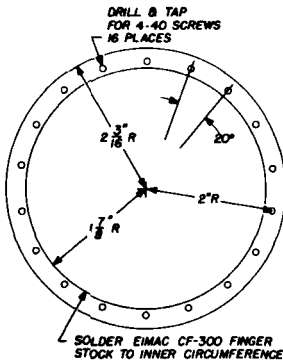
The complete anode assembly with the 3CX1000A7 in the socket resonates near the high end of the 432-MHz band. A small variable capacitance placed at the opposite end of the line from the tube permits circuit resonance across the complete amateur band.

The plate-tuning capacitor is built from a brass hinge purchased at the local hardware store. The hinge is attached to the chassis and supports an inverted L-shaped aluminum plate, which swings in close proximity to the strip line. When the top portion of the plate is parallel to the plate line, it is about $\frac{3}{8}$ -inch away from the line.

The moving part of the hinge is jumpered with a wide strap of flashing copper. The hinge provides mechanical stability—electrical conductivity through the hinge joint is provided by the flashing copper. The capacitor is varied by means of an eccentric arm driven by an under-chassis worm drive from the counter dial. A $\frac{1}{4}$ -inch diameter teflon rod insulates the drive system from the hinged capacitor plate.

The filament line is made of two 4-inch lengths of number-10 copper wire spaced $\frac{3}{4}$ -

fig. 3. Anode ring for the strip-line amplifier. Make from $\frac{1}{8}$ " copper.



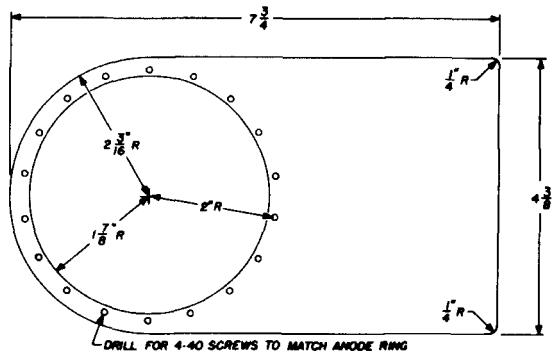
inch apart. They are bent back to reach the filament bypass capacitors mounted on an aluminum bracket in the corner of the chassis. The filament circuit tunes quite broadly and is relatively uncritical.

The rf drive point is attached to one leg of the line, near the midpoint, and may be juggled a bit to establish the lowest drive level after the amplifier is in operation. This is best accomplished with a directional wattmeter in the drive line. In the interest of con-

serving drive power, RG-225/U teflon coaxial cable is used to couple the driver to the amplifier.

The Eimac SK-870 air socket and chimney ground the multiple grid terminals of the 3CX1000A7. Four small mica capacitors are placed across the heater terminals of the socket to bypass the two sides of the heater. The 3-pF neutralizing capacitor is connected from one side of the heater to a short length of wire soldered to two adjacent bolts of the mounting socket.

fig. 2. Construction of the plate line. Make from $\frac{1}{8}$ " copper sheet and silver plate.



A small value of cathode bias—11 volts—is applied to the 3CX1000A7 to reduce the zero-signal plate current. In lieu of an expensive zener diode, a homebrew version was made by placing 13 bargain-counter 3-ampere stud rectifiers in series and using their summed forward voltage drop as zener bias. The rectifiers are rated at 50 volts PIV and provide enormous carrying capacity at minimum price. The diodes were mounted on a $\frac{1}{4}$ -inch aluminum plate with mica insulating washers and placed in the power supply unit. Alternatively, two 1N4561 50-watt zener diodes may be connected in series to provide a bias of 11.2 volts at a substantially higher cost.

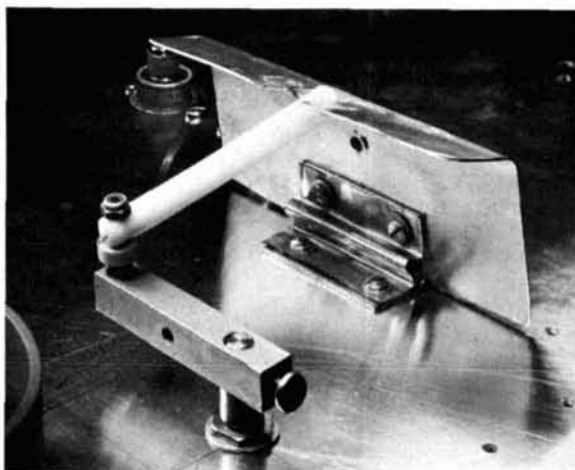
The output circuit is capacitively coupled to the plate line by means of a small, semi-variable capacitor made of a copper tab supported by the center pin of the coaxial antenna connector. The output circuit of the amplifier is designed for heavy-duty RG-17/U coaxial line. The mounting holes of the co-

axial connector are made oversize; antenna coupling may be adjusted by loosening the bolts holding the connector to the chassis and moving it about in the mounting hole.

amplifier adjustment

The filament line and plate line may be

The plate-tuning capacitor is made from an L-shaped aluminum plate and brass hinge. The hinge is shorted with copper foil for good electrical conductivity; flipper is driven with length of teflon rod and offset cam. The teflon rod is drilled at end and copper wire is passed through the rod and soldered to the plate with aluminum solder. Brass may be used for plate if desired.



grid-dipped to 432 MHz with the 3CX1000A7 in the socket and no voltages applied. Alternatively, the plate circuit may be tested cold by applying grid drive (no plate or filament voltage) and coupling a diode voltmeter to the antenna terminal. The resonance point should be achieved in the mid-range of the "flipper" capacitor. For "hot" adjustment, a dummy load and power output indicator are required since plate-current dip is not a true indicator of performance.

For initial adjustment, reduced plate voltage, 2000 volts or so, and reduced rf drive are applied to the amplifier. The plate circuit is resonated for maximum output, and the antenna coupling capacitor (C7) is adjusted for best power transfer. Coupling exists between the input and output circuits, and while the amplifier remains stable, grid

current varies abruptly with plate-circuit tuning and loading. Grid current should run about 40 percent of the loaded plate current at all times.

It may be necessary to experiment with the value of the feedback capacitor (C9) to obtain the proper ratio of grid-to-plate current at the full input level. Either a 3-pF or a 5-pF capacitor may be used. A variable capacitor is not recommended at this point because the internal inductance of such a unit is too high.

Once you have established a feel for the tuning, the plate voltage and drive may be increased to the values shown in the table. The filament voltage should be reduced to the minimum value that will provide full output—about 4.7 volts or so. This is because backheating tends to raise cathode temperature above normal. Standby bias is incorporated in the power supply and is removed for proper operation by shorting out the VOX terminals.

When antenna coupling is too heavy, resonance indication of the plate current will be very broad, and the output will be low. When coupling is too light, you'll find a sharp resonance combined with rather severe fluctuations in grid current as the plate circuit is tuned. When properly loaded, maximum power output will be achieved with the plate circuit slightly detuned from the apparent point of resonance, as noted on the plate meter.

operation

At a plate potential of 3000 volts and with 670 mA of plate current, power output into a dummy load was measured at better than 850 watts. This is an over-all amplifier efficiency of 41 percent and includes losses in the measuring circuit. Operating efficiency was estimated to be nearer 45 percent, with an actual plate-circuit power output of about 900 watts or better. Driving power was measured at 170 watts including circuit losses. Raising the plate voltage to 4000 volts permitted a plate current of 900 mA for an input of 3.6 kilowatts with a measured output of 1.75 kilowatt into the load and an over-all efficiency of about 48 percent.

A word of caution: the 3CX1000A7 has not

been rated for operation at this frequency nor for plate potentials above 3.5 kV. Experimental operations of this magnitude at this frequency exceed the warranty of the tube. In addition, it should be noted that full grid drive should not be applied without the plate circuit loaded and high voltage applied under any circumstances—otherwise the grid structure may be damaged. The data derived is based on an experimental test, and the Application Engineering department of the Eimac Division of Varian should be consulted

before using this information for final equipment design.

While this amplifier is an **experimental** unit, it points the way toward future amplifier design in the upper portion of the vhf spectrum. This design could be scaled down in size and power for a single 4CX250B running at the 500-watt level. Meanwhile, tests to determine the full capability of the 3CX1000A7 tube in the vhf region continue.

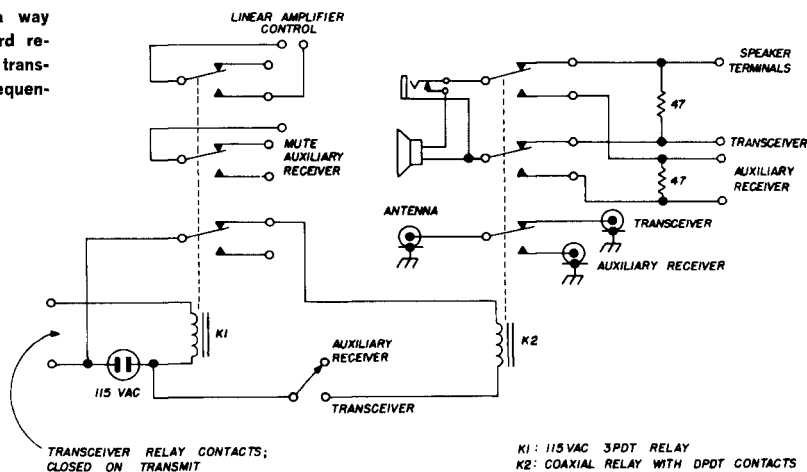
ham radio

■ using an outboard receiver with a transceiver

There are many times when it's helpful to have an outboard receiver available for listening on frequencies other than your transmit frequency. Remote VFO's accomplish this in most cases, but not always. A good example

board receiver may be switched onto the antenna and speaker at the flick of a switch. When the transceiver is switched to transmit, the speaker and antenna are automatically connected to the transceiver. Relay K1 is a

fig. 1. Here is a way to use an outboard receiver with your transceiver for split-frequency operation.



occurs when working DX stations on 80 meters. With your transceiver tuned to 3850 kHz, the preselector isn't peaked up on the European section of the band even if you have a remote VFO. Also, with a separate receiver, you can monitor other bands for activity. This is particularly useful when you're waiting for ten or fifteen meters to open up in the morning.

With the circuit shown in **fig. 1**, an out-

3PDT relay and K2 is a 115-Vac coaxial relay with auxiliary dpdt contacts. Before connecting this circuit to your transceiver, check the ratings on the external contacts of the relay in the transceiver; some of them are not designed for 115 volts. If this is the case, you'll have to go to a lower voltage 4pdt relay at K1 and use the extra contact to control the coaxial relay.

Jim Fisk, W1DTY

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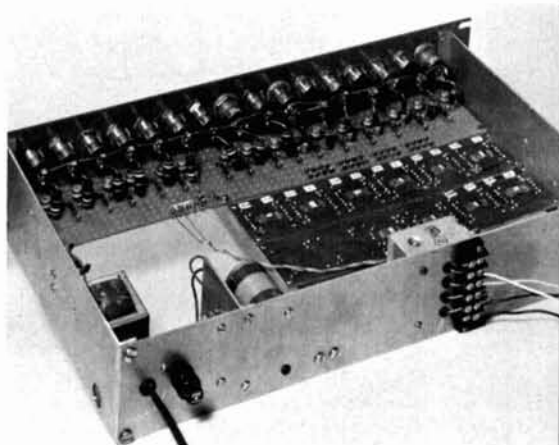


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wind-direction indication

Hank Olson, W6GXN, P. O. Box 339, Menlo Park, California 94025

If you know that the wind speed is on the increase, you can certainly lower your beam before the storm hits, but to many hams, the wind **direction** is important. This is especially true if you can't lower your beam below surrounding buildings and trees, and must rely on "turning the elements into the wind" for protection. Here is a digital wind-direction



indicator of relatively unusual design using integrated circuits.

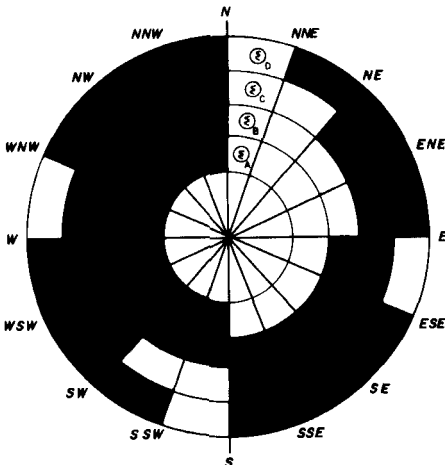
The usual method of remotely indicating wind-direction is to use a pair of selsyns. This scheme requires five wires between the selsyn mounted on the weather-head and the indicator in the shack. Selsyns, while normally very expensive, are available inexpensively as military surplus and are readily adaptable to wind-direction-indicator service. The fact that many surplus selsyns are designed for 400 or 800 Hz is not a deterrent to hams; they just use them on 60 Hz with lower voltages.

I used digital encoding for this indicator because I wanted to explore some new techniques. If you build this unit, you'll not only end up with a durable wind-direction indicator, you may also get a foothold on the rapidly expanding field of digital control and integrated-circuit logic.

digital shaft encoder

The heart of the indicator is the digital shaft encoder illustrated in fig. 1. This encoder is a very simple one that will indicate only sixteen discrete directions (points of the compass). This example is a binary-coded disc; note that there are four tracks on it. These four tracks cut the information-carrying region of any direction sector into four boxes. Since any box can be white or black, every direction sector has a different pattern of black and white boxes. If you doubt this, go

fig. 1. Binary-coded direction wheel.

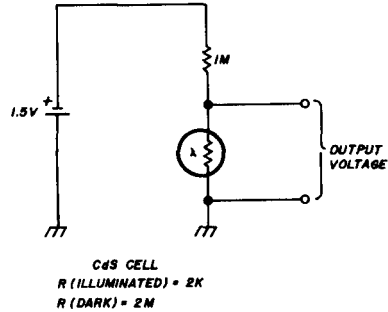


through them and check. These black and white patterns can easily be converted to voltages by lamps and cadmium sulphide (CdS) photocells and transmitted to a remote readout. The wires necessary would be six; one common, four for CdS cells, and one for the lamp-supply voltage.

As an example, the system could be connected as shown in fig. 2, with a + 1.5-volt cell to determine the logic-level. Assume a CdS cell "dark resistance" of 2 megohms and an "illuminated resistance" of 200 ohms. The output will be +1 V when a cell is covered by a black box and nearly zero when covered by a white (clear) box. This gives the series of outputs listed in table 1. You may recognize this series of numbers as the binary equivalent of our decimal numbers zero through 15, if the segments were numbered clockwise starting at N (north).

However, for direction indication, the ordinary binary-coded disc has a major drawback

fig. 2. Cadmium-sulfide cell circuit.



—if there are CdS cell misalignments, errors as great as 180° can occur. For this reason, we will use the Gray-scale coded disc shown in fig. 3. Notice that on this disc, misalignment of a CdS cell can't cause an error greater than one sector. The voltage outputs using this disc are listed in table 2. This electrical code will appear at the shack end of a multi-conductor cable and resistor-battery combination.

digital decoding

The binary code is used to actuate sixteen lamps that are labeled with the sixteen points

table 1. Output from a binary-coded disc.

Direction	Binary Output	Voltages
N	0000	
NNE	0001	
NE	0010	
ENE	0011	
E	0100	
ESE	0101	
SE	0110	
SSE	0111	
S	1000	
SSW	1001	
SW	1010	
WSW	1011	
W	1100	
WNW	1101	
NW	1110	
NNW	1111	

of the compass (N, NNE, NE, etc). The decoding will be done with integrated-circuit four-input gates. These particular gates belong to the DTL (diode-transistor logic) family, but any of the other logic families would be used in a similar way. The main reason I used DTL logic was because DTL units were available at a very low price.

The 930 series of DTL integrated circuits come as close to an industry standard as any digital IC made. This series is made by Fairchild, Motorola, Raytheon, Sylvania and Texas Instruments, as well as several others. With such industry-wide acceptance of the 930 series, and with so many companies

fig. 3. Gray-scale direction wheel.

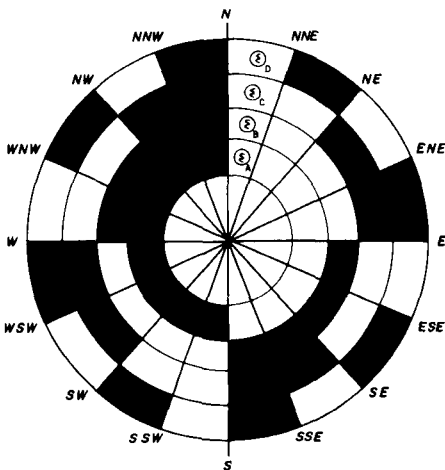
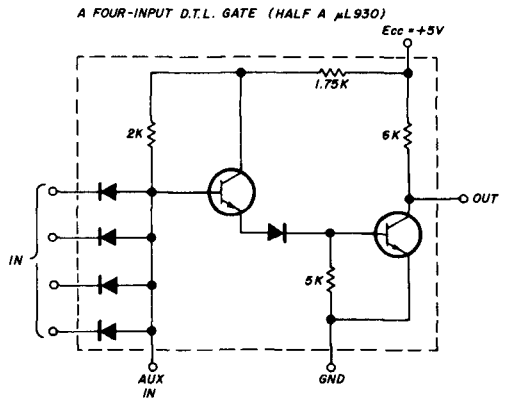


table 2. Output from a Gray-scale coded disc.

Direction	Gray-scale Output	Voltages
N	0000	
NNE	0001	
NE	0011	
ENE	0010	
E	0110	
ESE	0111	
SE	0101	
SSE	0100	
S	1100	
SSW	1101	
SW	1111	
WSW	1110	
W	1010	
WNW	1011	
NW	1001	
NNW	1000	

grinding them out to mil specs, it is only natural that a large quantity of rejects are available through the surplus electronics emporiums. Such is the case in the San Francisco Bay Area, where 930 DTL series units are produced, and where one local surplus store

fig. 4. A four-input DTL gate using one-half of a Fairchild μ L930.



offers them at ten for a dollar—mixed, unmarked, and of unknown worth. I built the wind-direction indicator shown in the photos from 930's and 946's that were gleaned from \$5.00 worth of such offerings.

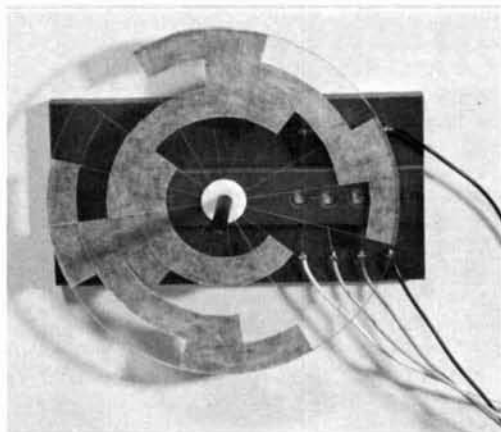
Several large mail-order companies who specialize in surplus semiconductors have 930 DTL series IC's for sale at 50c to \$1.50 each, depending on whether they are tested or not. While it's true that these are IC's

from a production line turning out units which sell for \$10 and up, they may or may not be bargains. The reason for this is the Motorola MC830P DTL line, in the plastic dual-inline package.

The MC830P line is electrically identical to the 930 series but it is not in a ceramic package nor is it tested to military specifications. The MC830P and MC846P are \$1.55 and \$1.65 respectively. This pricing makes the surplus mail-order units much less attractive because the MC830P series has defined specifications.

The way these DTL gates are used is the crux of this system. To understand it we must take a look at the basic DTL gate. The circuit of a 930 four-input gate is shown in **fig. 4**. With **all** four inputs open-circuited, or connected to +1.5 volts, the output level will be nearly zero since both transistors are conducting. If any one or more inputs are

Gray-scale code wheel on the cadmium-sulfide cell mount. If you look closely, you can see three of the CdS cells. The tape side of the wheel is turned down so that it appears reversed from fig. 3.

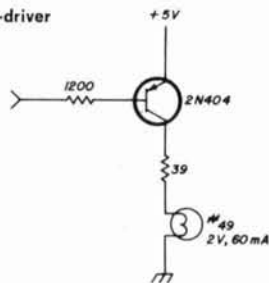


grounded, both transistors stop conducting, and the output level goes up to at least 3 volts. We can sum this up by saying that the output is low when **all** the inputs are "high," and the output is "high" when **any** input is "low."

The only exclusive state for the gate is when **all** the inputs are high, the "and" condition. That is, input number 1 is high **and**

input number 2 is high **and** input number 3 is high **and** input number 4 is high. This is called NAND logic (short for Negative AND) since the output has a logic state opposite from the inputs. Since the exclusive state will

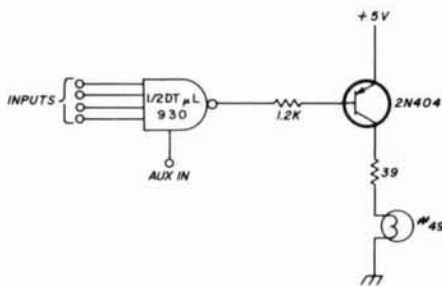
fig. 5. Lamp-driver circuit.



be the one that lights the indicator, a "low" gate output is used for the turn-on signal. The circuit of **fig. 5** will turn the lamp on when driven by a "low" gate output.

The circuit of **fig. 4** is generally abbreviated in logic diagrams. A combination of **fig. 4** and **5** is drawn in **fig. 6**. It is the circuit of **fig. 6** which is duplicated sixteen times—once for each indicator lamp at each point of the compass. Each one of these circuits has the same function: when all four inputs are "high," the lamp lights. If the sixteen indicator circuits all work the same way, how is the lamp that corresponds to the digital code-wheel position lighted? **Figs. 7** and **8** show the circuitry that does the job.

fig. 6. Lamp-driver circuit and four-input gate.



Unlike the example of **fig. 2**, no 1.5-volt cell is used to generate zero and 1-volt logic levels in this system. If a dark square is des-

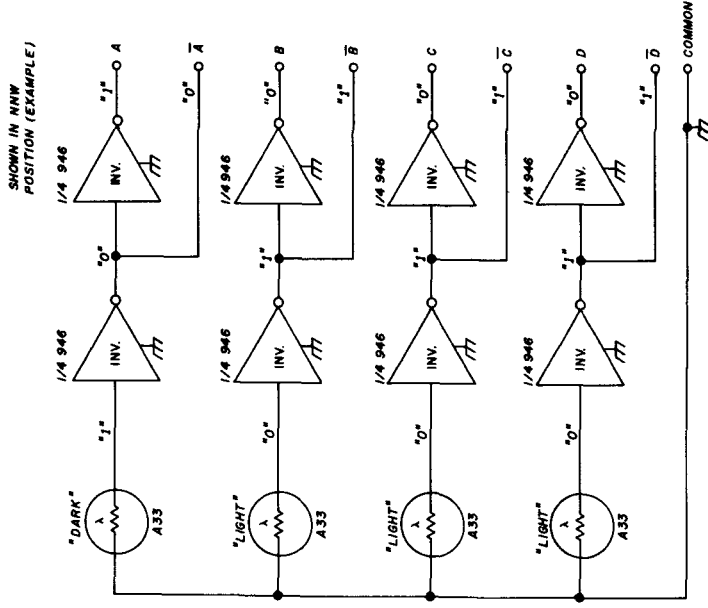


fig. 7. Circuit for generating a logic output from the gray-scale direction wheel.

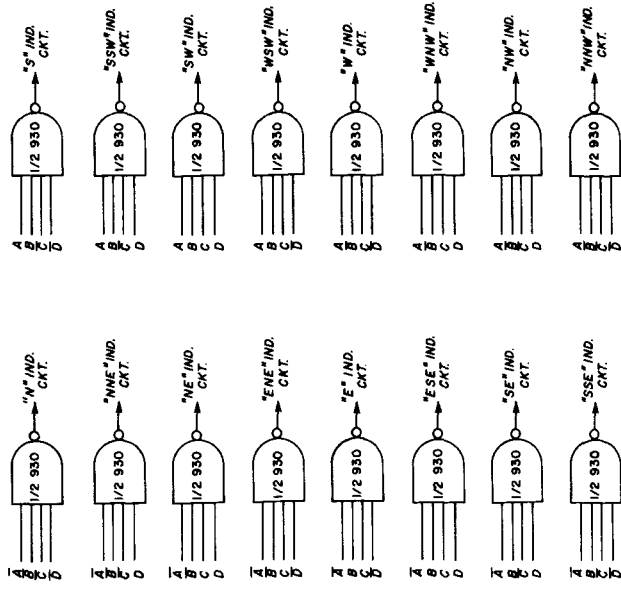


fig. 8. Decoding the logic output from the gray-scale direction wheel.

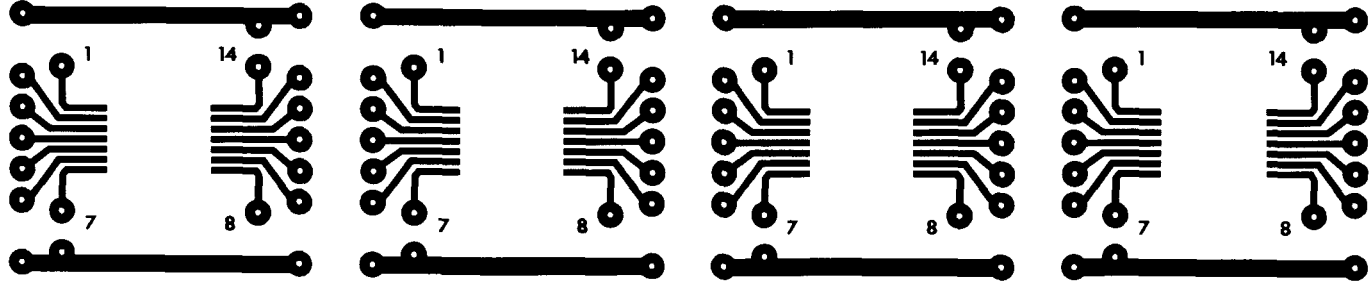


fig. 9. Layout of the etched circuit boards used to mount the integrated circuits. Wiring between IC's is done with small insulated wire.

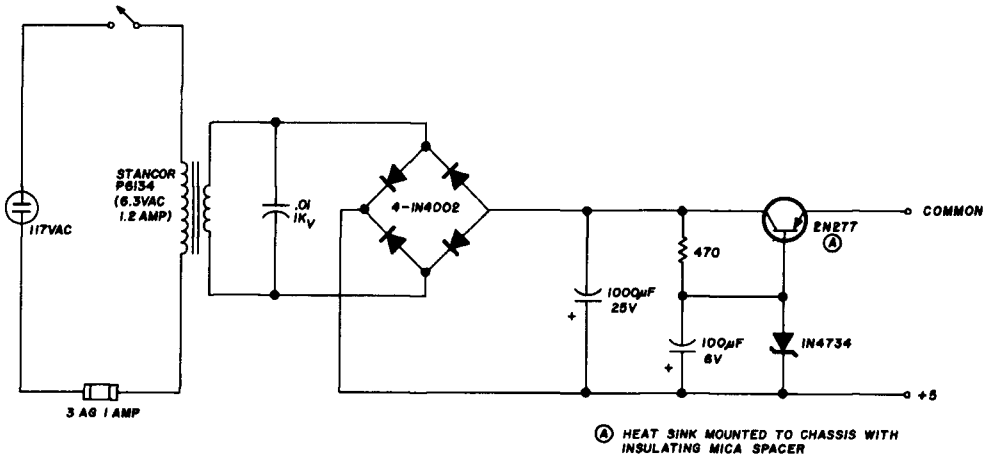


fig. 10. Regulated power supply for the direction indicator.

ignated as 1 and a white (clear) square as zero, the resistance of any of the four CdS cells is low for zero and high for 1. By connecting each CdS cell from input to ground of a DTL inverter (a DTL gate with only one input used) we will get an output with the correct DTL logic levels, but inverted in sense. Another inverter is added to restore the sense of the input signal.

If the CdS cells are designated as A , B , C and D , the eight outputs will be A , \bar{A} , B , \bar{B} , C , \bar{C} , D , and \bar{D} . \bar{A} is the inverse of A , or as the logic designers say, "not A ." So \bar{A} will be 1 when A is zero and zero when A is 1. \bar{B} , \bar{C} and \bar{D} are "not B ," "not C " and "not D ." By connecting A , \bar{A} , B , \bar{B} , C , \bar{C} , D and \bar{D} to the four-input gates as shown in fig. 8, the correct gate (and only the correct gate) will always end up with four 1 inputs when the wind vane points to the direction that corresponds to that gate. You can check this by methodically going through and checking (by penciling in zero and 1) for each possible input condition (direction) and determining which gate has all its inputs at 1.

construction

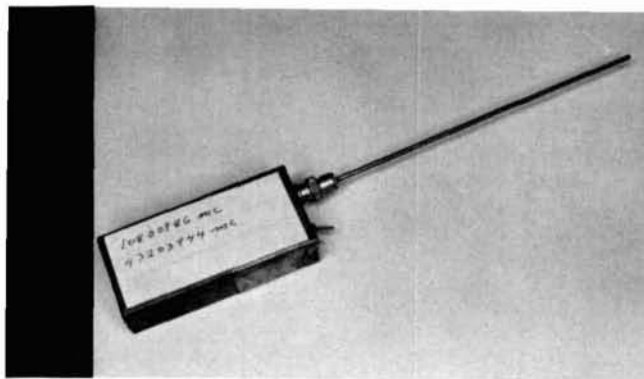
The details of construction are shown in the photos. The indicator unit is built around a 3-1/2-inch rack panel with odd bits of scrap aluminum to form a box-type arrangement. The power supply is located in the right-rear corner. The indicator circuits are built on perforated Vector board (64AA18)

in the front of the unit right next to the indicator lamps. The IC's are mounted on two six-section strips of etched-circuit board at the left center. The layout of these etched-circuit boards is shown in fig. 9. Note that an rf-interference filter was put in each of the four input signal lines to prevent false-triggering the DTL circuits with your transmitter. The particular filters I used are simple toroid-L, coaxial-C filters from the surplus store, but any good rf lowpass filter should do as well.

The Gray-scale encoder wheel is cut from 1/8-inch plexiglass, scribed, and the dark sections pasted on with masking tape. The masking tape is then colored with a black felt marking pen. The CdS cell assembly is made from odd pieces of phenolic; the cells are held in the holes with GE RTV Silastic which is widely available for caulking showers.

Although this remote-wind direction indicator only indicates sixteen point of the compass, it eliminates most of the moving parts associated with remote indicators—in addition, you don't have to worry about a long multiconductor selsyn cable. The sixteen indicator points have another advantage—if you want to tie an antenna rotator to the wind-direction system, small deviations in wind direction won't activate the rotating system. If you want more than sixteen points of the compass, a more complex coded wheel will do the job.

ham radio



calibrated
signal source
for
432 and 1296 mhz

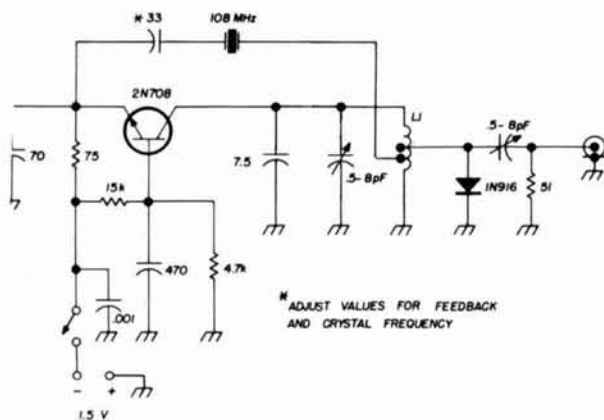
Find your way
on 1296
and 432 MHz

Del Crowell, K6RIL, 1674 Morgan St.
Mountain View, California 94040

A stable signal source for the uhf bands is a very useful item for all vhf and uhf experimenters. The circuit shown here is simple, has good stability and is very portable. It puts out a strong signal on 432, and when it was carried three blocks from home a strong signal was received.

The signal source is simply a 108-MHz crystal-controlled oscillator using a single 2N708 transistor with a 1N916 diode connected from the output tap on L1 to ground for generating harmonics. When the output is displayed on a Hewlett-Packard spectrum analyzer, the twentieth harmonic is still quite large. The spectrum chart in **fig. 2** shows the output before and after the diode was in-

fig. 1. The calibrated uhf signal source. L1 is 5 turns number 20, space wound on $\frac{1}{4}$ " diameter; tapped 1 turn and $2\frac{1}{2}$ turns from ground end.



stalled. The General Electric 1N916 was recommended by K6UQH, and I find it does very nicely as a multiplier. You can also use this diode in local oscillator chains for 432- and 1296-MHz receiving converters; for 50 cents, it's a very good varactor.

The signal source can be built into most any type of package—I used a home-made

sheet-brass box $2\frac{1}{2} \times 5 \times 1$ inch. Make all the leads as short as possible. A $6\frac{1}{4}$ -inch antenna made from number-12 copper wire can be used for both bands.

This is an excellent signal source for tuning antennas and adjusting receiver front ends. Stability is very good, and I use it on 432 MHz for frequency calibration. The

Construction of the calibrated signal source.

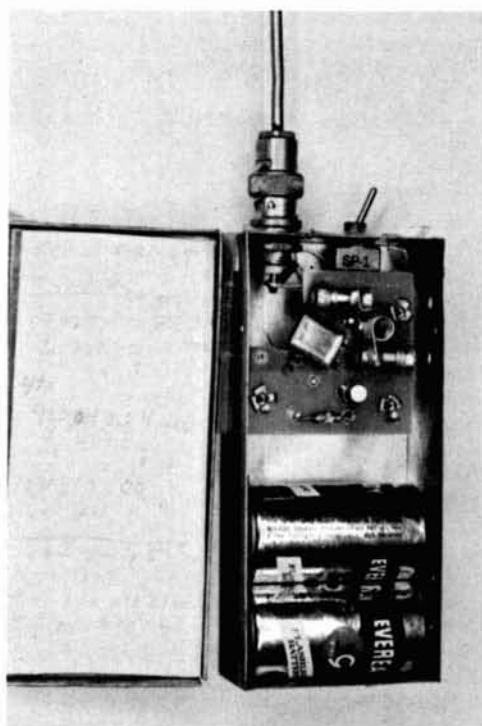
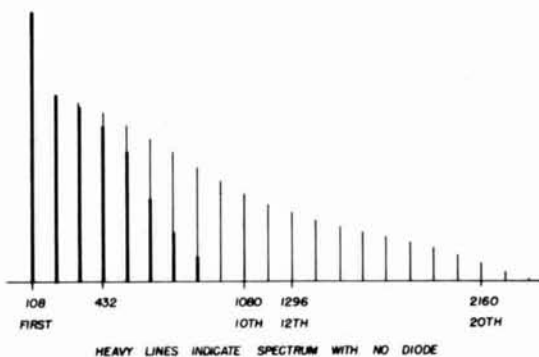


fig. 2. Spectrum output of the signal source as measured with a Hewlett-Packard 8551 analyzer. Heavy lines indicate spectrum without diode.



oscillator is checked periodically on a Hewlett-Packard frequency counter; usually it is within 1kHz at 432 MHz. Drift is very slight—with a stable BFO, I can only detect approximately 200-Hz drift during a 10- to 15-minute period. This is apparently due to slight voltage drops in the flashlight cells.

ham radio

frequency translation in ssb transmitters

After an ssb is generated,
it has to be
put on at least
one of the amateur bands;
here's how
it is done

Forest H. Belt, 2610 Whittier Avenue, Louisville, Kentucky 40205

Single-sideband signals for ham communications are almost never generated at the operating frequency of the ssb transmitter. For example, a transmitter output consisting of either the upper or lower sideband of 14.25 MHz is not actually generated at that frequency. No matter what the output frequency of the transmitter, sidebands are developed in the balanced modulator with a constant "carrier" frequency.

The fixed-frequency sideband is changed to the several operating frequencies through what is basically a heterodyne process. The sideband is mixed with a pure rf signal; they beat together and form a new sideband signal near the desired frequency. The process has several names. The most common is **frequency conversion**. But, in transmitters, to distinguish from the similar process in receivers, the term **frequency translation** is more accurate.

the simplest system

You can understand the basics of the process easily if you refer to **fig. 1**. The block diagram illustrates the simplest form of frequency translation.

A crystal oscillator generates the carrier for modulation. Its signal is mixed with voice signals in the balanced modulator, producing a double-sideband signal with the carrier eliminated. A sideband filter, either mechanical or crystal-lattice, trims off the unneeded sideband. All that is left is the one sideband of the initial carrier frequency.

To translate the desired sideband upward to an operating frequency, a heterodyne mixer is used. A variable-frequency oscillator (VFO) furnishes a signal that beats with the sideband from the filter and produces a sideband at the desired frequency. In the

process, the simple mixer can't avoid also producing a carrier at the VFO frequency and an image sideband (as far from the VFO carrier as the desired sideband, but on the opposite side).

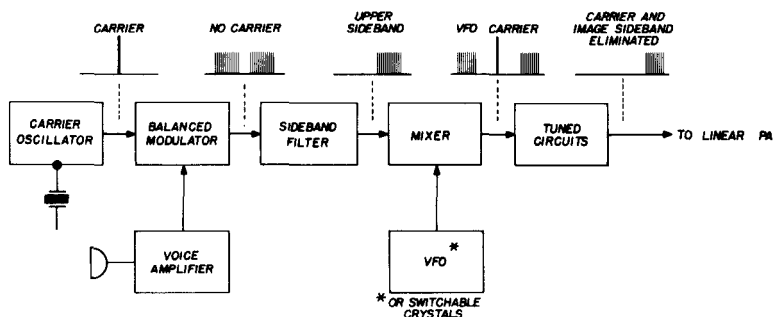
Ordinary tank circuits, tuned to the desired sideband, eliminate the carrier and the unwanted sideband—neither of which is very close to the frequency of the wanted sideband. The sideband, which is now the sideband of the operating frequency, is fed to the linear power amplifier.

The reasons for going through this process may not be obvious. First of all, the isolated sideband can't be raised in frequency by simple frequency multipliers, as in non-ssb transmitters, because they would lose their identity completely. In the second place, a

A typical double-heterodyne system is diagrammed in **fig. 2**. The diagram includes more detail of an actual transmitter than did **fig. 1**, yet it is still simplified. Also included are frequencies as they occur in one model of transmitter; they will help you understand exactly what's happening in a transmitter like this.

The carrier oscillator in this one is in the 455-kHz range. (Others include 1.65 MHz, 2.2 MHz, 3.3 MHz, 5.5 MHz, and 9 MHz.) To pick which sideband will be generated, the carrier frequency is shifted above and below the nominal 455 kHz; the two frequencies are listed on the diagram. I'll base my explanation of the system on generating a lower sideband (lsb) in the transmitter output; the carrier oscillator runs at 453.65 kHz.

fig. 1. Simplest means of translating a sideband from a carrier-generated frequency to an operating frequency.



constant carrier frequency in the balanced modulator means that the resulting sidebands can always be fed to the same filter. If there were a lot of different frequencies, a different filter would be needed for each one. It's much easier to heterodyne or **translate** the fixed frequency up to the various desired ones.

double heterodyning

Not many ham transmitters use the simplest single-translation version just described—only a couple of kit-type models, that I know of. Such systems are not very effective at producing high output frequencies. Therefore, in multiband ssb transmitters and in those for vhf use, something more elaborate is preferable. A double heterodyne arrangement can produce the higher output frequencies needed. It's the most popular frequency-translating system found in ham transmitters.

Mixed in the balanced modulator with the .1–3 kHz voice signals, the carrier produces a pair of sidebands. The lower sideband contains frequencies from 450.65 to 454.55 kHz (the differences between the lsb carrier frequency and the voice frequencies). The upper sideband contains frequencies from 453.75 to 456.65 kHz (the sums). The mechanical filter sharply chops off the upper sideband, leaving only a single sideband encompassing 450.65 to 453.55 kHz. For simplicity, this can be called the lower sideband of 455 kHz, even though there is some separation from that frequency. The carrier itself is eliminated in the balanced modulator.

The first frequency translation takes place in the first mixer. The VFO is tunable from 2.5 to 2.7 MHz; a frequency of 2.6 MHz (2600 kHz) is chosen for the example. Again, because of the heterodyne process, two side-

bands are produced, but they are far apart. The desired one encompasses the sideband frequencies from 3050.65 to 3053.55 kHz; the other is an "image," with frequencies from 2146.45 to 2149.35 kHz. The desired sideband is still a lower sideband even though it has an "upper" position with respect to the image. The desired sideband is the lower sideband of 3055 kHz ($2600 + 455$ kHz).

The tuned circuits that follow the first mixer get rid of the 2.1-MHz sideband, being tuned to the vicinity of 3 MHz. The VFO carrier doesn't appear in the output of this mixer, as it did in **fig. 1**, because the mixer is a **balanced mixer**. It's a close relative of a balanced modulator and cancels whatever rf carrier is applied to it. Translation therefore affects only the sideband that is applied to the balanced mixer.

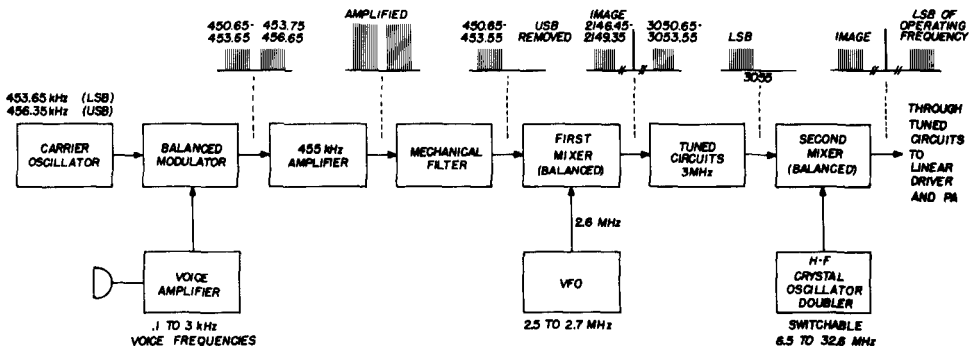
The 3055-kHz sideband must still be raised to the operating frequency. The second translation is handled much like the first. A switchable crystal oscillator supplies an rf

8.6775; but its oscillator is a doubler, so its operating frequency is 17.3550 MHz. The lower sideband of 14.25 MHz lies from 14.2470 to 14.2499 MHz. The sideband signal fed to the second mixer must therefore be the difference between those frequencies and the hf crystal frequency; that sideband covers from 3.1051 to 3.1080 MHz.

For the first mixer to produce that sideband for the second mixer, the VFO must be set at the difference between it and the input sideband from the filter. Calculating the differences, you can find that the VFO must produce an rf signal at 2654.45 kHz. (You can subtract 450.65 from 3105.1 kHz or you can subtract 453.55 from 3108.0 kHz; those are the limits of the lower sideband coming from the mechanical filter and the limits of the sidebands to be developed by the first mixer.)

On the front of the transmitter, the hf-crystal switch would point to the 14.2-MHz sector of the 20-meter band, and the VFO

fig. 2. Double heterodyning gives the operator a wide choice of frequencies, grouped into the several bands.



signal for the second balanced mixer. Beating with the sideband signals that were produced in the first mixer, the rf signal develops a single-sideband signal in one of the ham bands. The band depends on the crystal selected in the hf oscillator, and the exact frequency depends on the setting of the VFO. An example will show you how this works.

Suppose you want to produce the lower sideband of 14.25 MHz. You set the switch of the hf oscillator to the crystal that places the output frequency in that vicinity. The crystal for this happens to have a frequency of

dial would indicate 50 kHz. The combined readings would signify an operating frequency of 14.25 MHz. The transmitter output would be the lower sideband of that frequency.

frequency synthesis

Developing bands of frequencies by one translation and developing the frequencies within that band by another are excellent reasons for using double and triple heterodyne systems. Frequencies can be spread out wider than with any other system. Bands can even be sectored, and the VFO range used to cover

only a portion of each ham band—thus spreading the frequencies even wider and making it that much easier to tune a particular operating frequency.

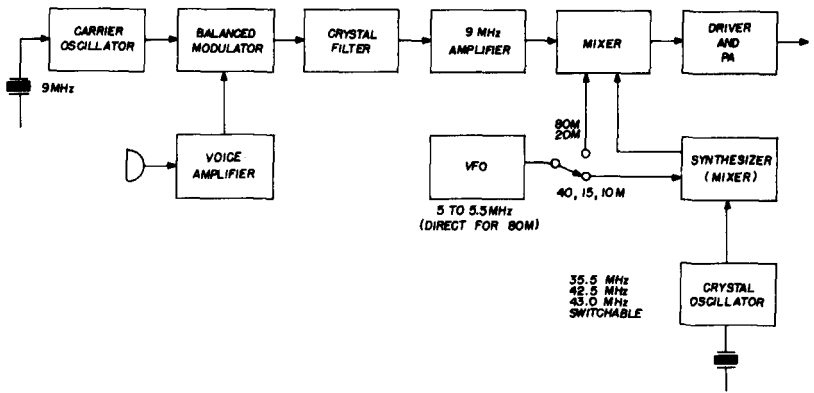
There's another way this can be done—by a method called **frequency synthesis**. The chief principle behind synthesis is illustrated in the transmitter diagramed in **fig. 3**. For simplicity, the frequencies are marked without reference to the sidebands; you know that what comes through the 9-MHz ampli-

If the desired operating frequency is to be, say, 28.9 MHz, the synthesizer must supply an rf signal at 37.9 MHz to beat with the 9-MHz sideband signal coming from the filter and amplifier.

How does the synthesizer create such a signal? It mixes the signal from the 43-MHz crystal with a signal from the VFO. To synthesize a 37.9-MHz signal, the VFO must be set to generate a 5.1-MHz signal.

From the panel of the transmitter, with

fig. 3. The basic setup for frequency synthesis. The translating signal itself is heterodyned up before it is mixed with the sideband for the translation process.



fier is actually one sideband or the other of 9 MHz. The same is true of the frequencies following the mixer.

The switching from band to band, and the tuning within bands, is all accomplished before the rf signal is mixed with the sideband signal. Developing all the various rf mixing signals artificially is where the term **synthesis** comes from. In commercial multi-frequency transmitters, it is done entirely with crystals; a few crystals can synthesize hundreds of individual frequencies by the heterodyne translation method.

How the transmitter in **fig. 3** works is not hard to figure out. The 9-MHz carrier oscillator is common in modern ham transmitters. After the balanced modulator, the sideband filter, and some amplification, the single sideband is applied to the mixer. There, the translation process is simple—just a single heterodyne. The synthesizer (sometimes called **heterodyne mixer** or **premixer**) must supply a signal that will heterodyne with the 9-MHz sideband to form the sideband of the desired operating frequency.

the viewpoint of the operator, it looks like this: the bandswitch knob is tuned to cover the segment of ham band from 28.5 to 29 MHz; this selects the 43-MHz crystal. The VFO dial is twisted until it reads .9; this represents 900 kHz (.9 MHz) on the dial and sets the VFO frequency at 5.1 MHz. The synthesizer mixes the 43-MHz and the 5.1 MHz signals. A tuned circuit that was selected by the bandswitch control picks off the difference between the two, or 37.9 MHz, which is fed to the main mixer. There, the 37.9-MHz signal beats with a sideband of 9 MHz; another tuned circuit picks off the difference, which is the sideband of 28.9 MHz—the desired operating frequency.

Other crystals and mixing schemes in this transmitter produce the other frequencies in the ham bands that are used for ssb. In some bands, the VFO is fed to the mixer directly to produce the desired operating-frequency translation.

triple heterodyning

From the words, you can figure out that

a transmitter with triple translation is one with three mixers. And, of course, it also needs three oscillators in addition to the carrier generator.

You can probably picture the arrangement in your mind. After the balanced modulator and sideband filter, the sideband signal goes to a first mixer where a crystal-generated signal is beat against it to produce a sort of intermediate-frequency sideband signal. At the second mixer, a VFO puts in a signal to tune the sideband signal **within** each band. A *third mixer*, usually with crystal switching, translates the sideband signal to bands or segments of bands. In other words, a triple-heterodyne system works like a double system with an extra stage of mixing in front of it.

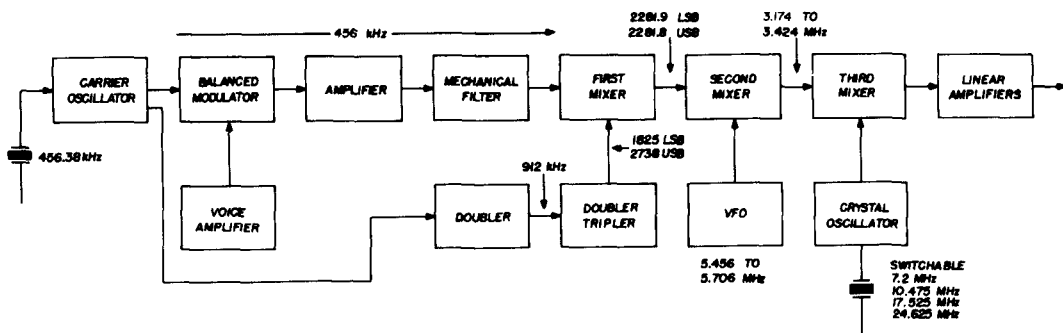
An interesting example of triple translation in a ham ssb transmitter is in the Sideband Engineers SB-34 transceiver. **Fig. 4** is a block diagram of it. An interesting thing about this one is the use of the carrier oscillator to also furnish the rf signal at the first mixer. By careful choice of the carrier frequency, the designer has also come up with a novel way to shift sidebands.

band near 2281.9 kHz. If the stage is operating as a tripler, the signal going to the first mixer is 2738.2 kHz. That translates the 456.38-kHz sideband signal to a sideband near 2281.8 kHz.

The VFO generates a signal that is tunable from 5456 to 5706 kHz. This translates the sideband to some frequency between 3.174 and 3.424 MHz—the exact frequency depending on the dial setting of the VFO. Whatever the VFO setting, the sideband developed is on the upper or lower side of the new frequency, whichever is selected at the doubler/tripler stage.

You've probably already figured out the third mixer, if you've been studying **fig. 4**. With the crystal selector set for the 7.2-MHz crystal, the range of difference frequencies tuned in the second mixer by the VFO is from 3.775 to 4.025 MHz. For the 10.475 crystal, it is from 7.05 to 7.3 MHz; for the 17.525 crystal, from 14.1 to 14.35 MHz; for the 24.625 crystal, from 21.0 to 21.45 MHz. Thus, the 80-, 40-, 20-, and 15-meter ham ssb bands are all covered. Naturally, the VFO dial is calibrated to show each of these band sectors.

fig. 4. Special case of triple translation. The first mixer gets multiplied signal from the carrier oscillator.



The 456.38-kHz carrier is modulated as usual, amplified, and filtered to produce the sideband signal. A sample of the carrier is also fed to a doubler to produce a 912.75-kHz signal. The stage following that is either a doubler or a tripler, depending on the setting of the sideband switch. With the signal frequency doubled, a signal at 1825.5 kHz is fed to the first mixer. There it beats with the sideband of 456.38 kHz, translating to a side-

mixers that translate ssb

In most single-sideband ham transmitters, the mixer circuits are ordinary tube or transistor mixers. In one transmitter I know of, a semiconductor diode mixer is used for translating the carrier frequency to an intermediate frequency. Typical tube and transistor transmitter mixers are shown in **fig. 5**.

These are not the only configurations used, by any means, but they are typical. Tube

mixers are usually pentodes in modern ssb transmitters; seldom do you find a triode used for this purpose. The two signals are merely coupled to the grid, mixed inside the tube, and fed along to the next stage.

In transistor mixers, common practice is to couple one signal to the base and the other to the emitter. In the transistor stage shown, the sideband is fed to the base, and the VFO signal to the emitter. The output frequencies are developed in the collector circuit.

Simple frequency conversion like this is okay for ssb transmitters, although it does

and a sideband of 3218.1 kHz (difference). Picking out the right one is the job of the tuned circuits following the mixer. In this example, a broadband tuned circuit centered around 3.3 MHz can do the job. Only the sideband of 3218.1 kHz gets through. The tuned circuit thus eliminates the new carrier that was generated as part of the translating process, as well as the original sideband and the new image sideband.

In first mixers, getting rid of the new carrier can be a problem because it is so near the sideband frequencies. It may even be troublesome to get rid of the image sideband unless the translation is a long step upward. Also, some of the original carrier may be lingering with the sideband, having slipped through the balanced modulator and the sideband filter.

The solution to all these possibilities is a balanced mixer, which was already mentioned briefly. An example of a tube-type balanced mixer is shown in fig. 6. A balanced mixer looks and operates just like a balanced modulator; the difference is that two rf signals are fed in rather than rf and af signals.

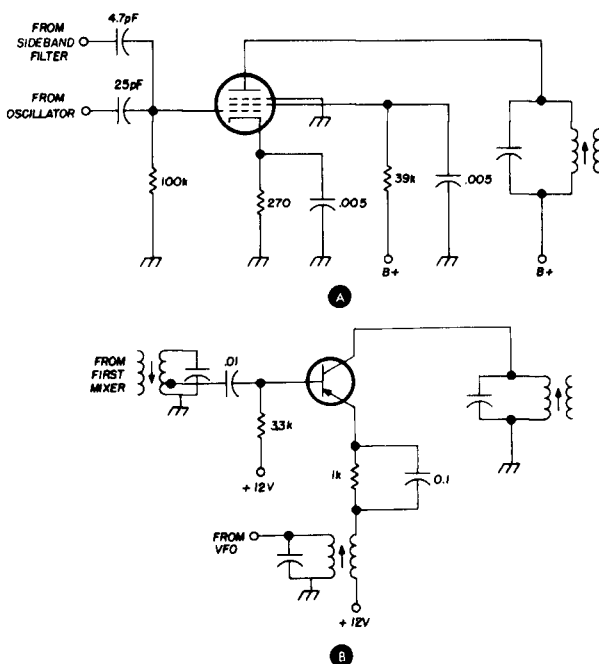
As in the usual balanced modulator stage, the signal to be canceled out is fed into the stage in parallel, and the output is taken in push-pull. Mixing is accomplished by feeding in the other signal—in this instance the sidebands—in the same mode as the signal is taken out of the stage. Thus, the VFO signal is fed simultaneously to the grids of both tubes (in parallel), and the sidebands from the mechanical filter are fed to the mixer grids in push-pull. The two 220-pF capacitors couple the VFO signal equally to the grids.

Balance is important in the two tubes, so a balancing-type cathode bias circuit is common to both tubes. During alignment of a transmitter using this system of translation, the balancing potentiometer is adjusted for a null of VFO signal in the mixer output.*

Bringing the sideband up to the operating frequency in a single-sideband transmitter is obviously not as simple as mere frequency multiplication. That approach would be im-

* The subject of ssb transmitter alignment, including how to adjust balanced modulators, is covered by Larry Allen in *repair bench* on page 58.

fig. 5. Two versions of frequency mixers used in amateur ssb transmitters.



create a problem. When two signals are beat together in a nonlinear mixer, the output consists of the two original frequencies, their sum, and their differences. As an example, suppose the VFO in the transmitter of fig. 4 is set at 5500 kHz, and the lower sideband has been chosen. The frequencies applied to the mixer (the transistor in fig. 5) are 2281.9 and 5500.0 kHz. The output consists of four frequencies: 5500.0 kHz, a sideband of 2281.0 kHz, a sideband of 7781.9 kHz (sum),

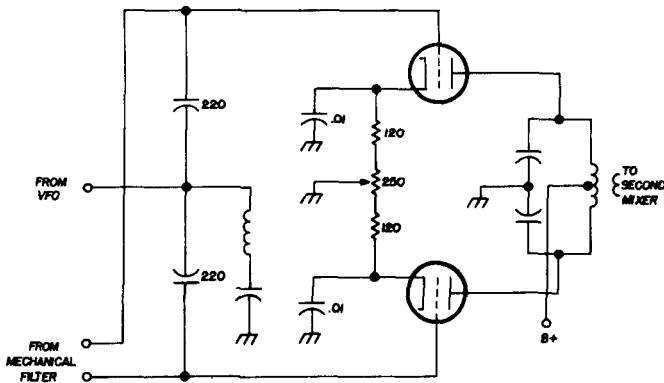


fig. 6. Balanced mixer used in some ssb transmitters to eliminate the carrier that is generated by frequency translation.

possible with sideband. An alternative, in hf ssb transmitters, is phase-shift generation of the sideband signal; the sideband can be produced right at the operating frequency. This method was discussed in the July issue of **ham radio**. Modern designs shy away from the phase-shift method because multiband characteristics are desirable in ham transmitters. Frequency translation seems to be the most practical way to raise the sideband frequency.

Next month I'll delve into another little-understood facet of the modern ham ssb transmitter: voice-operated transmission, better known as **VOX**. I'll explain various methods of accomplishing this type of hands-off operation. Also, we'll take a quick look at **MOX**—the manual version, usually called **PTT** or **push-to-talk**. **VOX** and **MOX** go together, in a way, and the up-to-date ham should understand both.

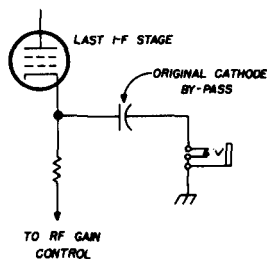
ham radio

the i-f cathode jack

Here is a very simple modification that will greatly increase the versatility of your communications receiver. Only one part is required: an ordinary closed-circuit phone jack. The diagram shows where the jack goes: in the lead between the i-f stage cathode bypass capacitor and ground. The jack may be mounted on the rear apron of the receiver chassis near the last i-f stage.

As long as nothing is plugged into the jack, it is a short circuit and the receiver works exactly as before the modification. When a phone plug is inserted, the i-f stage becomes a cathode follower, and provides a low-impedance i-f output for driving a Q-5'er, fm adapter, monitor scope, etc. An ac vtm can be plugged into the jack for precise indication of signal level. With a vtm plugged in, it is possible to make comparisons of antenna gain, measurement of front-to-back ratio, transmission line attenuation, preamp

gain, TR switch loss, image rejection, signal fading, skirt steepness ratio—practically any measurement requiring dB comparisons or rf signal levels. Be sure to turn the agc off.



Sometimes a cathode follower becomes regenerative if terminated in a capacitive reactance. If there is any sign of instability, the phone plug should be shunted with a suitable loading resistor.

Fred Brown, W6HPH

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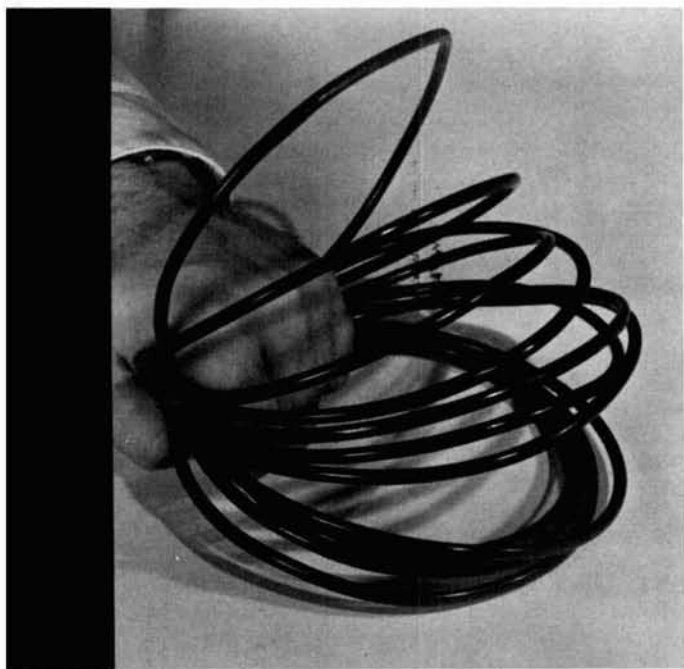
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ASK THE HAM WHO OWNS ONE



what you should know about standard RG-type coax cable

 Charles Camillo, W9JTB, Ampheno! Cable
Division, Bunker Ramo Corporation

Most amateurs who need standard coax cable will merely call out a catalog number, an RG designation and a length requirement—and patiently wait at the counter for the merchandise. To other hams, particularly those with vhf background, more thought is given to selecting the line. Unfortunately, this consideration never gets much further than the length of feedline required vs attenuation per foot. And even then they are likely to purchase inferior cabling.

behind the problem

To all too many, standard coaxial cable is a foolproof commodity which can normally be bought “blind.” Few realize that today a feedline made by one manufacturer can

exhibit completely different characteristics from that made by another—even though both cables carry the same RG designation.

Moreover, too free usage of the term "RG cable" has led to considerable confusion. The term RG actually connotes cable meeting latest revision specs of MIL-C-17; older versions of JAN-spec and MIL-C-17 cable do not. Unless the manufacturer clearly states this, it cannot be assumed that the latest spec is being adhered to.

Additionally, some manufacturers have blurred this distinction with meaningless terms such as "RG-type." It is essential that you be aware of these things—since your entire cabling system could fail as a direct result of buying the wrong coax for your individual application. Slow cable degradation, prime cause of gradually-deterioration signals, is extremely hard to detect.

In some instances, it might prove valuable to review military specifications. While many will be irrelevant, some reveal key parameters applicable to amateur needs.

For example, consider percentage of braid coverage in a typical ssb transmission system. To prevent signal leakage that might interfere with other services such as television, the percentage of braid cover should be quite high—at least 90 percent of the dielectric must be completely shielded. Yet, many cables presently being used in this kind of rf work exhibit only 65 percent coverage. Add to this other problems that fre-

quently develop, and the over-all evaluation process can be more clearly appreciated.

determine characteristics, spot problems

Some amateurs are confused by the significance of the cable dielectric. Simply stated, the dielectric quality of any coaxial line determines both long- and short-term attenuation as well as over-all power-handling capabilities.

How do you spot a poor dielectric? If a thick-wall coaxial line with silver-plated copper conductor has a dielectric that appears amber or gray when placed on a sheet of white paper, it is probably composed of inferior or scrap polyethylene. Inspect a sample of the cable you are replacing (or currently using).

Demonstrating the color check can be extremely helpful. Bear in mind, however, that wall thickness—which varies from one impedance to another (between 50-, 75- and 95-ohm types)—determines opacity; opacity, in turn, determines color hue. Also, conductor color can affect over-all hue.

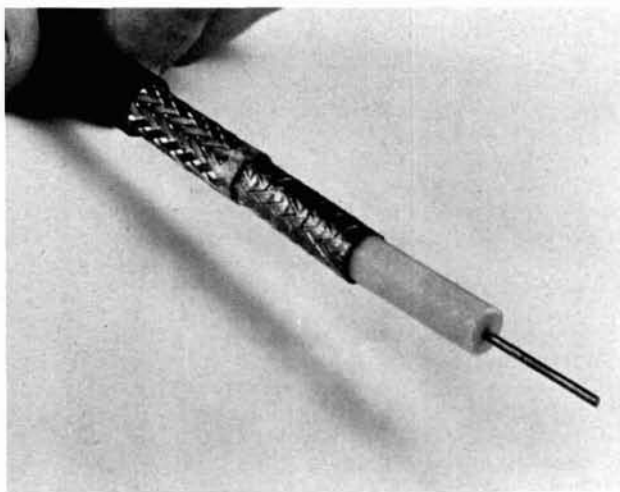
What about foam cables? Here, too, some evaluation can be accomplished visually. Bubbles should be of similar size and round in shape throughout. If a micrometer's handy, check the extrusion of the cable and check to see if the dielectric is tight on the conductor.

If you ask yourself a few key questions, you can shed a lot of light on both the operational efficiency of your system and the requirements that should be met with new cable. Does the line become brittle or fluid during periods of temperature extremes? Or, is it presently fluid or brittle? What about dimensional stability? Have gradual changes been noticed?

watch for capacity

Although few amateurs are aware of this characteristic, standard coax lines are actually extremely long capacitors—each exhibiting a pronounced effect on the tuned output circuit at each end (transmitter, antenna, etc.).

To cope with this problem, coaxial cables are rated in terms of dielectric constants. As



the constant approaches 1.00, the more nearly the capacity (and subsequent attenuation) approaches the low figure of open-wire lines. Knowledge of this figure allows you to analyze frequency-handling capabilities of the coax in question.

For example, cellular polyethylene types (foam cable) are rated at a dielectric constant of 1.5, compared to 2.26 for conventional solid polyethylene. A look at solid dielectric RG-8/U cable in terms of capacity shows an actual capacity of 29.5 pF per foot. This compares with 24.5 pF for foam lines of equal size.

seldom exceeds \$.02 per foot.

rf power attenuation

RG-8/U and RG-58/U cable should be examined at this point in terms of attenuation. Both solid-polyethylene and polyethylene-foam types are compared in **table 1**. The dB rating is per 100-foot length. These figures assume no cable degradation due to heat or general aging.

Suppose you have a 144-MHz vhf system. Given such information, it can be seen that with RG-58/U, more than 5 dB of rf power output is lost at 144 MHz. With RG-8/U

table 1. Attenuation and power ratings of popular RG/U coaxial cables.

type RG/U cable	dB per 100 feet nominal attenuation				power rating watts			
	MHz				MHz			
	10	50	100	200	10	50	100	200
RG-8/U solid	0.55	1.33	2.00	3.50	3500	1500	975	685
RG-8/U foam	0.32	0.77	1.18	2.07	3500	1500	975	685
RG-58/U solid	1.25	3.13	4.16	6.90	1000	450	300	200

If you want to buy a 100-foot length, the difference between one dielectric and another can be 500 pF—enough to severely degrade matching of the most well-designed 220- or 432-MHz transmission system.

solve migration problem now

While nearly all standard coaxial cables have black polyvinyl-chloride jackets, there are actually two sub-categories of jacket material that should be evaluated early in the game. If you don't consider this now, you may be plagued with high attenuation in a few months.

The first kind, known as Type I, is found only in older versions of JAN and MIL cables and can prove troublesome in certain applications. Depending upon age and environmental temperatures, it's possible for the polyvinyl chloride's "plasticizer" to migrate out of the jacket and into the cable.

Result of wrong application choice? Electrical characteristics will be drastically changed, to say nothing of attenuation. In nearly all instances you should use Type IIa polyvinyl-chloride jacketing material. Incidentally, the cost of this extra protection

foam, however, only 2.7 dB is sacrificed. This means you would have to generate almost twice the power with RG-58/U to achieve the same results as you would with foam-dielectric RG-8/U.

In lower right foreground, cable conductors enter an extruder. The extruder puts an insulation of polyethylene on the conductor wire—then the cable travels through a cooling bath.

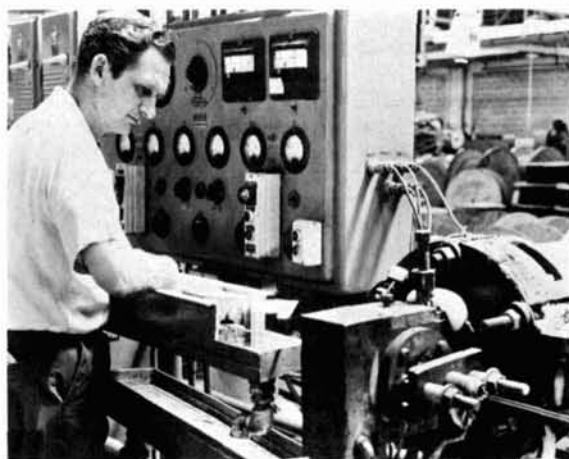


table 2. Sampling of popular foamed polyethylene dielectric coaxial cables.

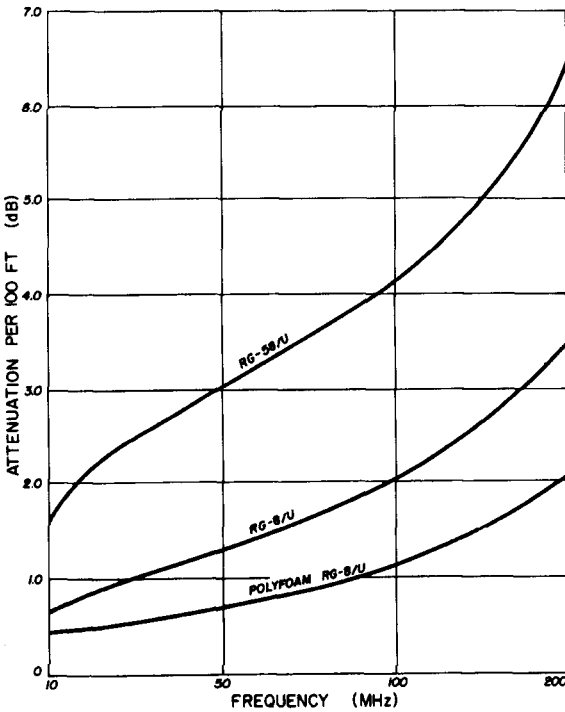
RG/U number	Amphenol number	jacket OD	jacket type	shield	dielectric OD	center conductor	VP %	capacitance pF/ft	maximum operating nominal	
									volts rms	impedance in ohms
8	621-111	.405	I	C	.285	7/19C	80	24.5	1500	50
11	621-100	.405	IIIa	C	.285	14 C	80	16.5	3000	75
59 type	621-715	.195	IIIa	C	.107	22 CW	80	17.0	500	72
59	621-186	.242	P	C	.146	20 CW	80	17.3	1000	75

C—copper, CW—copperweld, P—polyethylene, VP—propagation velocity

determine power requirement

Note the maximum power ratings for standard communications cables listed in table 1. It is important to know both the power-handling capability of the cable in question as well as the specific frequency

fig. 1. Attenuation curves for three popular types of coaxial cable.



intended for use. The ratings in table 1 assume a perfect match between transmitter and line, and line and antenna. If vswr is high, power-handling capability diminishes and losses run high as dielectric heating occurs.

check conductors

It is important to realize that in order to meet many of the standards and ratings discussed so far, other elements of coax construction come into play. Many of these—while seemingly obscure—may account for difficulty you are experiencing now with a particular brand of cable.

Again, if a sample can be obtained of existing in-use cable, much can be determined. Visually, for example, check whether the conductor is off center in the dielectric. If there is more than 10-percent error, serious problems can be expected. Are there as many strands in the center conductor as specified in the latest MIL-spec requirement? Though this might seem unimportant, it can be crucial in work above 10 meters.

In most good-quality standard coax lines, braid should fit tightly. If it doesn't, this can indicate a strong possibility of a change in electrical characteristics. Braid tightness, however, can vary; RG-8A/U, for example, has an extremely loose braid. It's wise to check cable specifications.

Be sure to inquire as to flexibility requirements. Maximum flexibility is achieved with stranded center conductors, although attenuation losses can be cut appreciably with solid conductor carriers. The answer to this question involves considerations discussed earlier—including frequency of operation and transmitted power in watts.

cable selection checklist

In addition to the above, an amateur should consider the following checklist:

- 1. Impedance:** What is the actual impedance and percentage of spec variation?
- 2. Frequency:** Is there any attenuation periodicity (high points) along the intended frequency curve?

table 3. Coaxial cable joint jacket characteristics.

MIL-C-17 designation	jacket type	temperature limits
type I	black polyvinyl chloride	-40° C to +80° C
type IIa	black polyvinyl chloride, non-contaminating	under 1/4", -55° C to +80° C over 1/4", -40° C to +80° C

- 3. Power:** What is theoretical corona voltage? Actual voltage?
- 4. Jacket composition:** Is it of non-contamination material, according to MIL specifications? What are the changes in attenuation at 432 and 1296 MHz after aging tests?
- 5. Jacket tightness:** Does it fit tight enough to show braid marks clearly, or is it loose, possibly indicating poor extrusion and instability?

- 6. Connectors:** Has the right coaxial connector been selected for use with the standard RG cable chosen? Is it designed specifically for this purpose, or does it require an adapter?

If you take these steps, and the cable you are using fills each requirement satisfactorily, you should have a minimum of trouble with your system. However, here are three more steps you can take to insure that you're getting the best cable available:

- 7. Vendor:** Have you seen the manufacturer's certification stating that the cable meets given RG specs?
- 8. QPL:** Is the manufacturer on the federal Qualified Products List (QPL)?
- 9. Vendor certification availability:** Have you seen similar certification? When you bought cable from another supplier, was it merely a restatement of catalog specs?

ham radio

lightweight headphones

Many DX'ers find that heavy headphones are too heavy on the head and ears for long periods. The solution to the problem is a bit complex and will be discussed here.

EP2BQ uses the TELEX twinset which has a headband plus eartips. On test, however, it was found that this type has a strong peak above 2kHz that causes problems with interference and voice intelligibility.

There is a group of TELEX secretarial-type under-the-chin units called **Dynaset**, **Monoset**, and **Tele-fi**. Although not offered to the mass market, Allied and Newmark also carry a "new 799;" this appears to be a Monoset built for additional impedances. These devices all use the same transducer with a response of 50 to 500 Hz and are very light, weighing from one-half to 1.25 ounces.

KH6IJ highly recommends the Dynaset HUP-01. This has the transducer in a phone plug, with plastic tubing leading up to a junction below the chin. This is available only in 15 ohms, however. It is frequently found in airliners.

Don Miller, W9WNV, has been pictured

wearing the Monoset unit, which has the transducer in the junction under the chin. Although these have been offered in 125 and 2000 ohms, Newark now shows them as the "new 799" in 15, 500 and 1000 ohms too. For those of you with 500-ohm equipment such as Collins, this appears to be the best solution.

The remaining design is the Tele-fi. This has the transducer in one earphone, with a tube leading to the other ear. This results in a half-ounce unit but puts a millisecond time delay in the line to the other ear. This promises good performance on music and voice, but we have no reports on the possible effect of this small time delay on code reception. However, it is available in the full range of five impedances, is extremely light weight and passes under the chin. Further, it appears to have small earmuffs that fit into the outer ear rather than plastic tips to fit the entrance to the ear canal as found in the other units.

ham radio



What's the BIG Idea?

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The manager of advanced development for a big communications company—an active ham since his early teens—asked the same question of another long-time enthusiast and nationally-known authority on solid state devices. They observed that effective application of the new technology—largely a product of the aerospace industry—demanded a high degree of engineering sophistication and a variety of technical capabilities not generally found outside of that industry . . . Their idea . . . why not organize a group of outstanding professional engineer/hams . . . to do the job . . . and develop their own new generation of no-compromise, ham gear?

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WHAT'S THE BIG IDEA?

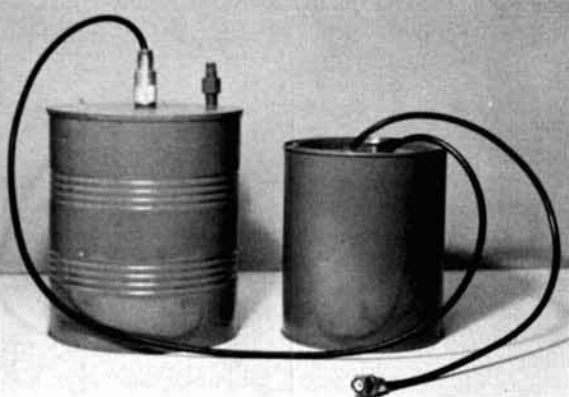
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experimental dummy loads

How to design
and build
dummy loads
that provide
good performance
well into
the vhf region

Bill Wildenhein, W8YFB, RD 2 Blanche Avenue, Elyria, Ohio 44035

As so often happens to the inveterate home brewer, I needed a good dummy load which was usable well into the vhf region. Like many others, I had some misgivings about the accuracy of homebrew loads. The following information is presented as a guide to the design and construction of dummy loads, as well as a description of my load and its characteristics as measured on a laboratory-type rf impedance meter.

The better commercial loads are built around deposited film or similar resistors where the film thickness is made small to minimize variations in bulk resistance due to skin effect. At low frequencies, rf currents penetrate more deeply into a conductor and as the frequency is increased, penetration becomes progressively less. Depending upon the type of resistor, other effects are also present. Since the average homebrewer can't afford an ideal resistor, we must do the best we can with ordinary carbon composition resistors.

In addition, at vhf the reactive effects can become rather formidable. Commercial practice is to construct the load shield in the form of tapered line sections to match the load impedance at all points as shown in **fig. 1**. Notice that the tubular resistor element is long compared to diameter. Also, the

resistor-to-connector lead is in the form of a cone that tapers from the outside diameter of the resistor element to the diameter of the connector center conductor.

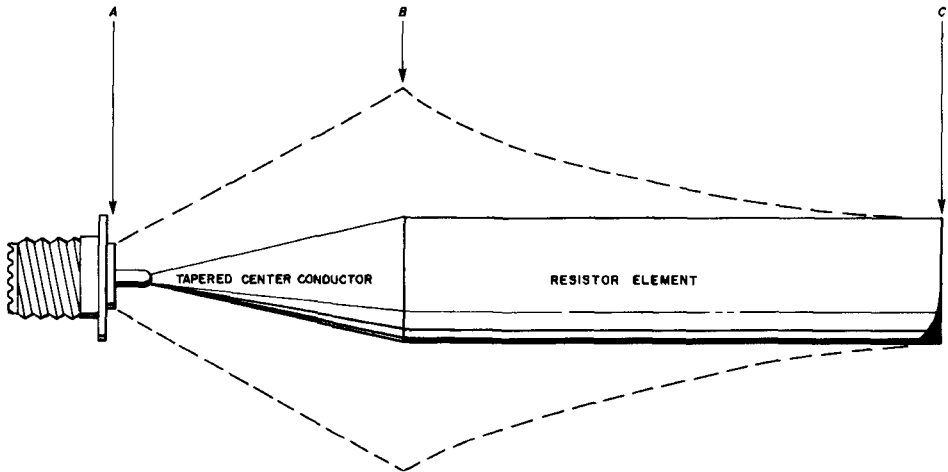
The coaxial structure between point A and B is a tapered line section which is designed so that its characteristic impedance is the same as the impedance of the resistive element at all points. The resistive element starts at point B, and from that point to point C the impedance diminishes uniformly to zero. At point C the outer section of the coax structure is fastened to the resistor to form the return leg back to the connector body.

I designed it for an air dielectric—it wasn't intended to be used with an oil coolant. The dimensions and values were chosen for a 50-ohm load.

The completed unit looked good enough to consider as a permanent load, so I mounted it in a coffee can and added oil coolant. If you intend to use cooling oil, the shield diameters should be increased about 30% to allow for the different dielectric constant of oil. The performance curve shown in **fig. 3** from 4 to 50 MHz shows some deviation from the measured dc value of 52 ohms.

The impedance meter indicated that the reactive effects through this range were small

fig. 1. Geometry used in commercial loads to maintain the load impedance at all points.



Any departure from the mathematically derived curve of the coaxial shield, or any departure in the linearity of the deposited resistive film on the ceramic tube which makes up the resistor element, results in something other than a purely resistive load; this becomes progressively worse as frequency is increased.

Since I've worked with and rebuilt precision dummy loads, I was dubious about the prospects of building anything useful with available components. However, since I had a large number of 390-ohm, 2-watt resistors with very short leads (I bought one of those "five for a buck" board deals) I built the dummy load shown in **fig. 2**. I didn't know if the load would be worthwhile or not, so

compared to the variations in resistance and were negligible below 30 MHz. At 50 MHz, resistance variations contributed approximately 1% of the impedance variation. The load, then, is resistive for all practical purposes to 50 MHz.

The changes in resistive value are a built-in factor determined by the use of plain old carbon resistors. Above 50 MHz, the load impedance begins to do some pretty wild things. This is shown in curve A of **fig. 4**. Again, the wildest excursions were due to variations in the resistive value, while the reactive value remained surprisingly reasonable. It was apparent that the over-all geometry of the load wasn't too bad, and the limitation seemed to be the resistors themselves.

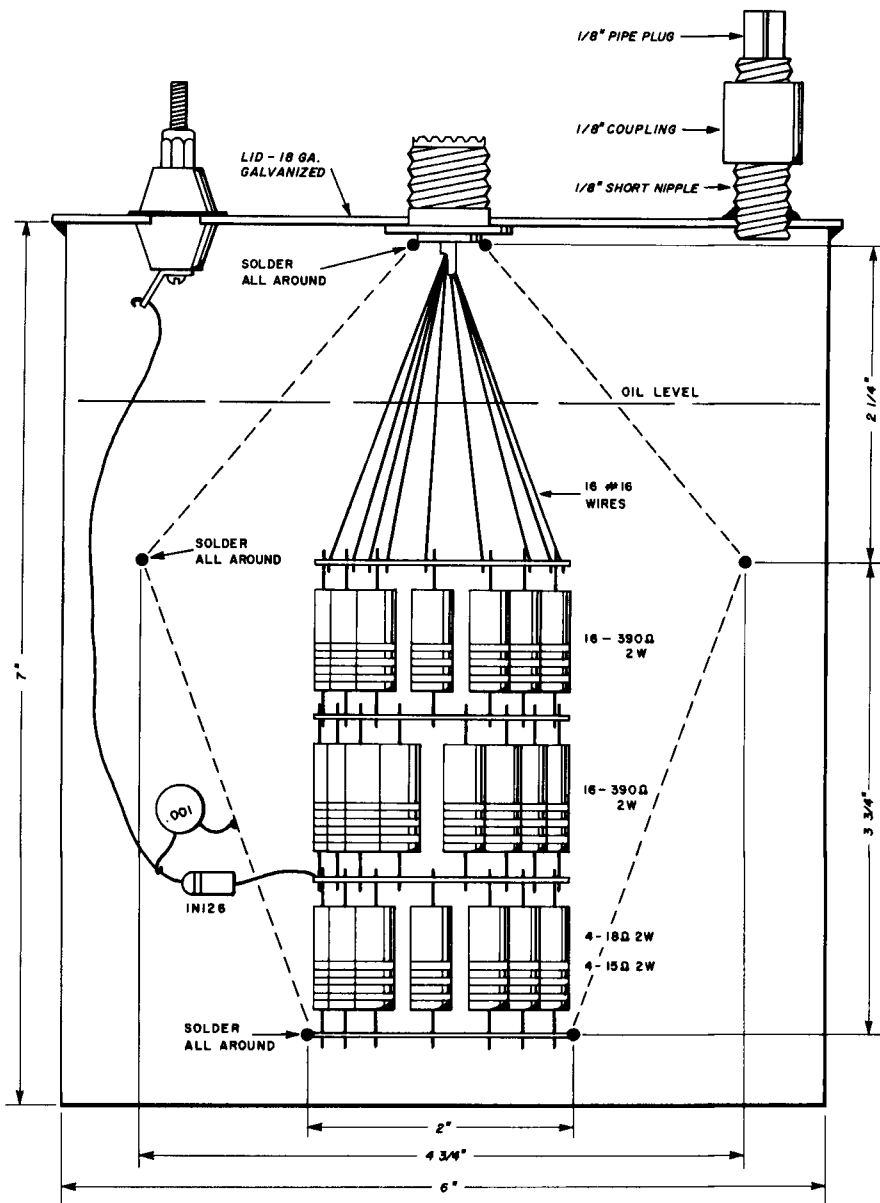


fig. 2. The air-cooled dummy load built by W8YFB. The performance of this load at various frequencies is plotted in fig. 3 and 4.

the load leveler

To improve the performance of the load, I built a load leveler. This is nothing more than a long length of coax—the longer the better. It's connected in series with the load. This piece of coax improves the load through attenuation.

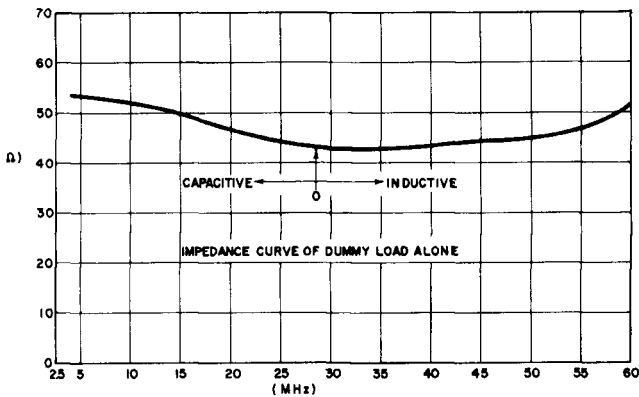
For example, suppose you were to feed a

vhf transmitter into an SWR bridge, then out to a long length of coax which could be open or short circuited at the far end. If the cable were long enough to represent a 10-dB loss and you fed 100 watts to the line, only 10 watts would actually arrive at the far end. The remainder would be dissipated in line loss. Since the 10-dB loss works both ways, the line would appear beautifully matched,

because the reflected signal would be 1/100 of the forward power indicated on the SWR bridge—and with an infinite mismatch at that! This, incidentally, is the reason why a vhf antenna should be measured at the antenna rather than the transmitter end of the line.

I coiled 80 feet of RG-58A/U into another coffee can, fed the ends through rubber grommets in the lid, and put coaxial connectors on the ends. With the load leveler, a recheck on the impedance meter showed the

fig. 3. Impedance of the dummy load from 5 to 60 MHz.



expected results—curve B in fig. 4. The improvement is already felt near 50 MHz; above 150 MHz the curve begins to flatten out. Since the impedance meter I used only goes to 250 MHz, this represents the end of the line for actual measurements.

However, because of the characteristics of the load leveler, the undulations of the curve will become progressively smaller with increased frequency; the curve tends to center around the mean impedance of the coax used in the load leveler. It should be noted that an appreciable part of the total power may be dissipated in the load leveler. At frequencies above 50 MHz, watch for hot spots in the load leveler cable. One way to do this is to use a length of coax from the transmitter to the load leveler—perhaps eight feet long. As you use this combination, run your hand along the 8-foot length of cable. Don't let it get very warm.

Similar hot spots will occur at half-wave intervals all along the line coiled up in the

can, and inside the can, heat dissipation is impeded. It helps heat dissipation if the inside layers of coiled coax go to the dummy load instead of the transmitter. This is particularly true if the inside layers are coiled on a form. I used a 1-1/2-inch wood dowel as a form.

The combination dummy load and load leveler certainly isn't the sort of thing needed for lab-standard work, but it does represent a fair resistive load for amateur practice. At worst, it exhibits an SWR of about 1.25, and over much of the range it is considerably better.

choosing the resistors

Since the impedance meter indicated that the maximum variations were a result of the resistive component, it appears that the basic geometry of the load allows considerable latitude in resistor choice as long as the basic configuration is maintained. It's economically unsound to buy all new resistors to duplicate my load when a Heathkit Cantenna is available for the same money. However, if you have a bunch of resistors stripped out of television sets and old equipment, it's worthwhile to see if you have a usable group of values. One such group is illustrated in fig. 5. Maximum power capability is available when all resistors are the same value. When using dissimilar values, group values together which are closest in resistance. To find the total resistance value, proceed as follows:

1. Find the equivalent resistance of each group of identical resistors; merely divide the value of one resistor by the number of resistors in the group.

2. Find the resistance of each paralleled group: this can be done rapidly on a piece of cross-ruled paper as shown in fig. 6 or calculated from $R_{total} = (R_1 \times R_2)/(R_1 + R_2)$

3. Add the values of each of the series banks to obtain the total for the load; this value should be between 50 and 53 ohms.

The next problem is to determine the wattage capability of the load:

4. First, determine the maximum voltage capability of the **second highest** group of three resistors—in fig. 5, this is group 5:

$$E = \sqrt{PR} = \sqrt{6 \times 40} = 15.4V$$

5. Determine the current through this bank from Ohm's law:

$$I = E/R = 15.4/40 = 0.385 A$$

6. Since this same voltage will appear across resistor bank number 6, determine the current through bank number 6:

$$I = E/R = 15.4/50 = 0.31 A$$

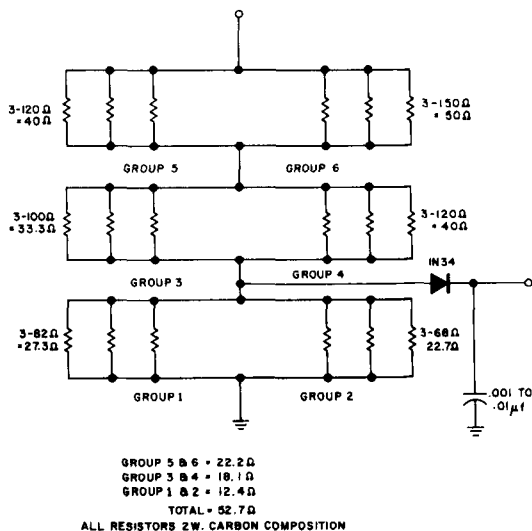
7. Determine the maximum current through the dummy load by adding the currents found in steps 2 and 3:

$$0.385 + 0.31 = 0.695 A$$

8. Determine total power capability for load:

$$P = I^2R = (.695)^2 \times 52.7 = 25.6 \text{ watts}$$

fig. 5. Method of setting up resistance values to obtain a satisfactory dummy load.

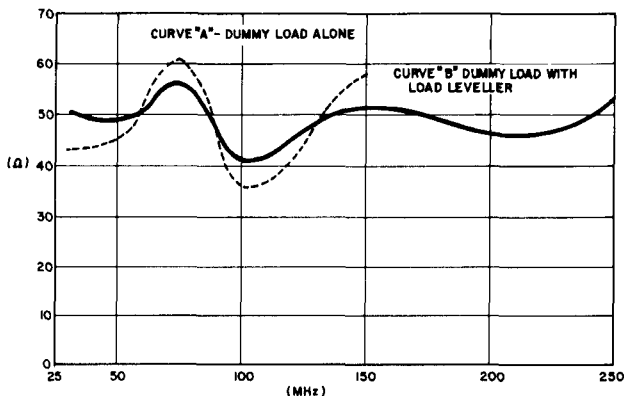


If you use an oil coolant, this number may be safely multiplied by four for short periods. This load would then have a peak capability of about 100 watts.

Some of you may question the value I obtained in step 5. Since we have eighteen 2-watt resistors, you may think that the load should be capable of dissipating 36 watts. However, remember that the current that flows through banks 1 and 2 also has to flow

through banks 3 and 4 and 5 and 6. If you raise the current through banks 1 and 2 to the point where they are dissipating their maximum power, the same current exceeds

fig. 4. Impedance of the dummy load between 30 and 250 MHz. The effect of the load leveler at the upper frequencies is quite evident.



the ratings of the higher resistance banks.

It is often helpful to determine the dissipation of each bank of resistors separately—or at least each paralleled bank. For example, banks 1 and 2 have an equivalent resistance of 12.4 ohms; from step 8 the power dissipation is calculated at 6 watts.

To measure the resistance of the load with a simple ohmmeter, it's best to buy one good resistor; preferably a 50-ohm 1% film resistor but at least a 51-ohm 5% carbon composition type. If you use resistors that may be on hand for the load, you may find that although the finished load is calculated to be 50 ohms, it will measure 55 to 60 ohms after assembly. You can measure the 50-ohm standard resistor with your ohmmeter, adjust the meter so that it reads exactly 50 ohms and connect the meter across the dummy load. Then connect resistors across the highest resistance bank in the load until you find a value that gives a reading of 50 ohms.

Under no circumstances use wire-wound resistors—even those that are claimed to be non-inductive. Be wary of resistors which look like molded carbon types but have one color band much wider than the others. They are wire wound.

construction

Make four rings of thin brass shimstock, dairy tin (light gauge steel which is tinned on both sides) or light galvanized iron. See **fig. 8** for details. Space the resistors around a circle so there is an 1/8-inch or more between

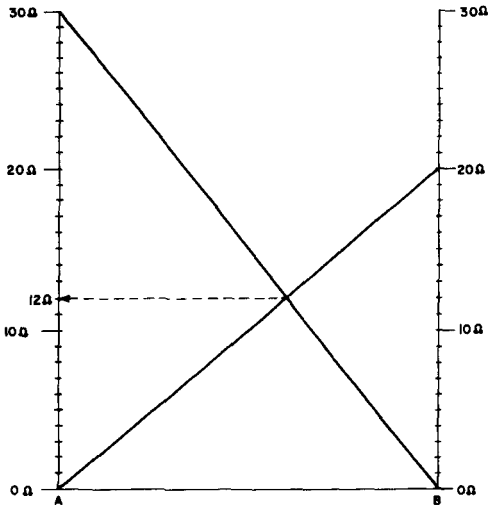


fig. 6. Nomograph for determining the equivalent parallel resistance of two parallel resistors. Example: what is the equivalent resistance when a 20- and 30-ohm resistor are connected in parallel? Plot a line from B on the right-hand scale to 30 on the left-hand scale and a line from A on the left to 20 on the right. The cross-over point of these two lines gives the equivalent resistance. This chart can easily be extended to higher values if desired since both scales are the same length.

resistors to allow coolant circulation. If a resistor has to be added later to make the entire string look like 50 ohms, it can be mounted in the center of one bank. The center hole in the rings is necessary for free flow of coolant through the resistor assembly.

If your resistor banks are made up of unequal groups of resistance values, alternate the resistors around the circle. The conical group of wires connecting the ring nearest the connector are installed next. A dozen pieces of number 22 or 20 wire make an adequate cone. Get the apex of this cone of wires as close to the center line of the holes in the resistor rings as possible. This will help in assembly of the outer coaxial shield later. When you have completed the cone,

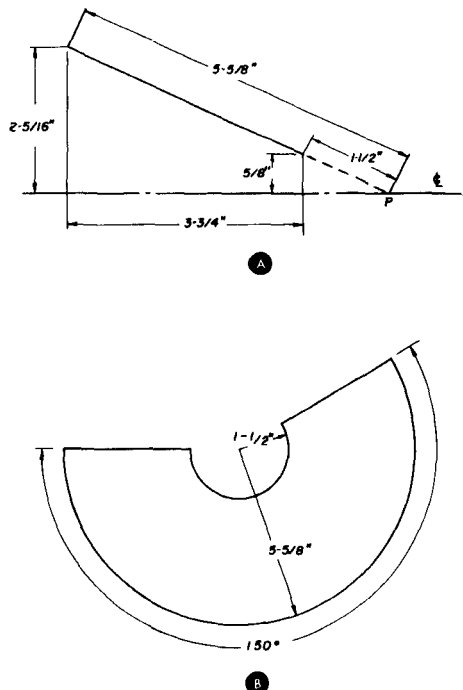
you can get an accurate measurement of the distance from the connector body to the first resistor ring.

the coaxial shield

Next, determine the dimensions and layout of the conical pieces. These are made from copper window screen. Don't use aluminum screen; it's difficult to make a solder joint. If any substitution is necessary, go to good clean galvanized screen.

The screen can be cut by trial and error to form the desired sections, but it's easy to make an accurate cardboard template and use this to cut the screen sections to exact size and shape. **Fig. 2** shows the overall assembly, but is drawn to illustrate the original load. We'll make one modification. The ratio of diameters shown for the screen diameter and resistor ring diameter were for air dielectric. In this design example, we'll make the ring diameter 1-1/4 inches to accommodate fewer resistors. Also, the ratio of this diameter to the screen diameter will be calculated for oil dielectric.

fig. 7. Laying out the outer shield.



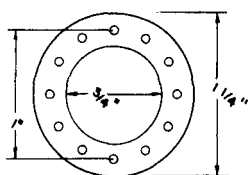
For oil dielectric, multiply the desired load impedance by \sqrt{K} , where K is the dielectric constant of the oil, to obtain the proper ratio of diameters using the conventional coaxial-line impedance formula. The dielectric constant of transformer oil is about 2.2 and the square root of 2.2 is 1.48. Multiply the desired 52.7-ohm load impedance by 1.48; 78 ohms. From the coaxial-line impedance formula:

$$Z_0 = \frac{138}{\sqrt{K}} \log_{10} (D1/D2)$$

we find that the ratio of diameters for an air dielectric is 2.38, while for transformer oil the ratio of diameters is 3.67.

Regardless of the diameter of the resistor banks, you only have to multiply the resistor-bank diameter by 3.67 to obtain the correct screen diameter for a 52.7-ohm impedance in

fig. 8. Resistor mounting ring construction.



oil coolant. In this case, screen diameter will be about 4-5/8 inches.

A method for making an accurate layout of these screen segments is shown in **fig. 7**. First, draw an accurate side view of one half of the screen cone as shown in **fig. 7A**. Extend the line representing the outside of the cone until it intersects the centerline at point P. This line provides the needed radii. Measure the dashed segment with a scale; it is 1-1/2 inches long—total length is 5-5/8 inches. Now draw two concentric circles with these radii on a piece of cardboard as shown in **fig. 7B**. To determine the angle of the cut-out portion:

1. Find the circumference of the finished cone from circumference = π diameter = $3.14 \times 4\text{-}5/8 = 14\text{-}1/2$ inches

2. Determine the circumference of the

outer circle you have drawn: $3.14 \times 11\text{-}1/4 = 35$ inches

3. From proportion, determine the number of degrees of circle needed:

$$360^\circ \frac{14.5}{35} = 150^\circ$$

Lay out this 150 degree segment on your cardboard template with a protractor; allow about 1/4 inch more for overlap. Cut the screen to match the template. Roll the screen up into a cone with the same amount of overlap. The loose ends of screen wire can be used to hook the thing together until it is soldered along the entire length of the screen. Be sure to solder it thoroughly because it is the return lead and will be carrying fairly heavy rf currents.

Make the top screen cone using the same method. Securely solder the inner conductor of the coax connector to the apex of the wire cone. Then solder the top cone to the body of the connector. Before soldering the lower cone in place, cut a hole in it for the rectifier diode. Mount the diode on the resistor ring and slide the lower cone into position.

Make a thorough solder connection all around the cone where it meets the lowest resistor ring. Then solder the junction of the two screen cones. The can lid can be made of galvanized iron or sheet brass. Be sure to include a vent that can be opened to release internal pressure when the load runs warm.

calibration

A word about the diode rectifier. It can be any germanium rectifier such as an 1N34A. Many of these diodes are rated in the vicinity of 40 volts or so. To prevent breakdown, it's wise to connect them to the resistor ring closest to the ground end of the resistor banks. A vtvm or 20,000 ohm/volt dc meter may be used to measure the relative output. Because of the nonlinear characteristic of the diode at low power levels, it's desirable to have a rough idea of power into the load. Otherwise, small power increases may look far better than they actually are. Any power calibration, of course, will be a rude approximation unless you have some lab gear available.

Calibration can be done at 60 Hz by feeding the load with a variable transformer or a

group of series-connected filament transformers. First, calculate a number of calibration points from $P = E^2/R$. With 12.6V applied to a 50-ohm load, for example, there is 3.2 watts into the load. After calculating a number of points, step through them with the variable transformer or other supply and record the meter readings. These points can be plotted on graph paper so the intermediate points can be read.

The finished load may be filled with mineral oil from the drug store, or transformer oil if you know someone in the utility business. Be sure the top resistor bank is submerged an inch or so below the oil level but allow some air space for expansion. This load shouldn't be run at maximum rating (about four times rated resistor capability) for much over three minutes without allowing a ten-minute cooling period before reapplying power. I've run mine at this overload level repeatedly for 5-minute periods, but extended abuse could result in long-term impedance variations that will eventually send the load to the scrap heap.

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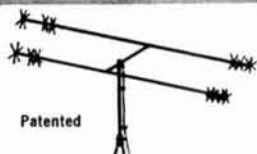
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solid-state screen clamp

Here's a
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of the clamp tube
that provides
high performance with
low-power dissipation
and
almost no heat

The clamp tube has been a popular circuit in amateur transmitters for a long time. Its primary purpose is to lower the screen voltage of the final in the absence of excitation and prevent excessive current in the final tubes. It has also been used as an a-m modulator. In this application it varies the final screen voltage—and thus the plate current—in response to an audio input. The third reason for using a screen clamp resulted in this article.

My transmitter uses fixed bias and grid-block keying. The final tubes are maintained at low plate current during key-up by this bias, but they're not completely cut off. As a result, when the transmitter is used with a T-R switch, current in the plate circuit causes objectionable noise in the receiver.

I originally installed a 6AQ5 clamp tube to cut off the final amplifier tubes completely and eliminate the noise. Before long, the tube circuit was discarded in favor of a transistorized version. To see why, let's discuss the principles of each concept.

tube vs transistors

As shown in **fig. 1**, the clamp tube may be represented by a switch and a resistor connected between the screen and ground. When the final tube is operating normally and excitation is present on the grid, excitation also puts bias on the clamp-tube grid. This bias is sufficient to cut off the clamp tube so that no plate current flows, and it looks like an open switch.

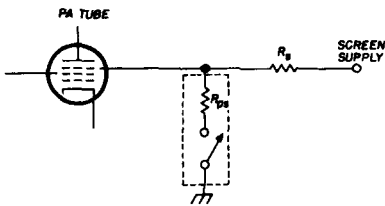
Screen current flows normally through the screen-dropping resistor, and the final tube operates as if the clamp tube were not present. If excitation is lost, the bias on the clamp tube disappears, and it conducts heav-

Chris Grant, WØLRW, 6623 South Grant Street, Littleton, Colorado 80120

ily. The screen voltage in the final amplifier drops to a low voltage and prevents the amplifier from drawing too much plate current. The clamp tube plate voltage (and final tube screen voltage) depends upon the voltage divider formed by the screen dropping resistor R_g and the clamp tube's saturation plate resistance R_{pg} . In my transmitter, this voltage is about 15 volts.

This clamp tube circuit has two serious disadvantages. First, since final amplifier screen voltage is not dropped to zero, some plate current may still flow. Secondly, a great deal of power is dissipated in the clamp tube itself and in the screen resistor, R_g , because of the high plate current through the clamp tube. This power is wasted and appears as unwanted heat inside the transmitter cabinet.

fig. 1. Clamp-tube equivalent.



These disadvantages may be overcome by devising a circuit which will work like fig. 2. Here a single-pole, double-throw switch alternately connects the screen to the supply and then to ground. When excitation is applied to the final, the switch connects the final amplifier screen to the screen supply. If excitation is removed, the switch moves to the opposite position, connects the screen to ground and prevents any final plate current.

Note that in contrast to the circuit in fig. 1, no supply current is drawn when the final screen is grounded. This can be accomplished quite simply by using inexpensive high-voltage silicon transistors in the circuit shown in fig. 3.

transistor circuit operation

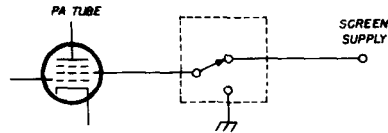
The diagram of the solid-state screen clamp I use in my transmitter is shown in fig. 3. The basic principles of operation are quite simple. If excitation is present on the final amplifier grid, rf is coupled through the 470-pF capacitor to the 1N60 diode. The

capacitor is charged during the positive half cycle of the excitation signal and puts a negative voltage on the base of the 2N3440, cutting it off. The collector voltage rises to a value determined by the 91k and 220k voltage divider.

The other 2N3440 is connected as an emitter follower. Its emitter voltage is about 0.6 volts less positive than the base. Therefore, the final amplifier is effectively connected to a source having the same output voltage as the junction of the 91k and 220k voltage divider. Note the additional benefit of this circuit: final screen voltage is relatively independent of screen current.

If excitation is lost, the negative voltage at the base of the first 2N3440 disappears. It is then biased into conduction by the current flowing through the 3.9M resistor to the screen supply. The transistor saturates and brings the base of the other 2N3440 down to a few tenths of a volt; the transistor cuts off and holds the screen at ground (I measured about a tenth of a volt on the

fig. 2. Transistor clamp equivalent.



screen). Therefore, there is no final-amplifier plate current.

design considerations

You have to go a little farther than basic operation if you want to adapt this circuit to transmitters with different voltage sources and screen requirements. The first consideration is transistor selection. Since the transistors don't handle high frequencies, audio types will do. The collector-to-emitter breakdown voltage, BV_{CEO} should equal or exceed the voltage of the screen source. The 2N3440's I used with a 300-volt supply in fig. 3 have a rated BV_{CEO} of only 250 volts. However, out of five units I tested, all had breakdown voltages of at least 500 volts.

The next step is to determine the voltage

divider resistances (91k and 220k in **fig. 3**). The resistors are chosen so that their junction will be at the recommended screen voltage when the screen supply voltage is connected. When the first transistor conducts, this junction is at ground potential, so the power rating of the 91k resistor must be computed using the full screen supply voltage. With the values in **fig. 3**, a half-watt rating was sufficient. The divider as shown draws about 1 mA from the screen supply during normal operation of the final tubes.

The transistor in series with the screen supply must pass all of the screen current. The 2.2k resistor is necessary to limit the power dissipation of the transistor by dropping more of the voltage difference between the supply and the screen as the screen current increases. Thus, as screen current (and transistor current) increase, the voltage drop across the transistor goes down. In the circuit shown in **fig. 3**, the maximum power dissipated by the 2N3440 emitter follower is less than 1.25 watts.

The 1N60 peak rectifier and filter (33k and 330 pF) are a simple way of switching the transistors with rf excitation from the final amplifier grid. The circuit is fast and sensitive enough to function properly even during ssb operation where excitation varies rapidly. If the final uses no fixed bias, the input end of the 33k resistor may be connected directly to the final grid, and the 470-pF coupling capacitor and 1N60 are not needed.

With a direct connection, it may be necessary to increase the value of the resistor to reduce loading on the grid circuit. The 3.9M resistor is chosen so that the first 2N3440 conducts and its collector goes to ground when there is no excitation on the final grid.

There are four other components which haven't been mentioned. The .01's on the screen and supply terminals are used as by-passes. The two 1N914's are necessary to protect the transistors from reverse breakdown at the emitter-base junction.

The design information in the preceding paragraphs is offered for those of you whose requirements differ greatly from the ones shown in **fig. 3**. If you have or are building a transmitter which has a low-voltage supply of about 300 volts and requires about 200

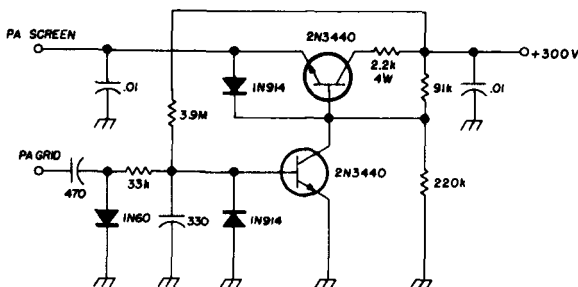
volts in the screen, the circuit may be used without changes.

operation

When the board is mounted and all connections are made, turn on the transmitter. Measure the voltage at the screen terminal with a vtvm. It shouldn't read more than several tenths of a volt. Before going on, switch the voltmeter to a range at least as high as the screen supply. Next, key the transmitter. If all is well, the voltage should jump to the level determined by the voltage divider.

In case of trouble check all connections first. Be especially careful of connections to the transistors and diodes. If the screen voltage does not drop with loss of excitation, try

fig. 3. Schematic of the solid-state screen-clamp circuit.



reducing the size of the 3.9M resistor. If the screen voltage does not rise with excitation, check for a low negative voltage at the base of the first transistor. If the voltage is present, it indicates the peak rectifier and filter are operating and that one of the transistors is probably defective. When correct operation is obtained, the transmitter is ready to use on all modes.

The transistorized screen clamp outperforms the old 6AQ5 and uses a small fraction of the power. As evidenced above, it has held up well in operation. I strongly recommend that you use this circuit or a variation of it if you're building any new equipment requiring a screen clamp; the savings in power and the reduction of heat are worth it.

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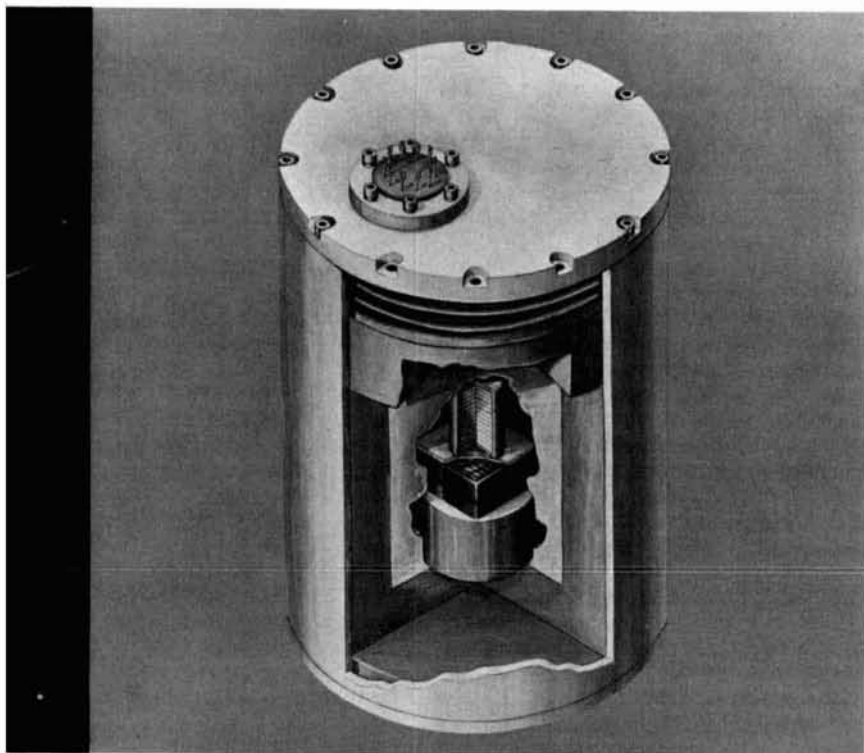
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a discussion of thermoelectric power supplies

Current prices rule out thermoelectric generators for hams, but in not too many years you may be using them on field day

With the growth of the atomic age, a new breed of power supplies has come into being —thermoelectric power generation. These are direct-conversion units with no moving parts, maintenance and lubrication free, silent in operation, and which provide one of the most promising long-lived, dependable power supplies for extended space trips.

Since they are so dependable, the Nassau Satellite Tracking Amateur Radio Society has chosen this type of power supply for the Moonray package they hope to put on the moon in 1970. Some day in the near future we will find these power supplies being used from the outermost fringes of our galaxy to twenty-thousand leagues under the sea.

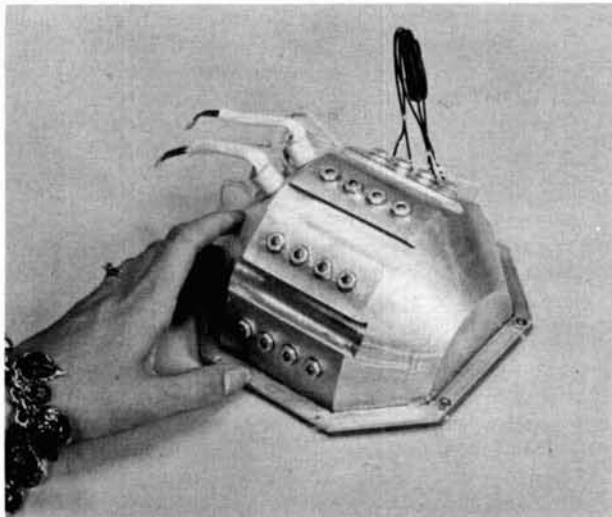
Jesse Bryant, K1AJE, 21 Winnhaven Drive, Hudson, New Hampshire 03051

Thermoelectric power supplies are not limited to costly space programs. As of now they are in use around the world in remote positions, above, in, and under the sea. That buoy you passed on your last weekend cruise could possibly have been powered by a thermoelectric generator supplemented by batteries, and operating at a considerable saving in cost over conventional systems.

Groups who are contemplating remote mountain-top repeater stations, microwave links, etc., would do well to consider this type of power supply where electricity is not readily available. For example, a commercially available thermoelectric power supply with approximately 12 watts continuous output (12 volts, 1 ampere) would require about 300 gallons of propane per year.

Wait! Before you stop reading, remember that although 12 watts is not much, it's enough to charge a ballast battery (preferably nickel-cadmium for long-life and low maintenance) during non-transmit periods. A 10-ampere-hour, nickel-cadmium, 12-volt battery could supply a 120-watt steady-state load for one hour—but our loads are only intermittent. With properly designed solid-state transmitting and receiving equipment, the peak-power duty cycle is very low. The 12-watt thermoelectric generator would continue to supply power day and night for

Commercial 1-watt thermoelectric generator. Total weight is 4.8 pounds; volume, 0.07 cubic feet.

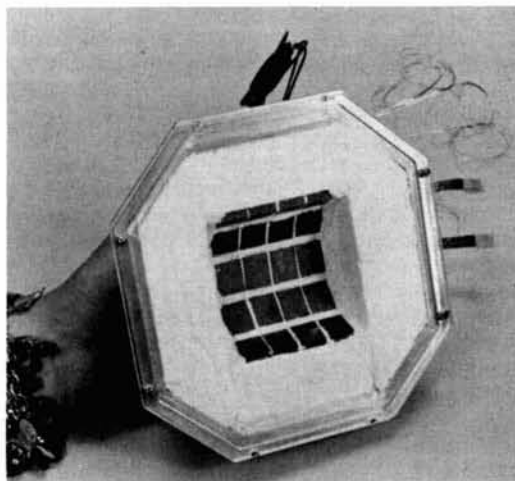


charging the ballast batteries—they would permit a relatively high-power drain during periods of peak power requirements.

theory

Direct conversion of heat into electricity has been with us for over 145 years. In 1822, Johann Seebeck reported he had observed that a magnetic needle was deflected when placed near a closed loop of two dissimilar metals when a temperature difference was maintained between the two junctions. A very basic thermoelectric circuit is shown in **fig. 1**. One end is hot and the other end is

Inside the 10-watt thermoelectric generator. Fuel is Plutonium-238.



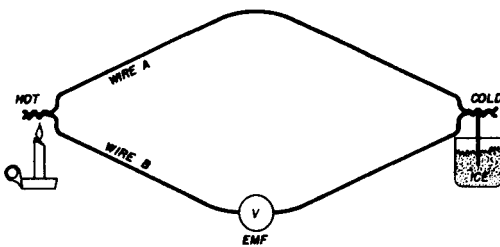
immersed in an ice bath; by using dissimilar metal wire in a closed loop, a small voltage is developed. With the exception of its usefulness in measuring temperature, Seebeck's discovery went rather unnoticed. In fact, you can buy a complete set of temperature vs emf tables for various dissimilar metal thermocouples from the U. S. Government Printing Office.

Very little progress was made until after the discovery of the transistor in 1948. The demand for ultrapure metals for semiconductors made it possible for further research on thermoelectrics. It was found that by doping certain pure metals, a positive or negative

valance could be achieved as in transistors and diodes. By making a closed loop and applying heat, at temperatures which were previously considered destructive, to one junction and cooling the other, considerable power could be generated. So much interest was generated in this type of thermoelectric that the Navy contracted Westinghouse for material research and development with the possibility of quiet power generation for ships.

A typical thermoelectric module is shown in **fig. 2**. By applying heat to one end and

fig. 1. The basic thermoelectric circuit.



cooling the other, power is produced. The voltage output will be very low, but the amperage output could be quite high, 25 amps or more, depending on the size of the module. The P and N legs are hooked in series; by hooking many more together, you can get an appreciable voltage output. As with battery circuits, voltage will increase as cells are added in series, but current is common to all.

The type of material and size of the legs determine the electrical resistance. This is normally very low, approximately 0.01 ohm per P and N leg. The thermoelectric materials are very susceptible to poisoning when at operating temperatures; therefore, they are usually sealed in an inert atmosphere or vacuum.

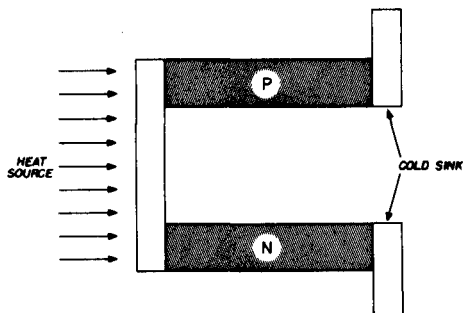
Some of the metals now used are silicon-germanium, bismuth-telluride and lead-telluride. They are usually selected according to operating temperature, ease of fabrication and cost per unit. As the state of the art improves in metallurgy, the temperature limit on the hot end is reaching 2000° F and above.

radioisotopic thermoelectric generators

With the recovery of radioisotopes from spent nuclear reactor elements, we have a constantly enlarging fuel supply for use as a source of heat in thermoelectric generators. The radioisotope is a byproduct of nuclear reactor operation and generates heat as it decays. Radioactive isotopes are rated in half lives; that is, they will decay to one-half their original thermal power when they reach their half-life. Typical half lives of several radio-active elements or isotopes that show promise in the thermoelectric field are listed in **table 1**.

With the present state of the art, the efficiency of thermoelectric generators is approaching 10%. If a generator output of 10 watts (thermal) is required at the end of 5 years (half-life of fuel), assuming 10% effi-

fig. 2. Typical thermoelectric module.



ciency, the initial fuel load for the generator must be 200 thermal watts. At the beginning of life, 100 watts of thermal heat must be dissipated. This is wasteful but necessary if the rated power of 10 electrical watts is to be maintained at the end of 5 years. This is called power flattening or dampening.

A radioisotopic thermoelectric generator for space applications is shown in the photographs. This generator has an unbelievable two-watts-per-pound power-to-weight ratio and is available in 5- to 50-watt output class. As shown in the graph in **fig. 4**, the generator has an optimum operating range where the load resistance matches the generator electrical resistance and provides maximum peak

power.

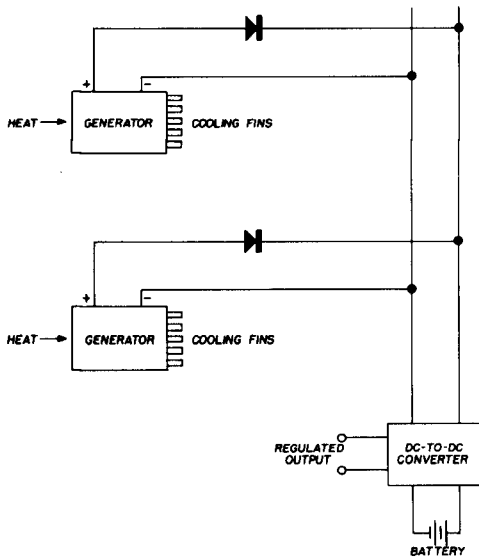
The high cost of efficient generators and the tremendous cost of the long-lived fuel, Plutonium-238, which is controlled by the Atomic Energy Commission, leads to caution when considering the use of isotopic generators. However, they can be used to charge batteries (made to last 5 years or more) which will supply peak power demands.

The battery pack will supply tremendous power for a short time and is rechargeable with a thermoelectric generator. At the end of the sensed battery life, the batteries may

operate repeaters and microwave relay stations could operate using batteries that are recharged by the generator. A typical low-power generator is shown in the cutaway drawing.

If you don't want to go nuclear, there are several manufacturers offering off-the-shelf generators with outputs of 100 mW to 50 watts and higher. These are usually heated with fossil fuel such as gas, oil or propane; any heat producer can be used. They are presently used to charge communications batteries on railroads, supply power for Coast

fig. 3. Method of putting several thermoelectric generators in parallel.



be disconnected remotely; low power would still be furnished by the thermoelectric generator until the end of several half-lives of the radioisotope fuel. For the Moonray project the thermoelectric generator has another benefit: when the cold lunar night sets in, the heat from the radioisotope fuel could keep the batteries from freezing.

For less sophisticated applications there is another breed of generators called terrestrial generators. These are for remote applications and use cheaper fuel and abundant shielding. They can be placed anywhere. Automatic re-

Guard buoys, supply power to fire towers in remote areas and in many many more ways.

Thermoelectric generators are unique in that the output can be short-circuited temporarily without harm to the elements. They are inherently self-protecting. The only real dangers are overheating and exposing the internal parts of the generator to air at operating temperature. Most manufacturers build in safety features to protect against overheating; they also seal the internal parts.

In most thermoelectric generators, the voltage is very low—3 to 5 volts output. This is

almost useless unless it's converted to a higher output voltage. Dc-to-dc converters are used to step up the voltage to a useable value such as 6, 12, 24 or 28 volts using highly efficient toroid-core transformers and solid-state circuitry.

It's my opinion that this can be improved by adding enough legs to get the voltage up to 13 volts or more. Paralleled generators would be possible with less power loss by using diodes to prevent the flow of current to the lower voltage generator.

Diodes have a 0.75-volt drop, and with low voltage units this is a considerable power loss. However, by increasing the initial voltage, 0.75 volt would be a small loss, and generators with different characteristics could be used as shown in fig. 3. By using thermoelectrics on both sides of the heat source, costly insulation would be replaced with another bank of thermoelectrics, thereby increasing efficiency.

other uses for thermoelectrics

Ok, so you don't want a unique power supply. How about cooling the hot final in that super-duper factory-built rig that threatens to melt under sustained key-down conditions or the VFO that never stops drifting? By passing current with less than 1% ripple through a thermoelectric module, you can have a cooler; one end will get cool, and the other end will get hot. In effect it's a heat pump. When you reverse the polarity, the

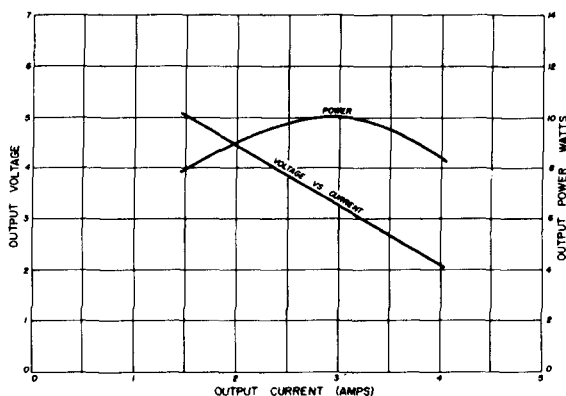
table 1. Typical half lives and power density of several radioactive elements that show promise in the thermoelectric field.

isotope	half life years	power density thermal watts/in ²
Thulium 170	0.3	100
Thulium 171	1.9	10
Strontium 90	28.0	6
Plutonium 238	89.0	25

hot and cold ends will also reverse. These are off-the-shelf items carried by most of the large electronic distributors and are fairly reasonable in price.

Thermoelectric power generation is still in its infancy, and advances are being made every day. In the next five to ten years, engineers will be getting the power up and the

fig. 4. Operating range of the thermoelectric module pictured at the beginning of the article.



price down—priced low enough perhaps to compete with batteries at the corner drug store. Can you think of a better way to power that field-day station?

references

1. Westinghouse Thermoelectric Handbook.
2. "General Principles of Thermoelectric Thermometry," Leeds and Northrup Company.

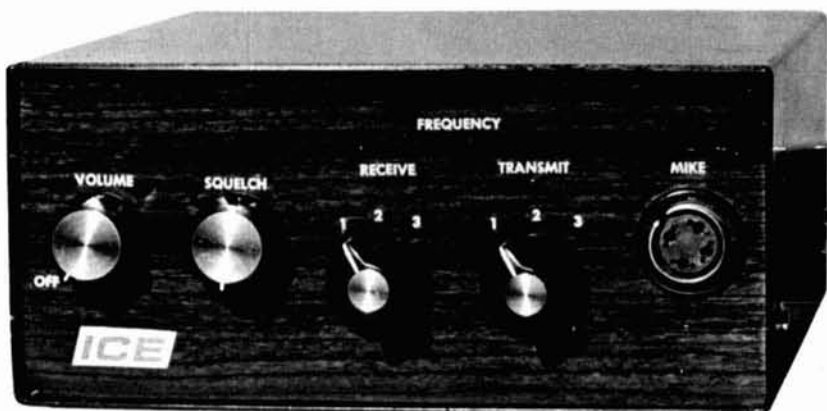
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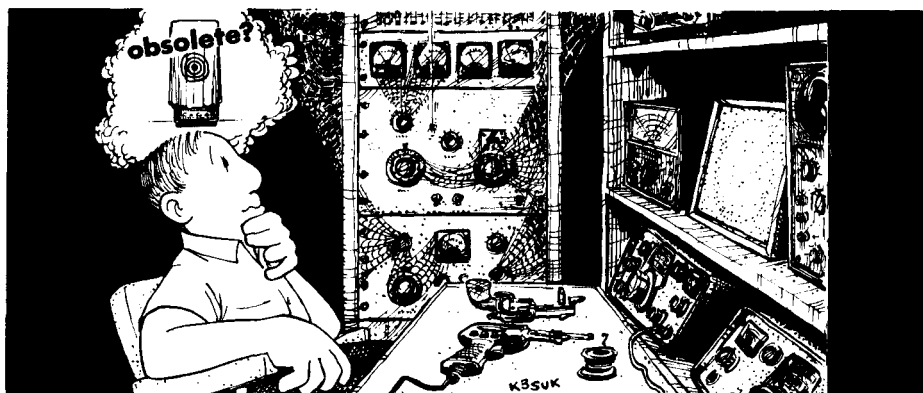
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some thoughts on
converting a-m
power amplifiers
to ssb service

Bill Blankenship, WA4GNW/W1RDR 6709 Green Meadow Court, Louisville, Kentucky 40207

In countless ham shacks across the country there are a great many a-m components and complete a-m rigs which you may think are obsolete, but in actuality, they can be used with outstanding success on ssb. After many years on a-m, I was faced with the same problem as most amateurs changing to ssb: whether to junk existing equipment, sell it at give-away prices and buy new gear, or modify the old rig for ssb operation. Most amateurs buy a commercial ssb transceiver which runs at power levels of 100- to 500-watts PEP. Ultimately, however, they feel the urge to tack on a linear for increased power.

Unfortunately, the average amateur doesn't have sufficient know-how or time to design and build a complete ssb transmitter or transceiver, so it makes good sense for him to buy a commercial unit. When it comes to the linear, it's a different story. Back in the attic among the cobwebs are many fine pieces of equipment which are equal or superior to anything commercially available. This is not intended as an indictment of the many fine commercial linears available; but if iron is more plentiful than silver around your shack, the following may prove of interest.

My first step in "going ssb" was to put together a Heathkit SB-401 transmitter. My old kW plate-modulated amplifier was an unlikely candidate for ssb service—at first glance. The basic circuitry didn't conform with the generally-accepted configuration of today's ssb linears. It used a pair of push-pull 250TH's in a tuned-grid, tuned-plate configuration with a split-stator butterfly tank capacitor with a plug-in coil and swinging-link plus a five-band coil turret in the grid

circuit tuned with a two-section capacitor. Sound familiar?

When I took a closer look at the amplifier, I couldn't see any reason why a push-pull circuit, with its inherent harmonic cancellation, wouldn't actually be better than a single-ended final. Similarly, a tuned-grid circuit should be as good or better than a broadband untuned input. The use of 250TH's is irrelevant, and many other tubes will perform just as well. If the a-m amplifier uses a single-ended configuration, so be it. The point is that many existing a-m amplifiers can be converted to linear service at practically no additional cost.

power supplies

The plate voltage supply consists of two 4400-Vct, 1.5-kVA hypersil-core pole transformers in parallel, a pair of 872A rectifiers, a capacitor-input filter consisting of a 2- μ F, 4000-V oil-filled capacitor, a 20-H broadcast-type choke, another 2- μ F oil-filled capacitor and a 100k-ohm, 200-watt bleeder resistor. The 3-kVA continuous-duty rating of the plate transformer is the key to the hard regulation I get with this supply without the big filter capacitance usually required. The plate transformer never has to deliver more than half of its rated output. From idling plate current to maximum current on voice peaks, voltage regulation is on the order of 3%.

Grid bias is obtained from a 2500-ohm, 100-watt resistor across the output of a heavy-duty full-wave supply. The resistor is tapped at the grid cut-off point of the 250TH. No grid-leak bias is used. An overload relay protects the tubes in case of excessive plate current.

amplifier modifications

I installed a small relay adjacent to the bias supply and connected the cut-off bias voltage to one set of relay contacts. A second slider tap on the bias supply resistor was positioned so the final draws the correct no-signal idling plate current when plate voltage is applied; this bias is connected to the normally-open relay contact. The armature is connected to a feedthrough capacitor which feeds the bias to the shielded grid circuit. The relay coil is connected to an

"external relay" socket on the exciter. When receiving, the final is biased to cut-off, and when the exciter's PTT-VOX relay is closed, ssb operating bias is applied to the amplifier.

A dpdt center-off switch replaces the original spst plate voltage switch on the front panel. This allows instant switching from ssb to high-level plate-modulated a-m, primarily for working DX stations still operating a-m on 10 meters. If you're only interested in ssb operation, this change is not needed.

Conventional control circuitry is used: one side of the ac line to the amplifier is connected to the high-voltage plate relay coil. The other side is wired in series through the door interlock, the overload relay contacts and one set of contacts on the dpdt front panel switch to the other side of the coil. With this wiring, high voltage to the amplifier is independent of exciter control. For a-m operation one side of a two-conductor control line from the a-m exciter is connected to the plate relay coil; the other side is connected as above except that it uses the other side of the dpdt switch so plate voltage can be controlled by the "transmit-standby" switch on the a-m exciter.

A switch in the modulator filament circuit and a switch which operates a relay to jumper high voltage across the modulation transformer secondary for CW operation are used for ssb. Therefore, three switches and a relay convert the amplifier from 1-kW plate-modulated a-m to 2-kW PEP ssb.

operation

The amplifier is loaded to 1200 watts dc input—3000 Vdc at 400 mA with the exciter switch in the "tune" position. Switching the exciter to upper or lower sideband, a steady tone is applied to the mike and the mike level control advanced until the amplifier plate current is 330 mA—990 watts dc input.

In regulating the amount of drive to the amplifier, the exciter should be tuned for normal maximum output. If this produces excessive drive to the amplifier, the exciter drive control can be reduced slightly. However, make sure that the drive to the exciter's own final is not reduced below the manufacturer's specifications for minimum drive. Otherwise, the alc circuit and carrier sup-

pression will be upset and signal quality seriously affected. If minimum permissible output from the exciter still results in excessive drive, use an attenuator.

construction

Little specific mention has been made about design and construction of the amplifier. Just remember that the dictates of good engineering practice should be followed when adapting an existing amplifier to ssb service. Briefly, the amplifier should have a well-regulated bias supply; a plate voltage supply with minimum voltage drop from idling current to full current drawn on voice peaks; adequate isolation of grid and plate circuits; good neutralization; complete shielding; impedance matching to the transmission line; a good low-pass filter and adequate cooling. My amplifier—which is typical of most well-designed a-m amplifiers—incorporates all these provisions.

One inexpensive way to match a swinging link to coaxial line is by using a 30- to 500-pF variable in series with the center conductor of the coax line. The capacitor is mounted inside the shielded enclosure on an insulated plate with its shaft extended through the front panel. To adjust the link correctly, mesh it to the point where the amplifier draws a very small amount of current and tune the loading capacitor for a peak current reading (usually an increase of a few mils). Then mesh the link until the amplifier is drawing the proper operating current.

If the link is correctly tuned, retuning the loading capacitor after the amplifier is loaded to the proper value will indicate no change in the resonant point of the link-tuning capacitor. Fairly wide frequency excursions within any given band are possible without retuning. The tuned-grid, tuned-plate, tuned-link configuration provides very

effective suppression of spurious radiations.

cooling

Proper cooling of the shielded plate-circuit compartment is very important. To quote a recent article by a well-known ham engineer, "The only time you have too much cooling is when the blast of air blows the tubes out of the sockets." I might add that you may also have too much cooling—or a bad bearing—when blower noise makes it hard to hear the receiver. My amplifier uses the pressurized chassis method; the chassis has a solid bottom plate.

The top of the chassis forms the floor of the plate-circuit compartment with twenty-four 1/4-inch holes drilled around the periphery of each tube socket. Mounted on the rear edge of the chassis are two 4-inch centrifugal blowers rated at 250 cfm each with zero back pressure. Since the total area of the drilled holes is considerably less than the combined area of the two blower nozzles, there's a certain amount of back pressure, but this is inconsequential. Blower noise is at a very tolerable level and do a nonsense cooling job.

An old axiom states that, "the proof of the pudding is in the eating." The old warmed-over pudding at this QTH is tasty indeed. I have been literally swamped with unsolicited signal reports lauding the quality and strength of the signal. So many reports have been made of, "strongest signal on the band," "beautiful pattern on the scope," "extremely narrow, but pinning the S-meter," etcetera, that I'm almost beginning to believe it. I hope this apparent immodesty will spur some of you to make your own conversions. Borrow a derrick and drag that old a-m rig out of the corner and join the fun!

ham radio

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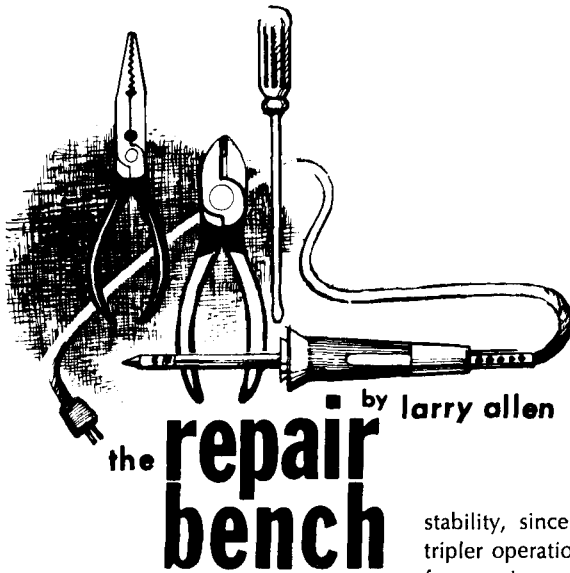
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aligning vhf transmitters

Speaking generally, aligning the broadband tuned circuits between stages in a vhf transmitter is a lot like aligning them in any transmitter. Yet, there are subtle differences you should know about. Take a simple 2-meter a-m transmitter as an example; one is shown in block-schematic form in **fig. 1**.

First, you need some sort of alignment indicator. Monitoring rf power output of the transmitter is inadequate for adjusting the early stages; the indications are too broad to be any help. A vtvm connected across the grid resistor of a late stage—say the driver—is usually effective; it measures class-C grid current. The manual that comes with the transmitter may tell you what point is best.

The first slight peculiarity arises in adjusting the oscillator. In many vhf transmitters, inexpensive 8-MHz fundamental crystals generate the carrier, as shown in **fig. 1**. However, the oscillator is also a tripler. Its output is at 24 MHz. Tuning its plate coil isn't exactly as simple as merely peaking it for maximum vtvm indication, but the proper technique is seldom spelled out in a manufacturer's alignment instructions.

For oscillator alignment, you should plug in the highest-frequency crystal you have. This permits you to set the plate coil for best

stability, since any instability of oscillator-tripler operation shows up first at the higher frequencies.

Most important is the way you do the adjustment. Key the transmitter, and tune the oscillator plate coil for a peak on the vtvm. Don't stop with that, though. Rock the slug back and forth a little; you'll notice the reading drops off more quickly on one side of the peak than on the other. With the adjustment peaked, turn the slug in the direction of the **slow** drop-off—just enough so you can notice the lower meter reading. Leave the slug set slightly on the "slow" side of the peak. That's the point of maximum stability. This adjustment is even more critical when the oscillator uses overtone- or harmonic-type crystals.

For the later stages, such as the driver and the final, it is better to monitor rf output. In vhf power stages, the plate-current dip doesn't usually occur exactly at the most efficient point. Therefore, for all output-stage adjustments, an rf wattmeter or other rf output meter is the most effective tuning indicator.

For interstage and output adjustments, use a crystal frequency near the center of the over-all range. Only the oscillator needs the special treatment at the high end. Just peak the interstage multiplier or amplifier adjustments for maximum rf output.

The order in which you make the final-stage and output-coupling adjustments can be a factor in vhf transmitters. Start with coupling as loose as possible (least rf output). If there is a plate-current meter, use it

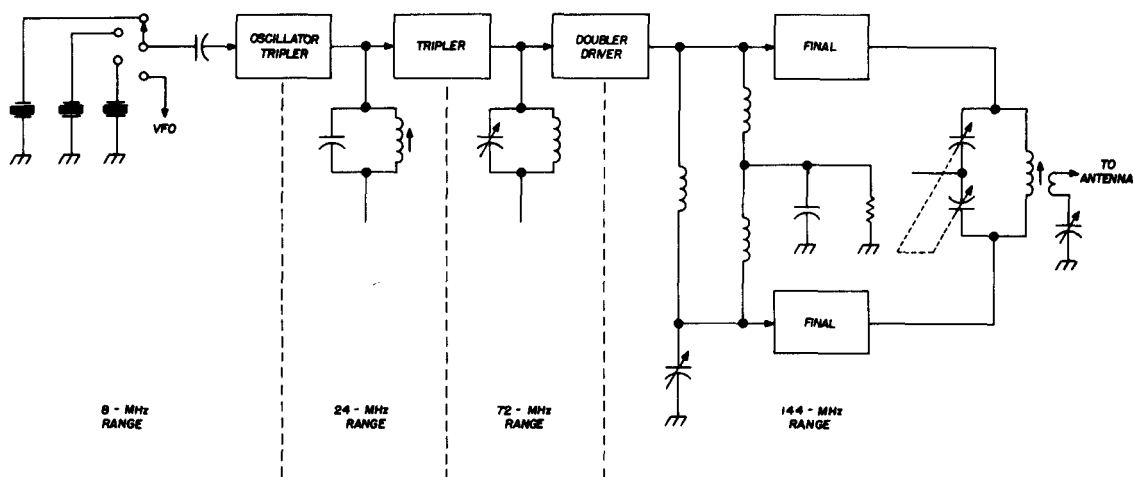
to dip the power-amplifier tank as a preliminary step; later, however, you'll retune the tank for maximum rf output. Drive in vhf transmitters is seldom adjustable; you just tune the driver stage for strongest drive to the final.

Start by increasing antenna coupling just enough to get some output reading on the rf wattmeter or whatever you're using as an output indicator. Ignore final plate current at this point. Now tune the plate tank for maximum rf output. Then peak the antenna trimmer, if there is one. In small steps, increase rf output to the rated wattage for the transmitter, repeating the sequence just outlined: increase coupling, tune plate, tune

indicator. At the grid of the final is a good place to hook a vtvm to check alignment results in the earlier stages. Loosen antenna coupling to begin with.

Plug in the highest-frequency crystal you have for either band. Key the transmitter and peak both slugs (top and bottom) in the first interstage transformer. Then go back and shift the plate-winding adjustment slightly to the stable side, as already described. Change to a crystal whose frequency is near the center of the bands. If necessary, use two: one near the center of each band. Peak both adjustments of the second interstage transformer for best drive at the final, measured by the vtvm.

fig. 1. Common configuration for an amateur vhf transmitter.



antenna. Don't allow plate current to exceed whatever is recommended for the tube; also, don't exceed the FCC-ordained plate power input. Don't tune the plate tank for a dip on the meter—tune for output, always. If plate current gets too high, back off the coupling so there's less rf output; then retune the plate and antenna controls for peak output.

a-m vhf transmitter

Another vhf ham transmitter is diagramed in fig. 2. This one is a two-band. Also, interstage coupling is by double-tuned transformer, although that makes little difference to alignment—just a couple of extra slugs to turn. Let's run through its alignment quickly.

Connect a dummy load and rf output

Switching from 2 meters to 6 meters changes the interstage coupling to the final (power amp), and also alters the mode of operation in the last multiplier stage. For 6 meters, it becomes merely an amplifier. There's no driver-stage adjustment for 6 meters, so switch to 2-meter operation. Peak the two series-resonant trimmers for maximum drive to the final.

Output coupling systems for each band are separate and different. Tune the 6-meter band first, which uses the pi-network output system. The standard alignment procedure applies, except that there is no coupling adjustment. Tune the plate trimmer (C1) for maximum rf into the dummy load. Then adjust the antenna-tuning trimmer (C2). Peak

them both, going back and forth a couple of times to make sure interaction doesn't affect the result. You'll have to refine them both a little when you connect an antenna.

Switch to the 2-meter band. Output coupling is a simple link, series-tuned. Peak the trimmer for maximum output, with either the dummy load or the antenna connected. If you use the dummy, retouch the adjustment when you change to the antenna.

Many vhf transmitters have the plate and antenna tuning controls on the front panel. They should be readjusted with each crystal change, or with each change in VFO tuning. Always peak them for maximum rf output.

single-sideband vhf

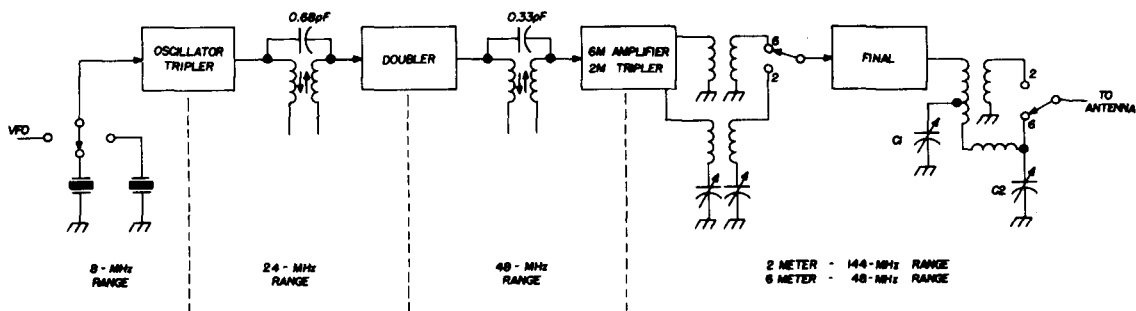
There are some differences between vhf transmitters for a-m only and those that operate in the ssb mode. The variations aren't major, but they are sufficient to alter the general alignment procedure. The block-schematic of fig. 3 shows the layout of one double-conversion vhf ssb transmitter—from the Gonset Sidewinder series. This one is about average among this class of transmitter, so it makes a fair example. The steps,

VFO, which is part of the transmitter, you can tune any frequency within the sector. The carrier oscillator is the 9-MHz type that is common in modern transceivers. It is this synthesis method of developing the vhf output frequency that dictates a special sequence for transmitter alignment steps. The procedure isn't as straightforward as in simpler a-m units with frequency multipliers.

The 9-MHz i-f amp and the 15-MHz band-pass i-f amp are part of the receiver, too (the Sidewinder units are transceivers). You can align the two stages with a signal generator; a sweep generator is recommended for the band-pass amplifier. However, you can also do a fairly close job with the transmitter keyed on, using the signals generated by the unit's own oscillators. Follow the manufacturer's instructions religiously if you have the test equipment; if not, the technique I'll outline here will work.

The first step is to align the 9-MHz section. Disable the tuning oscillator or VFO. Connect a vtm with rf demodulator to the output of the first mixer (the transistor collector, in this instance). Key the transmitter on, with its mode switch set for CW operation. Peak

fig. 2. Two-band vhf transmitter is only slightly more complex than the design shown in fig. 1.



and the order you do them in, can be adapted to other ssb transmitters in the vhf bands.

This transmitter uses frequency synthesis to tune the 6-meter band in sectors. Four 35-MHz crystals handle this. The tuning oscillator lets you use crystals from 5.5 to 6.5 MHz to select spot frequencies within the 1-MHz spread of each sector; with the

T10 for maximum. Move the vtm probe to the 9-MHz input of the mixer (emitter, here). Adjust the crystal trimming capacitor (C23) for exactly 65 mV on the vtm. Move the vtm probe back to the mixer collector again, and again peak T10. The trimmer capacitor puts the crystal signal at exactly the right spot on the response slope of the sideband filter, and you have peaked T10 to

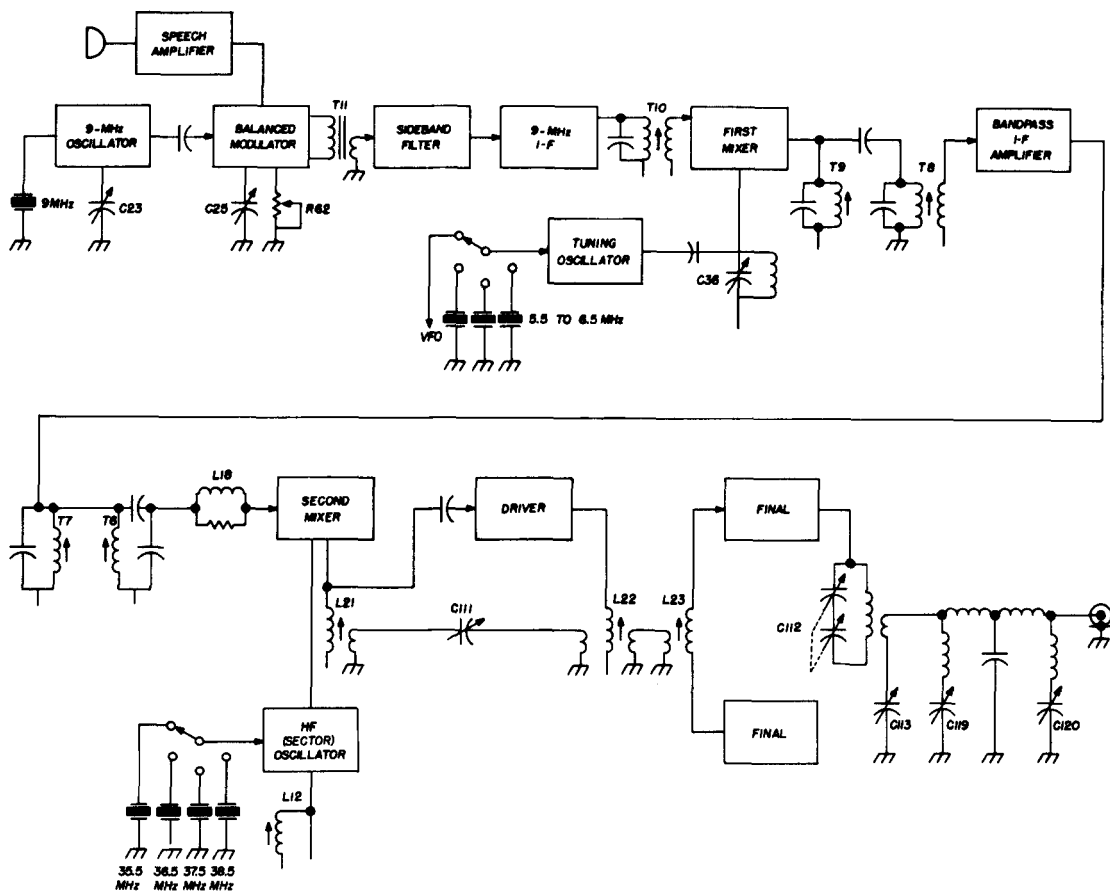


fig. 3. Adjustments and their sequence are complicated in this vhf single-sideband transmitter. Special equipment is recommended for adjusting some of the tuned band-pass circuits.

pass the resulting 9-MHz signal efficiently.

Reactivate the tuning oscillator or VFO. Move the vtvm probe to the input of the second mixer. Tune the VFO to the center of its dial, and peak C36 for maximum reading on the vtvm. That couples the tuning oscillator most efficiently to the first mixer.

Next, if you do have a frequency meter, check the frequency of each of the four sector crystals. If you don't have a frequency meter, you must simply take them for granted. If, later, you find one sector isn't accurately calibrated, its crystal is probably off frequency. It's unlikely that more than one will be off. If they all are off, you can warp them onto frequency with L12.

Move the vtvm probe to the output of the second mixer. Now align the band-pass i-f

amp. It must be stagger-tuned and is best done with a sweep generator. However, you can get by this way: key the transmitter and peak T9, T8, T7, T6, and L18 for maximum signal reading on the vtvm. Go over all five a couple of times to make up for any interaction. To get the bandwidth needed, go back to T7 and T6. Turn the slug of T7 **counterclockwise** one quarter-turn, and the slug of T6 **clockwise** one quarter-turn. This is a rough approximation, but you can usually get by. To check it, tune the VFO from one end to the other of its range. The vtvm reading will vary from one end to the other, but there should **not** be any sharp dropoff in the reading near either end of the VFO range. If there is, try correcting it with T8, T9, or L18. If you can't, realign them

at their peaks and try correcting the too-narrow band pass with T7 or T6.

Now go back to the front end and adjust the balanced modulator. Start with the transmitter still in the CW mode. Connect the vtvm rf probe at the input terminal of the sideband filter. Key the transmitter, and peak T11. Switch to the upper-sideband mode, and adjust R62 for a null (minimum on the vtvm). With a nonmetallic screwdriver, rock C25 slightly. If the rf indication on the vtvm can be reduced by turning C25 in either direction, turn it that way *only slightly* and re-null R62. Continue this in small increments—C25 for a slightly lesser reading and then R62 for minimum—stopping when turning C25 further starts increasing the meter reading instead of decreasing it. The meter should finally null at less than a millivolt.

Interstage coupling from the second mixer to the driver and final stages is also band-pass circuitry. Though they are not stagger-tuned, the circuits are heavily interdependent. Also, the neutralization network that includes L21 and capacitor C111 complicates adjustment. The manufacturer recommends alignment with a sweep generator, and it's a good idea. If you don't have a sweep generator, or don't know how to use it, **do not** bother the neutralization.

To align the driver-coupling circuits as best you can, connect a dummy load and rf output indicator to the antenna output jack. With the transmitter in the CW mode, and the sector switch set at the 51-52 range, peak L22 and L23 for most rf output. As you switch sector crystals over the entire 6-meter band, with the VFO set at center range, you should notice some variation in output power. If it varies more than 5 or 6 watts, try readjusting L22 and L23 to compromise between absolute maximum output and fairly even distribution of power over the band.

Tune capacitor C112 (the final-amplifier tune knob on the front panel) for maximum transmitter output, with the sector switch set to produce 51-52 MHz and the VFO at center. Also tune C113 (antenna tune on the front panel) for best rf output.

antenna adjustments at vhf

Any vhf transmitter should be connected to its working antenna for finishing adjust-

ments in the final amp and output network. Antenna matching is critically important in vhf operation, and the adjustments in the final stages can seriously affect this matching.

The most useful indicator is an in-line wattmeter or reverse-reading rf indicator. It connects in the coax line between the vhf transmitter and the antenna. Lacking this kind of meter, you can get by with a simple field-strength meter.

With the in-line rf indicator, you first tune the final tank for maximum output. If the output matching network has several adjustments as does the one in **fig. 3**, start by setting the panel control to the center of its capacitance range (the plates half-closed); you may have to open the cabinet to see where halfway is, if there are no stops on the capacitor. Tune the first trimmer (C119, here) for best output. Change the indicator to read reflected power and tune the last trimmer for minimum reflected power. Then touch up the first trimmer for minimum, too.

Recheck forward power. Touch up C113, and then—still measuring forward power—touch up C119 and C120. If you have to turn these two very much, and if turning them increases the reflected power indicated by the meter, the mismatch is serious and is at the antenna itself. You can't "adjust" it out.

If you are doing the output alignment with only a field-strength meter, adjust the final-amplifier tune control first, then the antenna tune. Set both for maximum rf radiation measured by the meter. Adjust the trimmers also for maximum radiated field strength. Go through the final-amplifier tuning, antenna tuning, and matching adjustments several times to overcome any interaction.

With any vhf transmitter, your best bet is always to follow the alignment instructions supplied by the manufacturer. If you need special equipment, buy or borrow it. For best communications, always try to do a thorough alignment job. It takes time and care, but the results are worth it. In a later column, if you want to know more about it, I'll tell you how to set up the scope and sweep generator to do band-pass alignments in receivers and transmitters (important in ssb transceivers). Drop me a note and let me know.

ham radio

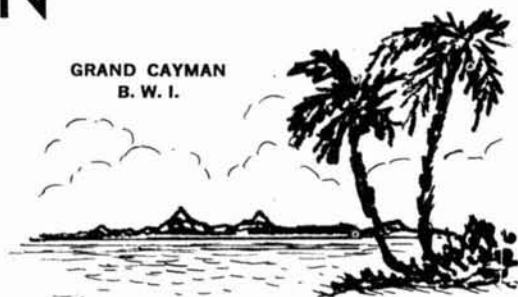
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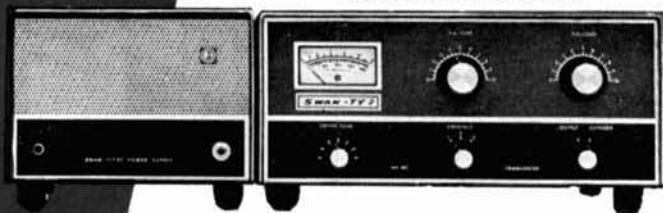
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the new heath HW-100 five-band ssb transceiver

One of the more intriguing new pieces of equipment to be placed on the market in recent months is the Heath HW-100—a five-band ssb transceiver selling in kit form for only \$240.00! This certainly bears looking into, and a more careful review uncovers a very interesting piece of gear.

A quick definition might call it a cross between the well-known Heath HW-series single banders and their well-respected SB-101 that sells for \$130.00 more. Nothing has been left out of the HW-100. It gives complete coverage of all amateur bands between 3.5 and 29.7 MHz in 500-kHz segments with selectable sideband or CW operation. Included are features sometimes considered extras—CW sidetone, VOX and a 100-kHz crystal calibrator.

When the kit arrived here at **ham radio**, we looked it over and were all quite impressed to find that the folks at Heath hadn't cut any corners quality-wise to give you a dependable piece of equipment. I'd certainly hate to try and put a similar group

of parts on my bench for the same price. Among the branded components were many well-known names including Westinghouse, E. F. Johnson and Potter & Brumfield.

assembly

The kit should go together in about 35 or 40 hours, although I did a bit better. Alignment and tune-up went very quickly and were not at all difficult. About another hour or so had the rig on the air in good order. As is usual for the larger kit manufacturers, the directions were very detailed and excellently prepared. There were one or two areas which might have used a bit of clarification, but I suspect that these will be cleared up in later kits. Ours was in the first group of production units shipped. None of these problems was serious, however, and by reference elsewhere in the instructions they were quickly clarified.

Most of the wiring was handled by nine circuit boards; a large wiring harness provided most of the interconnections between the boards, controls and other points. The only

Skip Tenney, W1NLB, Fayston, Vermont 05673

two major portions of the transceiver which are hand wired are the VFO and the final amplifier. Both of these were quite straightforward and went together very nicely. The manufacturer was very careful to provide ample warning regarding any operations requiring special care such as the use of heat sinks on semiconductor leads, etc.

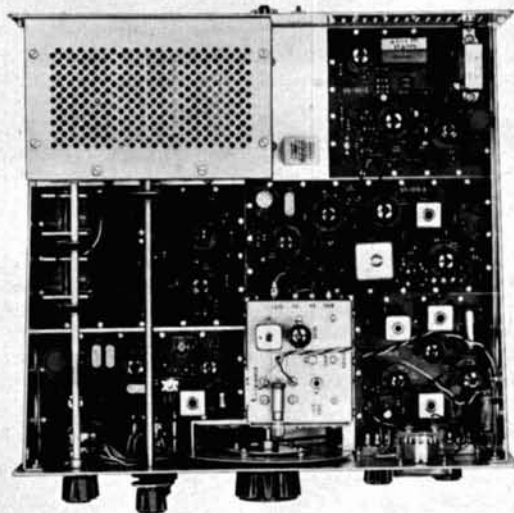
As mentioned earlier, the unit was amazingly easy to align. The receiver is tuned up first. This procedure requires only a vtm and another receiver (even a broadcast receiver will suffice). As you are aligning the receiver, most of the stages of the transmitter are tuned up, too, since the receiver and transmitter sections share many stages and tuned circuits; final peaking is done later in the transmitting mode to assure maximum drive and proper operation.

One point of interest which didn't affect our kit, but did trouble a neighbor, was the initial tune-up of the transmitter. The instructions assume that T1 at the output of the balanced modulator will be sufficiently close to alignment as shipped from the factory so some drive will be available before this transformer is touched. Our friend found this to be untrue and had to jump ahead four steps in the alignment procedure to obtain any initial drive whatsoever. This may well have been a random case, but we pass it along for what it's worth.

receiver section

A block diagram of the HW-100 is shown in fig. 1. You will see that it is a very straightforward unit with a double-conversion receiver of conventional design.

The incoming signal is first passed through a single stage rf amplifier/preselector and on to a crystal-controlled first mixer. The first intermediate frequency is tunable from 8.395 MHz to 8.895 MHz. The second mixer converts this signal to 3.395 mHz and uses the VFO/VFO-amplifier from the transmitter as a local oscillator: this insures common

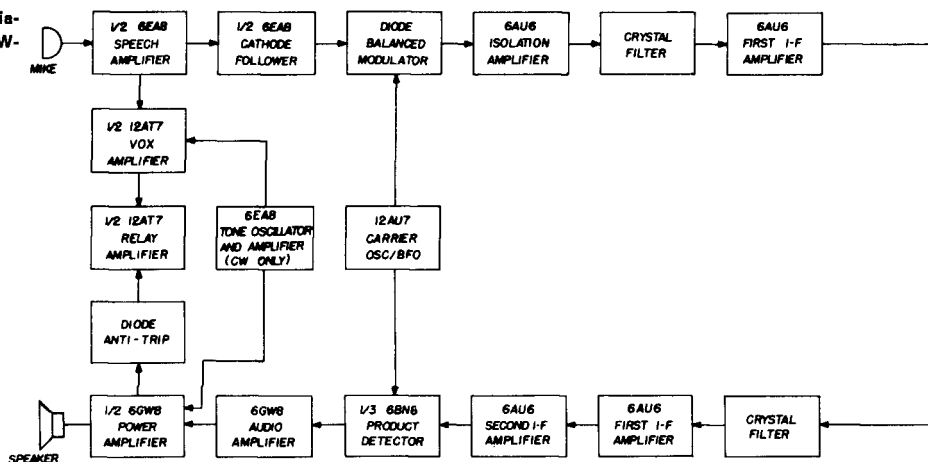


operating frequencies for both transmitting and receiving. It is at this frequency that a four-pole crystal-lattice filter is inserted between the second mixer and the first i-f amplifier.

This filter offers a shape factor of 3 to 1, which, although not outstanding, is more than ample for good selectivity, even under crowded conditions. After passing through two stages of i-f amplification, the 3.395-MHz signal is passed into the product detector along with the crystal-controlled output of the combination carrier-oscillator/BFO stage. The audio output of the product detector is passed through a two-stage audio amplifier to the speaker or headphone jack.

The operation of the receiver is identical

fig. 1. Block diagram of the HW-100.



in all modes of reception with the exception that the BFO frequency is varied for lower sideband, upper sideband or CW reception. Numerous a-m stations have been copied very successfully with this receiver using both the lower and upper sideband modes. The lack of an envelope detector for a-m certainly appears to be no disadvantage.

transmitting section

The transmitting section of the HW-100 is a typical example of a modern filter-type multiband ssb transmitter in the 150-watt class. In ssb operation, the audio signal, after passing through a speech amplifier and cathode follower, is mixed with the output from the carrier-oscillator/BFO in a four-diode balanced modulator to produce a double-sideband suppressed-carrier signal. During tune-up and CW operation the balanced modulator is unbalanced to allow the carrier signal to be passed on to the isolation amplifier which follows.

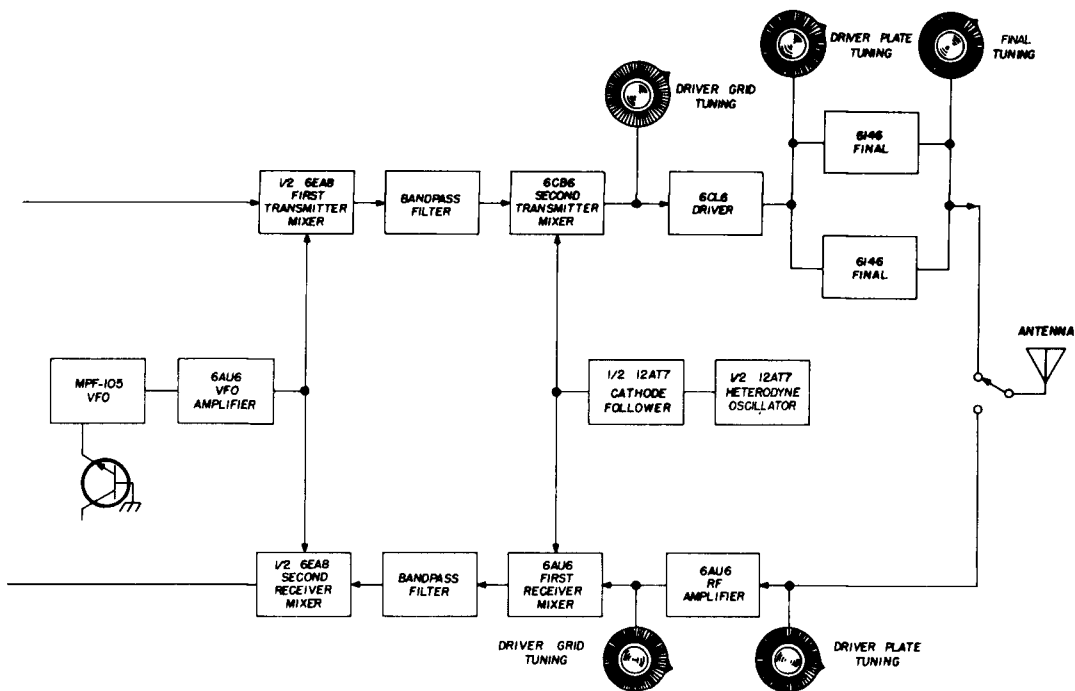
The same four-pole crystal filter is then used to both attenuate the undesired sideband and further suppress the carrier. For CW or tune-up operation, the frequency of the carrier oscillator is shifted to place it within the pass band of the filter. This filter does an adequate job of sideband suppression. Al-

though in running tests of my rig I have been able to copy the undesired sideband, it has been too weak to make any actual power measurements of it. I feel sure that the manufacturer's specifications of 45-dB suppression are being met and that under normal band conditions this sideband would never be detected.

In checking out carrier suppression, I was very pleasantly surprised. My test consisted of using a calibrated communications receiver with a short stub antenna. Coupling was adjusted to give a 60-dB S-meter reading with carrier inserted. When the HW-100 was switched to either sideband position, the S-meter reading dropped to below 5 dB—a difference in excess of 55 dB. I'm sure that this wouldn't be considered a completely accurate test by laboratory standards, but it certainly indicates carrier suppression well in excess of the manufacturer's claimed 45 dB. Local on-the-air tests have shown no sign of a carrier being transmitted.

From the filter the signal is given one stage of i-f amplification and is delivered to the first transmitter mixer. The VFO/VFO-amplifier is used as a local oscillator for this mixer.

The VFO is quite interesting and deserves some attention. It uses an MPF-105 FET in a



Hartley oscillator circuit. This solid-state design largely eliminates heat and the drifting which it causes. The engineers at Heath obtained regulated power for the MPF105 in a clever manner; they inserted a 2N3393 used as a Zener diode in the cathode leg of the 6AU6 VFO amplifier. This amplifier is run in a steady-state condition, so it provides a very stable source of low-voltage dc of ample power for a VFO circuit.

By varying the polarity of the voltage across a 1N191 diode, a carrier-shift capacitor is switched in or out of the VFO circuit. This causes a frequency shift in the VFO signal and maintains a constant output carrier frequency regardless of the sideband being used. This corrects for the carrier oscillator shift necessary when shifting modes to allow the proper portion of the signal to fall within the pass band of the crystal filter.

The output of the first mixer varies in frequency from 8.395 to 8.895 MHz. This signal is sent through a band-pass filter to eliminate harmonics or other spurious signals and is then converted by the second mixer to the final operating frequency. This second mixer is crystal controlled by the same heterodyne

oscillator that is used for receiving.

A 6CL6 driver delivers a signal with sufficient power to drive a pair of 6146's in the final to a maximum dc power input of 180 watts PEP or 170 watts CW (50% duty factor). The output of my transmitter varied from about 110 watts on 75 meters to approximately 80 watts on 10 meters when measured with a Waters reflectometer and a 50-ohm load.

The manufacturer is to be commended on the choice of a transmitting-type tube in the final rather than one of the popular TV sweep tubes. I am sure that the signal is much better for it*, but the temptation must have been great when you consider the price tag on this kit.

Other circuit features which should be noted include a CW sidetone generator. This is used to create an audible monitoring signal when sending CW. It is **not** used to generate the CW signal itself. The VOX circuit is conventional, taking either the output of the

* W. I. Orr, W6SAI, "Full-Blast Operation of TV Sweep Tubes in Linear Service," *ham radio*, April, 1968, p. 9.

speech amplifier or the CW tone oscillator, as the case may be, and using it to switch from receive to transmit.

harmonic drive mechanism

Another interesting area is the main tuning dial. This dial is a unique assembly using the Harmonic Drive™ principle developed by the United Shoe Machinery Corporation. To the best of my knowledge, this is its first application in a piece of amateur equipment, although mechanical assemblies designed around the Harmonic Drive principle have been used by both industry and the military for a number of years.

The principle of this unit is best illustrated in **fig. 2**. There are three basic parts to this drive. One is a fixed spline which does not rotate (shown in white). Next is a flexible concentric spline (shown in black) which has two more teeth than the first spline and is sufficiently larger in diameter to clear the internal spline. The variable capacitor shaft is attached to the outer spline.

The tuning knob fits over these splines. The bore of this knob has two flats (shown in gray) which force engagement of the two splines at two points 180° apart. As the knob is turned, the point of engagement of the splines will also see the same angular displacement. Since there is an unequal number of teeth on the two splines, relative motion will be created between them. Because the inner spline is fixed, the outer spline will turn in the same direction as the tuning knob. The reduction ratio of the tuning knob to the outer spline is the number of the teeth on the outer spline to the difference in number of teeth on the two splines. In this case there are 82 teeth on the outer spline and 80 on the inner. Thus, the reduction in the Harmonic Drive assembly is 82/2 or 41:1. Add this to a further gear reduction at the variable capacitor in the VFO and you end up with a tuning rate of 18 kHz per turn, a very pleasant dial indeed.

When this dial is first assembled, it may seem unsatisfactory. The "stiction" or break-away force required each time you start to turn it may be annoying. **Don't worry!** With proper lubrication as outlined in the directions, and a bit of use, this dial rapidly be-

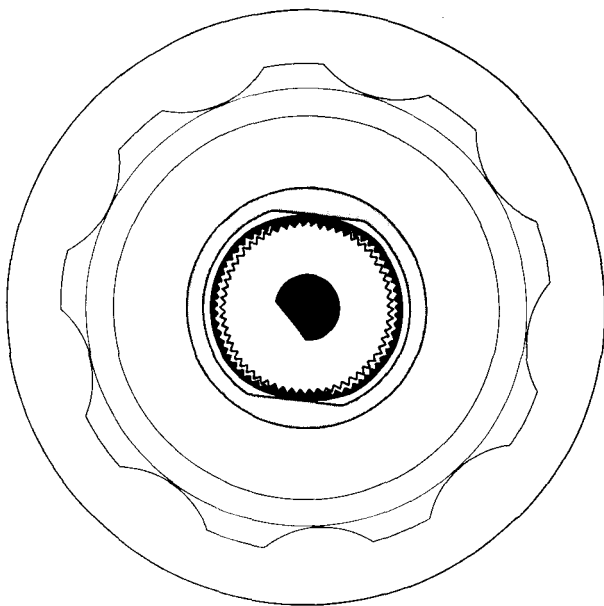
comes a real beauty. I have been well pleased with mine.

operation

In actual use, I was quite pleased with the VOX operation of this transceiver, and I happen to be one of those not usually given to VOX operation. The semi-break-in operation on CW is also quite satisfactory.

In summary, I can say that the HW-100

fig. 2. Internal construction of the Harmonic-Drive dial used on the Heath HW-100.



has given a fine account of itself on the air. Critical comments regarding audio quality have been excellent. Sensitivity of the receiver has been very surprising. It has been placed on the bench beside a very good receiver costing several times as much, and it has held its own with no trouble at all. The ease of tune-up and handling of this transceiver in all modes of operation are a pleasure. Whether you are considering your first plunge into sideband or are looking for a versatile transceiver for both home and mobile operations, you should be well pleased with the Heath HW-100.

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propagation

predictions for september

September is a month
of large seasonal change
in ionospheric behavior;
Perhaps now is the time
to consider changing
your operating schedules

During the past few months, amateurs in the Northern Hemisphere have enjoyed excellent twenty-meter propagation during hours of darkness, and fifteen meters has frequently remained open into the middle of the night. Ten meters has opened late, or not at all, and primarily to the south.

During September, all this will change. Fifteen and ten meters will open and close earlier and be open to higher latitudes. This is the month when median ionospheric predictions are least representative of conditions on any particular day during the month. The reason for this change is that the apparent solar latitude has dropped from 8° North to 3° South during the month of September. The resulting change in ionospheric propagation conditions has caught many an unwary amateur with his schedules down.

predicting the muf

Propagation conditions during September 1968 will be very similar to those that occurred during September 1967. The observed smoothed sunspot number for September 1967 was 96; the predicted smoothed sunspot number for September 1968 is 103. As a result, the predicted MUF's for September 1968 are only about 5% greater than those of September 1967.

With this in mind, I scaled a number of vertical incidence ionograms taken at Point Arguello, California, (35.5° North latitude) during September 1967 to determine the MUF for the east-west path with a control point at 35.5° North latitude. The transformation from vertical to oblique incidence propagation is not without its problems, and the MUF's derived this way are invariably low. This is due to a number of causes. However,

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

the MUF error for east-west F2-layer propagation in temperate latitudes is usually less than 8%.

The particular scaling method I used yields MUF values that are somewhat higher than those scaled for a fixed transmission distance of 2500 miles (4000 km). This discrepancy arises from consideration of reflection from virtual heights (apparent heights of reflection) as great as 400 miles which result in single F2-layer hops up to 3300 miles.

This type of propagation is not usually considered for commercial communication circuits, but amateur radio communications can stand the spreading loss that occurs with upper rays. The point is that the scaled MUF's come as close to those useful for amateur work as you'll see except from oblique-incidence ionogram data. Oblique-incidence data is not as available nor as well understood as vertical-incidence ionogram data.

The change in MUF with time of day that occurs during the month is illustrated in fig. 1. This chart shows the scaled values of MUF's for September 1 and September 30, 1968. The predicted MUF values for Point Arguello, as corrected by the ITS semimonthly revision factors (1.05 for the second half of September), are shown by the open circles.

Note the following characteristics which mark the change of MUF behavior throughout the month:

1. Night-time MUF's decrease and 14 MHz closes some nights.

2. MUF's rise faster during the morning and fall faster during the evening as the month progresses.

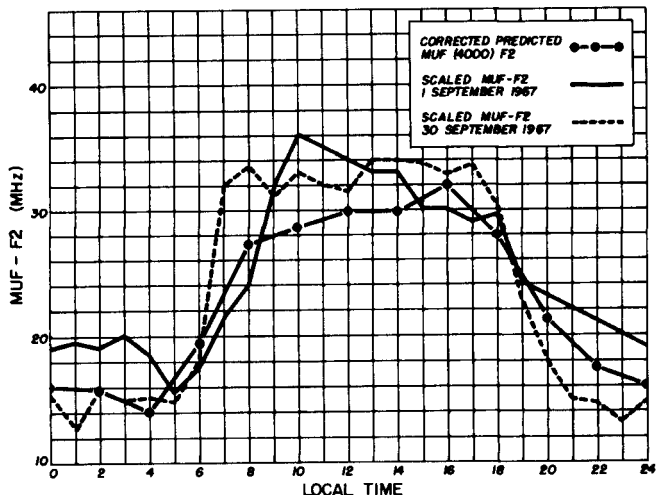
Note particularly that for a path with a control point at 35.5° North latitude at the beginning of the month, 21 MHz is open until 10 PM (at the control point) and 14 MHz remains open throughout the night. At the end of the month, the 21 MHz band closes by 7 PM, and 14 MHz closes occasionally during the night.

You can see just how much day-to-day MUF variation may be expected by looking at fig. 2. This chart shows the scaled MUF during the last ten days of September 1967. On at least five days, the highest scaled MUF during the day was greater than 40 MHz (25% higher than the highest predicted MUF). The peak MUF occurred between 2 PM and 5 PM local time.

September 20, 21, 28, 29 and 30 were disturbed days. The disturbances lowered maximum daytime MUF's until a few days after the disturbance, and raised minimum nighttime MUF's on the first night after the disturbance. Sporadic-E propagation was very evident on the 21st and 22nd and also occurred on the 30th.

The scaled MUF's are for ordinary waves. The extraordinary-wave MUF is somewhat higher, and depends on the orientation of the path with respect to the earth's magnetic field. It is about 200 kHz higher for a long east-west path in temperate latitudes and as

fig. 1. Maximum usable frequencies (F2 layer) scaled from vertical incidence ionograms taken at Point Arguello, California (latitude 35.5° N) on September 30, 1967 as compared with corrected predictions (see text).



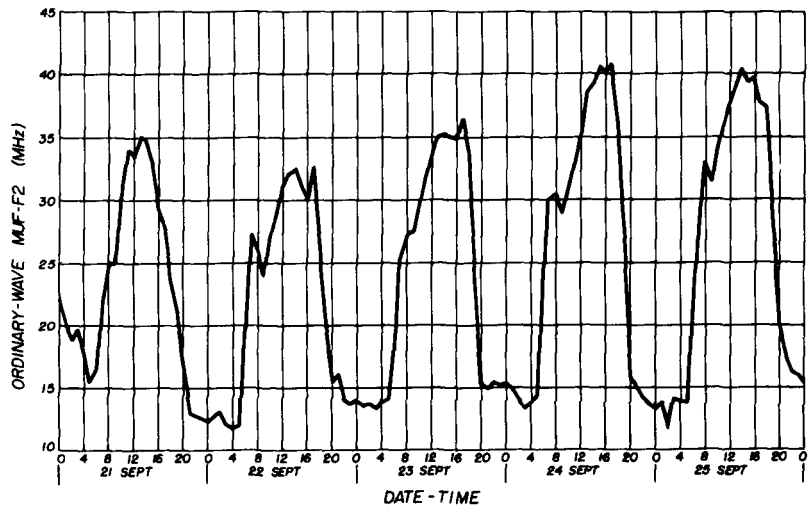
much as 1400 kHz higher for a north-south path at low latitudes.

Within 2000 miles of the magnetic equator (between 8° and 14° South latitude over South America), transequatorial forward scatter (TE) may extend the operational MUF to almost twice that obtained by ordinary re-

fect 50 MHz to open for a control-point latitude of 30° N at least once during the month, to 25° N at least five days during the month, and to 20° N at least ten days during the month.

However, just because the MUF is high

fig. 2. Maximum usable frequencies (F2 layer) scaled from vertical incidence ionograms taken at Point Arguello, California between September 1 and September 30, 1967.



fraction during the evening hours near the equinoxes. During disturbed conditions (usually early in the disturbance), TE may be worked at higher latitudes than normal. It is not known whether this extension is due to sporadic-E or a distortion of the normal F2-layer.

It appears that there are a number of forms of TE. One form has propagated signals at frequencies as high as 100 MHz between Hawaii and Raratonga. But I digress; a close look at the trends in **fig. 2** leads to a possible prediction means for propagation by ordinary refraction during stable conditions: if the nighttime MUF is higher than the night before, then tomorrow's daytime MUF will be higher than today's. On the other hand, nights with exceptionally high MUF's (during disturbances) are not usually followed by days with exceptionally high MUF's.

If the predicted variation of MUF with latitude holds (MUF's increase proportionately), then you could expect the daytime MUF's to be 13% higher at 30° N, 30% higher at 25° N, and 43% higher at 20° N than those at 35.5° N. In addition, if the MUF's are 5% higher this year than last, then you could ex-

pect 50 MHz to open for a control-point latitude of 30° N at least once during the month, to 25° N at least five days during the month, and to 20° N at least ten days during the month.

enough doesn't insure communications (even by backscatter), since 50 MHz activity is low in the proper places. By the time you receive this, it's almost too late to send equipment or line up schedules with amateurs in Costa Rica, Easter Island, Pitcairn, Tahiti or New Zealand. You may be able to work these places plus parts of South America if someone there is active on 50 MHz.

maximum usable frequency

It appears that paths with predicted MUF's in excess of 36 MHz could stand serious watching for occasional 50 MHz openings; paths with predicted MUF's in excess of 42 MHz won't need watching. So much for 50 MHz DX possibilities.

A time chart of the median MUF derived from September 1968 ITS predictions for 105° West longitude is shown in **fig. 3**. The MUF's derived from this time chart are believed to be accurate within 10% for the continental United States.

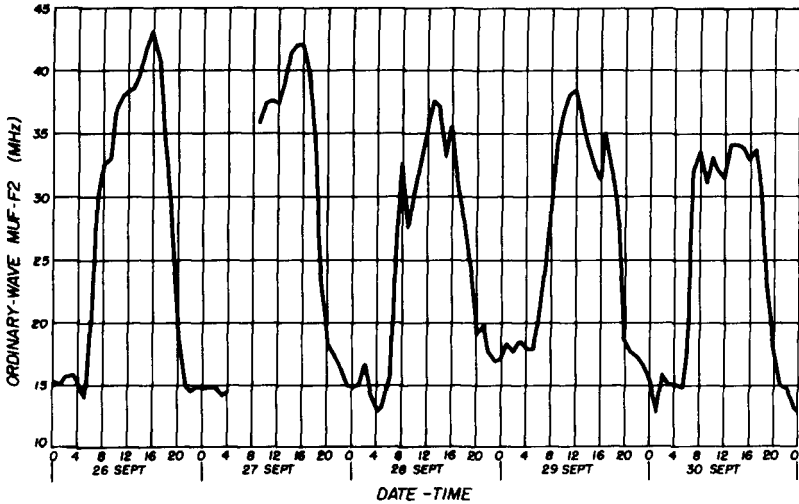
The time and latitude are those of your control point, 1250 miles away from your station in the direction of propagation. If only one control point is considered, all that is

guaranteed is that a frequency below the MUF will propagate to a distance of 2500 miles during half the days of the month. A signal at this frequency may, and probably will, propagate much further.

The time chart can be treated as a contour

maximum range

Time charts of maximum range determined by absorption and atmospheric noise levels are shown in **fig. 4** to **fig. 8**. These charts are based on an output power of 100 watts CW or 800 watts ssb with combined receiving

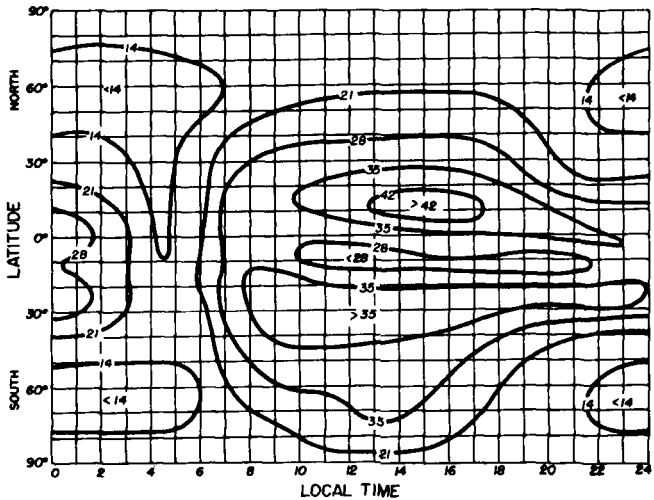


map of MUF over a limited geographical area. If a possible MUF error of 20-30% is not serious, the time chart could be used worldwide. However, much more accurate worldwide predictions can be obtained from the series of ITS MUF contour maps in **Ionospheric Predictions**.

and transmitting antenna gains compared to an isotropic radiator of -12 dB for 80 meters, 0 dB for 40 meters and 12 dB for 20 meters.

Transmission losses because of ground reflection on multi-hop paths are neglected as is any focussing gain. Reflection losses from poor grounds may be as great as 6 dB per

fig. 3. Time chart of predicted median MUF is for September 1968 for the American continent.



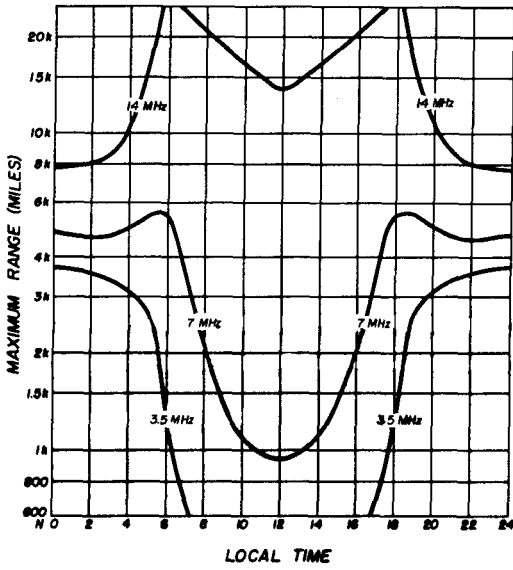


fig. 4. Maximum range due to absorption and noise vs local time from 38° N latitude to the north.

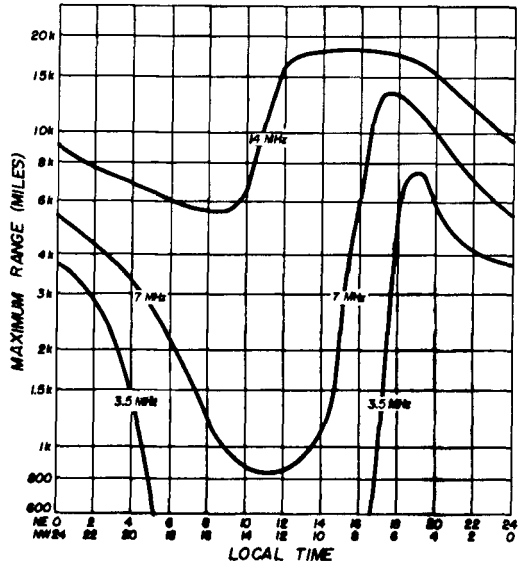


fig. 5. Maximum range due to absorption and noise vs local time from 38° N latitude to the northeast (top time scale) and to the northwest (bottom time scale).

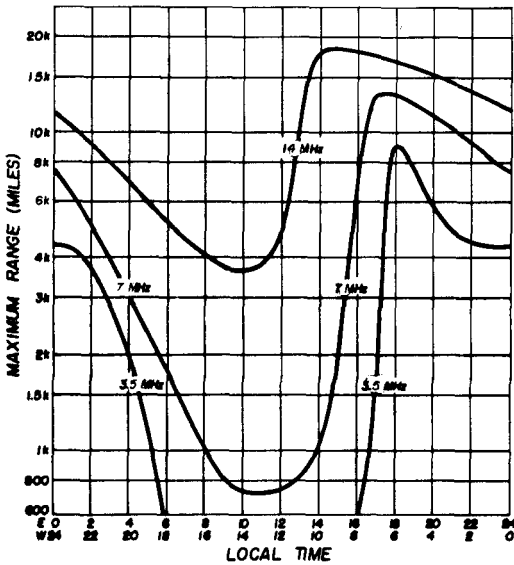


fig. 6. Maximum range due to absorption and noise vs local time from 38° N latitude to the east (top time scale) and to the west (lower time scale).

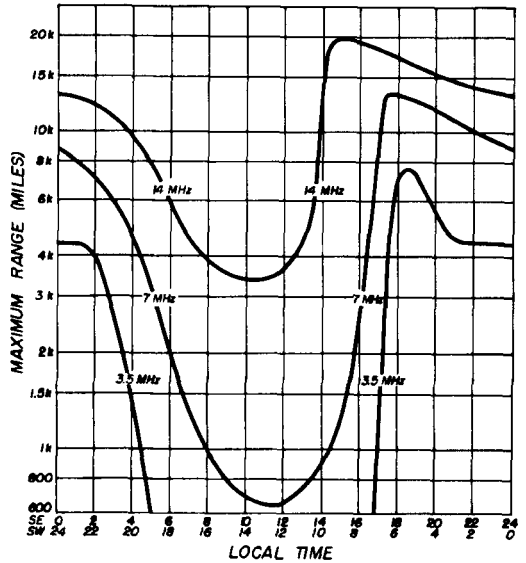


fig. 7. Maximum range due to absorption and noise vs local time from 38° N latitude to the southeast (top time scale) and to the southwest (bottom time scale).

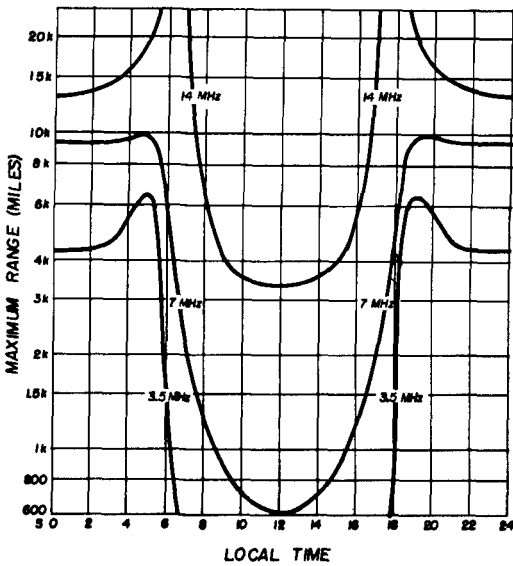


fig. 8. Maximum range due to absorption and noise vs local time from 38° latitude to the south.

reflection; losses from reflection from sea water are generally under 1 dB. Therefore, the maximum range on multi-hop paths completely over land will be less than that predicted.

Atmospheric noise varies considerably with location and time of day. The atmospheric noise levels used for purposes of these predictions are median noise levels over the continental United States for the autumn season. The variation of noise with time of day has been converted from six time blocks to an even function of time.

The peaks in the curve of 80-meter maximum range in fig. 5, 6, 7 and 8 are caused by the assumption that the noise level increases smoothly through the night until midnight instead of being constant from 6 PM to 10 PM and 2 AM to 6 AM. A 10-dB increase in system parameters on 80 meters would increase the nighttime range to the twilight zone. A decrease of system parameters of 10 dB on 80 meters would restrict the night-time range to 2000 miles.

propagation summary for september

80 and 40 meters. On these two bands, de-

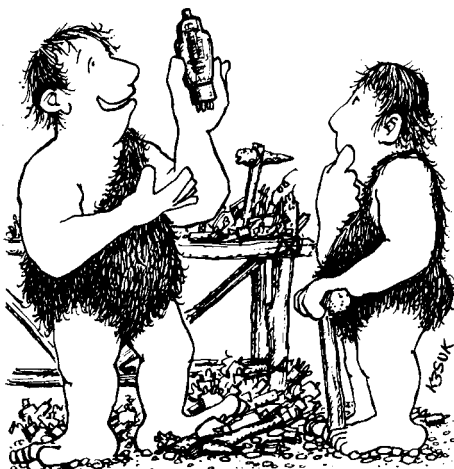
creasing noise levels and increasing hours of darkness will improve DX conditions in the Northern Hemisphere.

20 meters. Twenty will be open to somewhere most of the time, but mostly to the Southern Hemisphere during predawn hours. Polar paths will be more favored during daylight hours.

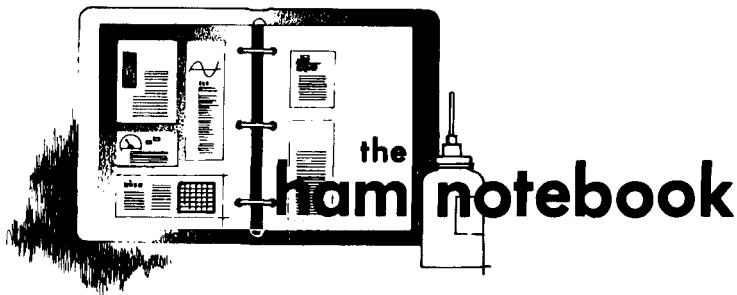
15 meters. Fifteen will open and close earlier as the month progresses. Polar paths should be workable from most of the United States during hours of daylight (at both ends of the path) most of the days of the month.

10 meters. Ten will open to the Southern Hemisphere and between East and West Coasts of the United States most of the days of the month. Openings between the East Coast and Europe and between the West Coast and Japan may be expected to increase in frequency as the month progresses.

ham radio



"After 806 failures—success!"



using industrial cartridge fuses

The cartridge fuses used by industry in electric-motor controls for machine tools and the like can be put to good use by the radio amateur. These fuses, since they are designed to handle the 440- and 550-Vac voltages used by industry, are ruggedly built. A typical fuse is about an inch in diameter by six inches long. Its body is made from either ceramic material or phenolic. The brass contacts, one at each end, will easily "take" solder.

These fuses can easily be made into coil forms for VFO's or small transmitters. With a little ingenuity on the part of the builder, a whole set of mobile loading coils could be fabricated from a few fuses. Finally, a length of resistance wire wound around the fuse body will make an inexpensive voltage-dropping resistor.

When using these fuses, two precautions are in order; make sure that the fuse body can handle the heat generated by a dropping resistor or transmitter coil. Secondly, make sure the fuse is blown—otherwise it will present a direct short circuit. A good source for blown fuses of this type is a small machine shop or industrial electrician.

D. E. Hausman, VE3BUUE

hook, line 'n sinker

Two items for the tool box that are valuable in the shack and out in the field are often overlooked—fishing line and sinkers. Not the kind used for sunfish, but heavy-duty nylon cord and some four- or six-ounce salt-water sinkers. These two low-cost items provide a way of getting a line across a tree limb to support a dipole. They may also be used to transport tools to the top of a tower or mast, or even up to the roof.

When they are used to haul tools up and down, a large heavy-duty battery clip tied on the end of the line in place of the sinker provides a quick means of attaching tools without the bother and risk of untying them with one hand at the top of the tower. I wish I could suggest a simple method for lifting a beam, but not every ham has a helicopter in his junk-box!

George Haymans, WA4NED

simple solder dispenser

Here's a neat solder dispenser some hams may not know about. It's made by simply punching a small hole in the side of the box that a one-pound roll of solder comes in. Put some tape over the hole for reinforcement. If the solder is 16 gauge (.062") or larger, the assembly can be used as a third hand when soldering.

Tony Felese, W2KID

makeshift test equipment

Since there are few tools and no test instruments available at sea, I've often used a long-wire antenna as a signal generator. It works fine, even in the audio stages of a receiver. Ashore, an 80-meter (or shorter) dipole will work due to the multitude of stations.

If you have access to another receiver in good operating condition, it can often be used as a signal tracer if you turn the audio gain up—just clip a lead to the grid of the detector stage or to the rf end of the detector

diode.

A receiver tuned to WWV or a multiplex signal, or beat against an internal crystal calibrator, makes a fine audio generator; just put a couple of clip leads across the speaker terminals. If you must know what audio frequency you're using, invest a dollar in a harmonica and mark the standard musical pitches on the blow holes with a sharp scribe.

Keith Olson, W7FS

vhf antenna switching without relays

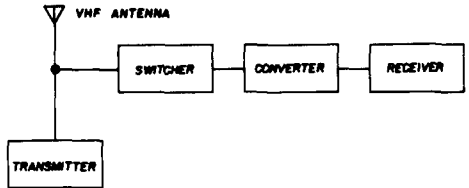
Fig. 1 shows an unusual circuit which—without switches or relays—permits a vhf receiver/converter combination to be permanently connected to a transmitting antenna without damage to the receiver when the transmitter is turned on.

The unit, simply called the electronic "switcher" around my shack, taps into the transmission line and allows low-level incoming signals to pass into the receiver. When the transmitter is operating, however, the high-level signal voltage on the line ac-

tivates the switcher and it blocks the path to the converter.

A few precautionary words however: the unit was designed specifically for 50-ohm coaxial feedlines and the power-handling capability drops off sharply as frequency increases. At 50 MHz, it will handle a peak of 500 watts; at 144 MHz, 350 watts peak; and,

fig. 2. Block diagram showing how the vhf switcher is installed.



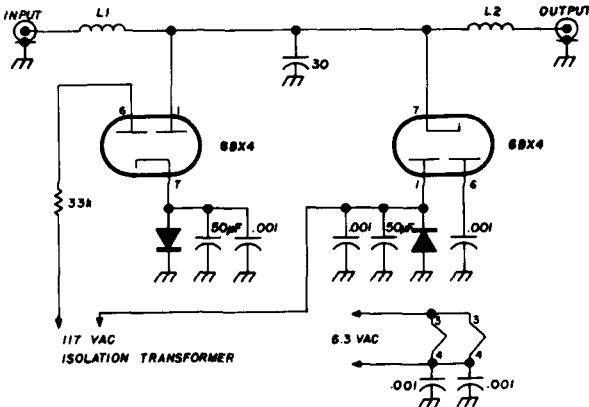
at 220 MHz, 125 watts.

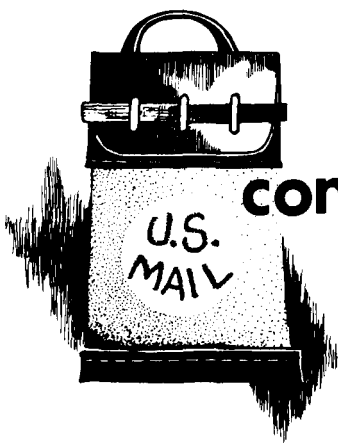
The advantages of such a system should be obvious to dyed-in-the-wool contesters who are frequently plagued with relay failures or too many mechanical switches.

Some trial-and-error may be required with coils L1 and L2 to obtain optimum performance and accurate low-level triggering. I found that for 220 MHz, both coils can be a 3/4-turn of number 18 wire, 3/4" diameter. At two meters, 1-3/4 turns of the same provided best results. For six meters, 6-1/2 turns of number 18 enameled wire on a 3/4" diameter form did the trick.

Bob Brown, K2ZSQ/W9HBF

fig. 1. The vhf switcher, a device that automatically isolates your converter from the antenna whenever the transmitter is turned on. Note that separate coils are required for each band you use. The 30-pF capacitor between the coils is a variable.





comments

Dear HR:

The reason that 146 and 440 MHz have developed considerable FM activity while 220 MHz has been bypassed is the very nature of the equipment available. I've been working my way through Syracuse University by servicing GE two-way FM gear. The high-band units are designed to cover the 450- to 470-MHz range. Movement into the ham bands is simply a matter of buying new crystals and retuning the rf stages in most cases, or at most, the addition of a small capacitance to a few tuned circuits in order to make them tune. No expensive rf modifications are required.

I recently attempted to put some gear on 220 for a RACES link. All rf and multiplier coils in the Pre-Progress receiver had to be rebuilt and the transmitter just wouldn't do. And this with a complete DATAFILE full of maintenance info at my fingertips backed up by a mountain of parts and commercial test gear! The frequency was just too great for the equipment and performance was seriously degraded when I finally got the receiver moved. Ten microvolts were required to open the squelch whereas an identical receiver on two meters has 0.5 μ V sensitivity for quieting.

Bill Santiff, WA2QKT

Dear HR:

Fantastic!!! Amateur Radio is alive and living in Greenville!

Ted Cohen, W9VZL/4

Dear HR:

I just received my first copy of **ham radio**, and read half of it before even opening the rest of my mail—it's great!

Of the fourteen articles on the "contents" page, I'm now researching or working on projects directly related to seven of them. The magazine couldn't be more timely.

Lawrence W. Banks, W1DYJ
Massachusetts Institute of Technology

Dear HR:

I had hoped that, since you took the word "ham" in the title of your magazine, your writing would set the communications world straight on the most probable and likely origin of the term. However, I was disappointed to see that your "second look" repeated the apocryphal tales broadcast by the general press.

I am enclosing photo copies of pages from "The Telegraph Instructor" that document the most likely origin of our beloved title. (**See text below. Ed.**) A letter from Paul Godley in the September 1965 **QST** is sufficient evidence alone that "ham" was a Morse man's term. . . . Somewhere in the past, I have heard a poor operator referred to as a "ham fisted bum."

Your magazine is great, keep it up.

Bob Jones, KH6AD

From pages 52 and 54 of "The Telegraph Instructor," by G. M. Dodge, Valparaiso, Indiana, copyright 1908:

Ham—A telegraph operator who is not proficient.

The "Correspondence From Members" section of QST for September 1965 sheds more light on the subject. W2PXR reports that H. C. Gawler, a radio inspector in Boston in 1912, was completely unaware of three young operators whose initials would coin the word "ham." Mr. Gawler also confirmed that in those days a ham operator usually referred to a railroad telegrapher whose code speed was approximately 10 words per minute and no more.

In the same issue of QST, Paul Godley, Ex-2EZ, wrote that in 1907 he had asked an old-time wireless operator with whom he was working why a particular operator was called a ham. The old timer replied, "He's got a ham for a 'fist'!"

Editor

Dear HR:

Back in 1910 around Boston we kids used to work each other at 5 wpm with our spark coils and transformer. I had a United Wireless coffin, a 20-wire aerial and ten-cent key, e.g., two dimes for contacts. NAD and BH called us "hams" and told us to get off their wave lengths. The English operators on the White Star and Cunarders never called us "hams." I used a silicon detector in 1910 and am still using one in 1968! What's new?

Bill Dickson, K4BQ
1914 call—BK



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dictionary of electronic terms

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This new dictionary, edited by Robert E. Beam, Ph.D., professor of Electrical Engineering at Northwestern University, includes definitions of over 4800 terms and hundreds of illustrations. The appendix provides data on schematic symbols, EIA color codes, abbreviations and letter symbols for electronic terms, Ohm's law formulas and Greek letters.

This book, priced at \$1.00, provides a low-priced, authoritative reference for everyone in electronics, including amateurs, engineers, technicians, students and experimenters. It is also an excellent electronics spelling guide. \$1.00 postpaid from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680.

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high power with new npn transistors

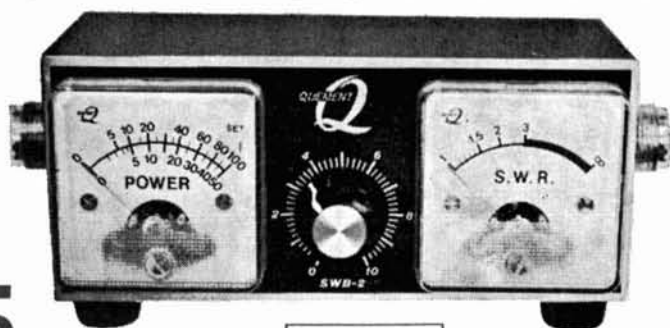
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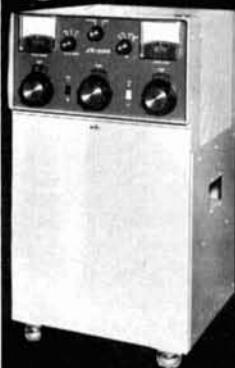
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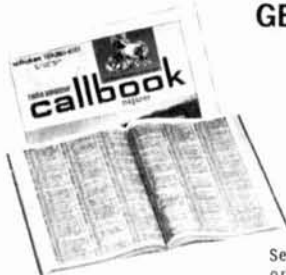


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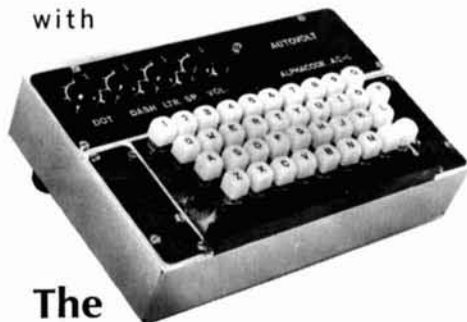
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FOR ASSISTANCE with your FM Problems particularly with Motorola equipment contact Neil McKie, WA6KLA, 5005 W. 21st Street, Los Angeles, California 90016

FOURTH MASSACHUSETTS QSO PARTY October 5 and 6. For details see QST, or write MIT Radio Society, W1MX, Box 558, 3 Ames St., Cambridge, Mass. 02139

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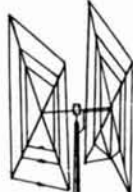
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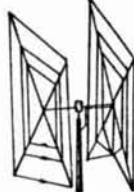
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400	<input type="checkbox"/> .16	<input type="checkbox"/> .40	<input type="checkbox"/> .50	<input type="checkbox"/> 1.50
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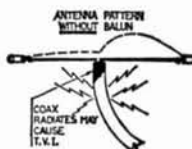
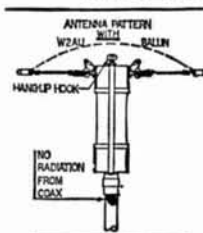
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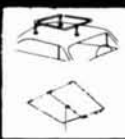
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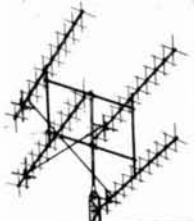
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WHEN YOU WERE A KID —

Did you ever plant a seed and then go out the next morning to find out what happened? Remember how pleased you were when the first tuft of dirt was pushed aside by the germinating sprout. My dad thought me impatient with all my questions, but Mom said that my garden would teach me many things. Certain it was that I learned not to expect too much too quickly.

In later life this lesson served me well, particularly in ham radio, where improvements were especially slow and hard to measure. Now as we approach the season when most of us are disposed to antenna work, I should like to explain that one device is inexpensively available that will easily permit an immediate improvement in your station's performance. I am referring to the balun — to be inserted as a replacement for the center insulator on your coaxially-fed dipoles.

There is nothing you can buy that will obtain as much relative improvement in your overall station capability as a proper balun. The 2AU balun at \$12.95 is the most popular.

Most rigs today have unbalanced outputs; that is, they have a pi network as the plate tank, and this gear is designed to work into loads between 25 and 100 ohms — **unbalanced loads.**

In my preceding ads I have repeatedly stressed the need for low VSWR. High reflected power causes more **anguish** and **frustration** than anything else. By balancing your antenna system, you may lower your VSWR, but more important is the fact that with the **balun**, your radiation efficiency is often higher. Normally, the grounded chassis side of the coax connects to one side of the dipole, and this side has very little induced electro-magnetic radiation from it (the dipole isn't a particularly good

transformer). Thus a balun, and especially a well-designed product like the 2AU, accepts the unbalanced input from the coax and permits equal distribution, and consequently equal electromagnetic radiation, from both halves of your antenna.

Any improvement from the use of a balun will give you more reach-out ability on receive as well as transmit.

One balun may be used with the same coax to feed two or more dipoles. All you require is a resonant length for each band you operate. You can't expect the balun, which in itself is a broad band device, to work properly when, for example, you feed 40 meter energy to it with just 80 meter resonators connected. In other words you actually need a separate set of driven elements for each band you operate — though they can all be connected to the same balun and feed.

The 2AU balun pictured here is its own lightning arrester and center insulator as well. It is furnished with a metallic eye that will permit you to support the entire center of your antenna from your tower or tree. It's side connections are good for a 600-pound pull. Condensation drips out from small holes in its bottom. It is rated for a full 1 KW. The 2AU balun may reduce TVI, improve antenna system efficiency, and protect against lightning, all at the same time.

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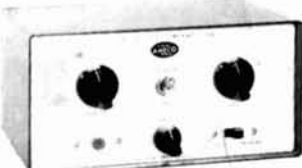
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