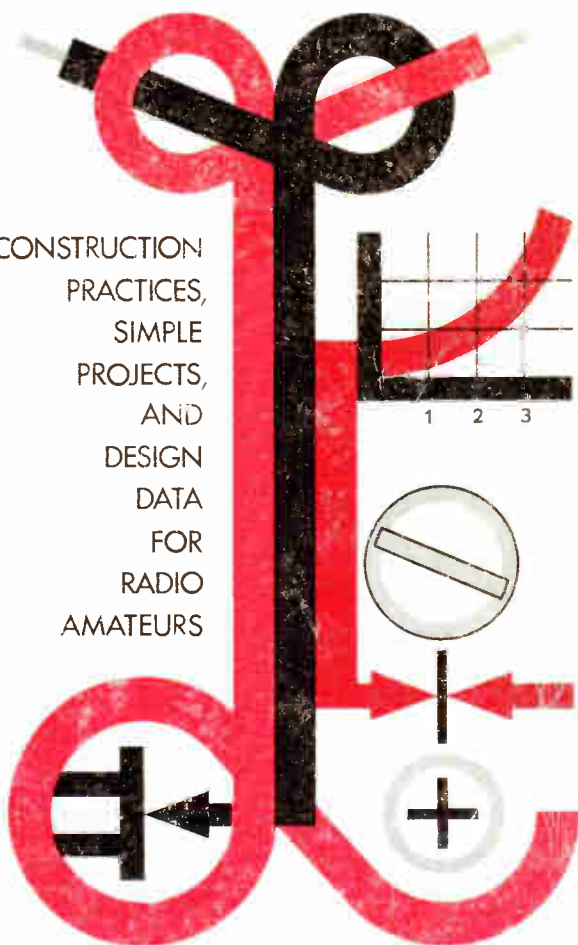


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CONSTRUCTION
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DATA
FOR
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AMATEURS



ham notebook

edited by

James R. Fisk, W1DTY

**Communications Technology
Greenville, New Hampshire 03048**

FIRST EDITION

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foreword

The amateur spirit is alive and well in the *ham notebook* column appearing monthly in *ham radio magazine*. Month after month, the contributions to this popular column reflect the qualities of the true amateur — inquisitiveness, persistence, improvisation, imagination and an open mind. *Ham notebook* features are generally not the products of textbooks and slide rules, but rather, the products of experimentation, experience and an almost intuitive feel for the art and science of amateur radio.

The amateur spirit has always been characterized by friendliness, helpfulness and an eagerness to share one's knowledge, tricks and pet circuits with others. *Ham notebook* has become a small part of this tradition, standing with the friendly ragchews on 75-meter sideband or six-meter a-m, or with the all-night basement sessions of the dedicated uhf enthusiast.

Here are hundreds of practical, bench-tested kinks and ideas gathered from the stations and workshops of hundreds of amateurs — just like you. I am sure you will find this book will quickly become one of the most used volumes in your workshop.

James R. Fisk, W1DTY

antennas and transmission lines

tiltover tower base

When you need a tiltover tower base, do you go out with a pocket full of loot and buy one? Most people do, and come back with a mighty thin wallet. How about using a little Yankee-Scotch ingenuity instead?

The length of the scrap piece of board shown in the photo depends upon the span of the tower legs. Hold it on the end of the tower and give it a good whack with a hammer. The marks left on the wood are used as a template by the pipe fitter when he makes up the U-shaped piece of plumbing that forms half of the hinge.

The coupling in the center of the U (A) is used to change the distance between the legs of the U to match the tower legs. The ends of the legs and the two uprights (B and C) are flattened and drilled for 3/8-inch bolts. The uprights fit into the legs of the tower and should be 11 or 12 inches long. They aren't bolted to the tower legs, just slipped in. This makes it easy to add another section of tower later on without a lot of work.

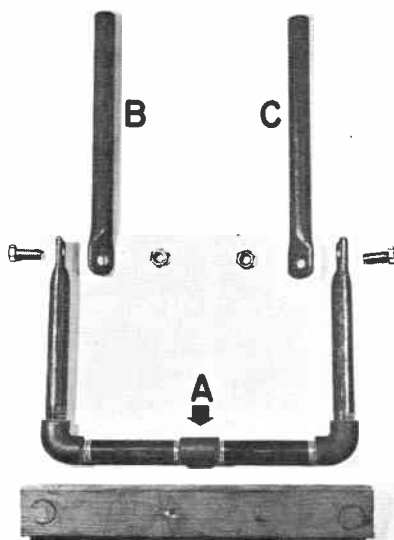
Dig a hole about two-feet deep and fill it with about sixteen inches of broken stone or gravel. Then bury the base of the hinge, the U-shaped piece, in a concrete slab six to eight inches thick. The larger the diameter of the slab, the greater stability you'll have. Don't forget the broken stone — this permits any water that collects to leach into the ground. Otherwise, water under the base may freeze and lift the concrete or crack it. To

make good cement, use three shovels of sand to one of cement.

Did you ever see the wind sweep a tower off the ground from the base? I never did either, so why go halfway to china with cement? You'd be better off to put all that labor into a good deep guy anchor or dead man that holds the top.

The total cost of the whole works is low — \$1.82, state tax included. Compare that to the price of a commercially-made unit!

Ted Woolner WA1APB



The tiltover lower base. The U-Shaped piece is buried in cement; the two uprights, B and C, are slipped into the tower legs. The hinge is formed by the 3/8-inch bolts.

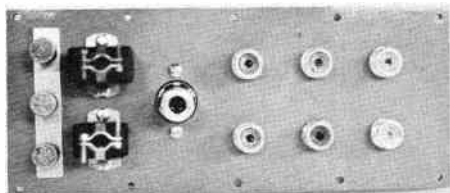
another use for coax relay coils

If you are using surplus coax relays with 24- or 28-volt dc coils, you might be able to use this idea. On a homebrew transmitter using push-pull 6AQ5s (the tubes could just as well have 6V6s), I found that the relay coil resistance was about the same resistance as the modulator cathode resistance should be. So I used the relay coil as the bias resistor. This worked okay, since the power supply B- was open on receive and grounded on transmit. Don't forget to bypass the coil with a 25-volt, 25- μ F capacitor.

Bil Eslick, KØVQY

coaxial feedthrough panel

Devising a simple and uncluttered method of bringing coax feedlines and rotor control cables into the shack is always a problem. After using a haywire lashup for years, I solved the problem neatly and efficiently with a small wall-mounted panel. Every wire except ac power is brought into the operating room through this panel.



The panel is made of 1/8-inch tempered masonite, 4 inches high by 9 1/2 inches wide. The coaxial feedthrough connectors are PL258 straight adapters (Amphenol 83-1J) mounted snugly in a hole made with a 5/8-inch punch. Epoxy cement is applied around the connectors on the rear of the panel and will hold them in place forever if the epoxy is properly and thoroughly mixed.

The two Jones plugs which show in the photograph are used for control lines

in antenna rotators; the small round plug is used for a remote swr meter. The metal strap is connected to a ground rod outside the house.

The panel may be painted with a spray can to match the decor or equipment. It is screwed to the wall over a 3 x 8 1/2-inch cutout in the wall near the radio desk; the coaxial cable and rotator control leads pass through the wall partitions to the outside. This panel is rather elaborate since it has six antenna lines and two rotor lines, but a smaller one for a simpler setup could be installed over a standard ac wall outlet box.

Edwin W. Hill, W3URE

using dipole antennas on non-harmonic frequencies

There are cases where it's desirable to use a dipole at frequencies on two bands which are not harmonically related. A good example is the phone man who wants to use a single antenna centered on 3.9 and 7.25 MHz. Since these frequencies are not harmonically related, the operator resorts to a compromise arrangement that leaves something to be desired on both frequencies.

Here's a system I use to obtain resonant operation with an off-center-fed Windom antenna. I cut the antenna as a full-wave dipole at 7.25 MHz as a starter; natural half-wave resonance is around 3.625 MHz. The resonant point is moved up to 3.9 MHz by the simple expedient of putting a fixed capacitor of the right value in the center. This is simple and can be used on *non-center-fed* dipoles operating on 10 and 15 meters or 15 and 20.

To compute what size capacitor you need in the center, consider the antenna as an isolated single-wire transmission line operating above a ground plane. Although the characteristic impedance of this line can be computed from the formula

$$Z_0 = 138 \log \frac{4h}{d}$$

where n is the height above ground and d is the diameter of the wire, an assumed value of 600 ohms is close enough for our purpose.

The difference between a quarter wavelength at 3.625 MHz and 3.9 MHz is 4.6 feet; at 3.9 MHz, this represents 6.55 degrees of 600-ohm line measured against a voltage node is $600 \tan 6.55^\circ$ or 69 ohms. Therefore, a 69-ohm capacitive reactance in series with each quarter wave will do the job; this is the same as 138 ohms in the center of the dipole. At 3.9 MHz, this is 290 pF; either 270 or 300 pF are close enough. A 1000-volt mica capacitor will be more than adequate for a full kilowatt.

To lower the frequency, the same approach may be used, but an inductance must be used as the reactive element.

Henry Keen, W2CTK

tips for keeping your beam

Amateurs driving past my house stop to look at the beam—it's black! The beam is wrapped with Scotch Electrical tape to protect it from the salt air blowing in from the ocean. The tape apparently has no adverse effects on beam operation; after years of use, the aluminum still remains bright. It takes about nine rolls of tape to wrap a 15-meter, four-element Cush-Craft beam, and this includes wrapping the boom and supports.

Down through the years, I have tried various methods to protect beams. Each new preparation was gleefully smeared on with high expectations and bad results. First came formvar varnish, boat paints, chrome varnish, Rusto, fibreglass, epoxy and metal paint. You name it and I've tried it. Most of the concoctions eventually craze and flake off due to the sun's rays.

Aluminum elements seems to have special problems all of their own—a white powder caused by electrolysis pushes up under the applications and the coating comes off. Even fiber glass and epoxy crack when moisture gets into the

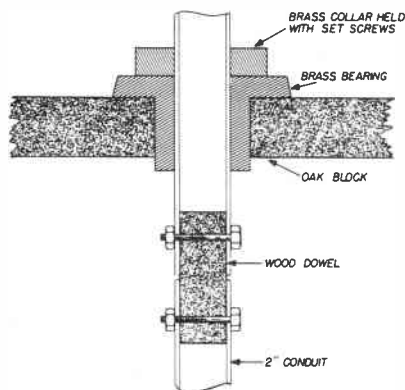


fig. 1. Simple antenna bearing mount is easily installed on top of a triangular tower.

cracks. After awhile, the joints of the tubing develop insulation properties and the beam has to be taken apart. These troubles prompted the experiment with the Scotch Electrical tape. The U-bolts and hardware were painted with *Derusto* paint where they could not be wrapped. However, with care, the nuts and bolts can be wrapped and smeared with RTV-102 GE Silicon Rubber.

My mania for protection went as far as wrapping the tower with electrical tape; at least, that portion which sticks up above the garage and catches the salt breeze. Many of the rungs on my first tower broke from the twisting caused by the beam blowing back and forth on the rotor on top of the tower. The next tower I put up had the motor mounted at the base, and a connecting piece of electrical conduit going up through the center. With this method of construction, the pipe twists and not the tower.

Sometimes the top of a triangular tower rusts out at the joints and the top section has to be removed. An antenna bearing mount can be made by using a triangular piece of two-inch thick oak and bolting it in place with lag screws. Boiling it in wax helps to weatherproof the block. While oak makes a good bearing, it is better if you can have a machine shop make a brass bearing and a collet to slip over the conduit as shown in fig. 1. All open ends of the pipe and aluminum should be sealed with corks.

Many fellows have used conduit clamps to join two pieces of conduit. These conduit couplers don't hold up, even if the nut is tightened and set with drift pins. It is better to insert a dowel in each piece of conduit with 1/4-inch bolts to hold the piece from twisting (see fig. 1).

Ed Marriner, W6BLZ

mono-loop antenna

Nothing new or startling is claimed for this antenna. However, like the open-wire-line fed flat top, the one-wavelength loop deserves a great deal more attention than it's received.

Almost everyone using a quad parasitic array is pleased with the results. The quad has a good capture area, and for modest

height above ground, a favorable low angle of radiation. A quad mono-loop has these same advantages, but with the disadvantage of being bidirectional.

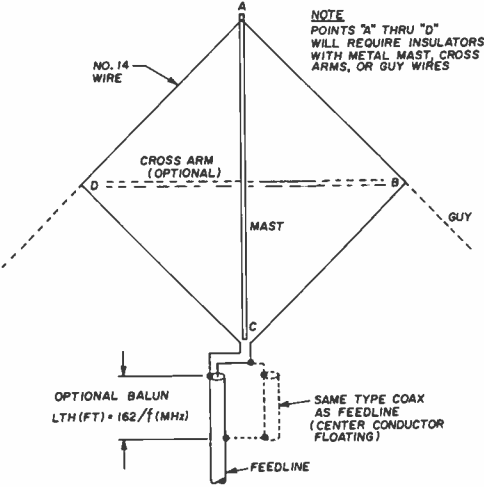
My new location precluded a tower and rotary beam. As an alternative I erected thirty feet of tv mast on the chimney of the house; and from the top, some forty feet above ground, suspended a 20-meter quad loop (fig. 2)*. Because of limitations in guying the sides at the proper height, the antenna looks like an ARRL diamond, but it works like a gem (pun intended)!

Reports broadside to the antenna are generally 6 to 12 dB higher than those received from a three-half-wave doublet (100 feet long, center fed; 40 feet high) favoring the same directions. In fact, except that the antenna is fixed, it performs just about as well as a two-element quad in a former location.

For a 20-meter loop, I suggest starting with sides 17 feet 10 inches long and trimming to optimize the swr. In my case, the regular formula length $250/f_{MHz} =$ side in feet resulted in a resonant frequency higher than desired; probably because the sides are elongated into a diamond shape.

Hams having towers high enough to accommodate a 40-meter loop might find it would outperform a conventional inverted-V doublet, even though the bottom current loop is close to the ground.

James A. Gundry, W8BW



f (MHz)	total length	length per side
14.0	71' 10"	17' 11 1/2"
14.1	71' 4"	17' 10"
14.2	70' 9"	17' 8 3/4"
14.3	70' 3"	17' 6 3/4"
21.0	47' 10"	11' 11 1/2"
21.1	47' 8"	11' 11"
21.2	47' 5"	11' 10 1/4"
21.3	47' 2"	11' 9 1/2"
28.0	35' 10"	8' 11 1/2"
28.5	35' 3"	8' 9 3/4"
29.0	34' 8"	8' 8"
29.5	34' 1"	8' 6 1/2"

fig. 2. The mono-loop antenna.

low-loss coax

When I apparently wasn't getting out of my own backyard with a new antenna installation on two meters I was a little suspicious of that "new manufacture" RG-11/U I was using. I used a 200-watt bulb to check out the coax — when connected directly to the transmitter output connector, it was lit to full brilliance; when the coax was put between the bulb

*Suggestions for crossarm braces and a balun are included. The balun is recommended for optimum efficiency. editor

and the transmitter, the bulb barely glowed.

The solution was a section of home-made low-loss coaxial line. A 20-foot length of ½-inch rigid copper pipe was used for the out conductor; a length of aluminum clothesline wire (about no. 9) serves as the center conductor; teflon washers spaced every 8 inches keep the center conductor centered in the pipe.

Impedance is about 72 ohms, and there is almost no loss. The assembly is strong enough to support small antennas (like a coaxial dipole) and is unaffected by rain; lateral support is provided by three nylon guys.

Gus Gercke, K6BIJ

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 saltwater 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

inverted-vee antennas

In my experience with an inverted-vee antenna for 40 meters, 67.5 feet long with the apex at 30 feet and the ends 15 feet off the ground, resonant frequency was 7.18 MHz. This suggests that inverted-vee length is:

$$\text{length (feet)} = \frac{485}{f\text{MHz}}$$

The resonant frequency was indicated by a very deep and sharp null with an Omega-T antenna noise bridge.

It may be of interest to comment on the affect of adding a parasitic inverted-vee director, 64.13-feet long (5% shorter than radiator), at the same apex and end heights, spaced 16 feet in front of the driven element. (Judging from Yagi data this is good spacing for gain, but not for bandwidth.) The director raised the resonant frequency 114 kHz to 7.294 MHz. This suggests:

$$\text{driven element (feet)} = \frac{492}{f\text{MHz}}$$

$$\text{director length (feet)} = \frac{467}{f\text{MHz}}$$

Element spacing was 0.11 wavelength.

The Omega-T noise bridge indicated a driving-point impedance of 27 ohms for this two-element inverted-vee Yagi. Two such antennas were used on 40-meter phone and cw during the 1950 ARRL Field Day with good success. A 38-ohm Q-section will match 27 ohms to 50-ohm coax, two 22.93-foot lengths of RG-59/U (length = 165/fMHz) were soldered in parallel to make a 37.5-ohm Q-section. According to the book, the Q-section should be made from twin lead, but with the double-coax Q-section swr was less than 1.4:1.

Another two-element inverted-vee Yagi with a director 3% shorter than the radiator resulted in the following formula:

$$\text{driven element (feet)} = \frac{490}{f\text{MHz}}$$

$$\text{director length (feet)} = \frac{467}{f\text{MHz}}$$

Element spacing for this array was also 0.11 wavelength.

Bob Hume, WB6AQF

ground rods

Driving of 6-foot ground rods is easy in good loam, but many of us live in areas

where the top soil is sandy or rocky, and you cannot find even a very poor ground with a 6- or 8-foot rod. It is necessary to get through hardpan or layers of rock to find moisture, to say nothing of water. I had to go down 18 feet.

Some small ground rods are made so they can be threaded together, but the driving cap furnished with them comes loose with each blow of the sledge and buggers the threads. Also, when threaded they unscrew underground.

The same is true of well hardware. Ordinary ground points tend to pack the soil around them, and thus insulate themselves from actual ground. Well points will do the same, but due to their greater diameter, they pack less. A 1½-inch well point can be driven into soft soil, but the threads get mashed even when using the fitting made for driving.

The easiest way in the long run is to get a 1½-inch well point and several sections of extension pipe. Drill a ¼-inch hole at the top of one pipe and hang two buckets of rocks from the top. I used 5-gallon paint cans. Fill the buckets with rocks and then fill the leftover space with water for weight.

Cut off the end of an old 3/8-inch hose and drop the end down into the well point. Use a pair of chain tongs and take about three or four turns in the assembly with the water running easy. Then take half a turn back; keep the water running.

The pipe should run down about an inch for each 4 turns. After running it into the ground as far as necessary to reach the ground-water level fill the pipe with copper sulfate and stick a funnel into the top. It will catch enough rain to melt about 5 pounds of copper sulfate a year. A 25-pound sack of copper sulfate costs about \$6.00 wholesale.

I finally ran my 18-foot ground in after only 10,000 turns or so!

Keith Olson, W7FS

simplified balun

An antenna with open ends is a balanced system. Feeding it directly with coax (unbalanced feeder) causes rf radia-

tion on the outside of the coax. The usual detuning sleeve (bazooka), a 1-to-1 transformer, requires a rather elaborate metal sleeve an electrical quarter-wavelength long. This balun has no effect on the impedance of the line or antenna. It only keeps radiation from coming back down the outer shield of the coax.

This sleeve can be considerably simplified by using kitchen aluminum foil wrapped with plastic electrical tape along its entire length. The sleeve is shorted to the coax shield at the end away from the antenna; the antenna end remains open.

Richard Mollentine, WAØKKC

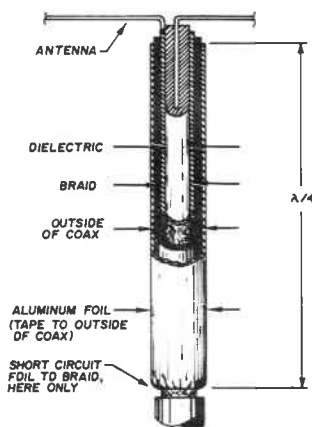
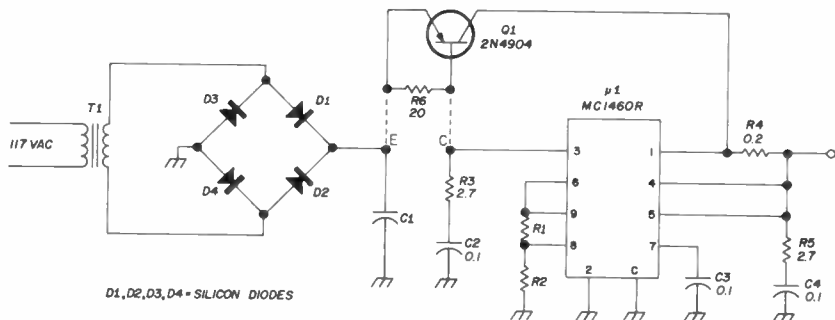


fig. 2. Simplified balun using aluminum foil.

coaxial cable connectors

Whenever I put a coaxial fitting on a piece of RG-48/U, I always seem to end up with a short circuit. At least one of the shield strands breaks, leaving a small piece to float around inside the connector until it shorts the whole thing out. A simple solution that works well on the high-frequency bands is shown in the photo.

All you have to do is saw two L-shaped slots in the reducing adapter (A in photo). Saw a slot down the center of the adapter almost to the threaded section with a hack saw; then saw or file a short notch at right angles to the main slot — do the same thing on both sides of



C1 filter capacitor (value depends on current drawn from supply)

R4 0.2 ohms (for use with Q1 at 2A maximum output)

R1,R2see text

T1 12 VAC,.45A (Stancor P-8392)

fig. 31. Schematic of the regulated supply. If an external pass transistor isn't used, terminals E and C are jumpered.

critical. You can obtain other fixed voltages by using a 400-mW zener diode for the desired voltage.

Don Nelson, WB2EGZ

a flexible voltage-regulated power supply

The power supply described here was originally designed for a 2-meter fm receiver. However, it can be used for any equipment requiring a regulated low-voltage power supply.

The integrated circuit is a Motorola MC1460R, which can handle up to 20 volts input at 600 mA maximum. For higher voltages, the MC1461R (35 V), or

the MC1561R (40 V) can be used. The case is a small diamond shape, which can be mounted in a heat sink.

the circuit

The power supply requires a minimum of parts (fig. 31). Resistors R1, R2 constitute a voltage divider that determines output voltage. With $R2 = 6.8k$, R1 will be $(2V_O - 7)k$ ohms, where V_O is the output voltage. R1 can be fixed or variable. Resistor R4 is the current-limiting resistor, which determines the short-circuit load current. If an external pass transistor is not used, values for R4 may be obtained from fig. 32 for various short-circuit load currents.

Since the IC transistors have high

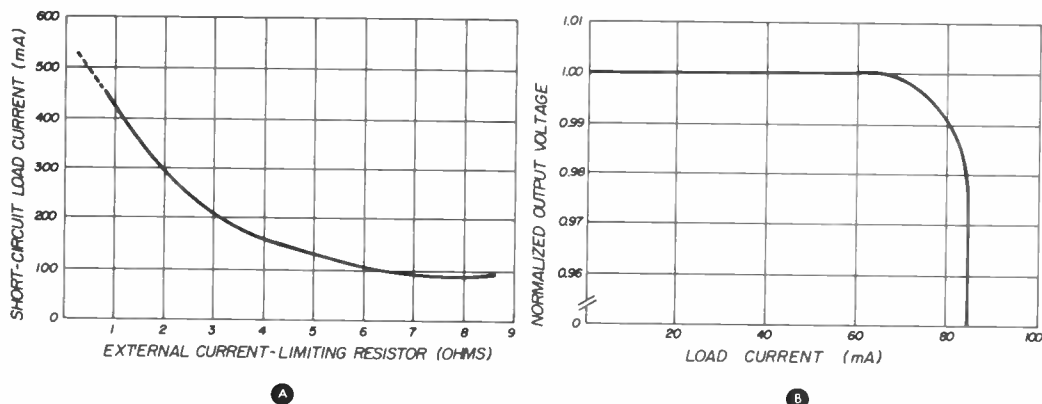


fig. 32. Curve A is used to determine the value of R4 if a pass transistor isn't used; curve B shows voltage and current when R4 is 6.8 ohms.

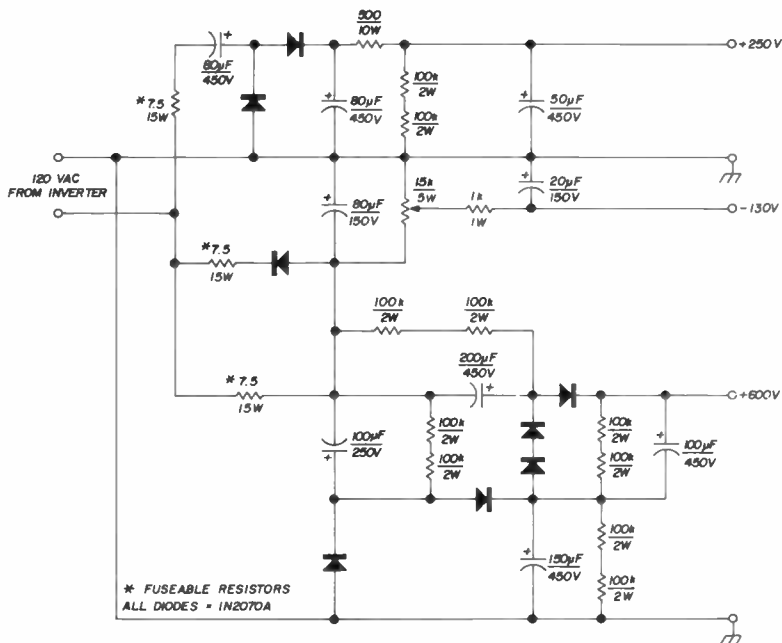


fig. 33. Low-cost mobile power supply uses mobile ac inverter.

frequency capabilities, there's a chance of oscillation with this device; therefore, some means must be used to suppress this tendency. Networks composed of R3, C2 and R5, C4 form suppressors for input and output respectively.

Because of the vhf transistors in the IC, care must be used in wiring the circuit. Hand wiring can be used, but it's not recommended unless vhf-type construction techniques are used. This means extremely short leads and vhf grounding methods.

The heat sink can be made from a piece of aluminum with a 1/2-inch hole to pass the IC leads. Make sure the leads don't touch the heat sink. The size of the heat sink will depend on the power to be dissipated.

Ron Vaceluke, W9SEK

low-cost mobile supply

Recently I wanted to try mobile operation with my Heath HW-32A, but I didn't want to invest in an expensive

mobile power supply. Since I had a dc-to-ac inverter on hand, I decided to build a transformerless voltage quadrupler and doubler circuit using the inverter as the ac source. The circuit is shown in fig. 33. Output voltages are +600 volts, +250 volts and -130 volts bias.

Henry Frink, W4GEG

rectifier terminal strip

Recently, while trying to find a suitable terminal strip for a bridge rectifier that used 8 diodes (2 in each leg), I decided to use an inverted octal socket. These sockets are fairly compact, and the diodes can be physically arranged similar to their schematic equivalent. This simplifies connections and eliminates costly wiring errors. One caution, however: do not mount the socket directly to the chassis; use a bracket or standoffs, since the contacts on some sockets can slide out the top of the socket and short the diodes to the chassis.

Robert G. Wheaton, W5PKK

receivers

frequency spotter

Many amateurs have a general-coverage receiver as well as an amateur-band-only receiver or transceiver. This is a desirable item, filling in those in-between band spots for casual listening or for checking for out-of-band radiation. A weak point for most such receivers is that they're usually of a less expensive type and thereby lack the precise frequency read-out you'd like to have.

This lack of precise frequency calibration complicates the use of the more common type of calibrator. A 100-kHz calibrator can cause more confusion than it can resolve when dial calibration on a receiver is such that you can't identify which of the marker signals represents the desired reference signal. In such instances a 1-MHz spotting signal is much more convenient to use. Most often, too, 1-MHz markers are sufficient for setting the bandset dial before using the calibrated bandspread dial.

There are many circuits for low-frequency crystals (100-200 kHz) and even more for the high-frequency spectrum. These circuits, unfortunately, don't always perform satisfactorily in the medium-frequency range.

This article describes a circuit especially adapted to a 1-MHz crystal. It oscillates dependably and produces profuse harmonics to the upper limit of the high-frequency band. Best of all, it can be built with junk-box parts with no requirement to purchase some hard-to-find component.

construction

As the schematic shows (fig. 1), this circuit borrows liberally from the Miller, Colpitts and Pierce. The coil in the collector-to-emitter circuit suggests the Miller; the use of a capacitive voltage divider in the collector-emitter-ground string is strongly reminiscent of the Colpitts; the crystal is placed between the collector and base, as in the Pierce. Which takes ascendancy, I don't know. But it oscillates with no hesitancy, and that's the reason why I selected it.

At this frequency, the arrangement of components isn't critical. I used a small piece of perf board and mounted parts with no thought to short leads. The only item I mounted with care was the inductor, which I placed away from and at right angles to the rf choke coil. The transistor was straight out of the junk box; before hitting the junk box, it lived on a printed-circuit board native to some unidentified equipment. The resistors and capacitors also came from various print-

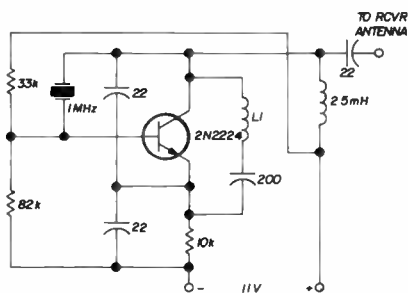


fig. 1. Schematic of the frequency spotter. A 1-MHz crystal works best in this application. L2 is 80 turns no. 32 scramble wound on a 1/2" form.

ed-circuit boards. The rf choke, I regret to say, seldom appears in present-day surplus offerings; I had to buy mine, a distinctly unpleasant transaction! The coil was a junk-box progeny, too. The ½-inch form came from an old tv set, and the no. 32 wire was salvaged from an ancient electrodynamic speaker field coil. I doubt that the value of any component is critical; most likely, any nearly alike substitute would function equally well.

The receiver in which I placed my calibrator is a Radio Shack offering, the DX-150. This receiver lends itself well to the addition of a frequency spotter. On its left-hand side (looking from the front of the receiver) there's ample room to mount a miniature toggle switch under the headphone jack. This switch must be mounted for a back-and-forth movement (not up and down) to have sufficient clearance. This coincides nicely with the movement of the slide switches.

Only the receiver's bottom plate needs to come off to permit mounting the switch and the oscillator board. There's plenty of room in the underside of the cabinet to mount the oscillator board. The chassis serves as the negative power lead, and the positive lead can be tapped off the bus at the main filter capacitor. As the set's voltage is a little higher than that demanded by the spotter, I used a 2000-ohm resistor in series with the positive power lead. This serves the dual purpose of dropping the voltage and insuring against zapping the receiver's power supply in case of a short circuit in the oscillator board.

Whatever receiver you use, you'll no doubt find a ready means of mounting your frequency spotter. If your receiver uses vacuum tubes, don't overlook the ready source of transistor power available from the cathode end of the cathode-bias resistor in the audio power amplifier stage. It's well filtered and usually just about the voltage you're looking for.

This is a pleasing little device. You'll enjoy building it, and you'll find it a much-used adjunct to your receiver.

Carl C. Drumeller, W5JJ

improved transceiver selectivity

The advantages of operating with transceivers are well established. However, most transceivers on the market lack adequate selectivity for CW reception. I've seen many circuits for narrowing receiver audio bandwidth, but they are either too complicated or entail too many modifications, which reduce the transceiver's trade-in value.

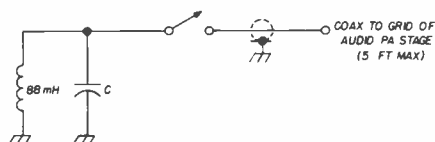


fig. 2. Simple filter for CW selectivity.

I use a parallel-resonant circuit between the audio-driver and power-amplifier tubes in my transceiver to improve selectivity. Don't be misled by the simplicity of the circuit (fig. 2), as it works very well. The toroid coil, which is available from surplus outlets, and capacitor C form a tuned circuit at the desired audio-tone frequency. A value of C that will give the desired audio frequency may be chosen from table 1.

Only two connections need be made at the transceiver — nothing to disconnect or modify. Simply install the tuned circuit between the audio power-amplifier tube grid and ground through a switch. The assembly may be remotely located or mounted at the operating position. The tuned circuit and switch may be mounted in a minibox, with a shielded cable to the transceiver. Thus no holes or other alterations are needed that will degrade the transceiver.

table 1. Capacitor values for desired tone

tone (Hz)	value of C (μF)
1700	0.1
1000	0.33
750	0.5
550	1.0

As with all passive networks, this filter causes some loss in audio power, especially as C is increased beyond 0.5 μ F. This means you'll have to crank up the volume control a bit higher than normal. This shouldn't pose a problem, because most transceivers have volume to spare.

With this simple filter you'll find that interference from other signals and man-made noise will almost disappear. CW signals a few hundred Hz apart can be separated easily, and the main tuning dial becomes very sensitive.

J. Donato, VE3BWD

communications speakers

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

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

how to do it

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

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

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

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

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

Bill Rogers, WA5RAQ

improved superregenerative receiver

The superregenerative receiver is a simple low-cost system often used in portable equipment operating on the amateur vhf and uhf bands. Although this circuit offers high sensitivity, it suffers from poor selectivity, high noise level, oscillator radiation and hangover. Hangover results in blocking that limits sensitivity because the receiver is swamped by its own residual signal.

The high sensitivity of the superregenerative detector is due to the use of

an alternating quench voltage, usually between 20 and 300 kHz. The regeneration control is set so the detector goes into oscillation on each positive peak of the quench voltage: on each negative swing the oscillator is cut off.

The superregeneration principle can be applied to any oscillator circuit; a grounded base Hartley circuit is shown in

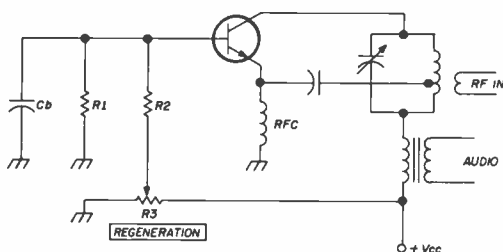


fig. 3. Simple superregenerative circuit using the grounded-base Hartley circuit.

fig. 3. If the bias is gradually increased with the *regeneration* control the circuit will break into oscillation. When the amplitude of oscillations overcomes the base bias voltage on the negative portion of the swing, rectified current through the base-emitter diode charges capacitor C_b , putting a negative bias on the base that runs off the transistor. When C_b has discharged through resistor R_b the circuit begins to oscillate again. Hence, the quenching frequency is determined by the base-bias voltage and the time constant of $C_b R_b$. In fig. 3 R_b is the effective resistance from base to ground and consists of R_1 in parallel with R_2 and part of R_3 .

An analysis of one cycle of quenching is shown in fig. 4. The oscillation is triggered by the incoming signal and builds until it overcomes the bias potential, then is quenched. When there is no input signal, internal circuit noise acts as the trigger with a resultant hiss in the audio output. Hangover is due to the fly-wheel effect of the tank circuit, and the higher the Q , the more troublesome the hangover.

K2ZSQ recently suggested a simple

circuit addition to limit the affects of radiation and hangover. The circuit change consists of adding a germanium diode across the tank circuit as shown in fig. 5. With this diode undesired energy is immediately dissipated after the oscillation burst. This eliminates hangover effects during the remaining period of the burst. Radiation is also lessened because the damping action of the diode lowers amplitude and shortens the duration of radiated pulse.

practical receiver

The modified superregenerative receiver shown in fig. 5 has been used on 50 and 144 MHz with similar results. The added diode is not critical, and any germanium point-contact seems to work well. The transistor should be a vhf type for best operation, but this isn't too critical either. I used Japanese types 2SC372 ($f_T = 150$ MHz) and 2SC387 ($f_T = 900$ MHz), but Motorola equivalent types HEP55 and HEP56 will perform as well.

Smoother operation is obtained by tapping the collector down on the final tank coil as shown in fig. 5. Oscillation can also be improved by moving the

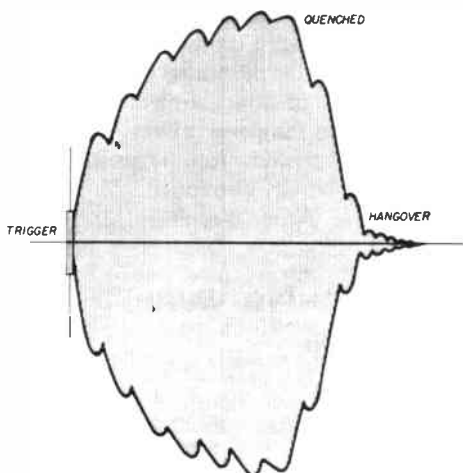
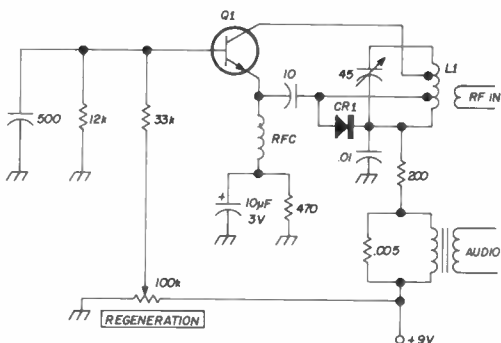


fig. 4. One cycle of superregenerative operation shows the input trigger, oscillation, quenching and hangover.



50 MHz

L1 7 turns no. 20, 9/16" diameter, 3/4" long. Collector tap at 2 1/2 turns, emitter tap 1/2 turn from ground

Q1 HEP 55

RFC 10 μ H

144 MHz

L1 4 turns no. 16, 3/8" diameter, 9/16" long. Collector tap at 1 1/2 turns, emitter tap at 1/2 turn from ground

Q1 HEP 56

RFC 20 turns no. 26, closewound on 1/4" form

fig. 5. Practical vhf receiver for 6 and 2 meters using the modified superregenerative circuit.

emitter tap down, but effective damping action requires the emitter to be tapped as high as possible on the coil.

Since this circuit is basically an oscillator, radiation was decreased very little by the addition of the germanium diode. However, the hangover effect was considerably improved. No noticeable decrease in sensitivity was found.

Atuyuki Iwakami, JA1BHG

inexpensive wwv converter

Here's a simple circuit that will convert WWV signals above 5 MHz to amateur bands between 80 and 15 meters. The converter consists of a simple crystal mixer using a single npn transistor (fig. 6).

A variety of crystal frequencies can be used, depending on which WWV signal is

table 2. Crystal frequencies that can be used to convert WWV to the various amateur bands.

ham band	WWV frequency (MHz)			
	10	15	20	25
80	6-6.5	—	—	—
40	2.7-3.0	8-7.70	—	—
20	4-4.35	—	6-6.65	—
15	—	6-6.45	—	3.55-4.00

desired. See table 2. The exact output frequency may be determined by finding the difference or sum of the desired WWV frequency and the crystal frequency.

No tuned circuits are used, so other signals will be received in addition to WWV. For example, if a 6.5-MHz crystal were used to convert WWV at 10 MHz to 3.5 MHz, all signals on 3.5, 3.0 and 10 MHz would be picked up. However, WWV is usually strong enough to be picked out easily. If necessary, tuned circuits could easily be installed between the antenna and the converter.

WWV transmits much interesting data in addition to time ticks. Schedules and types of information are published in the ARRL handbook.

Doug Pongrance, WA3JBN

general-coverage preselector

The circuit in fig. 7 was designed for receivers tuning between 0.5 and 30 MHz. Active devices are an fet in a common-gate, source-input circuit and an npn silicon transistor in a standard common-emitter circuit. The transistors are inexpensive. The fet, a GE FET-1, costs \$2.25; the 2N3563 transistor is available from Poly Paks at four for a dollar.

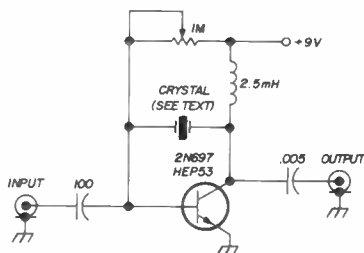
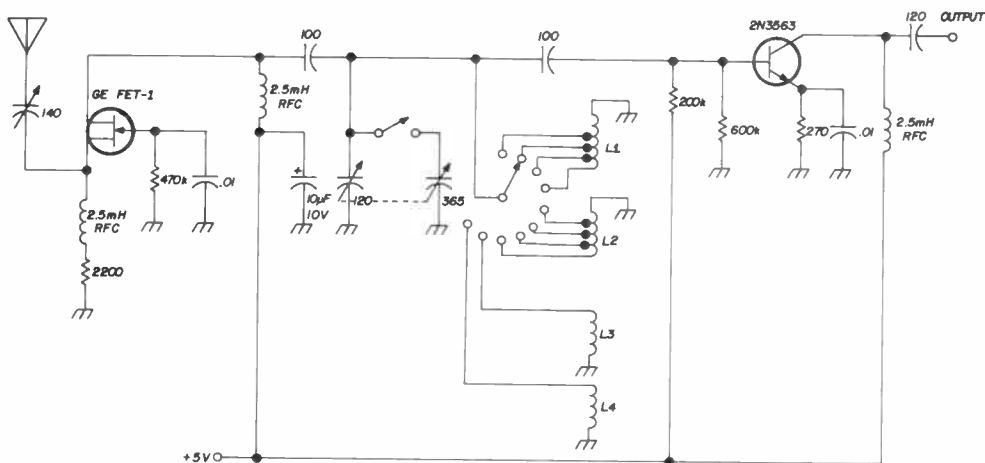


fig. 6. Simple mixer circuit for converting WWV signals to amateur bands.



L1 31 turns no. 18 airwound, 5/8" diameter, 1-7/8" long, (B&W 3007) tapped at 2, 11 and 27 turns (tunes 4-40 MHz)

L2 59 turns no. 18 airwound, 1" diameter, 1-7/8" long, (B&W 3016) tapped at 7, 28 and 38 turns (tunes 1.75-18 MHz)

fig. 7. Preselector for use with general-coverage receivers.

L3,L4 80 turns no. 26 closewound on 1/2" slug-tuned ceramic form, brass slug (L3 tunes 1.05-2.2 MHz, L4 tunes 0.5-1.25 MHz)

The preselector has fairly uniform gain. Measured at the receiver, preselector gain is 20 dB from 2 to 30 MHz, with a rising characteristic toward the low end of the broadcast band, where the gain of most receivers seems to be down.

The low-impedance source input of the fet matches low-impedance antennas. The fet is used mainly as an impedance-matching device and has little gain when used alone. High output impedance of the fet and high input impedance of the npn transistor result in low tank-circuit loading; thus tank-circuit Q remains high. With a 5-volt power supply, total current drain is less than 2 mA.

Another fet could have been used instead of the npn transistor; however, the 2N3563's gain characteristic, together with its low price, made it a desirable choice.

The broadcast band is divided into two segments. A two-gang capacitor, with provisions for paralleling, is used. This gives some flexibility, but it isn't entirely necessary for satisfactory operation. The variable capacitor in the antenna circuit is used to vary input coupling on the lower-frequency bands, since overloading causes cross modulation.

The transistor sockets and related circuit components are mounted on a 2 x 4-inch piece of perf board. This board, plus the larger parts, are mounted on a 4 x 7-inch piece of wood.

With my DX-150, which has spotty sensitivity and some image problems, the preselector improves reception on the low end of the BC band and on 160 meters. With a 25-foot-length of wire for an antenna, the preselector-receiver combination performs very well.

George Hirshfield, W50ZF

the simplest audio filter

Many audio filters have appeared in magazine articles for both CW and ssb. The more popular of these have been centered around 44- and 88-mH toroids available for as little as 30 cents each from surplus sources. This article describes such a filter for single-tone CW reception, but with all the nonessential frills removed. Also these filters have been tested on an audio-frequency spectrum analyzer, so their response using actual speaker loads is known quite exactly.

A CW filter must have one prime qualification to be considered good, namely very narrow bandwidth. Since 20 wpm represents only about 10 Hz actual information bandwidth, the filters of use to hams can be very sharp indeed. However, we can't practically use a 10-Hz filter, because transmitter plus receiver drift would require constant retuning. Also the filter would ring, producing disagreeable-sounding code. The ringing could be removed by damping, but the pulses would still have rounded, sine-wave-like shapes. Therefore, the filter should be somewhat wider than the minimum bandwidth based on information theory.

Assuming ssb-grade frequency stability, the combined short-term drift of transmitter and receiver will seldom exceed 50 Hz. This bandwidth, being five times wider than the minimum necessary to pass all the information, will pass pulses with a square-wave shape. So a practical CW filter should not be narrower than about 50 Hz; and to be called good, it should not be wider than about 100 Hz.

Another desirable feature for a general-purpose filter is that it operate in the speaker leads to eliminate the necessity for modifying resellable hardware.

If you're using an ssb receiver, all of which have linear product detectors, there is absolutely no signal-to-noise advantage whether the filter is in the predetection (i-f) or postdetection (audio) circuits. This isn't true for a-m receivers with their conventional detectors, however. For these, a Q multiplier in an i-f stage will give the ultimate in weak signal reception. But the audio filter is equally useful in a-m receivers for interference rejection. Why? Because the nonlinearity of the a-m detector becomes serious only on weak signals, where this nonlinearity produces noise. On weak signals the audio signal-to-noise ratio is considerably degraded from that which prevailed in the i-f amplifier before detection. This weak-signal degradation doesn't occur in an a-m receiver under normal interference conditions, since

then the signals are strong enough to operate the a-m detector above the degradation point.

However, you can't have everything; and even with an ssb receiver, a very narrowband filter in the speaker leads won't prevent a strong adjacent signal from generating strong agc, thus weakening the signal you're selecting with the CW filter. This can be annoying if the strong adjacent signal is CW, because receiver gain will flip up and down in response to keying. The desired signal will sound like one subjected to severe flutter or fading conditions. The only remedy for this in ssb receivers is a filter ahead of the agc rectifier. In a-m receivers, filtering in the i-f amplifier will prevent this problem.

circuit Q

The Q and hence the bandwidth of any practical filter is a function of the quality of components and circuit loading. Circuit Q defines the bandwidth as a percentage of the operating frequency. Thus a filter with a Q of 10 operating at 1 kHz would have a bandwidth of 100 Hz; and at an operating frequency of 100 Hz, the same coil with a larger tuning capacitor would still have a Q of about 10 but a bandwidth of only 10 Hz. Obviously if your desired signal had interference from a nearby signal, the lower you set the beat note, the easier it would be to

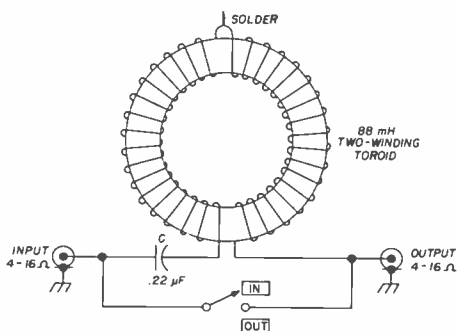


fig. 8. The simplest audio filter. Bandwidth is about 70 Hz at 1.2 kHz; circuit Q is approximately 17.

C (μ F)	Frequency (Hz)
0.22	1200
0.5	720
1.0	520
2.0	380

separate the two signals with a given filter. With an ssb receiver, about 300 Hz is the lowest you'd care to go, because the audio circuits have very little gain below this frequency. You can therefore choose a frequency between 300 and about 1500 Hz, depending upon personal tastes.

the simplest filter

The simplest filter circuit having all the desirable features just described is the series-resonant type of **fig. 8**. The capaci-

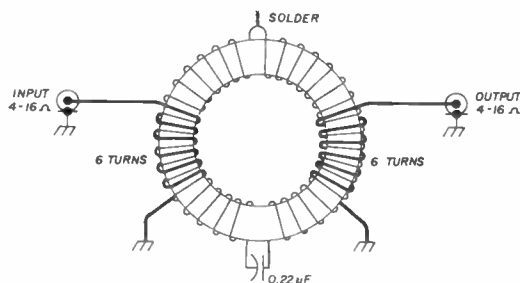


fig. 9. A filter with 40-Hz bandwidth at 1.2 kHz.

tor for this circuit should be $0.22 \mu\text{F}$ for a 1-kHz operating frequency, and $1.0 \mu\text{F}$ for 500 Hz. This filter has a bandwidth of about 70 Hz at 1.2 kHz and proportionally more or less at other frequencies. The Q of the filter is thus about 17, which consists almost entirely of the coil Q . Since this is the case the Q , and hence bandwidth, won't vary much if the speaker impedance is between 4 and 16 ohms.

For those who might want a very narrow filter but don't want to operate at a beat frequency near 500 Hz, the circuit of **fig. 9** will give about 40 Hz bandwidth at 1.2 kHz.

The 88-mH toroid in these circuits is an unpotted type available from surplus, and the low-impedance windings of **fig. 9** are each 6 turns of wire wound over the existing windings. Put the input winding on one side of the core and the output winding on the opposite side for the narrowest bandwidth. However, neither the number of turns nor their placing is very critical.

selectable-frequency filter

As I mentioned at the beginning of this article, these filters are designed for bare-bones simplicity. However, if you'd like more versatility at the expense of a few more components you can make a number of tuning capacitors switch-selectable. A selection of filter frequencies will allow the CW operator to copy signals that may otherwise be lost in extremely heavy interference. A schematic of a multiple-frequency filter is shown in **fig. 10**.

The insertion loss of all filters of this type is about the same: approximately 20 dB. This means 1 volt out for 10 volts in. This sounds high but really isn't. The audio-volume control can be cranked up to compensate. This insertion loss results because we're using a practical toroid instead of a perfect inductor. The Q of 17 means the coil has an effective ac resistance, including wire loss and core loss (and speaker load) of about 35 ohms. The insertion loss is the ratio of coil ac resistance to load resistance, which can't be improved for that particular coil. The speaker I used was a 3.2-ohm unit, which gave the 10:1 ratio of load resistance to total resistance reflected in the 10:1 input-to-output voltage readings in my measurements. This is explained so you won't feel something is lacking to cause

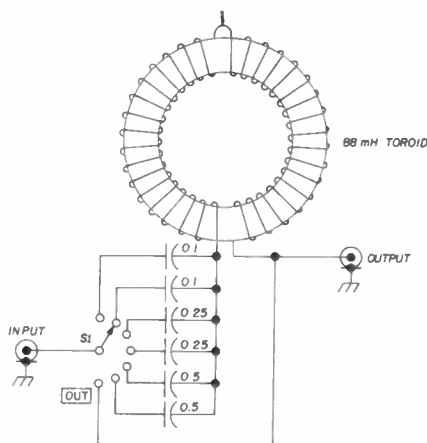
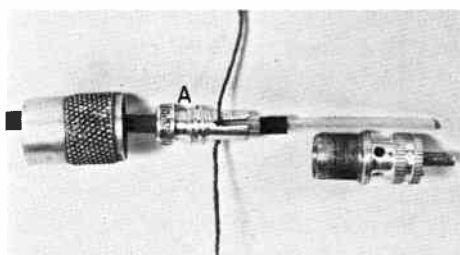


fig. 10. Frequency-selectable filter for improved CW reception. S1 is a shorting-type switch that connects the capacitors in parallel.

the adapter. Tin the notches on each side with a hot soldering iron.

Prepare the coaxial cable by pushing it through the adapter; strip off the outer jacket, unravel the shield and twist it into two leads opposite from each other. Slip the two pieces of twisted shield down the slots and solder them into the notch. If you've done a good tinning job, this shouldn't take too much heat. Now cut off the excess pieces of braid. With a little care and minimum solder, you should have no trouble screwing the adapter into the rest of the plug.

Save a small section of the outer



jacket — about a quarter inch long. This can be pushed into the reducing adapter and fills up the space around the dielectric, providing final electrical isolation. The inner dielectric is stripped off 1/16 inch in front of the adapter. When soldering the inner connector to the plug, be sure to slant it downward to keep any solder blobs from running down inside and shorting everything out.

Ted Woolner, WA1ABP

economical beam for ten meters

You've heard this old saw many times, but it bears repeating: without a beam antenna, it's futile to compete seriously for DX. I was listening to everybody working all the goodies on ten meters recently and recalled the antenna I used in the mid-fifties during the last sunspot cycle peak. It was a simple piece of plumbing that allowed you to get on ten meters in a hurry. In its basic form, it's

called a Wonderbar, or bow-tie antenna. Why not try it again, only this time crank some gain into it?

For those who may have forgotten, or who have recently joined the ham fraternity, I'll describe the procedure I used to adapt the original design¹ into an inexpensive beam for ten meters.

construction

The main source of material was an old biconical tv antenna. My beam was modeled after the original Wonderbar design using these materials and some hardware from my junkbox. The basic Wonderbar antenna that resulted is shown in fig. 3.

I dismantled the tv antenna completely. I cut two 30-inch crossbars from the old elements. Each end of the crossbar was flattened and drilled to accept 3/16-inch machine screws (I used a 13/64-inch drill). Next, the crossbars were attached to the open ends of each of the two elements. This forms a couple of isosceles triangles, or wing-shaped elements.

I used a handy piece of 3/4-inch pine board, 13-inches long by 10-inches wide, for the base. Any material can be used that's sufficiently rigid to hold the assembly. Standoff insulators, female coax connectors, and a length of 5/8-inch OD heavy-wall plastic tubing (for spacers) were produced from my junkbox.

assembly

Place the wing-shaped elements on the floor over the base. Space them about three inches apart. Drill six 13/64-inch holes (fig. 3) through the elements and completely through the base. (This will ensure alignment during final assembly.) Mount the female coax receptacle as shown near the top of the base. Place the standoff insulators and spacers as shown, and assemble them loosely with 3/16-inch machine screws. Don't tighten the screws on the standoff insulator where the coil will be attached.

the loading coil

If you don't have a B&W 3013 mini-inductor handy, it's easy to wind your own

such an apparently large insertion loss.

The filter in fig. 8 doesn't have dc continuity. Some ssb receivers, with no output transformer in their audio power amplifiers, require a dc path through the speaker. My Galaxy V is an example. A 4- to 16-ohm resistor across the input to the filter will solve this problem.

You've probably seen more complex filters for CW, but none will outperform these circuits in *actual use*. The loss of these filters increases 6 dB for each 40- or 70-Hz excursion from design center frequencies, and response is 40 dB down 200 or 350 Hz away.

The simplicity of the filter will allow it to be placed almost anywhere in your rig. The toroid measures about 1 by ½ inch, and the capacitors can be physically small. The filters will perform as stated if wired into your speaker leads if you'd rather not disturb the wiring of an expensive transceiver.

E. Dusina, W4NVK

a tunable audio filter for cw

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

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

Fig. 11 shows the filter response at 1000 Hz using the circuit of fig. 12. Gain at center frequency is approximately zero dB (gain of one), and the tuning range is from 750 to about 1600 Hz. The 3-dB bandwidth is 140 Hz.

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

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

The ICs require a dual-polarity power supply. My supply makes use of the trick known as "zero" or "common" reference to the IC. A single 9-volt transistor battery with two resistors is used, as shown in fig. 13. While this supply is adequate, slightly higher voltages will allow the filter to handle larger input signals.

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

operation

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

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

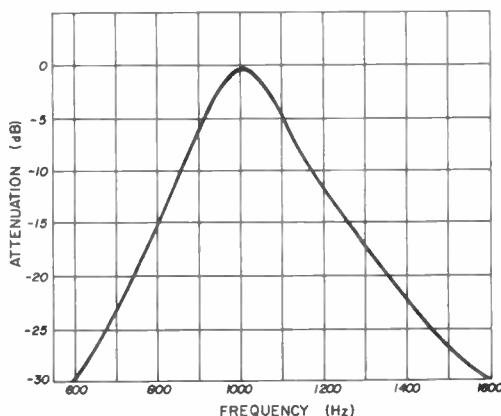
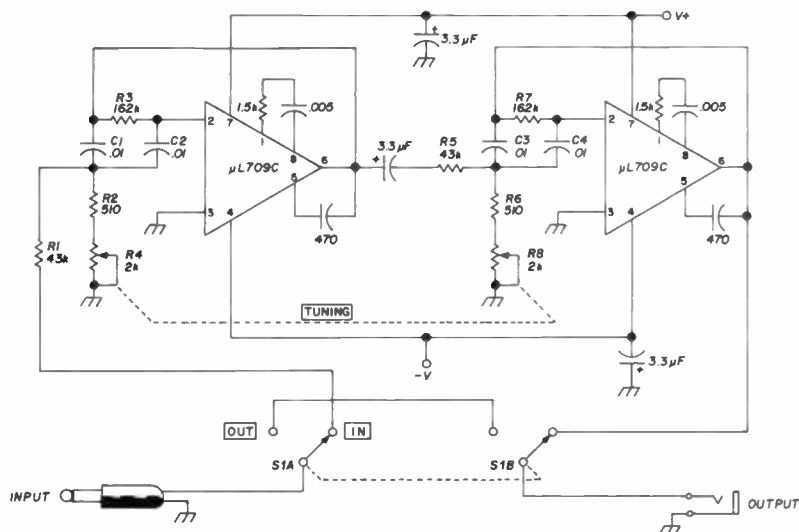


fig. 11. Filter response with a center frequency of 1000 Hz.



C1,C2, 01 μ F, matched to 5% or better
C3,C4

R1,R5 43k, matched to 2% or better

R2,R6 510 ohms, matched to 2% or better

R3,R7 162k, matched to 2% or better

R4,R8 2k ganged potentiometers (IRC no. PQ11-110 and M11-110)

fig. 12. Schematic of the tunable audio filter. Circuit features selective audio tone for CW reception.

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

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

N. J. Nicosia, WA1JSM

improving overload response in the collins 75A-4 receiver

Having obtained only limited results trying to improve the performance of commercially designed electronic equipment, I came to the conclusion that the designers knew what they were doing all along. Even so, the temptation to modify equipment is hard to resist, and my latest modification attempt resulted in a 13-dB improvement in strong-signal-handling

capability of the 75A-4 receiver. However, this was achieved by second-guessing Collins engineers some fifteen years after the set had been marketed!

I decided to modify my 75A-4 because several nearby stations caused front-end overload. This causes a decrease in weak-signal strength, even though the interfering signals are 25 to 50 kHz away. It's all but impossible to copy a weak CW signal under these conditions.

The remedy is easy: about five dollars worth of parts and a little time. Here's how to do it.

First replace marginal tubes, then align the receiver. Overload response may be checked using the setup in fig. 14. Set signal generator 1, representing the desired signal, to 3 microvolts at 14,050

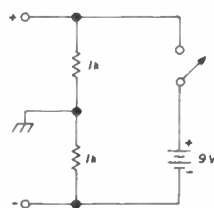


fig. 13. Transistor battery power supply that provides dual-polarity output for the filter.

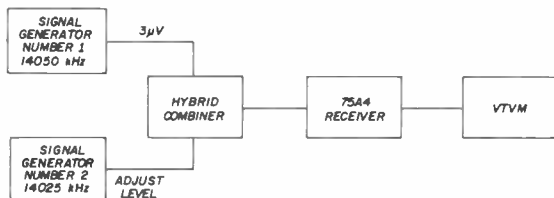


fig. 14. Test circuit for measuring effectiveness of receiver modifications.

MHz. Turn the receiver avc off and the bfo on. Now increase generator 2's level (at 14.025 MHz) until the desired signal, measured by the vtvm, decreases by 3 dB. This will require about 13,000 microvolts. All subsequent measurements are referenced to this level.

first mixer modifications

Replace the 6BA7 first mixer with a 12AT7. The modifications, fig. 15B, require only one 470-ohm $\frac{1}{2}$ -watt resistor and the 12AT7. Remove R14, R15, C35, C36. Revise the heater circuit as shown.

Next replace the 100-pF coupling capacitor with a 15-pF silver mica, then connect another 15-pF silver mica between grid and ground. This forms a capacitive voltage divider that reduces

signal level to the first mixer grid. (The photos (fig. 17 and 18) show a modified and unmodified set; actually, these are photos of two different receivers.)

After these changes have been made, peak the mixer grid and crystal-oscillator circuits for each band; also peak the mixer plate circuit. Only the capacitors should be peaked: C23, C26, C28, C30, C31, C32 and C17 in the mixer grid and C53 in the plate circuit.

The high-frequency crystal oscillator tuned circuits must also be retuned. Peak the tuning slugs on L11 through L17. Until the oscillator has been peaked, you'll probably find that the 21-MHz and higher-band crystals won't oscillate.

After the first mixer has been modified, a 24,000 microvolt signal will be required to cause a 3-dB decrease in the desired signal. I made this test before the two 15-pF capacitors were installed.

second mixer modifications

The next step is to replace the 6BA7 with a 6DJ8. Before and after circuits are shown in fig. 16. The new tube plus four new parts are required: a 1k, 1.2k and

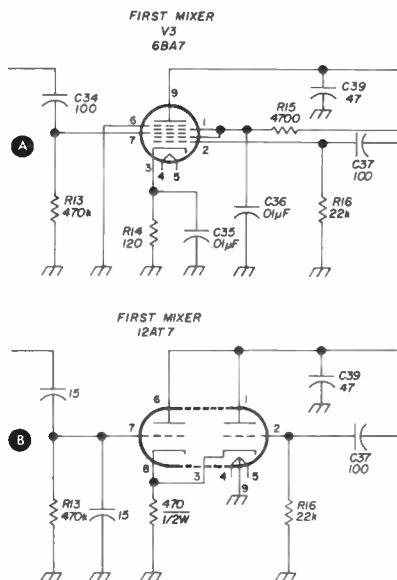


fig. 15. The 75A-4 first mixer before modification, (A), and after, (B). New parts consist of the 12AT7 and the 470K resistor.

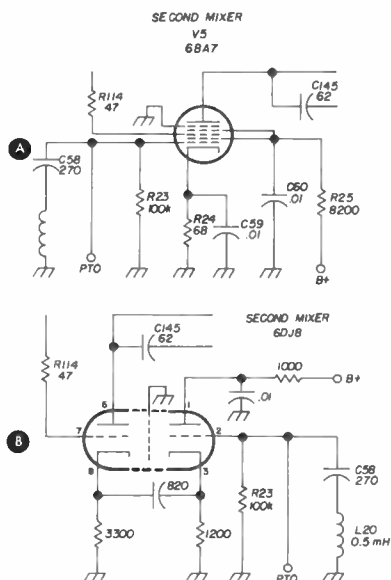


fig. 16. The 75A-4 second mixer. Original circuit is shown in (A); modifications in (B). Only five new parts are required for the change.

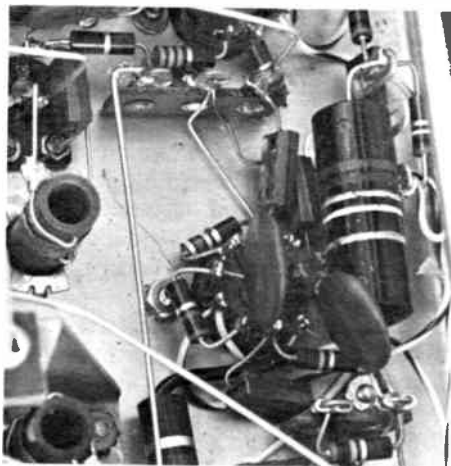


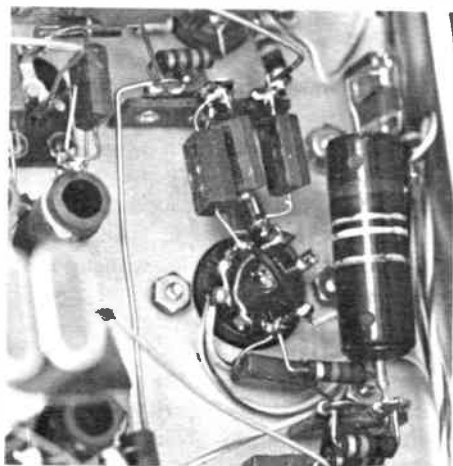
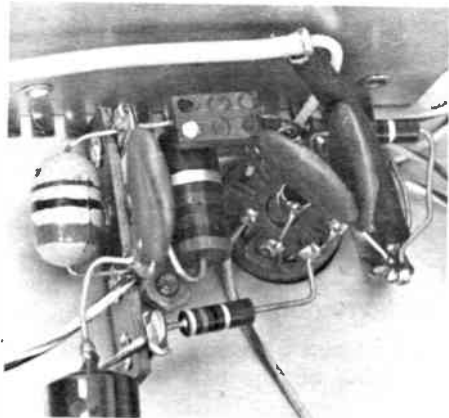
fig. 17. Wiring of unmodified 75A4 first mixer.

3.3k $\frac{1}{2}$ -watt resistor and an 820-pF silver mica capacitor. If you can't find an 820 pF capacitor, anything between 680 and 1000 pF will do. Remove R24, R25 and C59. Heater wiring is unchanged in this circuit. Complete the revision shown in fig. 16B. Peak the grid input circuits by tuning C56. *Do not* change inductor tuning.

After making the changes to the second mixer, check receiver performance again. A 40,000 microvolt signal from generator 2 will be required to reduce the desired signal by 3 dB.

In my receiver, the capacitive voltage divider was installed after modifications to the second mixer. A final check showed that an undesired signal of 60,000 microvolts was required to reduce

fig. 18. Unmodified 75A4 second mixer.



Modified first mixer; note reduced number of parts.

the desired signal by 3 dB (voltage ratio of 4.5).

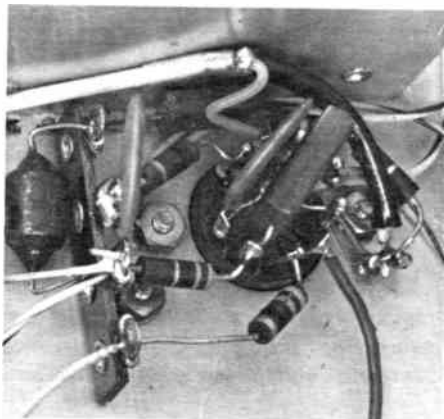
Measurements at 28 MHz showed an improvement of 3 dB in signal-plus-noise ratio. Overall receiver gain was somewhat lower, but this was more than compensated by the receiver's response to weak signals in the presence of local signals. More than enough gain was still available, however.

A type 6922 tube can be used instead of the 6DJ8. The 6922 has the same characteristics as the 6DJ8 but costs about three dollars more.

The modifications described are easily applied to the Collins 75A-4 receiver. The improvement in performance is well worth the investment in time and money.

Raymond F. Rinaudo, W6ZO

Component layout of modified second mixer.



75A-4 modifications

To anyone who has owned and cherished a 75A-4 receiver for many years, as I have, the discovery that it has developed sensitivity and frequency-stability problems is like discovering a trusted friend to be unfaithful. Here are some solutions to these problems, plus a hint to improve the receiver's audio response.

insensitivity

Loss in sensitivity first appeared as a loss of one or two S-units after about an hour of operation. (I use the *calibrate* signal as a sensitivity reference for a specific S-meter reading at 14.2 MHz, with a 50-ohm dummy antenna.) Sensitivity loss gradually increased until it was 6 or 7 S-units after 15 minutes of operation.

After several frustrating weeks of signal tracing and a new set of tubes, I was about to give up when I stumbled onto the answer. The *rejection tuning*, which is a bridged-T filter, has a sharp, deep null when properly adjusted. I noticed that when the set was first turned on, the *rejection tuning* behaved normally, but after warmup the null deteriorated and finally became useless. It seemed as though a comparatively low resistance was across the bridged-T inductor, L26.

The schematic shows one-half of V7, the Q-multiplier tube; some resistors; a choke; and C71, a 1000-pF capacitor in series across this inductor. Checking with a vtvm between ground and either side of L26 showed a *positive* voltage, which varied between 2 and 5 volts. I disconnected C71 from the inductor, but left the other end connected to the plate of V7 (and hence, B+). When I touched the vtvm probe to the free end of C71, I found the positive voltage to be even higher and still varying.

I checked other grids in the receiver with the vtvm and found four other leaky capacitors. When these were replaced, the sensitivity problem cleared up completely. A vtvm must be used, however. A

20k-ohms-per-volt vtvm simply shorts the leaky voltage to ground and gives no indication.

When checking for leaky capacitors the set must, of course, be on and the rf gain control positioned fully clockwise. If the rf gain control is backed off, the higher bias voltage will swamp the leakage voltage on some of the grids. I can only conjecture as to the cause of the capacitor leakage; but my friend Ray Wood, W9SDY, suggests migration of the silver coating on the mica. Until a better explanation comes along, I'll accept Ray's theory. Migration is supposed to accelerate in the presence of a dc voltage. However, some unused mica capacitors of the same type and vintage showed leakage to some degree; new units showed no leakage at all.

frequency instability

The frequency instability problem became evident in exactly the opposite manner. The pto frequency jumped around for about ten or fifteen minutes after the set was turned on, then settled to its usual rock-steadiness.

According to an article in *QST* by W3AFM, instability in the pto can be attributed to several factors, including capacitor C205, a 51-pF mica of the same type as those giving sensitivity problems. When I originally read the *QST* article, I assumed C205 might be changing value. However, after the previous experience with leakage, I found that a vtvm check between pin 1 of V15 and ground showed the telltale positive voltage.

After the set had been on for 10-15 minutes, the positive voltage disappeared, and the pto became stable. I suspect that, since a current flows through the capacitor when the pto is operating, the leakage path burns off after several minutes, then regenerates when the set is turned off. At any rate, the pto remained stable after installing a new capacitor for C205.

audio response

I became increasingly annoyed by an emphasis on the bass of the audio re-

sponse as both the 75A-4 and I grew older. I realize that as we grow older our ears become less sensitive to the higher frequencies, but retain their response to the lower frequencies, thus accenting the bass.

A newly acquired audio generator provided the opportunity to check the 75A-4 audio response. A surprisingly high peak (10 dB) appeared at 100 Hz, which decreased sharply on either side, to a level output between 300-3000 Hz.

W3AFM suggested that feedback resistor R71 be removed. I tried this when I first read his article, but I didn't like the increase in audio, which required riding the af gain control; nor did I like the unpleasant audio quality.

The audio-signal generator showed that, with resistor R71 disconnected, the 100-Hz peak disappeared and the entire audio response became a broad peak centered around 3200 Hz. Substituting a 100-k resistor for the 33k originally used for R71 eliminated the objectionable 100-Hz peak and smoothed the entire audio response. A 1-meg resistor in series with the af gain control improved its action.

less bass response

Another source of bass emphasis is the cathode network associated with the noise-limiter diode, V12. Since the limiter is useless for CW and ssb, I removed V12 and inserted the leads of a 0.005- μ F capacitor into the tube socket (holes for pins 2 and 7). This connects the detector output directly to the af gain control and eliminates the cathode network. If you would like even less bass, use a 0.001 μ F capacitor.

Albert G. Shafer, W4SD

a new cw monitor

Rf-actuated keying monitors certainly aren't new. They've been described in many articles, and there is even a commercial version available. However, the use of simple, inexpensive ICs, plus a different method of audio keying, make the monitor described here a worthwhile

and interesting project.

The main advantage of rf-actuated monitors, of course, is that no direct connection to the transmitter is needed: the monitor uses internal battery power and is actuated by the rf field from the transmitter. Another advantage to such a monitor, which is less precisely explainable but nonetheless very real — as most users of such monitors will testify — is the clean keying that results. Clicks and other distortion present with most directly actuated monitors are completely eliminated.

This rf-actuated monitor is unique in several respects. It uses only ICs and is a very simple project for anyone who would like to get acquainted with these devices. The IC is the Fairchild μ L914. It is becoming one of the most versatile ICs any amateur can use. As its designation (μ L for μ -micrologic) implies, it was developed for digital logic circuits but it serves very well as a linear IC that will amplify continuously varying af or rf signals up to several megahertz. It's available from large mail-order supply houses for less than a dollar.

Another interesting feature of the monitor is that the transmitter rf field is not used to activate a transistor switch to turn on an audio oscillator, as is commonly done. The audio oscillator in this monitor runs continuously. The rf field controls a μ L914 that is used as an enabling gate or switch. Audio passes through the gate only when the rf field closes or *enables* the circuit. The distortion and transients which occur when an audio oscillator is initially turned on with each transmitted keying character are thus avoided.

Particularly clean-sounding audio keying results from using this method. The audio oscillator shown in the circuit (another μ L914) is of fixed frequency; but if you have another audio oscillator available, perhaps of variable frequency, it could be used in place of the oscillator shown simply by feeding it directly into the enabling gate. Exactly how this can be done is covered later.

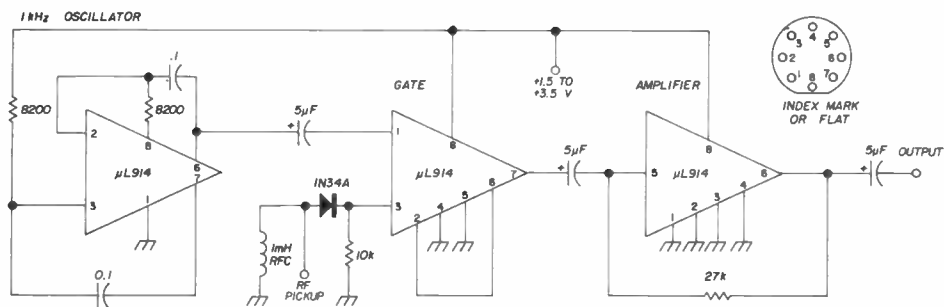


fig. 19 Schematic diagram of the integrated-circuit CW monitor.

circuit description

The μL914 consists of a pair of dual grounded-emitter transistor stages. External connections determine how the stages are used in any particular circuit function. Physically, the μL914 appears as an ordinary epoxy-cased transistor, except it has eight leads. The numbering of these leads follows the scheme shown in fig. 19.

It's particularly important to note that the *index flat* on the μL914 case denotes the 8th terminal lead. Anyone who is used to working with tubes will probably regard the index mark as denoting the 1st terminal lead, as I frequently did when first working with the μL914 . The result, of course, can very easily be ruined ICs.

The monitor circuit is shown in fig. 19. The first μL914 is connected as a free-running multivibrator. The oscillation frequency with the circuit components shown is approximately 1 kHz, but the frequency may be changed over a wide range by varying either the value of both resistors or both capacitors.

For variations over a small range — if you can't get used to the idea of hearing the same tone all the time — you can replace one of the resistors with a 20k-ohm potentiometer.

The second μL914 is actually connected to perform a logic function. In this case, however, the result is as though an spst switch, connected between terminals 1 and 7, were controlled by a dc-control voltage on terminal 3 (generated by rectification of the transmitter rf field).

operation

The 1N34A rectifies the rf pickup to produce a positive control voltage at terminal 3 of the second μL914 . The rf pickup may be obtained by connecting a short piece of wire (about 1 foot) to the rf pickup terminal, which can be placed anywhere in the vicinity of the transmitter power amplifier.

The last μL914 , which is used as an audio amplifier stage, actually makes very poor use of the capability of the μL914 . Only one transistor in the μL914 is active as an amplifier. A single audio transistor can be used for this unit if desired. Surprisingly enough, the cost will be about the same as that using the IC.

The amplifier stage output is sufficient to drive a medium-impedance (400- to 600-ohm) headset with adequate volume. If loudspeaker operation is preferred, the last μL914 can be followed by a transistor power amplifier of conventional design.

No special precautions need be followed in construction, except perhaps to note that the leads concerned with the rf pickup circuit should be kept as short as possible. The circuit of fig. 19 can be assembled on a 1-by-1½-inch piece of Vector board. The circuit can be mounted either inside a separate enclosure or inside the cabinet of your transmitter or transceiver.

The supply voltage is between 1.5 and 3.5 volts. The monitor will work well within this range, but at least 3 volts are necessary if the last μL914 drives a pair of headphones directly. The current drain

is very low and can be handled for an extended period by two penlight cells in series.

John J. Schultz, W2EY

miniature receiver antenna tuner

Most modern, high quality receivers have an input impedance of 50 ohms and are designed for use with matched, transmitting-type antennas. Unfortunately, these receivers must sometimes be used with random-length wires. This often leads to inadequate performance, especially poor sensitivity and degraded image rejection.

A simple antenna tuner can be used to correct this situation. A miniature tuner which will match a short high-impedance antenna to a receiver is shown in fig. 20. The version shown, which was suggested by a tuner built by K4BXO, has proved useful for matching 10-to-25-foot antennas between 10 and 30 MHz. Changing the type and size of core, number of turns and tap positions will modify the impedance matching and tuning range.

The tuner can be built in a very small minibox, since few parts are required and those that are needed can be miniature ones. A toroid similar to the one I used is available from Amidon. A plastic-dielectric variable capacitor from Lafayette is excellent for C1; its capacitance is not critical. I used a miniature Alco rotary switch.

The tuner is very easy to use. Simply tune the capacitor through its range for each switch position. There should be a significant improvement in signal strength

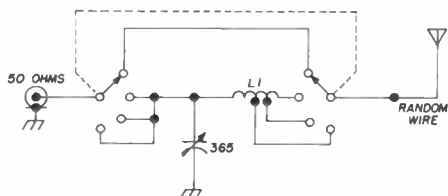


fig. 20. Receiver antenna tuner for random length antennas. For the range 10 to 30 MHz, the coil is 35 turns no. 26 on 5/8" diameter toroid, tapped at 10 and 20 turns.

and image rejection as this is done. If no improvement is noted, the antenna length or number of active turns on the core can be changed.

Paul Franson, WA7KRE

antenna impedance transformer for receivers

Amateurs spend a great deal of time and effort to obtain an impedance match between transmitter and antenna. The reason is obvious, of course. They want maximum power transfer with minimum loss; besides, it's just plain good engineering practice.

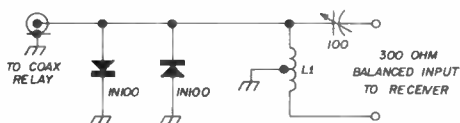


fig. 21. Receiver antenna coupler. L1 is 10 turns no. 12 enameled, tapped at the center.

While visiting several local stations, I was impressed with the fact that many hams neglect the other end of the circuit: Matching their antenna to the input impedance of their receiver. Receiver manufacturers have no way of knowing what type of antenna will be used, so a compromise is made in the receiver input circuit. The receiver is often designed for a balanced 300-ohm load.

If you run a length of 50-ohm coax from your antenna relay to this type of input circuit, an impedance mismatch of 6:1 will result. A little figuring will show that the input signal, with this mismatch, will result in a substantial loss.

The little gadget shown in fig. 21, mounted on a small phenolic board and attached to your receiver input terminals, will increase your receiver sensitivity by as much as 6 dB (voltage ratio of 2).

If the coil and capacitor combination has a Q around 10 or so, you can adjust the trimmer for maximum response, and the circuit will work without retuning on each band. If the Q is higher, of course,

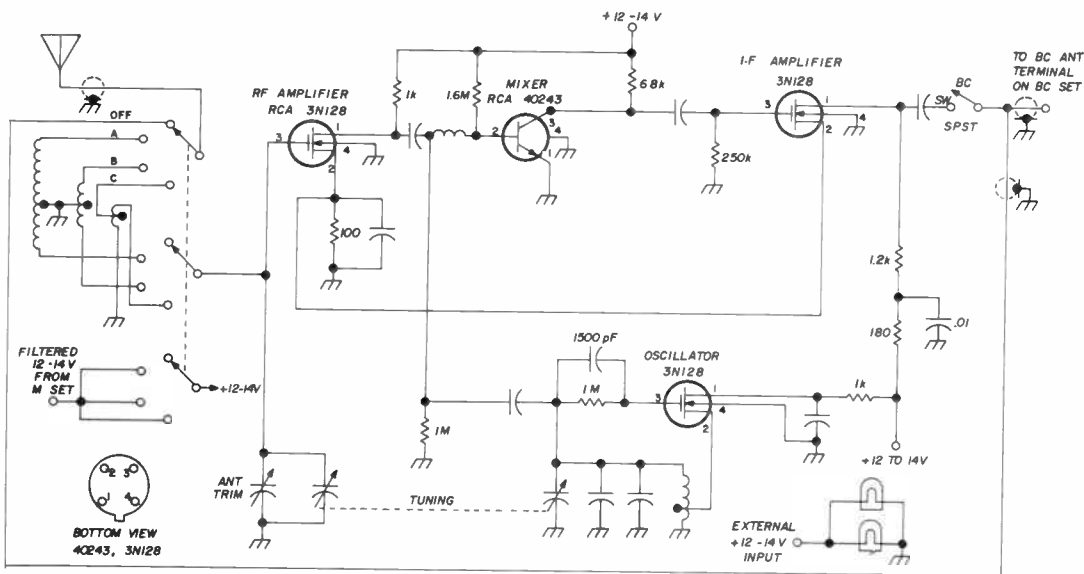


fig. 22. The solid-state Gonset converter. Mosfets and an npn transistor eliminate tube problems of noise, drift and high power drain. Components without values are unchanged from the original converter.

you'll have to readjust the trimmer for greater frequency excursions.

A note of caution: Most coax antenna relays have quite high isolation (up to 40 dB) between transmitting and receiving positions. However, with this transformer in the circuit of a receiver having an fet front end, the additional increase in input signal could damage the fet; relay isolation notwithstanding. A couple of diodes, placed as shown in fig. 21 will protect the fet from overload.

Alf Wilson, W6NIF

solid-state modification of a mobile converter

I was thinking of installing an old Gonset converter in my car, but the car's 12-volt system presented some problems. The converter tubes had 6-volt filaments, which would require a dropping resistor or heavy-duty zener. Also a power supply of +180 volts or more would have been necessary. Both these items would have represented considerable wasted power, more expense, installation problems, much more work, noise pickup problems,

has filtering, etc. Besides, I had a new all solid-state BC radio.

I felt that fets and/or transistors would eliminate all these problems, and the small amount of filtered dc power required could be readily supplied by the BC set after it had been adequately filtered to eliminate input noise. Additional benefits would be instant warmup, no oscillator drift, no tube-pin noise from road shocks, and no worry over failure of tube filaments, element shorts, etc.

If mechanical redesign could be eliminated and electrical redesign limited to changing a few resistors and capacitors, the overall expense and effort would be drastically reduced even further. With this in mind, a most satisfactory modernization was achieved with equal or better performance than when the converter left the factory.

construction

A most troublesome problem after removal of the tubes was finding a place to mount the transistors, since clearance was negligible between the tube socket connection ears and the bottom of the enclosing chassis. I finally decided to "spread-eagle" the short wires of the

solid-state devices, use ¼-inch pieces of spaghetti, then solder the wires directly to the socket ears where old and new components would also be convenient for connection. An all-fet converter was originally planned. However, the great improvement in conversion gain afforded by an npn transistor justified its use as the mixer.

The modifications are shown in fig. 22. I found it necessary to put in a slide switch to disconnect the antenna circuit from the output fet (during bc reception only), otherwise the loading effect was sufficient to ruin the Q and gain of the BC set. But this was the only mechanical change. There was plenty of room on the back apron of the converter to install this switch.

Where values of components are not given on the schematic, the value is the same as originally furnished by Gonset. Of course, the filament wiring, screen-grid wiring, screen-grid bypass capacitors and dropping resistors may be cut out of the chassis; and good riddance. Note that the pilot lights were changed from 6-volt 47s to 12-volt types and disconnected from the line feeding the collector and drains. They may be separately fed from a battery clip and wire attached to the instrument panel light to lessen the IR drop from the BC-set hash filter. In this way, when the BC set is turned off, the converter is "fail-safe" turned off too, irrespective of converter switch position. Wire from the on-off switch on the BC set should be shielded and fed to the shielded B+ line of the converter via phono pin and jack, or similar means, so each may be separated at will.

John R. Schuler

simple cw monitor

Did you ever try to send code without using a CW monitor? To have consistently good sending characteristics without a monitor is like trying to drive an automobile blindfolded. Hundreds of articles on CW monitors have been published, but many of them are elaborate in design and expensive to build. The *minimonitor* is

quite the opposite — it's the ultimate in simplicity. It costs pennies to build, is extremely reliable, and has a clean, pleasant-sounding tone.

The complete circuit is shown in fig. 23; components are mounted on a 1¼" x 2¼" circuit board. Although I use the *minimonitor* with an automatic keyer, it can easily be used with straight or semiautomatic keys. It can also be used as a code practice oscillator; volume is sufficient for group listening.

Perhaps the most desirable features of the *minimonitor* are its small size and easily obtained components. It is small enough that it can be built into practically any keyer already completed by fitting it into a convenient space.

Choice of parts is not at all critical. Virtually any npn and pnp transistors can be used. The capacitor value can range anywhere from .01 to .05 μ F. The resistor depends upon the desired tone. The diode is not critical; I used a 200 PIV unit. If you want, a 50k potentiometer can be used in place of the resistor to provide variable tone adjustment. Power to the unit may range from 3 to 6 volts, and any small 4- or 8-ohm speaker may be used.

If you're skeptical about using the *minimonitor*, dig the parts out of your junkbox and connect them on top of the kitchen table. With a small 4.5-volt battery for power you'll get an idea of how the oscillator sounds. You'll also be able to select parts for the circuit board since tone quality depends on component choice.

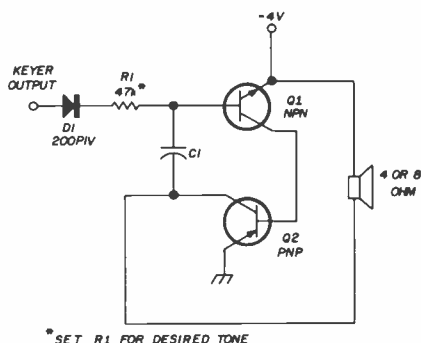


fig. 23. Schematic diagram of the simple CW monitor. Capacitor C1 is 0.01 to 0.05 μ F.

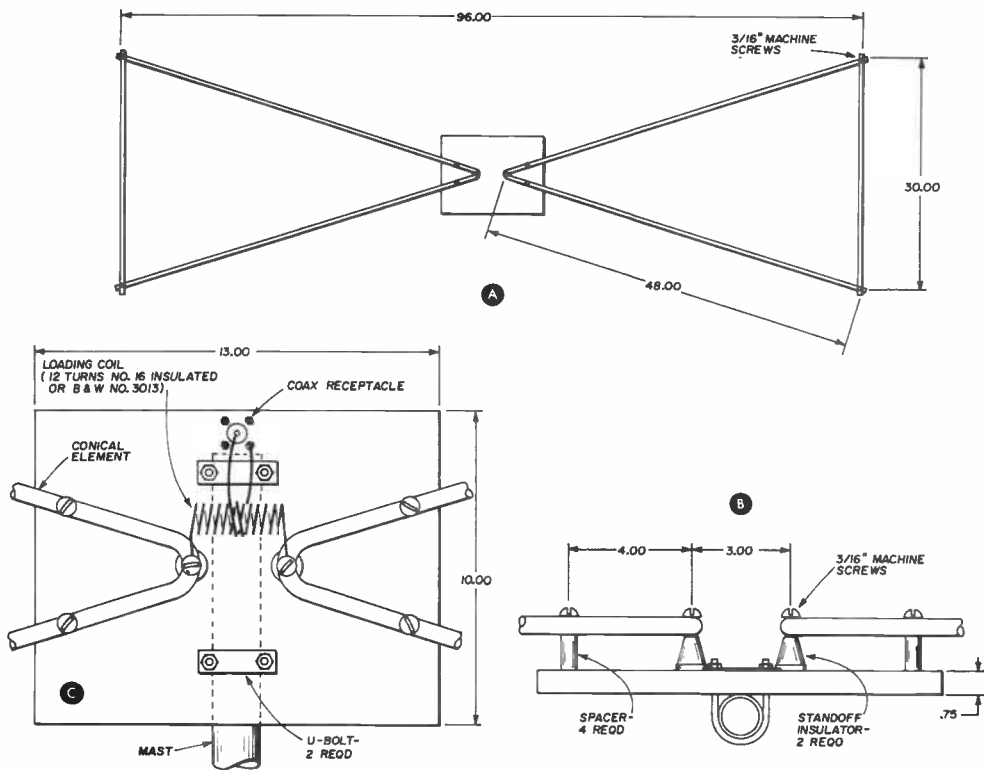


fig. 3 Dimensions, A, and mounting details, B, of the Wonderbar beam driven element.

coil. I used what was available: number 16 insulated solid copper wire. Considerable latitude can be used here. Just make sure the coil is sufficiently rigid to be self supporting. I wound my loading coil around a 1-inch dowel, using 12 turns, close spaced. Then I removed the dowel and stretched the coil until it was about 3 inches long. Next I wound 2 turns of number 12 solid copper wire around a 1¼-inch form for the primary. This coil was removed from the form and slipped over the loading coil.

Attach the loading coil to the elements by securing each lead to the screws at the apex of the V formed by each element. Place a solder lug between the head of the screw and the element. Now solder the primary coil leads to the coax connector. Tighten all machine screws.

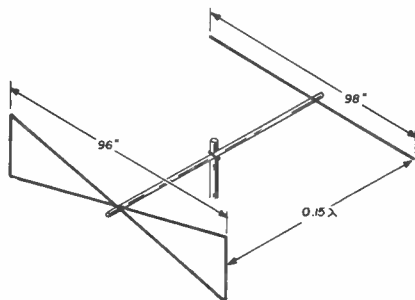
mast mount

Four holes are drilled to accept U bolts, which will secure the antenna to

the mast. I used 5/16-inch holes, positioned over the centerline of the base. The first two were immediately below the coax connector, and the second two were about one-half inch from the bottom of the base. Use your own ideas here to fit your available hardware.

preliminary tests

I attached a ten-foot piece of 1¼-inch conduit to my antenna for initial tuneup. I raised this assembly, with a piece of

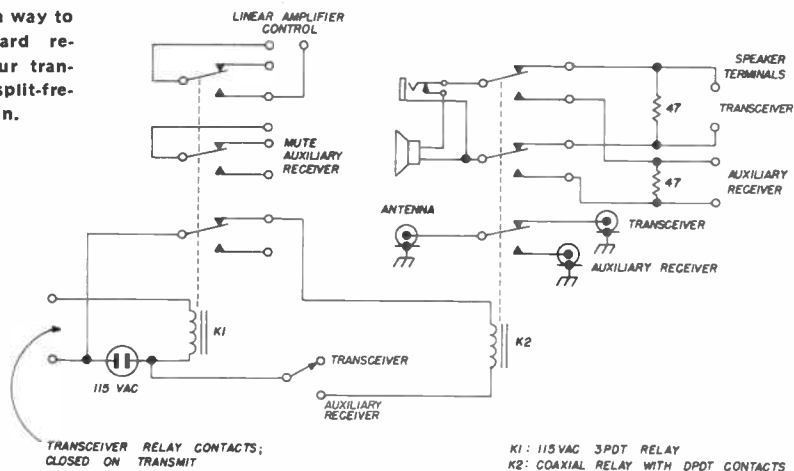


Overall view of the completed Wonderbar beam.

James Wiesmueller, WA9OHR

If the af gain-control in your transceiver is a 500k pot, R1 and R2 will be 1

fig. 26. Here is a way to use an outboard receiver with your transceiver for split-frequency operation.



simple wwv receiver

The simple fixed-tuned WWV receiver shown in fig. 25 uses a crystal filter and an IC i-f amplifier designed by W6GXM. Rf and audio amplification are provided by two inexpensive experimenter's modules available from the International Crystal Company. Selectivity with the crystal filter is excellent — on the order of 6 kHz — more than sufficient for reliable WWV reception.

Doug Pongrance, WA3JBN

improved sidetone operation for the SB301/401

After using my new Heath SB-301/401 combination for several weeks, I found one thing that was evidently overlooked — the sidetone monitoring system works through the speaker only, even with the earphones plugged into the phone jack. To make the sidetone monitoring compatible with earphone operation only involves the installation of a jumper from the 500-ohm *anti vox* jack to the *spare* jack diagonally across from it on the rear apron of the SB-301 receiver. The bus-wire jumper should be installed inside the chassis.

To complete the modification, it is also necessary to change the intercon-

necting audio cables that are used between the receiver and transmitter. The cable from the speaker is now connected to the *speaker 8 ohm* jack on the SB-301. The *speaker* jack on the SB-401 is connected to the newly wired *spare* jack on the SB-301. The *rcvr audio* jack on the SB-401 is no longer used. Other interconnecting cables are the same as before. With this simple modification, it is possible to use the sidetone monitoring system through either the earphones or the speaker. This modification should also work with the earlier SB-300/400 combination.

Don Bennechi, W1WLZ

using an outboard receiver

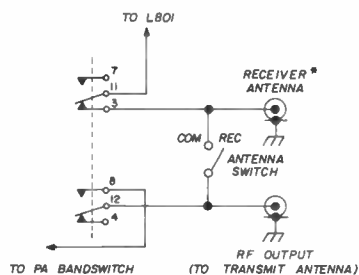
There are many times when it's helpful to have an outboard receiver available for listening on frequencies other than your transmit frequency. Remote vfos accomplish this in most cases, but not always. A good example occurs when working DX stations on 80 meters. With your transceiver tuned to 3850 kHz, the pre-selector 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. 26, an outboard 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 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.

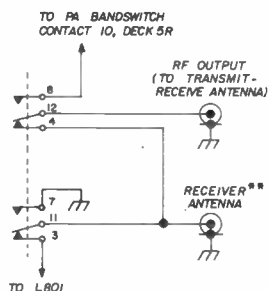
Jim Fisk, W1DTY

using an outboard receiver with the SB-100

The following idea is offered as an addendum to the above article. In my



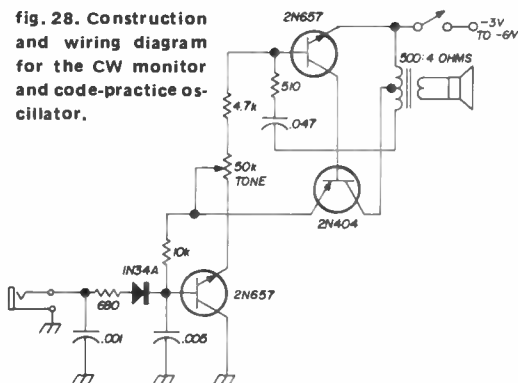
*THIS JACK ORIGINALLY USED FOR OPTIONAL CONNECTION OF A SEPARATE RECEIVING ANTENNA.



**THIS JACK NOW USED FOR CONNECTION TO OUTBOARD RECEIVER'S ANTENNA POST

fig. 27. Original, (A), and modified circuit, (B), showing SB-100 T-R wiring for outboard receiver; no new parts are required.

fig. 28. Construction and wiring diagram for the CW monitor and code-practice oscillator.



case, it was a simple matter of rewiring two sets of contacts on my Heath SB-100 antenna T-R relay so I could use a separate receiver with the unit. I wasn't concerned with switching the two audio outputs to a single speaker, as shown in the previous article, so no provision was made for this. The before-and-after circuits are shown in fig. 27.

Bill Clements, K4GMR

code-practice oscillator and monitor

If you ever operate CW, you know the necessity for a good CW monitor. Here is one which can be used with any transmitter that uses grid block keying. It can also be used for CW practice. Although many CW monitors have appeared in print, this one is very practical and simple (see fig. 28).

The oscillator uses two transistors in a modified multivibrator circuit. A small 500-ohm to voice-coil transformer drives the speaker; it also cleans up the tone for easy listening. The keying is done by a 2N657 transistor in the voltage return line from your transmitter. A diode isolates the transmitter keying bias from the transistor. Although I used two 2N657s (nnp) and one 2N404 (pnp), almost any junk-box transistors will work. Just use npn and pnp devices where I did. The transformer I used was salvaged from an old transistor radio. You can use either the receiver speaker or headphones.

Battery voltage can be anything in the

range from 1.5 to 9 volts. With three D-size flashlight cells series-connected to give 4.5 volts, the volume is about right for normal conditions.

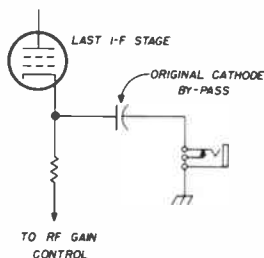
This is a pretty versatile unit; in one case it was even used to activate the vox input for semi-break-in-CW with a Gonset GSB-100 transmitter. It has also been used for code classes — for higher volume just add some more batteries.

Del Crowell, K6RIL

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 (fig. 29). 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.

fig. 29. I-f cathode jack.



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-5er, fm adapter, monitor scope, etc. An ac vtvm can be plugged into the jack for precise indication of signal level. With a vtvm 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

s-meter readings

If you have ever wondered what the S-meter readings actually indicate, the following chart may be of some help. This chart is based on an input signal of 50 microvolts at the antenna terminals for S-9 and 6-dB steps between each S-unit.

This is the standard used by most receiver manufacturers.

s-meter reading	signal strength (μV)
1	0.18
2	0.37
3	0.75
4	1.5
5	3.1
6	6.25
7	12.5
8	25.0
9	50.0
9 + 10 dB	158
9 + 20 dB	500
9 + 40 dB	5,000
9 + 60 dB	50,000
9 + 80 dB	500,000

That 80 over S9 report you just got means that you have a half-volt signal into the other fellow's receiver.

Bear in mind, however, that the S-meter is not an absolute signal-level indication. Due to nonlinearities in the preceding circuits the scale is often not linear. Also, the S-1 reading is often the agc threshold of the receiver, and can vary from less than 0.5 μV up to 2 μV , depending upon receiver design. With a 1.5 μV agc threshold as S-1, for example, the steps between each S-unit are approximately 5.0 dB.

Jim Fisk, W1DTY

transmitters

fet transmitters

The field-effect transistor has recently been finding its place in radio communications equipment. Although its performance as an rf amplifier and mixer is well known, the fet's merit in other functions seems to have gone unnoticed. The fet can function well in dc amplifiers, audio amplifiers, switching circuits, oscillators, multipliers and phase modulators. This article presents a low-powered vhf transmitter that uses field-effect transistors in every stage.

Although the fet deserves consideration when designing vhf transmitters it will not quell the nightmares of the solid-state vhf transmitter designer. It has its advantages: inexpensive, simple class-C biasing, low feedback capacitance, good efficiency with a 12-volt supply and relatively high power gain. The fet also has its disadvantage — low power dissipation. Commercially available field-effect transistors were developed primarily for small-signal use and typical power dissipation ratings are on the order of 0.4 watt. Since these ratings are based on no external heat sinking it is possible to decrease the case temperature and operate the fet beyond published ratings. However, for low-powered transmitters or low-power stages of higher powered transmitters, field-effect transistors can be used as is.

I have tried several experiments with transmitting circuits using fets, including fundamental and overtone crystal oscillators, frequency multipliers, rf amplifiers and phase modulators. The two transmit-

ters described here use experimental circuits. Both n-channel junction fets and n-channel depletion-mode mosfets were used. Mosfets and jfets are usually interchangeable but in most cases the simplicity of gate-leak bias for class-C circuits favors the jfet. The same results can be obtained by adding a diode from gate to source in the mosfet. In most cases performance is sufficient and the use of a mosfet doesn't warrant the additional component.

two-meter fm transmitter

The two-meter fm transmitter in fig. 1 uses five junction field-effect transistors. It is intended to be used with a portable vhf receiver as an fm walkie-talkie. Using a Heathkit GR88 vhf monitor receiver (tuned below its normal range) a range of over one-half mile was obtained between two walkie-talkies. An additional ampli-

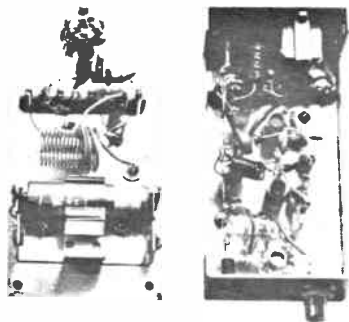


fig. 1. Two-meter transmitter on right uses five field-effect transistors; schematic is shown in fig. 2. Ten-meter fet transmitter is on left.

fier, tube and transistor, could be added for additional power for more serious work.

The cost of the fet transmitter is quite low. Homemade coils and chokes were used along with bargain-variety transistors. Devices were selected for optimum performance in the 72-MHz tripler and the 144-MHz doubler stages. Excluding the crystal the total cost was below \$10.00.

In this transmitter Q1 operates as an 8-MHz Pierce crystal oscillator which drives the phase modulator. The phase

drive a higher powered amplifier.

The transmitter was originally designed for a 9-volt power supply to conform with the monitor receiver. However, fet multipliers, like their vacuum-tube counterparts, require the highest supply voltage permissible for maximum efficiency. The fets used in the transmitter had a maximum V_{DS} of 20 volts and, as expected, best operation was obtained with a supply voltage of 20 volts. Two 9-volt batteries in series were used for portable operation; 9 volts for the receiver and 18 for the transmitter. A small

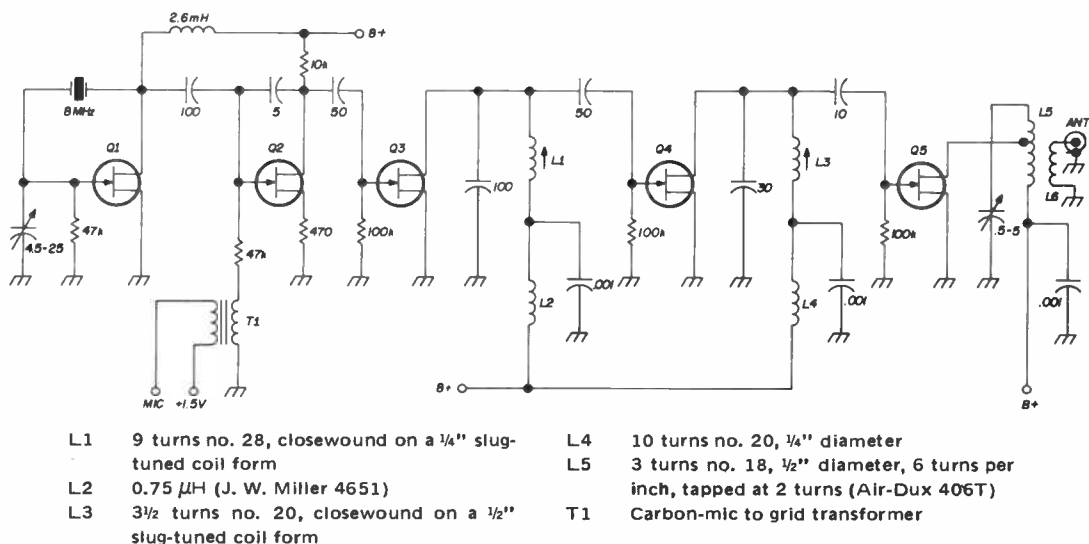


fig. 2. Schematic diagram of the two-meter fet phase-modulated fm transmitter.

modulator was designed after a circuit used in an antique Link high-band mobile transmitter. The Link modulator provided 20-kHz deviation at 144 MHz using a 3.0-MHz crystal. I had no difficulty obtaining 5-kHz deviation from an 8-MHz crystal in the fet version. The modulator drives a single tripler tuned to 24 MHz, which drives another single-tuned tripler and a doubler. The efficiency of the higher frequency multipliers decreases with increasing frequency, and it was necessary to select fets for optimum performance. Since the single-tuned multipliers offer little rejection of unwanted harmonics it is desirable to double tune the 144-MHz doubler if the unit is to

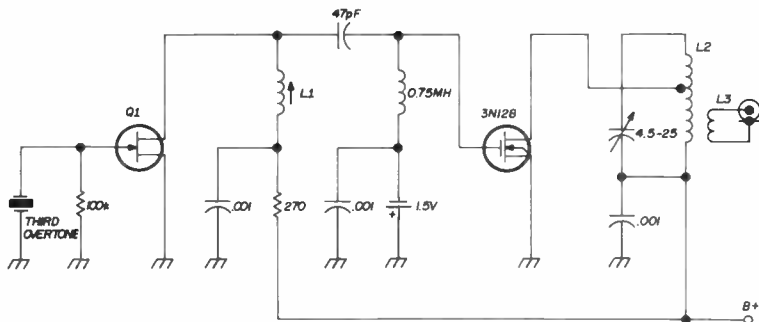
dry cell was used for the carbon microphone.

ten-meter transmitter

The transmitter in fig. 3 is capable of 200-mW output into a 50-ohm load at 29 MHz with an overall efficiency of 50%.* The transmitter is a two-stage affair consisting of a jfet overtone oscillator and a class-C mosfet amplifier.

I have had great success with fet crystal and self-excited oscillators. Not one fet oscillator failed to oscillate, and all the experimental circuits exhibited

*Overall efficiency is the total transmitter power input divided into the output power. This includes the power for the oscillator.



L1 12 turns no. 36, closewound on 1/4" slug-tuned form

fig. 3. Ten-meter transmitter using a mosfet power-amplifier stage. Power output is about 200 mW.

excellent frequency stability. In the tuned drain-oscillator in the ten-meter transmitter it was possible to operate fundamental crystals in the third overtone, and to operate overtone crystals in the fundamental simply by changing the tank resonance.

The final amplifier in the ten-meter transmitter uses a single 3N128 mosfet with a 1.5-V dry cell for fixed bias. Since the drain on the bias battery is essentially zero, a small 1.4-volt button-type mercury cell could be used since its shelf life is several years.

Two fets were tried in parallel with a small increase in power output but a reduction in efficiency. Previous experiments with parallel mosfet amplifiers at 50 MHz showed some hope, but most circuits, when pushed beyond one-watt output, resulted in burned out mosfets. Because of the difference in mosfet characteristics, one transistor does all the work while additional parallel devices only decrease efficiency. The single 3N128 required no neutralization and provided almost 200 mW output with a 12-volt supply.

The purpose of the ten-meter transmitter is the same as that of the two-meter fm unit. Many inexpensive 30- to 50-MHz monitor receivers can easily be tuned to 29 MHz, converted to a-m, and used with the fet transmitter. From my experience with 100-mW citizens-band transceivers on 10 meters, the range of such a combination should be quite respectable.

Albert D. Helfrick, K2BLA

L2 13 turns no. 18, tapped at 8 turns (Air-Dux 408T)

L3 3 turns hookup wire around cold end of L2

parasitic suppressor

Finding the right parasitic suppressor for a particular circuit can be a frustrating task. As a builder of a homebrew linear amplifier that had severe parasitic problems, I know what I'm talking about. I spent many days trying to suppress the parasitics and I finally came up with a coil-resistor combination that did the job on 80, 40 and 20 but wouldn't come through on 15 or 10. Since three out of five is better than nothing, I left the choke in. After operating for a few months and missing all the action on 15 and 10, I decided to try something else.

At the suggestion of a friend, I installed ferrite beads in the plate leads of my linear; I could not detect the slightest parasitic! Instant suppression. The magic of ferrite does not stop here. It can be used for rf shielding and decoupling; grids can be shielded from strong rf fields, thus reducing instability (see fig. 4).

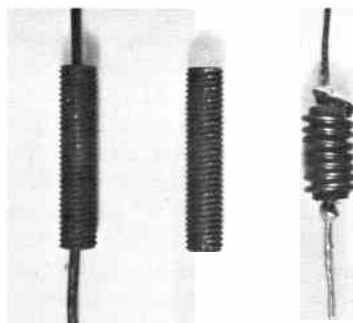


fig. 4. Ferrite beads from tv slug-tuned coils.

The surprising thing about these little giants is that almost anyone with even a small junk box has some: they're used in coils and coil forms to increase inductance. The slug itself is about 3/16 inch in diameter and 1 to 1½ inches long.

When you slip a piece of wire through the slug, you create a low-Q inductor. The wire itself has a very small but finite inductance that is multiplied many times by the permeability factor of the slug, which is 900 or so. Although the inductance is still quite small, it's enough to suppress vhf parasitics.

Jim Barcz, WA9JMY

solid-state screen clamp

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 dis-

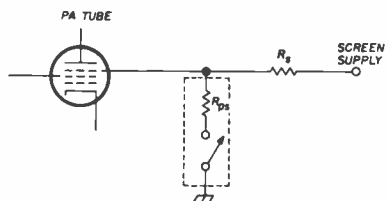


fig. 5. Clamp-tube equivalent circuit.

carded in favor of a transistorized version. To see why, let's discuss the principles of each concept.

tube vs transistors

As shown in fig. 5, 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

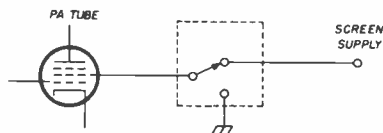


fig. 6. Transistor equivalent of the clamp circuit.

on the clamp tube disappears, and it conducts heavily. 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_S and the clamp tube's saturation plate resistance R_{PS} . 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_S , because of the high plate current through the clamp tube. This power is wasted and appears as unwanted heat inside the transmitter cabinet.

These disadvantages may be overcome by devising a circuit which will work like fig. 6. 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. 5, 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. 7.

transistor circuit operation

The diagram of the solid-state screen clamp I use in my transmitter is shown in fig. 7. 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 screen). Therefore, there is no final-amplifier plate current.

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 2N3440s I used with a 300-volt supply in fig. 7 have

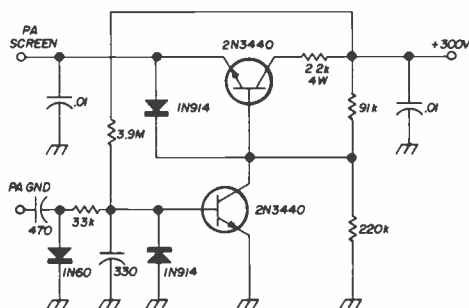


fig. 7. Schematic of the solid-state screen-clamp circuit.

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. 7). 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. 7, 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) increases, the voltage drop across the transistor goes down. In the circuit shown in fig. 7, 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 0.01- μ F capacitors on the screen and supply terminals are used as bypasses. The two 1N914s 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. 7. 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 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.

Chris Grant, WØLRW

half-watt solid-state cw transmitter

I have a friend who ridicules solid-state equipment, even after I've worked both coasts from mid-U. S. A. on six-meter phone with 100 milliwatts! So I told him I'd build and give him a solid-state CW transmitter. Its schematic is shown in fig. 8.

I made it as simple as possible, using junkbox parts and surplus or bargain transistors. It has only one tuned circuit, the output. The complete circuit evolved from several other circuits plus some ideas of my own.

At first, some of the base loads were rf chokes, but due to feedback or chassis ground loops, the final took off when detuned, so resistors were tried (with the same dummy antenna load). The resistors remained, and the unit is very stable.

The rf chokes are peaking coils from an old tv set; transistors are from IBM or similar boards sold everywhere. The coil is from an old surplus receiver.

The unit may be run at 15 or 16 volts, but all transistors should have heat sinks. It would be a good idea to change the 15-kilohm resistor in the oscillator base if you plan to run more than 12 volts. (Try

RG-8/U coax attached, in a vertical position and firmly lashed it to a picnic table. I found the best loading by tapping down on the loading coil; 10½ turns seemed to be optimum. I made a permanent connection at this point by soldering. The lowest standing wave ratio (about 1.4:1) occurred at about 28.95 MHz.

With 65 watts input, I made contacts with two WØ stations, and got 5-9 plus reports. The next day, I worked a KV4 and a couple of Gs. This simple antenna did indeed put out a good signal. But I wanted it to put out a better signal, so I added a reflector.

At this point you can enjoy this inexpensive antenna without further embellishments. It will provide a good signal on ten meters, it doesn't cost much, and you'll work some DX. However, if you like to experiment a little, as I do, you'll want to improve its performance. A simple reflector placed behind the Wonderbar antenna will produce from 3 to 5 dB gain over a reference dipole. This will effectively double your radiated power over the Wonderbar alone.

Handbook data showed that the shortest spacing for a reflector to improve performance was 0.15 wavelength (a little more than 4 feet on ten meters). This meant I could use the boom from the old tv antenna by merely adding a short extension. I decided to depart a bit from convention, for ease of assembly, and attached the boom and reflector immediately below the point where the Wonderbar was attached.

I made the boom extension about 5 inches longer than required. Then I put a bend of approximately 110 degrees radius in the extension about 4 inches from the end. I drilled two holes through the shorter leg of the boom extension and through the mast. This made an easy means of attachment. You could use the more conventional method of attaching the driven element and reflector at opposite ends of a one-piece boom. It would look prettier, perhaps, but wouldn't work any better. I wanted to use the materials on hand, so I used a short extension on the old tv boom.

the reflector

This element is simplicity itself. I cut my reflector from the remaining pieces of the old tv antenna tubing. It is 98 inches long (fig. 2). I used a pair of sheet metal cross braces I'd stripped previously from the tv antenna boom to attach the reflector.

how far, wonderbar?

That's it. A simple, low-cost beam antenna made from materials on hand. The whole thing cost less than ten dollars. Results; I've worked fifteen different countries, many of them several times, including a ZS6. All this was done with a 60-watt a-m transmitter and my Wonderbar beam only twenty feet above ground.

references

1. E. T. Bishop, K6OFM, "The Wonderbar Antenna," *QST*, November, 1956.

R. A. Clymer, W1FPF

coaxial cable supports

The opener flaps on the new aluminum beer and soft-drink cans make very strong supports for coax cable. Bore a small hole in the flap for a screw or a nail.

Don Farrell, W2GA

fixing a sticky AR-22 rotator

Many amateurs use the CDE AR-22 rotator with great success, especially where there is little cold weather or icing. However, in areas subject to ice storms, operators may experience trouble with a sluggish rotator, or one that only goes part of the way around. When the AR-22 is strained by heavy loading, the rather husky motor tends to bend over the teeth in the drive gears.

The three drive gears, part number TRA-39, consist of three thin iron gears sandwiched together on one spindle. They drive the TRA-18 ring gear which goes around the perimeter of the rotator. If one of the gears in the sandwich

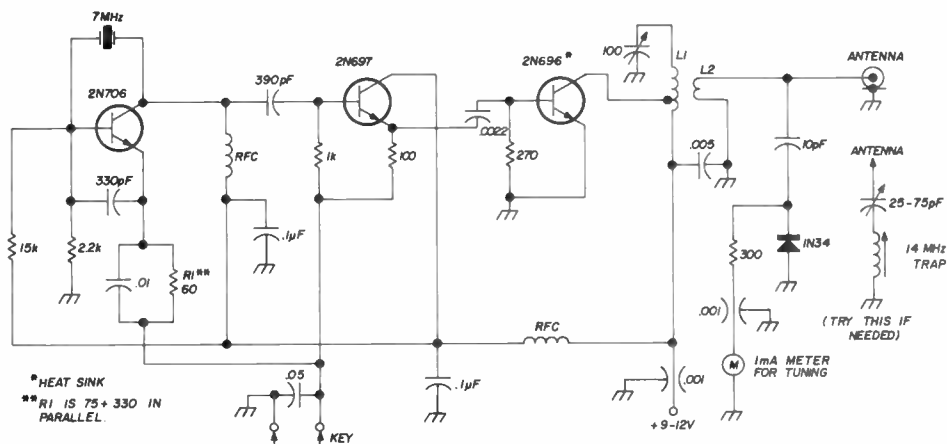


fig. 8. Solid-state CW transmitter built from junkbox parts; it puts out a half-watt with a 12-volt power supply.

22 kilohms to 27 kilohms.)

The unit was connected to a Bird wattmeter and indicated 200 mW at 9 volts and 500 mW at 12 volts, with a total current drain of around 100 mA at 12 V. You'll probably get more output with a 2N697 in the final, because it has more gain than the 2N696. In my case, the 2N697 worked better in the driver stage. This will light a number 47 bulb to about normal brightness.

If the second harmonic is bad, try the trap shown. Dip it with a gdo, or insert a 15-MHz crystal in the unit and adjust the trap for no output, using your receiver. (The trap wasn't installed on my transmitter.)

Bill Eslick, KØVQY

adding incremental tuning to your transceiver

There you are, sitting on 14040 kHz with your transceiver humming. You're waiting for a call from VU2JN, along with six-dozen other guys in the pileup. He's working them 2-kHz down from his transmitting frequency, and your dial hand has muscle fatigue from cranking back and forth — move down 2, give him a quick call, back up 2, listen . . . down again and call, etc. What you need is

receiver incremental tuning; no problem with the little circuit shown here.

What you want is a system where you can set your main tuning dial and leave it; then, by using an independent control, be able to vary operating frequency slightly to either side of the main-dial setting, either while receiving only (RIT), while transmitting only (ITT), or while both transmitting and receiving (IRTT). With these added features you can check the band around your main frequency, and return to the original frequency with the snap of a switch. If you want, you can move your QSO off a busy channel without losing the original channel setting, or listen to two QSOs without touching the main tuning dial. In fact, you can do all the things only the more expensive transceivers allow for.

Let's look at what is involved as far as including incremental tuning in your transceiver. A typical oscillator tank circuit is shown in fig. 9. After L1 and C1 have been set for the desired frequency

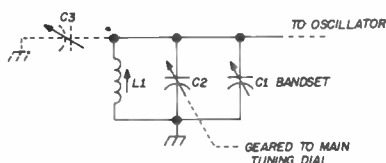


fig. 9. Simple, though impractical, incremental-tuning scheme uses added trimmer capacitor C3 across the vfo tank circuit.

limits of the oscillator, capacitor C2 is used to adjust output frequency. Therefore, C2, which is connected to the main tuning dial, is used to control the receiving and transmitting frequency of the transceiver.

If you put a small variable capacitor (typically 10 pF) across the tank circuit as shown by the dotted lines in fig. 9 (C3), set this additional capacitor at half mesh, and adjust the oscillator for the proper frequency limits, the circuit would operate exactly as before, with one exception: After setting the main tuning dial, the operating frequency could be varied slightly to either side of the main dial setting with C3. When C3 was returned to the half-mesh position, the operating frequency would be restored to the main-dial setting.

Although this is a simple way of obtaining incremental tuning, it is impractical. The same job can be done much more conveniently with a variable-capacitance diode or varactor. The varactor diode is essentially a variable capacitor that is controlled by a dc voltage instead of a rotating mechanical shaft. The capacitance range of these

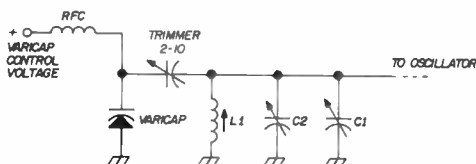


fig. 10. Varactor diode in place of C3 (see fig. 9) provides practical means of obtaining incremental tuning.

diodes, when operated over their rated voltage range, can be greater than 100 pF. If a 100-pF varactor was installed across the oscillator tank circuit, the result would be the same as an extra bandset capacitor. However, the effect on frequency change would be far too great.

This small problem can be solved by placing the varactor in series with a small trimmer capacitor as shown in fig. 10. The trimmer capacitor is then used to control the amount of effect diode capacitance has on oscillator frequency. If you use a 10-pF trimmer, and a 200-pF varactor, the maximum capacitance added to the tank circuit can be no greater than 10 pF.

To incorporate incremental tuning into your own transceiver, locate the bandspread and bandset tuning capacitors. Then install the varactor diode, trimmer

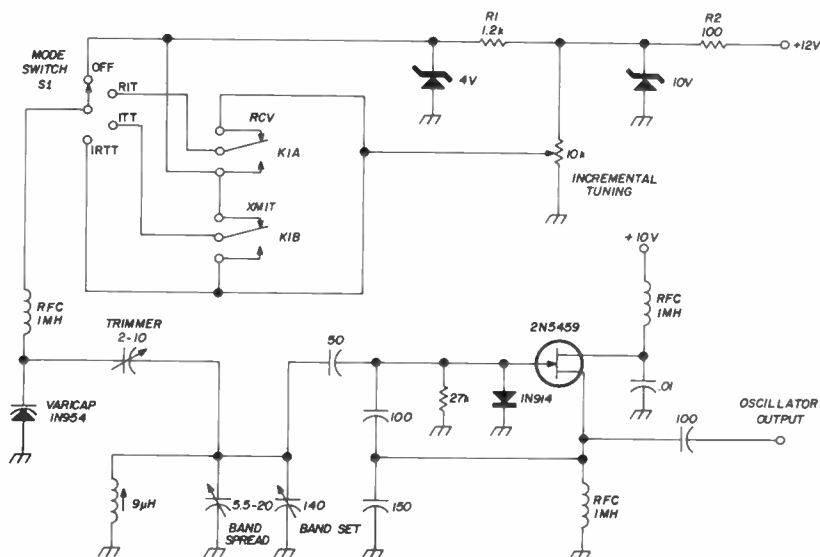


fig. 11. Complete incremental tuning circuit as added to a 5.0- to 5.5-MHz vfo. Resistors R1 and R2 must be calculated for proper voltage drops if B+ is greater than 12 volts. Relay K1 is shown in the receive position; it must be switched with the antenna changeover relay. With S1 in position 1, incremental tuning is off; position 2, receiver incremental tuning; position 3, transmitter incremental tuning; position 4, receiver and transmitter incremental tuning.

capacitor and rf choke as shown in fig. 11. Make sure the new components are mechanically rigid — you don't want to add instability to your transceiver's output frequency.

When soldering the diode into the circuit, protect it with heat sinks on the leads. Also, when installing the diode, make sure the anode of the varactor is grounded; the device must be reverse biased to operate correctly. The cathode end of the diode is usually marked with a white band.

When all the components have been installed and checked, adjust the varactor bias voltage slightly more negative than the center of the recommended operating voltage range. This will lower the oscillator frequency slightly, so you must adjust the bandset capacitor to re-establish main dial calibration. *Do not* touch the oscillator coil.

The varactor's change of capacitance with voltage is not a linear function, so the "mid-point voltage" must be on the low side of the diode's voltage range mid-point, i.e., 4 volts for a zero to 10-volt varactor. Also, since the mid-point voltage of the varactor is used as a reference level for re-calibrating your main tuning dial, the bias source should be zener regulated as shown in fig. 11. This is especially important, because any change in transceiver operating voltages will effect the output frequency.

The complete incremental-tuning circuit, with all regulation and control circuits, is shown in fig. 11. This is the circuit for a 5.0 to 5.5 MHz solid-state oscillator that I used to test this incremental tuning scheme. The potentiometer in the circuit should be a *linear* type to provide a linear change of frequency with rotation; use a high-quality potentiometer to eliminate problems with noise and frequency instability.

alignment

Vary the varactor bias voltage across its full range and see how much frequency variation you get. A total of 10 kHz is ideal. If you get too much variation, decrease the series trimmer capacitance; if

you don't have enough variation, increase trimmer capacitance. Each time you change the trimmer setting you must reset main-dial calibration with the band-set capacitor.

By alternately adjusting the trimmer and bandset capacitors, you can obtain the amount of incremental tuning you want, while retaining main-dial calibration. A digital frequency counter makes this adjustment very easy, but a BC-221 or external calibrated vfo will work as well.

Michael J. Goldstein, VE3GFN

six-meter tunnel-diode phone rig

Those of you who would like to try your hand at flea-power tunnel-diode hamming should get a big kick out of the rig shown in fig. 12. With a tunnel diode, the little transmitter can be built into the tiniest housing, yet still provide adequate rf output to make local contacts.

L1 consists of 4 turns of number-16 copper wire spaced 5/8" and wound 5/8" in diameter. L2 is the coupling to the antenna and can be a 1- or 2-turn link. Use a 26-MHz third-overtone type crystal.

To tune up the flea-power rig, hook a vom across the 150-ohm pot and adjust for minimum resistance. Now apply power to the transmitter and slowly advance the pot until oscillation occurs — at approximately 0.20 volts. At a bias voltage of 0.18, you'll notice a slight voltage upsurge which corresponds

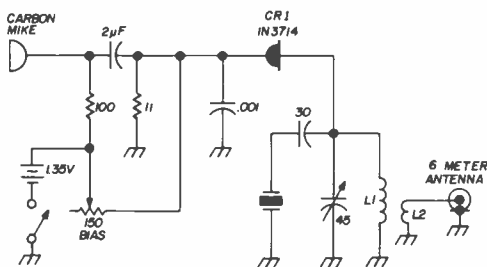


fig. 12. Simple 6-meter tunnel-diode transmitter. Crystal is a 26-MHz overtone type; output can be coupled to any 50-ohm coaxial line. The battery is a hearing-aid type.

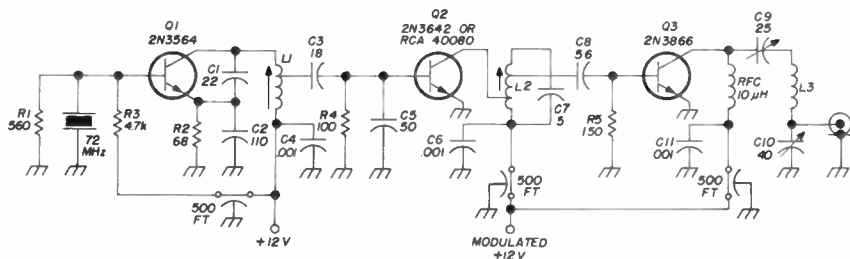
to an audible blast from a monitor receiver tuned to the transmit frequency. Adjust so that the tunnel diode stops breaking in and out of oscillation – at this point place a shaft lock on it. Tune the 45-pF variable for maximum output. Hook on a carbon microphone, and you're ready to give out your first CQ.

Bob Brown, K2ZSQ

from tv sets as well.

The slug for the doubler coil was sawed in half to obtain sharper tuning. If the unit won't load into your antenna, add a 10- to 20-pF capacitor across C10.

Heat sinks are available most anywhere. Mine were salvaged from old IBM circuit boards. Wakefield fin types are fine. Mine is the slip-on type, which is held to



- L1 4 turns no. 20, spaced 1 wire diameter, tapped $\frac{3}{4}$ turn from hot end.
L3 5 turns no. 18, $\frac{1}{2}$ " ID, 9/16" long

- L2 4 turns no. 20, spaced two wire diameters, collector tap $1\frac{1}{2}$ turns from cold end; C8 tap $\frac{3}{4}$ turn from top end

fig. 13. Schematic of the two-meter transmitter. L1 and L2 are modified tv i-f transformer coils. An a-m modulator should provide at least 125 mW of power to adequately modulate this rig.

solid-state two meter transmitter

I wanted a portable 2-meter transmitter, so I built one described in another magazine. I wasn't satisfied with its output nor with the critical tuning to obtain upward modulation, so I decided to "roll my own." The rig shown here is the result.

It produces 250-mW output measured on a Bird wattmeter. Unfortunately, I wasn't able to use inexpensive transistors in all stages. After trying several types in the final amplifier, the only device that provided upward modulation was the 2N3866.

The schematic is shown in fig. 13. I'd suggest staying away from PC boards on this band. I built several 2-meter units on PC boards, even on ground lines. A metal chassis is recommended for this rig.

I made the coils from old i-f transformers salvaged from a tv set. The transformer is the type whose slug has a hex slot (for a tuning tool) instead of a threaded brass screw. Most of the resistors and capacitors can be scrounged

the transistor with three set screws. I once blew the modulator transistors by using them without heat sinks. I replaced these with 2N1374s from some surplus boards from Radio Shack, and they work fine.

A suitable modulator for this transmitter is the Birnback modulator available from Round Hill Associates. Other modulators that provide at least 125 mW of audio power can also be used. On a 6-meter transmitter, I had a hard time keeping rf out of the modulator. A clue to this was a low audio howl from the modulator output transformer, and the transistors ran very hot. Shielding, rf chokes and bypass capacitors were required to tame the six-meter rig.

adjustment

Each stage should be checked with a grid-dip oscillator. The output from the 72-MHz stage and that from the doubler (145 MHz) was sufficient to peg the meter on my Millen gdo when in the diode position.

A resistor connected between B+ and Q3's base will change Q3's operating

angle. For example, a 10k-ohm resistor will make Q3 operate at something less than class C, or close to class AB. I didn't find this necessary for proper operation of the transmitter, however, I mention it for those who might wish to experiment.

With the transmitter in the basement, and a type 49 lamp as a dummy load, I worked as far as two miles fairly well. A whip antenna increased reports. Audio reports were crisp and clear.

One thing that might have caused the modulation problems with different final-amplifier transistors is that the modulator's 500-ohm output impedance is too high. But in homebrewing, you try to use what's available in your junk box.

W. G. Eslick, KØVQY

pi and pi-L networks

These graphs may be used to determine the component values used in pi and pi-L networks. The graphs cover the most generally used operating Qs, load resistances and antenna impedances. To use the charts, it's only necessary to know the plate voltage, peak plate current, desired operating Q and transmission line impedance of your transmitter or amplifier.

using the pi-network charts

To use the pi-network charts shown here, the following steps are taken:

1. Choose the amplifier tube(s) to be used. Select the plate voltage and determine the plate current for normal operation from the data sheet.

Assume, for example, that a pi network is being designed for a pair of 3-400Z tubes operating at a plate potential of 2500 volts and a PEP input of two kilowatts. Peak envelope plate current is determined by:

$$\frac{\text{peak envelope}}{\text{plate current}} = \frac{\text{PEP watts}}{\text{plate voltage}} \quad (1)$$

$$= \frac{2000}{2500} = 0.8 \text{ ampere}$$

2. Determine the approximate resonant load resistance from:

$$R1 = \frac{\text{plate voltage}}{2 \times \text{plate current in amperes}} \quad (2)$$

For the case of the 3-400Zs, the load resistance is: $2500 / (2 \times 0.8) = 1560$ ohms.

3. Choose the operating Q. Good practice calls for a Q between 10 and 20. A Q of 15 is recommended for linear amplifiers.

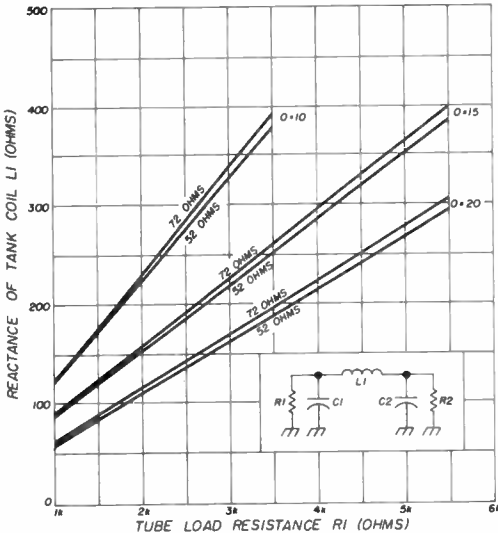


fig. 14. Tank coil reactance as a function of tube load resistance for pi networks. R1 is the tube load resistance and R2 is antenna resistance.

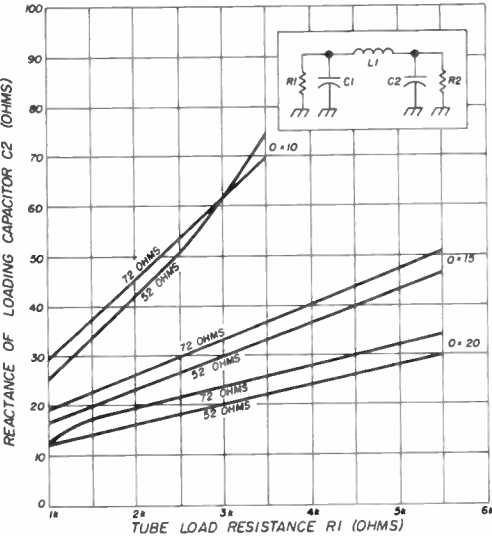


fig. 15. Reactance of the loading capacitor C2 as a function of tube load resistance for pi networks.

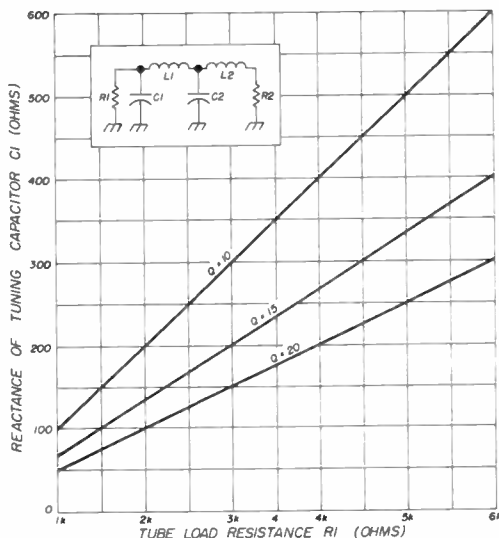


fig. 16. Reactance of the tuning capacitor C_1 as a function of tube load resistance for pi and pi-L networks.

4. Choose the antenna transmission line impedance (R_2). The charts shown here are designed for either 52- or 72-ohm loads because coaxial cables for these impedances are generally available.

5. Find the reactance of the pi-network coil from fig. 14. For the case of two 3-400Zs operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the coil is approximately 120 ohms.

6. Find the reactance of the loading capacitor (C_2) from fig. 15. For the case of 3-400Zs operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the loading capacitor is about 20 ohms.

7. Find the reactance of the tuning capacitor (C_1) from fig. 16. For the case of 3-400Zs operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the tuning capacitor is about 100 ohms.

For two 3-400Z tubes operating at a plate potential of 2500 volts with a peak plate current of 0.8 ampere (two kilowatts PEP) and a Q of 15, the values of the pi network plate circuit are: tuning capacitor (C_1) = 100 ohms; loading capacitor (C_2) = 20 ohms; pi network coil

(L_1) = 120 ohms. As a quick check, note that the sum of the reactances of the two capacitors is equal to the reactance of the inductor.

8. Determine the capacitance and inductance values for the pi network. Fig. 20 and 21 show reactance values of inductors and capacitors commonly used in rf circuits in the hf amateur bands. For the reactances determined for the 3-400Z tubes, the circuit components may easily be determined for each amateur band. In the case of the 20-meter band, for example, the values are: tuning capacitor (C_1) = 100 ohms = 113 pF; loading capacitor (C_2) = 20 ohms = 565 pF; pi network coil (L_1) = 120 ohms = 1.36 μ H.

using the pi-L network charts

Figs. 16, 17, 18 and 19 are used to determine pi-L network components.

1. Choose the amplifier tubes to be used. Select the plate voltage and determine the peak plate current for normal operation as outlined under step 1 for pi networks.

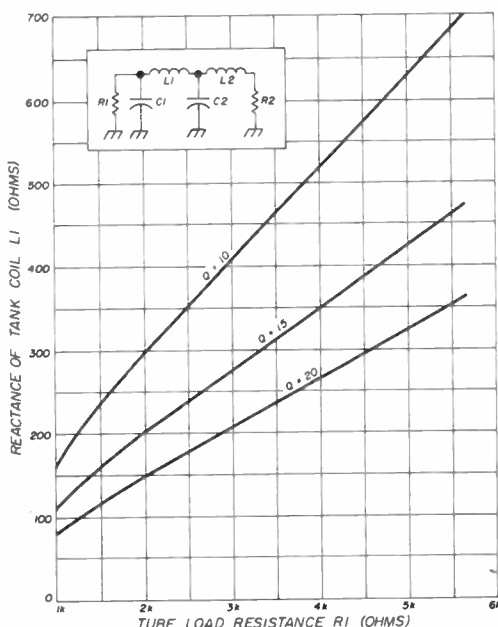


fig. 17. Tank-coil reactance (L_1) as a function of tube load resistance for pi-L networks.

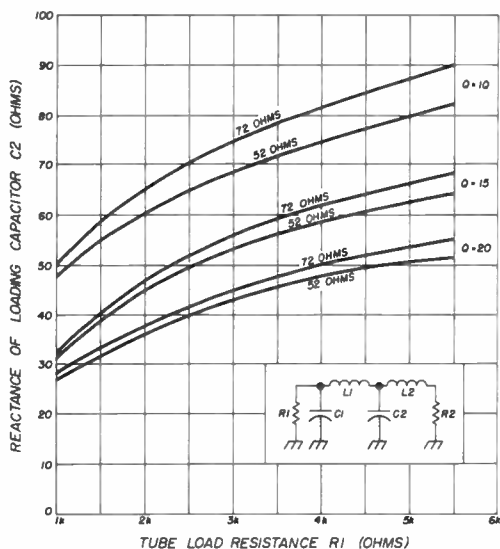


fig. 18. Reactance of the loading capacitor C_2 as a function of tube load resistance for pi-L networks.

Assume for example, that you want to design a pi-L network for a single 3-1000Z operating at a plate potential of 3000 volts and a PEP input of two kilowatts. Peak envelope plate current (eq. 1) is:

Peak envelope plate current = 0.667 ampere

2. Determine the load resistance, as outlined previously in eq. 2: load resistance (R_1) = 2250 ohms.

3. Choose the operating Q (let $Q = 15$).

4. Choose the transmission-line impedance (let $R_2 = 52$ ohms).

5. Find the reactance of the tank coil (L_1) from fig. 17. For the case of the 3-1000Z operating with a load resistance of 2250 ohms, the reactance of the coil is approximately 215 ohms.

6. Find the reactance of the loading capacitor (C_2) from fig. 18. In this case, the reactance is about 47 ohms.

7. Find the reactance of the tuning capacitor (C_1) from fig. 16. In this case, the reactance is about 150 ohms.

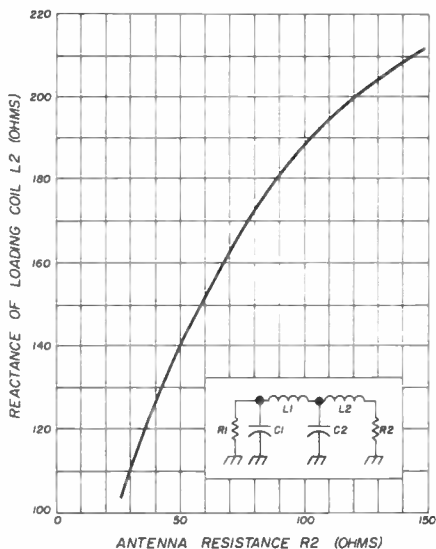


fig. 19. Reactance of the loading coil L_2 as a function of antenna resistance (R_2) for pi-L networks.

8. Find the reactance of the loading coil (L_2) from fig. 19. In this case, the reactance is about 140 ohms.

For a single 3-1000Z operating at a plate potential of 3000 volts with a peak plate current of 0.667 ampere (two kilowatts PEP) and a Q of 15, the value of the pi-L network plate circuit components is: tuning capacitor (C_1) = 150 ohms; loading capacitor (C_2) = 47 ohms; pi network coil (L_1) = 215 ohms; L network coil (L_2) = 150 ohms.

9. Determine the values of the capacitance and inductance for the components of the pi-L network. Fig. 20 and 21 show reactance values for inductors and capacitors in the range commonly used for rf circuitry in the high-frequency amateur bands. For the reactance values determined for the 3-1000Z tube, the circuit components may be easily determined for each amateur band. In the case of the 80-meter band, for example, the values are: tuning capacitor (C_1) = 150 ohms = 275 pF; loading capacitor (C_2) = 47 ohms = 900 pF; pi network coil (L_1) = 215 ohms = 9 μ H; L network coil (L_2) = 140 ohms = 6.5 μ H.

Capacitance values are for resonance

with a nonreactive load. It's suggested that the tuning capacitor have about 50% greater capacitance and the loading capacitor, 100% greater capacitance than indicated.

William I. Orr, W6SAI

10-dB power attenuator

Occasionally a situation arises when it is preferable to provide interstage isolation with an attenuator rather than reduce drive to a following stage in a transmitter. For example, an ssb exciter may operate at maximum linearity at some level, say 100 watts. However, when used as a driving source for a transmitting converter, the exciter may cause distortion by overdriving the converter. Reducing input to the exciter will eliminate distortion in the converter at the expense of reintroducing it in the exciter. The buffer attenuator described in this article will resolve the problem, because it allows both units to operate at points of least distortion.

Also, many amateurs use a simple reflectometer to measure vswr on the antenna transmission line. The reflectometer is reasonably accurate provided the source impedance doesn't change in response to varying load conditions. Most commercially built rigs and not a few homebrew ones are touchy about transmission-line vswr. When you're trying to match an antenna to such transmitters, a tip-off that something evil is at work is a change in apparent vswr when power level is changed. This is what happens when changing from a-m to CW on some rigs. A 10-dB pad between transmitter and reflectometer will correct this situation by stabilizing the transmitter load. In addition, power level is reduced for matching purposes.

The circuit is a symmetrical double network (fig. 22). Its symmetrical configuration allows the attenuator to be inserted into the circuit without regard for *input* or *output* orientation.

A prototype circuit was derived on the basis of a characteristic impedance, Z_0 , of 50 ohms. The nearest values that could be

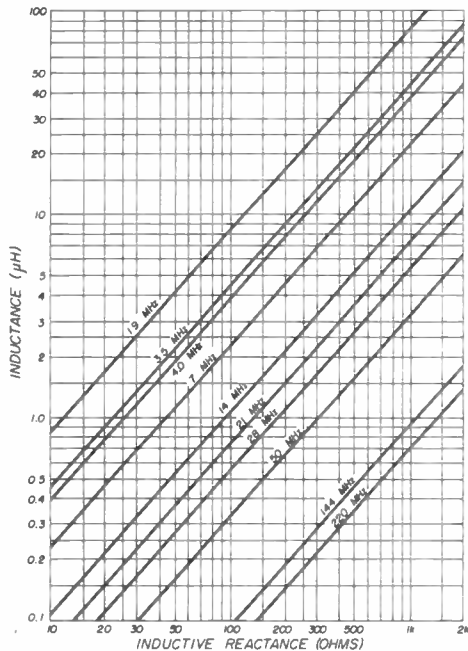


fig. 20. Reactance of inductors commonly used in the amateur bands from 1.9 to 220 MHz.

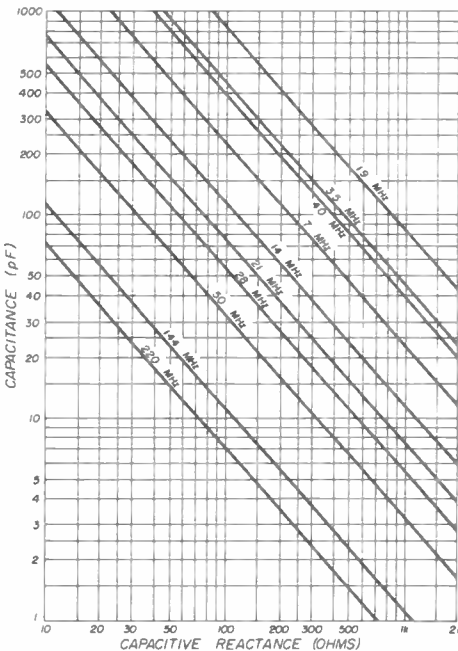


fig. 21. Reactance of capacitors commonly used in the amateur bands from 1.9 to 220 MHz.

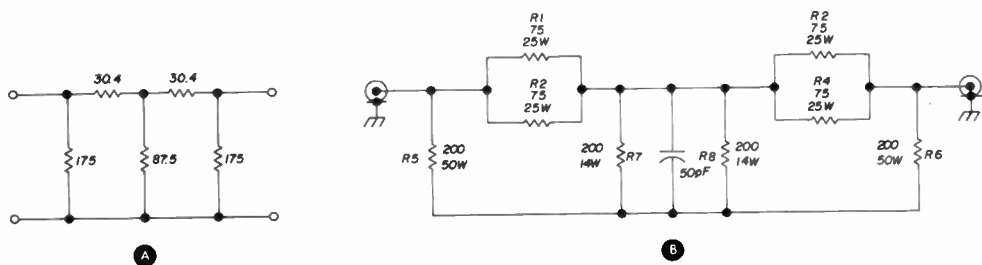


fig. 22. Attenuator prototype (A) was built with ordinary power resistors; final attenuator design in (B) uses non-inductive resistors and has a Z_o of 55.6 ohms. The small capacitor is needed to tune out residual inductive reactance.

R1-R4	75 ohm, 25 W (Sprague 472E7505)
R5, R6	400 ohm, 50 W (Sprague 475E2015)
R7, R8	200 ohm, 14 W (Sprague 459E2015)

obtained from components listed in the catalogs were used. The final version has a Z_o of 55.6 ohms. Resistors were selected so that the attenuator would dissipate 100 watts in either open- or short-circuit condition.

Theoretically, an attenuation of 10 dB in the line will hold vswr to less than 1.25 regardless of the load impedance. This was achieved fairly closely in practice. It should be noted that the design uses resistors with nearest-available values and is *not* intended for precision applications.

The first unit (fig. 22A) used ordinary Sprague Koolohm™ power resistors. The inductive effects were quite noticeable on the higher-frequency bands. However, it was usable as a load buffer on 160 and 80 meters (vswr less than 1.5). Probably the reactance could have been compensated on 40 and 20 meters and higher bands.

Noninductive power resistors were used in the second unit (fig. 22B). This unit had an uncompensated vswr of less than 2.0 through 10 meters. The vswr increased to 2.56 on 6 meters (fig. 23). Although the resistors are rated as non-inductive, there was some residual inductive reactance on 6 meters. A small capacitor was shunted across the network to tune out the inductive reactance. The vswr then decreased to 1.5 on 54 MHz, with no measurable change on 160 and 80 meters. Measured characteristics of the final circuit are presented in table 1.

The attenuator is housed in a Bud CU3013A enclosure; however, any box measuring 1 3/4 x 2 1/4 x 10 inches will be

satisfactory. Coaxial receptacles are type SO-239.

The attenuator is sensitive to the immediate environment and could not be compensated with the cover removed. The unit was compensated originally with a small variable capacitor which was replaced with a silver-mica fixed capacitor when the final capacitance value was determined.

The attenuator was operated continuously with an input of 120 watts for over an hour. It was on a workbench, in the open air, and could be handled without discomfort. The package form factor has a high surface-to-volume ratio, which

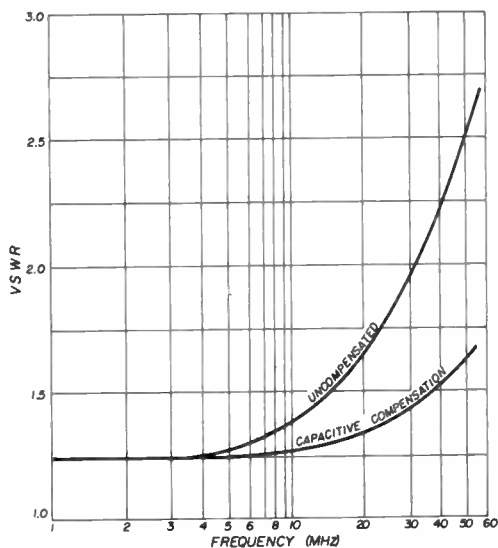


fig. 23. Effect of capacitance compensation on vswr.

table 1. Measured characteristics of the 10 dB attenuator.

Band	frequency (MHz)	vswr	excess phase delay* (deg)	insertion loss plus mismatch (dB)
160	1.8	1.22	10.2	10.55
80	3.6	1.24	11.0	—
40	7	1.27	12.2	10.63
20	14	1.33	14.4	—
15	21	1.35	15.2	10.64
10	29	1.48	19.5	10.72
6	54	1.56	21.9	10.8

*Inductive compensating capacitor could be larger.

helps dissipate heat. However, if the unit is to be enclosed, a heat sink or forced-air cooling should be used.

A note of thanks is due to Bill Cobum, W1ELP, who built the attenuators and adjusted the compensating capacitor. Also, I'd like to acknowledge the help of Lee Tibbert, K1YOZ, who made the insertion loss, phase delay and vswr measurements.

Chet Smith, K1CCL

universal flea-power transmitter

With QRP societies claiming record memberships and ham magazines reporting flea-power DX over increasingly long expanses, many amateurs are tinkering with low-power circuits. The basic problem with most published designs is that the oscillator circuits are crucially dependent upon the L/C factors and related components.

The result is a circuit which will produce outstanding performance on 80 me-

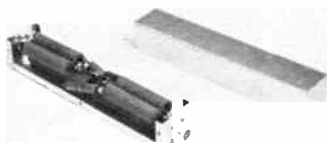


fig. 24. Construction of the 10-dB power attenuator.

ters, but simply refuses to oscillate — without major circuit overhaul — on 160 or 20. The universal circuit illustrated in fig. 25 overcomes this long-standing obstacle to true amateur-band flexibility.

By creating a parallel resonance, fundamental mode, a simple transistor oscillator can be built which will perform reliably over the range from 1 to 21.5 MHz. If you add tuned circuitry at the frequency desired, you can be on any ham band in that range.

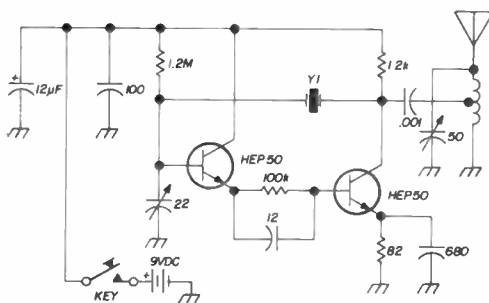


fig. 25. Circuit diagram for a low-power transmitter for 160 through 20 meters.

By using an extremely simple oscillator — with *no* tuned circuits — frequency drift due to aging and external ambient variations has been eliminated. The 22-pF variable stabilizes initial frequency shifts; it's a preset adjustment and doesn't have to be accessible from the front panel.

Similar in design to a Pierce oscillator with an emitter-follower inserted between the base of the second transistor and the crystal, the wideband approach increases input impedance at the crystal input to the transistor. This impedance is normally quite low when compared with crystal characteristics. As frequency increases, the two emitter bias circuits aid in reducing gain attenuation.

The only way this wideband oscillator can be upset from the stability standpoint is if the battery supply voltage fluctuates (you can easily cure any problems here by inserting a zener diode).

Referring to fig. 25, note that the output circuit is tuned to 14 MHz. This is included here only for practical purposes. The tiny transmitter can be put on the air

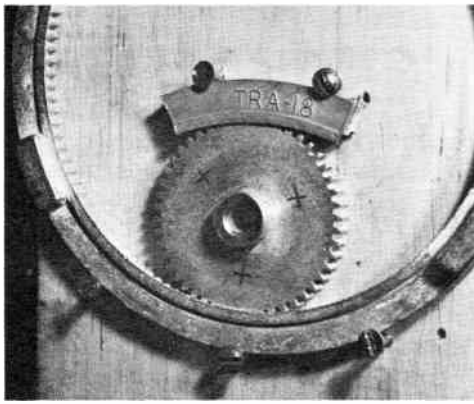


fig. 4. To drill the rivet holes, the TRA-39 drive gear is held down with a TRA-18 ring gear and sawed-off section from a damaged ring gear. The small crosses mark the rivet-hole locations.

happens to be a little larger, it will take all the strain and can wear to the point where the other two will wear unevenly. When all three gears are sufficiently worn, they start to bind. This usually starts at one particular point in rotation, but eventually spreads to the entire 360° and may cause the aluminum ring gear to snap in two.

The cure for this malady is rather simple: rivet the three stamped gears together. This way one gear won't take all the loading, and the three gears will wear evenly. Remove the TRA-18 ring gear and TRA-39 drive gears (use your instruction book for guidance). If your rotator has been binding, order a new TRA-39 drive gear from CDE.* Even a slight bend in these gears will lead to eventual trouble. Also inspect the TRA-18 ring gear for damage. Wash it off in solvent and look carefully for hairline cracks. This is also an opportunity to look over the rest of the parts in the rotator; if any are broken or badly worn, replace them.

Since each of the gears in the TRA-39 assembly was stamped out, each has a slight burr on one side. Do not try to file them flat. When mounting them on the spindle be sure to place them with the burrs toward the bottom; otherwise the gear stack will be too thick.

To drill the rivet holes in the TRA-39 gear, mount the ring gear on a piece of wood with screws as shown in the photo. Mesh the TRA-39 with the ring gear and hold it in place with a section of damaged ring gear. The beveled heads of the wood screws will force the gears together and hold them firmly in place. Drill three equally-spaced holes in the TRA-39 drive gear. If you have a drill press, use it; it will insure that the holes are perpendicular to the face of the gears. Use soft-iron rivets to hold the gears together; rivets 3/8-inch long are just about right. The rivets should be snug in the drill holes for maximum strength.

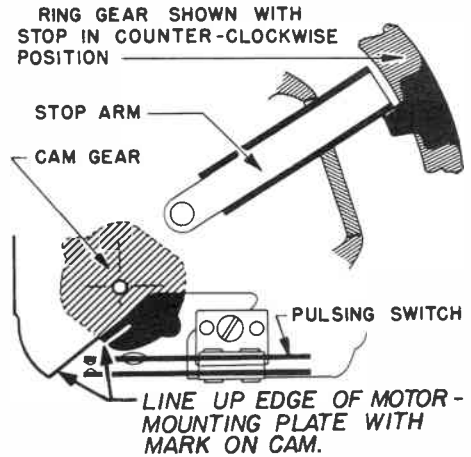


fig. 5. When putting your AR-22 back together, the line on the cam should line up with the edge of the motor-mounting plate as shown here. The ring gear should be against the stop arm in the counter-clockwise position.

Remember when reassembling that the ring gear should be placed in position last. Make sure that the line stamped on the cam gear is parallel with the edge of the motor-mounting plate as shown in fig. 5. In this position the pulsing-switch points should open; rotate the gears until this happens. The stop arm should be pushed to the left (counterclockwise) as far as it will go. In this position the stop lug on the ring gear should be up against the stop arm.

Be sure that all the ball bearings are in

directly from this diagram — or you can adapt the output for any band you want.

Note that the transmitter is dependent on *fundamental-frequency* crystal oscillation. Doubling or tripling action is not recommended, unless a more elaborate output stage is added.

Robert M. Brown, K2ZSQ

television interference: an effective remedy

I've been an active ham for 36 years. During this time I've built many low-power rigs. I guess I enjoy working *on* a transmitter as much as working *with* it. I like all bands, so I'm partial to a long-wire antenna. Recently I began using an indoor wire. This antenna provides excellent DX reports, but it doesn't discriminate against TVI.

A few months ago I retired from homebrew rigs and purchased a Swan 350C, a transceiver with many good features and a lot of power for the money. I soon discovered that when operating on 20 meters, my favorite band, the Swan caused interference with my wife's tv programs, particularly on channel 2. (She watches a portable in the next room, about 12 feet from the rig and antenna.) On the other hand, 15 meters was perfectly clean on all low tv channels, but TVI showed up to a lesser degree on channel 9 and higher.

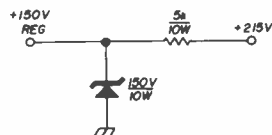
After discussing the problem with fellow amateurs, I realized that something had to give, and it turned out to be me. I would have to sacrifice some power. Fortunately this was not difficult to accomplish, and the difference in signal reports was insignificant.

the cure

It's quite possible that the modifications I made will apply to equipment other than Swan. Here's what I did, in the order of effectiveness, to cure the TVI problem:

1. Input to the driver stage, a 6GK6, was reduced, which decreased harmonic output.

fig. 26. Zener diode added to the Swan 215-volt power supply for final and driver screens to reduce input power for TVI reduction.



2. Input to the final, which uses a pair of 6LQ6/6JE6 tubes, was reduced from 360 to 200 watts.*

3. Capacitors (1 kV disc type, 0.001 μ F) were placed across filament, high-voltage, medium voltage and bias terminals — right at the Jones plug. (These are leads between the transceiver and power supply.) Also, I shortened the unshielded cable between the rig and power supply, which reduced radiation from the cable.

The Swan has three supplies: 850, 275 and 215 volts. This last voltage is obtained through a dropping resistor. The Swan schematic shows that the 850 volts is the sum of two supplies in series: 275 and 575 volts. I separated these supplies and used them independently. The 575 volts was applied to the final plates (instead of 850 volts); then I added a 150-volt zener, fed from the 215-volt supply (fig. 26). The 150-volt supply is now used for the final and driver screens. The "new" and "old" voltages are listed in table 2.

Lest these changes seem drastic, I'll reference an article appearing in *Sylvania News*. This article discusses several tubes, including the 6JE6. The following optimum values are suggested for the 6JE6:

plate voltage	500V
screen voltage	125V
bias voltage	- 85V
plate current	222mA

These values are recommended in the *Sylvania* article for class-C CW operation. For class AB1, the same maximum plate

*This is a decrease of about 2.5 dB, which is barely perceptible under *ideal* conditions. It's unlikely that the receiving station could tell the difference between 360 and 200 watts under actual band conditions. *editor*.

table 2. Voltages for the Swan 350C before and after the TVI modifications.

	driver		driver	
	plate	screen	plate	screen
new	270	150	525	150
old	270	270	850	210

and screen voltages are recommended, plus

idle plate current 40 mA
bias voltage - 44 Vdc

Note that these are still slightly below those for the "new" values given in table 2.

When the changes described above were incorporated, TVI was greatly reduced. I can now monitor a portable tv receiver with its single collapsible rod antenna about two feet from my indoor transmitting antenna. On channel 2, barely visible lines appear; but previously the picture was completely wiped out! With the tv set in the next room, no lines are visible at all. It's almost impossible to tell when I'm on the air. The loss in transmitter power is less than 2:1, so the sacrifice is small.

If you wish to reduce transmitter power still further, disconnect the input capacitor to the filter in the 275-volt supply. The voltage will be decreased to about 200 volts. Input to the driver will be much lower, reducing harmonic output still further; and the dc input to the final will be about 125 watts. The lower voltage will also be applied to the tubes in the receiver section; however, I didn't notice much loss in sensitivity. If you have a really severe TVI problem, this certainly should do the trick.

I. Queen, W2OUX

grid-current measurement in grounded-grid amplifiers

Measuring the grid current of a cathode-driven amplifier can be a delicate and exasperating task; it's a ticklish job to

"unground" the grid sufficiently to use a metering circuit and still hold the grid at rf ground. The inherent inductance of most bypass capacitors lets the grid circuit float above ground at some high frequency, and as a result, the amplifier exhibits instability and parasitics.

This problem can be avoided with the measuring circuit shown in fig. 27. The control grid is grounded through a 1-ohm composition resistor that is bypassed with a .01- μ F disc capacitor. The voltage drop generated by the flow of grid current through the resistor can be measured easily with a millivoltmeter which is calibrated in terms of grid current. Individual grid current for each of a parallel pair of tubes may be measured with the circuit of fig. 28. The maximum current which can be measured is determined by the internal resistance of the milliammeter plus the series resistance R1.

Suppose you want to read grid current on the order of 150 milliamperes; the meter should read 0-200 milliamperes. Since the original markings on the 0-1 mA meter scale can easily be multiplied by 200 to obtain the actual value of current, this is very convenient. Now, when 200 milliamperes flow through 1 ohm, a potential of 0.2 volt is developed across the resistor. Therefore, the meter should read 0.2 volt full scale to correspond to 200 mA of grid current.

Assume that the meter is a Triplet model 221-T, which has an internal resistance of 55 ohms. The voltage drop across the meter itself is 0.055 volts when one milliampere flows through it. To convert

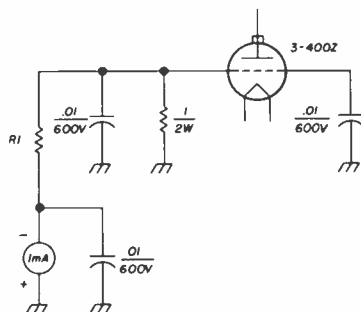


fig. 27. Measuring the grid current of a single grounded-grid stage.

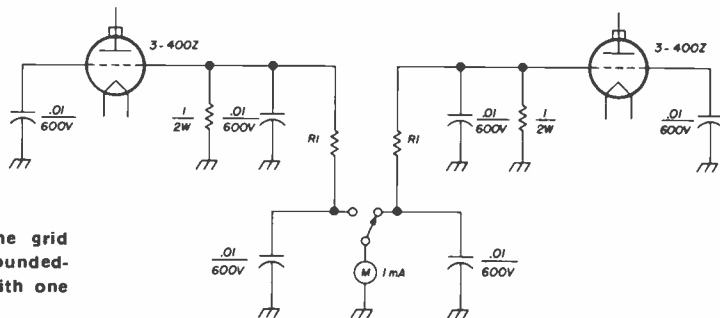


fig. 28. Measuring the grid current of a pair of grounded-grid amplifier tubes with one meter.

the milliamperes to a voltmeter reading 0.2 volt full scale, you must add a series multiplier. A voltage drop of 0.2 volt is developed across a 200-ohm resistor when one milliamperes of current flows through it. Therefore, the *difference* between 200 ohms and 55 ohms, 145 ohms, must be added in series with the meter to convert it to read 0.2 volt full scale.

On the other hand, if you put the meter across the 1-ohm grid resistor without the series multiplier, it will provide a full-scale reading corresponding to 55 milliamperes. If the maximum grid current is below 55 mA, no series resistor is required. Conversely, high values of grid current will produce a greater voltage drop across the 1-ohm resistor and larger values of series-multiplier resistance are needed.

William I. Orr, W6SAI

blower maintenance

Many items of electronic equipment have blowers that move air across heat-generating components. The squirrel-cage blower seems to be popular in transmitters and transceivers, probably because these blowers are relatively quiet. Over a period of time, however, dirt builds up on the impeller in squirrel-cage blowers. Blower efficiency can gradually deteriorate without your being aware of it if dirt is allowed to accumulate. The result can be disastrous, especially if the blower is used to cool high-power transmitting tubes.

One remedy is to inspect the blower periodically and remove the impeller to

give it a good cleaning. This usually involves considerable work, especially if the blower is located in an inaccessible place. Here's an effective preventive maintenance measure that not only keeps dirt buildup to a minimum, but eliminates the chore of removing the blower impeller for cleaning.

The intake port for the blower on my transceiver, a National NCX1000, is on the rear of the cabinet and flush with the vertical surface of the rear wall. After several hours of operation, I noticed that the blower impeller vanes were loaded with dirt. I used an old toothbrush to remove most of the grime, but it was necessary to remove the impeller from the blower to do a good cleaning job.

The answer to the problem consisted of a filter made from a piece of polyurethane foam about ¼-inch thick, which I cut to cover the blower intake port. Poly foam is available at retail sources that supply material for do-it-yourself furniture makers. It's used for filling seat cushions, and is available in several thicknesses. I secured the filter over the blower intake port with masking tape so it could be removed easily for cleaning.

Some types of poly foam have a more dense structure than others, so it's a good idea to test the filter before taping it into place. If the blower speed decreases appreciably, the foam is either too dense or too thick. A little experimentation will produce the right combination for effective filtering consistent with maximum blower efficiency.

I checked my blower and filter after about 90 hours of operation. The outside

of the filter was coated with a rather thick deposit of dirt, but the impeller and blower housing were spotless. It was a simple job to remove the foam filter and wash it in warm water and detergent. The entire operation took about ten minutes, and the filter can be used indefinitely.

Alf Wilson, W6NIF

32S-1 cw modification

The Collins 32S-1 as originally designed uses a 2.1-kHz-bandpass mechanical filter centered on 455 kHz and two bfo crystals, one above 455 by 1350 Hz and one 1350 Hz below, operating on 456.350 and 453.650 kHz respectively.

In CW operation, a tone oscillator at 1350 Hz is fed through the mike amplifier, balanced modulator and mechanical filter to create a CW carrier. The carrier is at 455 kHz in the low i-f by virtue of the relationship of the bfo crystal frequency (456-350 kHz) and the modulating frequency of the tone oscillator (1350 Hz) so that $456.350 - 1.350 = 455$ kHz.

The quality and stability of the diodes available at the time the 32S-1 was designed and manufactured was far below those available today. It is difficult to maintain carrier suppression better than 50 dB over a long length of time with the original components and circuitry. However, a very easy alternate method of vastly increasing carrier suppression is possible.

A 457.55 kHz crystal* is stocked for those amateurs who operate RTTY using a 2125/2975 tone keyer; this places the audio response between 1500 and 3600 Hz. Of specific importance is the fact that the crystal is removed from the skirt of the mechanical-filter bandpass by 1500 Hz rather than the normal 300 Hz, so an additional 40 to 50 dB carrier suppression is obtained. The 1350 Hz tone oscillator no longer puts the "carrier" at 455 kHz,

and its second harmonic could pass through the mechanical filter with little attenuation. Therefore the tone frequency must be raised to, ideally, 2550 Hz ($457.55 - 2.550 = 455$ kHz).

Refer to the 32S-1 schematic diagram. Locate switch S8F, just below V2B, the 6U8A bfo tube, and to the right of the tone oscillator. The upper sideband crystal, Y15, is shown wired to the most clockwise position (CW); a jumper goes to the adjacent contact, which is the *usb* wiper, and to the *tune* and *lock-key* wipers. The jumper should be removed so that Y15 is connected to the *usb* wiper only. The jumper from the *tune* and *lock-key* wipers should be connected to the CW wiper. Do not solder yet, as the new 457.55 kHz crystal should be connected between the CW wiper and pin 3 of V2B (the pin where the other crystals are connected).

To raise the frequency of the tone oscillator, C111 should be changed from its original value of 430 pF to 220 pF, and C110 should be reduced from 390 pF to 180 pF. The exact frequency of 2550 Hz might not be realized but the resultant frequency is sufficiently close to meet requirements.

Those amateurs who have operated modified 32S-1s report that they don't object to the higher pitched sidetone, although many use the sidetone from their electronic keyer so the frequency difference is not a consideration. After this modification, carrier suppression should be on the order of 80 dB. This is sufficiently below the white noise output to be undetectable. On-the-air reports have been excellent.

Jim Fisk, W1DTY

elmac chirp and drift

Even though the Elmac AF-67 is fifteen years old, it still has many fine attributes. In fact, I just bought one. It worked fine on a-m, and it's tailored speech characteristics are as desirable today as when it originally was designed. However, in its original form it had two serious flaws that limited it's usefulness: a

*The 457.55 kHz crystal is available from the Collins Radio Company, Cedar Rapids, Iowa 52406. Order part number 290-8709-00; \$5.50 plus sales tax, postage and insurance. The price is subject to change at any time without prior notice.

chirpy CW signal and unacceptable drift on 15 and 20 meters.

The chirp was cured in my unit by regulating the supply to the buffer/multiplier stages. A 420-ohm dropping resistor and a 10k bleeder along with an OB2 regulator in series with an OA3 provide the proper operating voltages (see fig. 29). Keying with the new regulated supply is clean and chirpless.

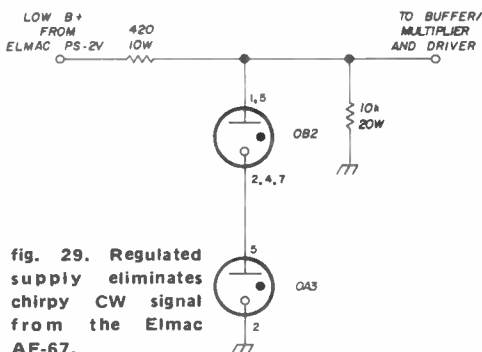


fig. 29. Regulated supply eliminates chirpy CW signal from the Elmac AF-67.

Frequency drift was noticed only on 15 and 20 meters. A 35-pF variable capacitor, C15, is unique on these bands, so it was suspect. Sure enough, when it was replaced with an ordinary 35-pF trimmer the drift disappeared. I mounted the trimmer on the bottom plate of the vfo module because the original position was almost inaccessible.

I carefully examined the faulty capacitor and found that only one leaf of the rotor shaft contact spring was soldered in place — the other leaf maintained electrical contact through pressure. I suspect that some type of diode action was occurring between the pressure spring and the mounting nut, but because of the inaccessibility of the original mounting position, I didn't put the unit back in the transmitter to run further tests.

George Hirshfield, W5OZF

spot switch for the heathkit sixer or twoer

A momentary pushbutton switch and a one-watt, 47k ohm resistor are all that are needed to put a spot switch on the Heathkit Sixer or Twoer.

The resistor is in series with the switch, and they are mounted across the two middle terminals of the terminal strip behind the neon lights. Fig. 30 shows the connections.

This puts about 50 volts on the B+ line to the driver stages of the transmitter when in the receive position. This is enough to get a carrier indication. The final doesn't draw current, because the cathode is open until the transmit switch is operated.

The resistor drops 210 volts at a measured 5 mA. The dissipation is one watt. Because it's not on for any length of time, I used a one-watt resistor without allowing for safety factor.

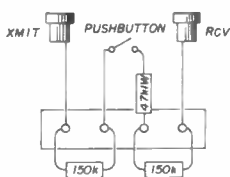
Edmond DeCloedt, WA6FNR

incremental tuning for the heath HW-100

The addition of incremental tuning to a ssb transceiver is an ideal and often needed refinement. It allows the receiver to be shifted a few kHz above or below the transmitting frequency. This helps when someone has trouble zero-beating your frequency or when you're in a roundtable where someone is inevitably off frequency. It can eliminate the embarrassing problem of tuning for the off-frequency station and then suddenly finding yourself equally off zero beat. It is especially handy where there may be a highpitched voice — you can tune for personal listening comfort without affecting your own transmit frequency.

In the receive position, with the incremental tuning switch on, varying the dc bias on the varicap diode causes the capacitance to change within the range of the incremental tuning control. With the transceiver in the tune position (or trans-

fig. 30. Spot switch for the Heath Sixer or Twoer. Pushbutton switch and 47k resistor are mounted across two middle terminals of strip behind neon lights.



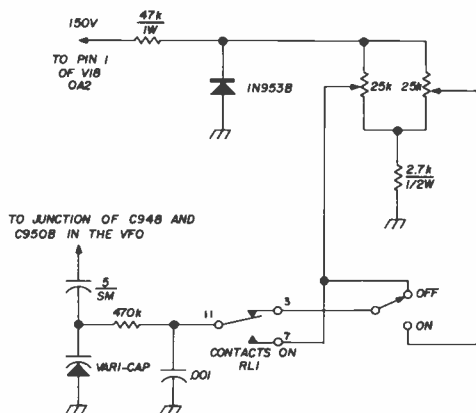


fig. 31. Receiver incremental tuning for the Heath HW-100. The two 25k pots should have linear tapers.

mitting), the antenna changeover relay (through contacts 3, 7 and 11) disconnects the manual tuning voltage and inserts a calibrate voltage in its place. This sets the capacitance of the varicap at the center frequency position of the incremental tuning control. A shift of 2 kHz either side of center can be obtained with the values shown in the schematic in fig. 31.

The bias voltage is derived from the +150 V regulated supply. The voltage is dropped to +12 V and regulated by the 47k, 1-watt resistor and a 1N963B zener diode; this keeps the voltage steady on both transmit and receive. The photo shows how the components may be mounted in an already well packed area.

The calibrate pot is mounted just to the side of the final tank cage in an existing 5/16-inch hole. (This hole will have to be enlarged to 3/8-inch to accommodate most pots.) Mount a 2-lug terminal strip (Cinch-Jones 52) on the underside of the chassis near the pot. A 47k, 1-watt resistor is placed between pin 1 of the OA2 (V18) to the closest tie point with the zener diode going from the same tie point to ground. The incremental tuning pot on the front panel is mounted to the right of the meter at the same height as the tuning controls to maintain a neat panel layout. You can get an extra knob from Heath to match the original equipment (part number 462-191.

\$.70). A sub-miniature spdt switch is mounted above the pot. Decal numbers may be added to indicate 1- and 2-kHz deviations from the center.

A five-conductor color-coded cable can be made up to wire between the chassis and the front panel controls. This cable can be passed through an existing 1/4 inch hole in the i-f circuit board.

Mount a one-terminal Cinch-Jones 51A terminal strip to the left of the backlash gear on the vfo capacitor; straighten the ground lug. Mount the varicap from the ungrounded lug to ground and mount the 5-pF mica capacitor from the same ungrounded lug to the junction of C948 and C950B.

Mount a two-terminal Cinch-Jones 52A terminal strip just in back of the vfo capacitor on the stud holding the vfo coil. Mount the 470k, 1/2-watt resistor between

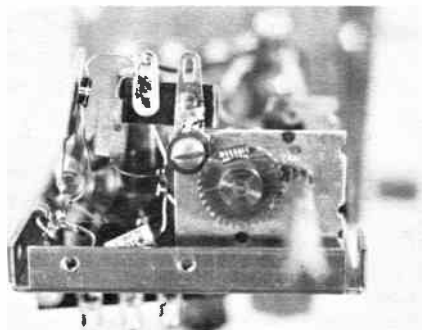


fig. 32. Varactor installation in the HW-100.

ungrounded terminals, bypassing one end of the resistor with a .001 disc ceramic to ground and connecting a wire between the other end of the resistor and the terminal strip mounted on the vfo. Leave enough slack in the rigid wire to clear the rotor. Connect a wire to the bypassed end of the 470k resistor and pass it through a hole in the top of the vfo chassis to a Cinch-Jones 53 terminal strip mounted on top.

To calibrate the two pots for the same receive and transmit frequency, set the front panel pot (number 1) to its center position. Set the number 2 pot (mounted near the final tank cage) to its approxi-

Bill McCracken, K1GUU

improving alc response in the SB-400 and SB-401

This problem is caused by a voltage spike passed to the alc network when switching out of the spot mode. The final amplifier bias changes suddenly from -100 V to -50 V, which creates a 50-volt pulse. This pulse must bleed off the alc line before the transmitter can operate.

I solved this problem by installing a silicon diode as shown in fig. 33. However *a word of caution is in order*. Adding the diode places it in series with a 4.7 megohm resistor in the voltage divider on

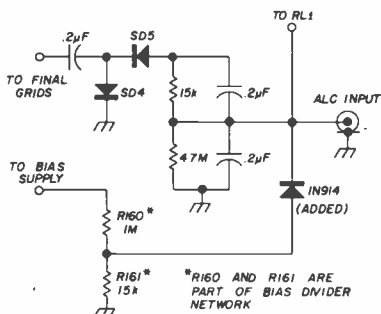


fig. 33. Adding a diode to improve alc response in the SB-400/401. The added diode should have near infinite back resistance.

For example, assuming a back resistance of 47 megohms in the diode, the 3V bias on this diode will produce 0.3 V on the alc line. Of course, a lower diode back resistance will put a higher permanent voltage on the alc line. So make sure the diode measures near infinity in the backward direction!

With a good diode, alc voltage will be 3 V. This doesn't interfere with alc operation in my particular rig, but I can't guarantee it won't create a problem in other transmitters.

David Wojcinski, WA9FDQ

crystal control for the HW-100

Crystal control is an option not offered by most transceiver manufacturers. Granted, most vfos are stable and accurate enough not to require the steadiness of a rock, but there are times when crystal control is a definite advantage. Mobile operation is enjoyed by a lot of amateurs, and here a stable, known frequency is most welcome. Many of my friends operate on a single frequency most of the time; there's someone monitoring the frequency nearly 24 hours a day, and this can be reassuring to a lone mobileer. Crystal control is also useful for net operation.

The crystal oscillator shown in fig. 34 provides more drive than necessary and must be padded down with the 45-pF trimmer. The circuit is very inexpensive to build — less than the cost of the crystal itself. A positive supply voltage is picked off the zener diode that biases the vfo amplifier tube. With the switching circuit shown, it is possible to select: vfo-controlled transmit *and* receiver; crystal-controlled transmit and vfo receiver; or, crystal-controlled transmit *and* receive.

When ordering the crystal for this circuit, specify that it will be used with a 32-pF load. The vfo tunes from 5.5 to 5.0

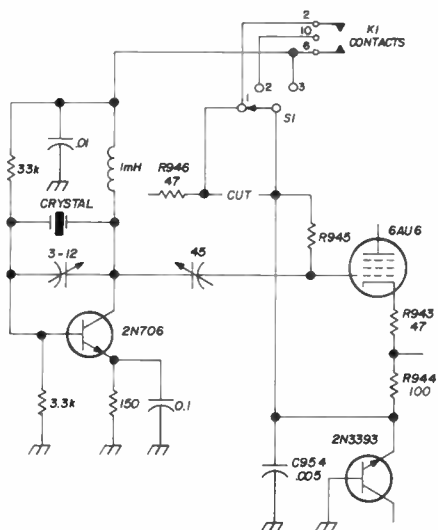


fig. 34. Crystal control for the Heathkit HW-100. In switch position 1, the vfo controls both transmit and receive; position 2, crystal control on transmit, vfo control on receive; position 3, crystal control on both transmit and receive.

MHz (low to high end of the band). Therefore, to operate crystal controlled on 3999 kHz for example, you need a crystal cut for 5001 kHz. The 12-pF trimmer is used to put the crystal exactly on frequency. If you plan to operate near a band edge, it would probably be a good idea to spend the extra money for a commercial-grade crystal.

The crystal oscillator is built up on a terminal-strip board and bolted to the back of the vfo chassis. An insulating strip is placed under the board to keep it insulated from the chassis. Drill a hole beside the 6AU6 to accommodate an insulated feedthrough bushing: the 45-pF trimmer is mounted between the circuit board and the feedthrough insulator. Additional holes are drilled through the top of the vfo chassis to accommodate the wires needed to operate the crystal circuit and incremental tuning unit.

The 47-ohm resistor is disconnected from terminal strip FF (pictorial 7-2 in the HW-100 manual). A wire is soldered to this resistor and passed through the top of the chassis to terminal 2 of the rotary vfo/crystal switch; a wire soldered

to terminal 1 of terminal strip FF is connected to the common terminal of the vfo/crystal switch; a wire coming from terminal 3 of the switch is connected to the B+ tie point of the crystal oscillator. Three more wires are connected to the switch as shown in the schematic and connected to terminals on relay K1.

Bill McCracken, K1GUU

loose HW-100 tuning knob

Many builders find that the main tuning knob on the HW-100 is loose even though the dial mechanism is properly assembled and working well. The remedy is very simple.

Remove the tuning knob and place a washer on the end surface of the flexible spline. The washer must have a hole large enough to pass the collar on the spline on the vfo shaft (see fig. 35); otherwise the knob will not fit. The added washer fills up the space between the spline and the inside of the knob, applying an outward force on the knob that keeps it from wobbling. With just the right washer thickness, and an even coating of silicone grease on both sides, the action is "silky" smooth. If the washer is too thick, the knob will either not go on, or tuning will be very stiff. Several very thin washers may be needed to get the right feel.

Al Lightstone, VE3EPY

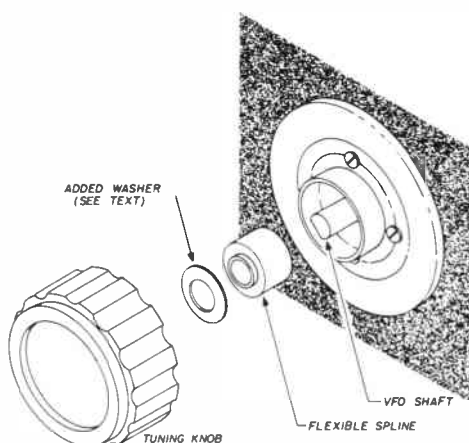


fig. 35. Simple remedy for a loose tuning knob on the Heath HW-100.

vhf and uhf

ssb input source for vhf, uhf transverters

Many amateurs require a good vfo for use in the vhf/uhf bands. The 28-MHz output of an ssb transceiver or transmitter can be used as an input source for an up-converter to obtain the desired vhf/uhf frequency. It's highly desirable to maintain the original modulation system of the transceiver or transmitter.

For output on 144 MHz, you could use 166 and 28 MHz; for 432-MHz output, the conversion can be made by mixing 404 and 28 MHz. The best method to obtain output on 1296 is to use 28 + 518 MHz, then mix the 546-MHz resultant signal with 750 MHz. This method will avoid all frequency components except the desired one.

It's possible to obtain the 28-MHz source from the transmitter or transceiver by (a) running the output into a dummy load and feeding a certain amount of energy through a small capacitor to the transverter, or (b) feeding the output to the transverter through an attenuator. These methods are inefficient and wasteful of power. Therefore, I've taken a new approach to the problem.

I use a type 6HM5/6HA5 tube in a cathode follower (fig. 1). This tube is very small, like the 6AK5. The circuit is installed on a little subchassis, which fits nicely in the 6JB6 input compartment of my TR4. The output BNC connector is at the rear, close to the 6JB6 output compartment, and is mounted with the switch on a small panel. When the switch is on, the 6JB6s are disabled, and no power is wasted.

The output impedance is given by $Z_{out} = 1/S = 1/0.02 = 50$ ohms, where S is the tube transconductance. With 135 V on the plate, this tube has a transconductance of 20k μ mhos. This results in a normalized value of 50 ohms output impedance.

Because the input impedance of a cathode follower is high, only a very small amount of capacitance is needed to feed the 28-MHz signal from the driver stage, and no misalignment will occur.

Using this method I can obtain 1-10 volts or CW or ssb signal; it also works well on a-m. The stability of the vfo is the same as in the hf bands. The overall stability is determined by the crystal oscillator. I've been using this addition to my TR4 for two years on 144 and 432 MHz with very good reports. The idea can be used in other ssb equipment with equal results.

Jacques Mainardi, F8MK

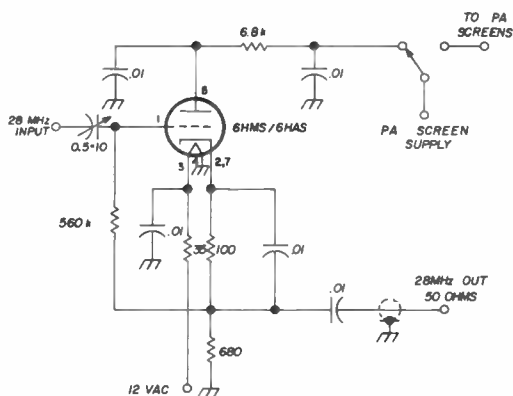
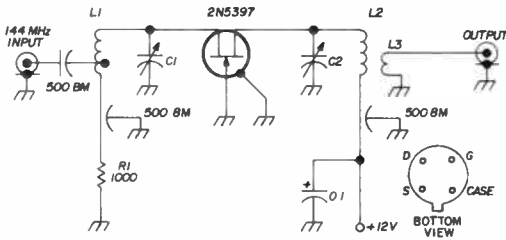


fig. 1. Cathode follower for efficient power transfer from transceiver to transverter.

super two-meter preamp

The new Siliconix 2N5397 fet makes noise figures of 1.5 to 2 dB commonplace at 144 MHz. Although the 2N5397 is priced at \$8.50 in small quantities, its performance and ease of operation make it well worth it. The 2N5397 was designed for vhf use, and it works well beyond 500 MHz, but that's another story. I was mainly interested in 144 MHz



- C1, C2 1.7-14.1 pF air variable (E. F. Johnson 189-505-5)
 L1 5 turns no. 16, 3/8" diameter, 3/4" long, tapped 1 1/2 turns from cold end
 L2 4 3/4 turns no. 16, 3/8" diameter, 1/2" long
 L3 2 turns no. 20 insulated around cold end of L2

fig. 2. High performance grounded gate fet preamplifier for 144 MHz. The 2N5397 is mounted in an IERC mounting clip type TXBE-019-021B or TXB2P-019-028B. Case and gate leads of the fet are soldered together.

operation because a 1.5 dB noise figure is needed for moonbounce.

At the suggestion of the designer of the device, I tried it in a grounded gate configuration (fig. 2). Much to my delight, the circuit performed better than anything I have ever had the opportunity to use, or measure. One glance at the schematic will show you that it is as "super simple" as you can get.

Since I already have a high-performance two-meter converter, the 2N5397 preamplifier was designed to mount as close to the antenna relay as possible. This is the reason for the type-N input connector.

I built my preamp on a 2 x 2-5/8-inch piece of copperclad printed-circuit board

with a 1 3/4 x 1 1/2-inch shield, and then fitted it into a 2 3/4 x 2 1/2 x 1-5/8-inch minibox. I prefer this type of construction because it provides good grounds (see fig. 3).

It's my opinion that the fet mounting is important to obtain optimum performance. IERC makes a TO-18 heat sink that can be soldered directly to the chassis; this is an ideal way to ground the fet case. The gate lead is soldered to the case lead to provide a good ground for the gate.

Tuneup is very simple. First apply +12 volts, and adjust R1 for 5 mA drain current. Next, feed in a 144-MHz signal and tune C1 and C2 for maximum output. Carefully adjust the position of L3 for maximum output, and readjust C2.

The gain of this fet preamplifier should be greater than 10 dB, and more likely, 15 dB. If you have a way of measuring noise figure, adjust the tap on L1 for best noise figure. If you don't have any noise-figure equipment, the tap shown in fig. 2 will give good performance.

Ken Holladay, K6HCP

simple 2-meter antenna

There are many commercial two-meter antennas available for home use, but fig. 4 shows a simple one you can put

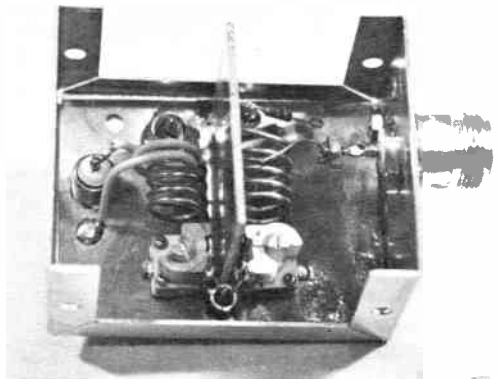


fig. 3. Two-meter preamp is built on a piece of copper-clad board and mounted in minibox.

place in the retainer spring clips. Thoroughly grease the gears with a good silicone grease such as Dow-Corning 44 or Lubriplate, a white lubricant available in many hardware stores; an 8-ounce tube is sufficient. Also, when replacing the top cover of the rotator, make sure the lugs in the ring gear rest in the recessed sockets provided inside. When the rotator is assembled in this manner it is in the *North* position, against the stops from a counter-clockwise direction through *East*. The control box should be oriented to coincide with this setting.

Ted Woolner, WA1ABP

*Send all parts orders to Cornell-Dubilier Electronics, Rotor Parts Department, Desplaines, Illinois 60018, *not* to their factory in Fuquay Springs, North Carolina.

80-meter beam antenna

Just "getting out" on 80 meters calls for nothing too elaborate in the way of antennas. But to *really* get out, to the tune of working a world-wide DX as a matter of course, calls for something a bit special. The antenna described here is the result of a fair amount of experimentation and construction on the part of Bill Shearman, VE1AX, whose DX countries total on 80-meter ssb is well over the magical 100 mark.

To say it's a two-element beam conjures up ghastly thoughts of huge towers and monstrous drooping elements, but this beam is vertical, which makes all the difference in the world. For one thing, it's unobtrusive. It sits right in the middle of a subdivision and is not particularly noticeable. Being a quarter-wave vertical, the elements are only about 60 feet high. Although this may seem a lot, by using small-diameter masts the problems of height and weight become quite reasonable.

elements

The antenna at VE1AX consists of two identical 48-foot aluminum masts topped with a 12-foot fiberglass whip of

the type used for marine transmitters. Any suitable combination of aluminum or steel tubing, tv masts, etc. could be used; but both verticals must be identical. The masts must be insulated from ground, which Bill did by using commercially built tripod mounts. But here again, traditional ham ingenuity should come to the rescue with such alternates as wooden posts, power line standoff insulators, or even Coke bottles, as I remember seeing in another article.

broadbanding

To make the antenna broadband, a three-wire cage encloses each mast. Fig. 7

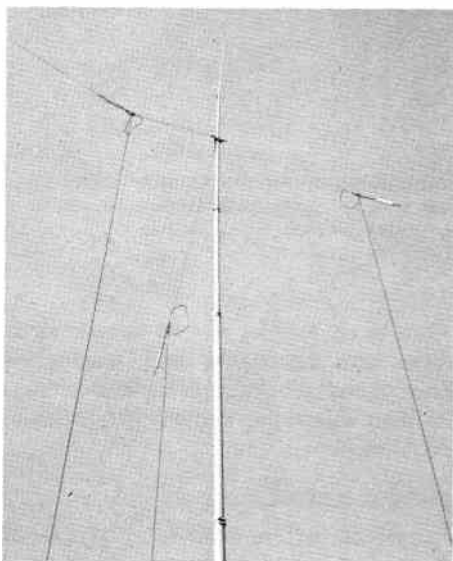


fig. 6. One of the vertical beam elements. Loops near insulators are bridge wires connecting horizontal and vertical cage wires.

and the photo show that these wires extend out from the mast about eight feet, and are connected to the mast at a point about fifteen feet from the top. (The exact point isn't critical.) Then they extend downward to the base of the mast. The upper arms of the cage connect to strain insulators, which are connected to the guys. The lower ends of the wires are also connected to strain insulators, which are in turn tied to the base

together economically in just a few minutes. It has a low angle of radiation and some gain, and can be mounted on a wood 2 x 2 with wood braces and plastic insulators. The antenna is fed at points A and B with a simple half-wave, 4:1 balun constructed of 26 inches of RG-58/U or RG-8/U.

Ed Marriner, W6BLZ

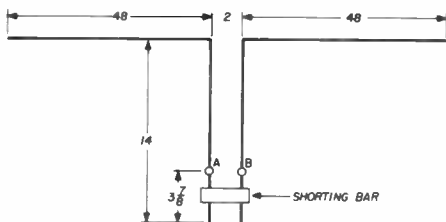


fig. 4. Simple two-meter antenna.

fet preamp for 432 MHz

If you're trying to get the most out of your receiving gear for 432 MHz, you should look into the new 2N5397 and 2N5398 field-effect transistors manufactured by Siliconix Incorporated, 1140 West Evelyn Avenue, Sunnyvale, California 94088.

These new devices are usable up to 800 MHz and their performance on 432 is nothing short of fantastic. Manufacturer's claims are many times on the optimistic side, but according to K6JYO and K6KV, the 2N5397 performs as advertised — at 450 MHz the noise figure is 2.5 dB typical (3.5 maximum), and power gain in the manufacturer's 450-MHz circuit in fig. 5 is 18 dB. The 2N5397 is \$8.50 and the 2N5398 is \$6.00.

In the common-source amplifier shown in fig. 5, the drain and gate leads should be shielded from each other to reduce feedback phase shift. The neutralizing coil should be low loss for best gain and noise figure. One other construction tip: tune patiently!

Another circuit for the 2N5397 is shown in fig. 6; this is a common-source mixer. When building this circuit, bypass the drain lead as close to the transistor

case as possible. Maintain a high impedance at the intermediate frequency for maximum gain and maintain a signal ground on the source. Tune the tank circuit for best gain and noise figure.

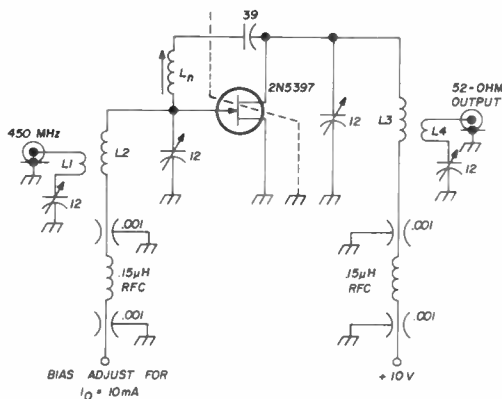
Jim Fisk, W1DTY

vhf antenna switching without relays

Fig. 7 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 activates 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-



- L1 1.4" piece of no. 22 enameled, spaced 0.1" from L2
- L2 1.1" piece of no. 16 solid copper
- L3 1.3" piece of no. 16 solid copper
- L4 1.4" piece of no. 22 enameled, spaced 0.3" from L3
- Ln 3 turns no. 22 enameled on 1/4" slug-tuned ceramic form, aluminum slug

fig. 5. 450-MHz common-source amplifier provides 18-dB gain.

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, 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.

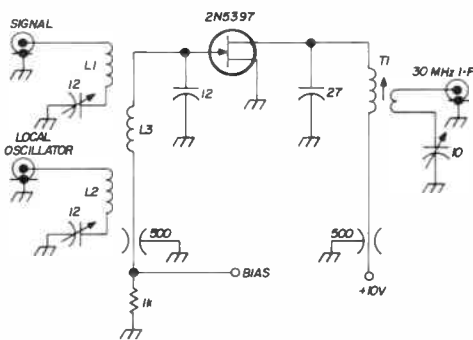
Some trial-and-error may be required with coils L1 and L2 to obtain optimum performance and accurate low-level triggering. I found that for 200 MHz, both coils can be a $\frac{3}{4}$ -turn of number 18 wire, $\frac{3}{4}$ " diameter. At two meters, $1\frac{1}{4}$ turns of the same provided best results. For six meters, $6\frac{1}{2}$ turns of number 18 enameled wire on a $\frac{3}{4}$ " diameter form did the trick.

Bob Brown, K2ZSQ

two-meter converter

The converter shown in fig. 8 receives on two frequencies. I use it with an auto radio to receive my vhf club stations on 145.35 MHz and an fm repeater on 146.94 MHz.

Most of the parts were salvaged from uhf television converters and transistorized BC receivers. Crystal frequencies were chosen to put the converter output into a



- L1 1.4" piece of no. 22 enamelled, close coupled to L3
- L2 1.6" piece of no. 22 enamelled, close coupled to L3
- L3 1.75" piece of no. 16 copper
- T1 Primary, 13 turns no. 22, closewound on $\frac{1}{4}$ " slug-tuned form; secondary is 3 turns no. 22, closewound over primary

fig. 6. 450-MHz common-source mixer circuit.

clear spot in the broadcast band.

The unit is built on a piece of circuit board and mounted in a small Bud box for complete shielding. The board should be insulated from the box if polarity is disregarded. (My box has a negative ground, so I didn't insulate the board.)

The output coil, L3, is an oscillator

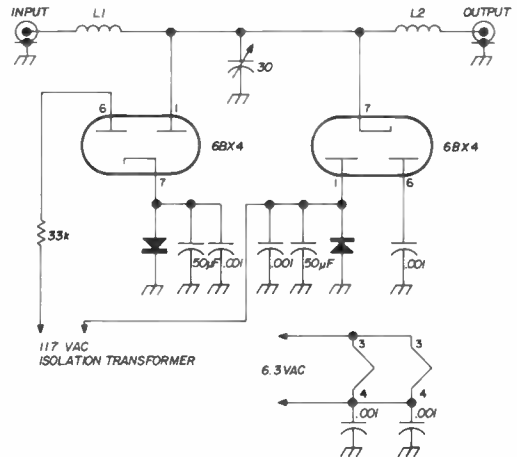


fig. 7. This vhf switcher circuit automatically isolates your converter from the antenna when transmitting. Coil data is given in text.

coil from an old transistor BC-band radio. The tap was the emitter tap for the autodyne mixer. The unused winding was connected between the collector and i-f transformer. You may have to try different values of C1 to resonate your coil.

The antenna coil, L1, tunes rather broadly. It should be peaked between the two desired signals. Spread or squeeze L1 to make C2 tune to resonance.

The crystals are third overtone types. The oscillator may be operated on either side of the input signal. For the low side, the crystal frequency will be

$$f_x = \frac{f_s - f_i}{3}$$

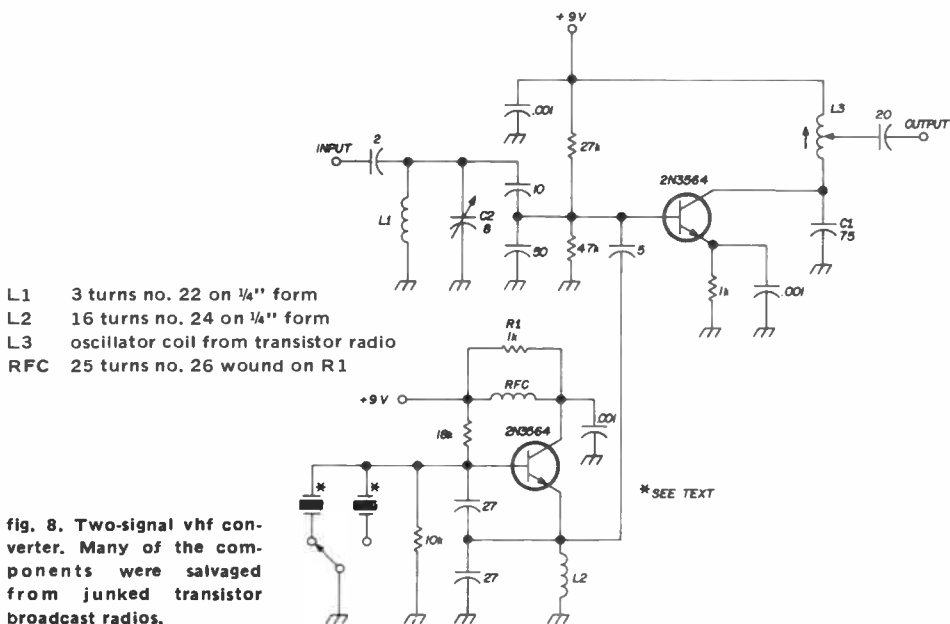
where

f_x = crystal frequency

f_s = signal frequency

f_i = intermediate frequency

To operate the oscillator on the high side of the input signal, merely add the signal frequency and intermediate fre-



quency, then divide by 3.

I mounted a slide switch on the box to select the crystals. No signal feedthrough occurred since the converter and auto radio were well shielded.

This converter operates on 9 volts. If you use it in your car, I'd suggest you bypass the B+ line with a 25- to 50- μ F capacitor. A dropping resistor and zener should be used to reduce the car's primary voltage.

W. G. Eslick, KØVQY

coaxial-line resonators

Transmission lines made of coax, twin-lead or "plumbing" are often used in vhf,

uhf and microwave equipment. They can be used as (or in) filters, as tuning elements in amplifiers or oscillators and as replacements for inductors or capacitors in tuned circuits. At frequencies over 100 MHz or so, transmission lines can be of practical lengths, and they exhibit high efficiencies and excellent performance and reproducibility, unlike some circuits using conventional lumped capacitors and inductors.

Unfortunately, it is sometimes hard to remember which type of transmission line exhibits which characteristics. To help out, fig. 9 summarizes the properties of shorted and open lines in terms of elec-

ELECTRICAL LENGTH, l →	l LESS THAN $\frac{\lambda}{4}$	$l = \frac{\lambda}{4}$	$\frac{\lambda}{4} < l < \frac{\lambda}{2}$	$l = \frac{\lambda}{2}$
SHORTED 	 INDUCTIVE	 PARALLEL RESONANT (HIGH Z AT F)	 CAPACITIVE	 SERIES RESONANT (LOW Z AT F)
OPEN 	 CAPACITIVE	 SERIES RESONANT	 INDUCTIVE	 PARALLEL RESONANT

fig. 9. Electrical properties of transmission-line resonators.

trical length. The table lists wavelengths only between zero and $\lambda/2$, but the addition of $\lambda/2$, or any multiple of $\lambda/2$, does not change the properties.

The electrical length, l , of a transmission line is easily found:

$$l = \frac{l_p}{V}$$

Where l_p is the physical length in meters, and V is the velocity factor (1 for air, 2/3 for common coax, 4/5 for small 300-ohm twinlead). Or for lengths in feet:

$$l = \frac{3.28l_p}{V}$$

The actual wavelength in meters is also easily determined:

$$\lambda = \frac{300}{f}$$

Where f is the frequency in megahertz. For example, a coax line shorted at one end is 1 meter long. What is its electrical length at 6 meters? Since V is 2/3 for common coax,

$$l = \frac{l_p}{V} = \frac{1}{2/3} = 3/2$$

The electrical length is 1.5M, so is one-fourth wavelength ($\lambda/4$) and acts as a parallel-tuned resonant circuit.

The transmission line can be loaded with a capacitor for tuning or to shorten the necessary line length. The effect of the capacitance depends on line impedance, location of the capacitor, fre-

quency, etc. Either experimentation or consultation of a reference containing details will help out here.

Paul Franson, WA7KRE

calibrated signal source for 432 and 1296

A stable signal source for the uhf bands is a very useful item for all vhf and uhf experimenters. The circuit shown in fig. 10 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.

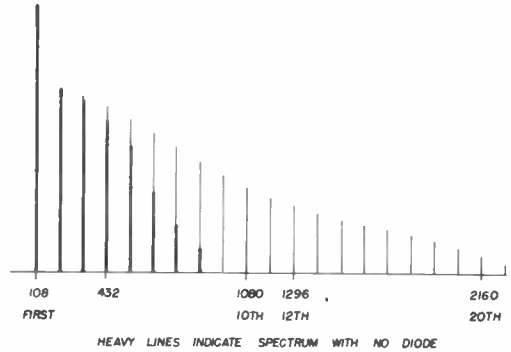


fig. 11. Spectrum output of the signal source as measured with a Hewlett-Packard 8551 analyzer. Heavy lines indicate spectrum without diode.

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 L_1 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. 11 shows the output before and after the diode was installed. 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 1269-MHz receiving converters;

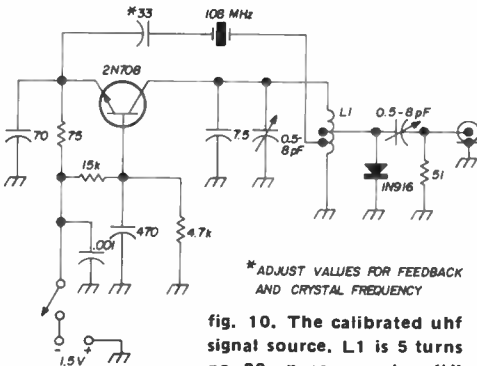


fig. 10. The calibrated uhf signal source. L_1 is 5 turns no. 20, space wound on $1/4$ " diameter, tapped 1 and $2\frac{1}{2}$ turns from ground end.

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 oscillator is checked periodically on a Hewlett-Packard frequency counter; usually it is within 1 kHz 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.

Del Crowell, K6RIL

using noise generators

A noise generator can be used to improve the performance of any vhf or uhf converter. An effective device I use in my shack is shown in fig. 12. A silicon diode should be used (a 1N21 or 1N23) and held in place with a common fuse-clip. The time you spend in constructing

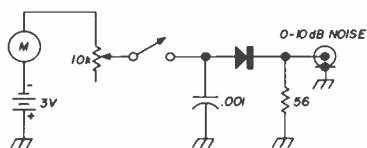


fig. 12. Simple silicon-diode noise generator.

the noise generator will be more than made up by the increased performance you'll get out of your receiving system.

General procedure for tuning up (or optimizing) a converter is to adjust it first for maximum gain — with a signal generator tuned to the frequency most used — and then measure the noise figure with a laboratory-type noise generator. This serves as point of reference. Let's say it's 9 dB.

Next, the silicon diode noise generator is connected to the receiver. The receiver is adjusted as described below until optimum performance is obtained. Then, when rechecked against your reference, you'll have an indication of how poorly your system was adjusted. A laboratory-type noise generator isn't required if you're not interested in this statistical comparison; you should be able to audibly note a marked improvement in overall performance.

The technique is actually quite simple. Output from the noise generator is fed into the converter simultaneously with a signal source. This can be from a signal generator or another ham transmitting across town. First, the noise generator is turned off and the converter is peaked for maximum signal strength. Then slowly adjust for minimum noise when the source signal is removed. Turn the noise generator on and continue adjusting for minimum noise level.

By adjusting both for maximum signal strength and minimum noise, you will find that for optimum differential, S-meter readings will not be as high as you had expected. Remember: Sensitivity has nothing to do with signal strength; it is the noise level which determines the difference between mediocre performance and outstanding converter sensitivity.

Incidentally, repeat this procedure periodically. Equipment is known to age rapidly during warm summer months and tubed-converters (yes, there are still many around) require frequent "optimizing." In using the meter, by the way, 1 mA of reverse crystal equals roughly 10 dB of noise. You can change the intensity by varying the 10k pot.

Bob Brown, K2ZSQ

simple untuned crystal mount

Many times in amateur uhf and microwave work you need an untuned crystal mount for detection of small levels of rf energy. Usually a small diode and bypass capacitor are simply haywired into the

circuit, but this method usually causes some problems. First of all, if the frequency of operation is high enough and the lead lengths are too long, a lot of the available energy will be radiated. Furthermore, the parasitic reactances, as a result of the lead lengths associated with the capacitor and diode, can cause some rather mysterious results.

The crystal mount in fig. 13 is based upon the use of two standard coaxial cable fittings, a type-N jack, the UG-23D/U, and a type-BNC jack, the

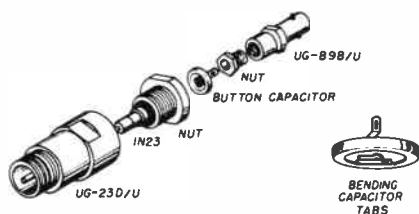


fig. 13. Untuned crystal mount for vhf measurements.

UG-89B/U. Both of these fittings are available at reasonable prices. To build the mount, discard the braid-clamping washer and rubber gasket from the type-N connector. Then take the female pin and carefully squeeze the large end so that it is a snug fit around the pin of a 1N23-type diode.

Ream out the cable-retaining nut so that a 1000-pF button capacitor may be force fit into the opening. Bend one of the tabs on the capacitor over so it will provide a spring contact to the diode when the nut is screwed into the connector. Bend the other tab so it comes straight up from the surface of the capacitor.

Remove the cable-retaining nut from the BNC connector, and place it on top of the type-N nut. Now, solder the whole works together, the N and BNC nuts and the feedthrough capacitor. Make sure that the N nut has been reamed out suffi-

ciently so the BNC nut sits flush and level. The easiest way to solder the parts together is on an electric stove or hot plate.

After this assembly has cooled down, solder a short piece of wire about 5/8 of an inch long to the upright tab on the button capacitor. Place the female pin from the BNC connector over the end of the wire and put the BNC connector on to test for proper wire length. The wire should be trimmed so that the end of the female pin is flush with the center insulator in the connector. When you find the right length, solder the female pin to the wire.

To use the completed mount, connect a sensitive dc microammeter across the BNC connector, apply some rf energy to the type-N connector, and you're in business. In most uhf and microwave work, the type-N connector will fit right in with existing equipment.

Jim Fisk W1DTY

hardware for uhf use

When building vhf and uhf components such as resonant cavities and strip-line amplifiers, the need occasionally arises for nylon screws and nuts. Many of us don't have a supply of such items and look for substitutes.

A source of raw material is the flexible plastic handles used on ordinary cotton swabs known (appropriately) as *Q TIPS*. The handles measure 0.10 inch in diameter — just the right size for a 4-40 die. The *Q-TIP* handles will thread nicely for use as screws. All that remains to be done is to saw a slot in one end to accept a screwdriver. The cotton swabs are available for a few cents in any drug store.

Insulating nuts or washers can be made from sheet plastic. Drill and tap a 4-40 hole in the sheet, then punch out the material around the hole with a paper punch. The disc out of the punch becomes the nut or washer.

Ted Swift, W6CMQ

shack and shop

repairing high-voltage transformers

The plate voltage transformer is pretty much taken for granted by most hams. It represents a pretty large initial investment and is expected to perform dependably with no servicing other than general cleaning and removal of accumulated dust. A transformer rated to provide power for a kilowatt rig can cost anywhere from about \$40.00 to over \$100.00, depending on how much horse trading you're willing to do. It is therefore not unreasonable to expect trouble-free operation of the device, providing the usual overload circuits are incorporated in its installation.

Despite good installation practice it's possible for power transformers to fail, sometimes for no apparent reason. Regardless of whether the transformer is a secondhand pole pig obtained from the power company, or a brand-new one from a well-known manufacturer, there is no absolute guarantee that something can't happen to cause a failure (Murphy's third law).

I have had two sad experiences with large power transformers. Both are commercially made units bought from reputable firms. The first failed as a result of rectifier tube flashback, and the second just plain short-circuited in the secondary. Both problems occurred despite overload protection circuits.

Each transformer was enclosed in a steel case, and each was wound with the secondary over the primary. One was even impregnated with tar. The tar is

supposed to protect the windings from moisture penetration. (It doesn't.) The unit that failed because of rectifier flashback was easiest to repair. The other was a more involved problem and entailed complete disassembly to locate and repair the trouble; the secondary had to be completely rewound on this unit.

Both transformers appeared to be hopeless cases as far as being repairable. However, I like to operate the rig and really couldn't stand the expense of replacing a \$75.00 piece of equipment. I figured I had nothing to lose anyway, so I decided to see if the transformers were worth repairing. The following paragraphs describe how these transformers, which would otherwise have ended up as expensive junk, were doctored and restored to service.

causes of transformer failure

Most power transformers fail because of high-voltage insulation breakdown. It is rare for a transformer to develop an open or short circuit in the primary winding. If problems do occur in the primary (e. g., a short between turns, open circuit, short to the case), then you are faced with removing anywhere from 3000 to 4000 turns of secondary wire, usually about number-22 AWG, just to expose the primary. This, of course, is assuming the primary is wound under the secondary as is the case with many large power transformers. The primary, once you get to it, is easier to repair than the secondary since the primary consists of 300 to 400 turns of fairly large wire (about number-16 AWG).

preliminary checks

It is not necessary to remove the case for an initial check, because all you want to do at this point is try to get a clue as to where the trouble is and how serious it might be. Not much can be done with a bad transformer while still in the power supply, so the first thing to do is set it on the workbench and make some resistance measurements. First test for a short between primary and case, then between secondary and case. Next measure the resistance across the secondary and between each end of the secondary and center tap.

The secondary resistance should be around 300 ohms between each end and the center tap if there is no short between turns. If the meter shows a resistance *differential* of more than 30 or 40 ohms between each end of the secondary and the center tap, then you know the problem is somewhere in the secondary.

The next step is to remove the case. Remove the bolts and save the lockwashers and nuts. You'll find upon reassembly that you must pull down the nuts on the case bolts quite tightly to minimize core hum. The lockwashers are necessary on these monsters!

The primary and secondary leads should now be exposed. These are just long enough to reach the terminals without much slack to spare, so you'll have to use reasonable care while working with them. They will stand some flexing, but it is possible to break one off where it disappears into the winding. Power transformers are pretty rugged otherwise and will withstand ordinary handling during disassembly and assembly.

Unsolder the leads from the lugs on the insulators, then dig out any tar. (Comes out in chunks — no problem.) Inspect the winding area for charred spots. These will be pretty obvious if the short was caused by rectifier flashback. If the failure was caused by some other problem then the trouble will be harder to locate, and you will have to disassemble the core.

flashback failure

My transformer caught fire when one of the rectifier tubes flashed back. For some reason I'll never know, the overload circuit just flat failed to protect the transformer. There was a charred spot where one of the secondary leads was brought up out of the transformer innards. Chances are pretty good that, with a rectifier flashback failure, the short-circuited area will be near the coil surface, and probably in the first secondary winding layer. In this type of failure, extreme heat caused by the high current surge destroys part of the insulation, allowing one or more secondary wires to touch.

Carefully peel away the insulation and clean off the carbonized material with an old toothbrush dipped in alcohol. Don't attempt to replace any of the original insulation, even if it appears to be okay. Apply at least six coats of red Glyptal varnish all around the exposed wire, and work it well into the surrounding layers of insulation. It is essential that every bit of carbonized material be completely removed before applying the Glyptal. Allow five full hours for each coat to dry.

Next, wrap a layer of Scotch number-33 electric tape (rated at 10,000 volts per layer) over the exposed wires, then apply two more coats of Glyptal over the tape and surrounding area. It's a good idea to slather Glyptal all over the outside of the windings, again observing adequate drying time between coats. You just can't put too much Glyptal on these things.

After another ohmmeter check, as described previously, the unit can be reassembled. It's really not necessary to pour melted tar around the windings. As a matter of fact, my transformer had only an inch layer of tar at the top and bottom of the case where the end bells join, so I didn't bother to replace the tar. This transformer has been working perfectly for the past five years. (By this time maybe you're getting the message that I don't have much faith in tar-impregnated power transformers. I have a strong

suspicion that this stuff does more harm than good.)

The repaired transformer should now be given a “smoke test” before reinstalling it in the power supply. The simple circuit shown in fig. 1 is used for this check. You can substitute several 115-volt light bulbs for the Variac, but they should be arranged so they can be progressively shorted to increase the primary voltage. Close the switch and slowly turn up the Variac. It should hardly be necessary to emphasize that you’re dealing

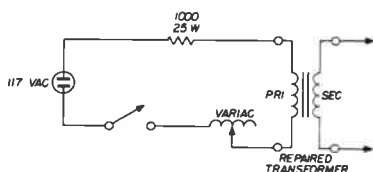


fig. 1. Circuit for smoke-testing the repaired transformer. The 1000-ohm resistor is for current limiting in case one of the windings is still shorted.

with upwards of 3000 volts of instant death, so use utmost care during this test. If you’ve done a good job of cleaning and insulating, your transformer can now be reinstalled in the power supply, and you’re back in business.

Why should an expensive, well-made power transformer mysteriously develop a short circuit? Commercially built transformers are conservatively designed, and engineers specify high voltage insulation with large safety factors. Nevertheless, a transformer can be working fine one day, and suddenly a failure can occur in the insulation, causing a short circuit. Apparently moisture somehow penetrates into the secondary and causes a high resistance path to build up. Finally, one day the insulation just breaks down.

When you inspect the external winding area after one of these “mysterious” failures, chances are you won’t see anything as obvious as in the flashback case. But don’t give up — all you have to do is remove the core, then unwind the secondary to expose the short circuit. This task isn’t really as formidable as it might appear. First, let’s take a look at how the core is put together.

core arrangement

The core consists of E I sections of silicon steel about 0.01-inch thick. The E sections are placed over each other back-to-back, interleaved with I sections at each outer edge (fig. 2). You will find that the laminations have been driven quite tightly into the coil. A little strong-arm effort here will work one of the end laminations loose, then the others will come out easily. You will also find upon reassembly that it will probably be impossible to drive all the laminations back into the coil. My rewound transformer has about ten E I sections missing, but the transformer doesn’t seem to know the difference.

the solenoid

As the coil is unwound, you’ll have to look carefully for evidence of insulation breakdown. It’s a good idea to go over each layer with a magnifying glass, because the area of failure will probably contain extremely small holes in the insulation. Be sure to count the number of turns taken from each layer and keep track of them. While unwinding my coil, I made a rough sketch of each layer, indicating the winding area and number of turns per layer. It is not worthwhile to rewind with the old wire, because no matter how careful you are, it’s virtually impossible to avoid nicking the wire. Even one tiny nick will wash out all your hard work.

If you have to remove the entire secondary winding, you’ll probably find a

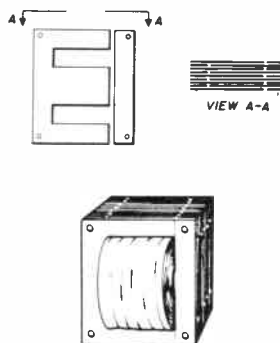


fig. 2. Core buildup and assembled transformer.

layer of paper-thin copper between secondary and primary. This is a shield to prevent electrostatic coupling between the two windings. Its purpose is to keep higher order line frequency harmonics out of the d-c plate supply. This shield isn't really necessary for ham power supplies and just causes added insulation problems. If you decide to keep the electrostatic shield, make certain it is reinstalled exactly as it was originally. My recommendation is to throw the thing away.

secondary wire

If your transformer is more than ten years old it is probably wound with enamelled Formvar. I searched all over trying to find a source of this wire, but found it's practically nonexistent these days. I rewound my transformer secondary with HPTX-200, made by Essex Wire Corporation. This material has a high temperature insulation of some type of polymer. It costs about \$2.00 per pound for number-22 AWG; it weighs about the same as Formvar, so you can determine how much you'll need after you remove the old wire.

When you've located the short circuit, splice on the new wire, wrap the splice with Scotch number-33 electrical tape and give the entire area the red Glyptal treatment. You are now ready to rewind.

rewinding the secondary

I built a jig out of a piece of mop handle to hold the solenoid for the rewind job. The solenoid, which has a paper tube core, was force fitted to one end of the jig and held with a couple of wooden wedges. I turned down the other end of the jig to fit into the chuck of a large portable drill motor. I mounted the motor on the bench with the solenoid protruding about eight inches over the edge of the bench. The spool of new wire was mounted onto a U-bracket fastened to a wooden cleat. This assembly was placed on the floor beneath the solenoid.

The gears in the drill motor provide just enough drag so that the solenoid can be turned by hand during the rewind

process. Maybe it's possible to apply power to the motor and rewind the coil, but I found that to control the windings you have to rotate the solenoid by hand. This takes about an hour per layer for us old folks. As you rewind, put on the same number of turns per layer that were removed — no more, no less. Don't attempt to hop up the transformer by changing the primary-to-secondary turns ratio. You are only asking for more problems, and an additional 500 volts or so on the secondary won't make that much difference to your signal anyway. (Remember, you have to *double* your transmitter power just to gain a 3-dB increase in signal strength.)

Wrap each rewind layer with Scotch number-33 electrical tape and apply two coats of red Glyptal varnish on each layer. You are now ready to reassemble the core laminations.

If you have rewound the coil as tightly as possible, you should have no trouble reinserting the core. It may be necessary to squeeze the solenoid slightly to accommodate the laminations. With one layer of tape around each layer of wire, plus the Glyptal, the solenoid buildup should be very close to the original. It may be necessary to tap the last few laminations home with a mallet. Don't worry about slight bends in the laminations. When you pull the bolts down in the case, the laminations will flatten out. Now give the transformer the ohmmeter check, then the smoke test as described earlier.

baking

An added refinement, though not necessary, is to give the repaired transformer a baked varnish treatment. For \$7.50 I had this done at a local electrical shop that specializes in rewinding motors and generators. (Incidentally, while there I asked them for a quote on how much they would charge to repair my transformer. Would you believe \$65.00, *not* including material?) For the baking treatment, the uncased transformer is immersed in a vacuum tank, and insulating varnish is forced into the windings. The unit is then baked for several hours. I was

mounting. This arrangement keeps the cage neat and taut. All the cage wires are connected electrically to the base of the vertical element and fed in parallel with it.

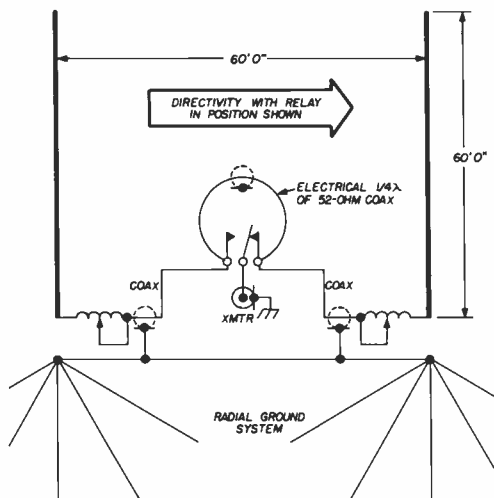


fig. 7. General construction of the array. Loading coils are made from commercial coil stock or home wound, approximately 35 turns, 2½ inches in diameter, close-spaced.

A small loading coil is connected in series with the coaxial cable feeding the base of each vertical, and a movable tap is used to adjust the element for optimum performance.

ground system

Like all vertical antennas, the secret of this one is the ground system. The old axiom of, "the more, the better," couldn't be more true. Bill's philosophy is to lay out every bit of spare wire he can get his hands on, and the ground under and around the array is a maze of everything from hookup wire to braided ground strapping. At this writing, well over 1000 feet of radials have been laid out, but there will likely be more by the time you read this! Each little bit helps put the signal where it will do the most good — low down, and a long way out.

orientation and feed

The two verticals are mounted in line

with the direction most wanted. In other words, to work European stations from the East coast, the elements should be mounted in a northeast/southwest line.

Each element is fed with 52-ohm coax. Both pieces are cut to exactly the same length. However, a third piece, an *electrical* quarter-wave-length, long, is inserted in series with one vertical, thus making it act as a reflector. This produces a significant gain in the forward direction and a very considerable front-to-back ratio. By using the relay or coax switch, this extra piece of coax can be switched from one vertical to the other, therefore instantly reversing the directivity of the array. Bill has found this feature to be very valuable and quite disconcerting to those on the other end of the contact.

tuning

Tune-up of the array was checked with some General Radio test equipment, but

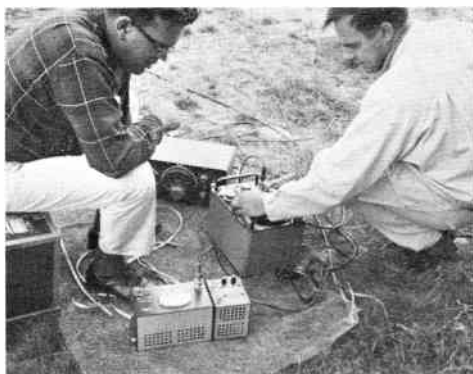


fig. 8. Final check by VE1RK (left) and VE1AX. Test equipment is GR 1606-A rf bridge with 1212-A null detector and a 1330-A bridge oscillator.

for those with more meager means (almost all of us), each element can be tuned separately by using an swr bridge and adjusting the tap on the loading coil for minimum swr. This should be very close to 1:1 at the chosen operating frequency. The loading coils are mounted in small water-tight metal or plastic containers at the base of the elements.

a little concerned as to how the baking would affect the temperature resistance of the Scotch electrical tape, but apparently it does no harm. This baking treatment is probably an improvement over tar impregnation.

You now have spent about \$25.00 for material, are back on the air, and have had the privilege of being introduced to the all-but-forgotten art (at least in ham circles) of transformer winding. Hey, you old timers — kind of tugs at the heart strings, eh?

Alf Wilson, W6NIF

printed-circuit labels

Often it's desirable to permanently label components, terminals, and the printed-circuit board itself. These items may be labelled with copper by using dry-transfer letters. The letters resist the metal etchant in the same manner as tape, etch-resist pencils, ink, or paint placed on the copper-clad board.

Dry transfer labels for electronic equipment are made by Datak Corporation, Passaic, N. J. Many amateurs use them to label panels or chassis. The transfers resemble decals, but are more convenient to use. The labels are transferred by lightly rubbing over the characters with a ball-point pen. The sheet of characters is gently lifted to assure a complete transfer. This technique isn't convenient for use with boards using photo-resist chemicals, however.

Earnest A. Franke, WA4WDK

switch-off flasher

Portable solid-state equipment that is designed for low power drain often excludes the use of "high-drain" pilot lamps. The circuit in fig. 3, designed by S. Thomas and described in *The Radio Constructor*, flashes the light *after* the switch is turned off. The circuit is applicable to any transistor equipment that uses a 9-volt battery. When switch S1 is *on* the capacitor is charged through the diode; when the switch is turned *off* the 6-volt lamp is connected across the capacitor,

causing the bulb to flash brightly. The circuit draws negligible battery current; in Mr. Simon's unit the leakage current drawn by the capacitor was measured at 6 μ A.

The capacitor cannot discharge the equipment circuits because of the diode. However, if the equipment has a large-value capacitor across the power supply, it will discharge through the diode into the lamp, increasing the length of the flash. With most equipment it would probably be possible to eliminate the diode and connect the capacitor and lamp directly to the negative supply lead, although this should be checked experimentally.

Jim Fisk, W1DTY

frequency-sensitive resistors

Deposited-film resistors operating in circuits at hf aren't always to be trusted if precise resistance values are needed.

While at work, I needed a fairly husky precision resistor for a phase-sensitive network in a 100-kW transmitter. When I learned that the resistor order would be delivered in the distant future, I decided to check our stock of deposited-film resistors to see if I could use an acceptable substitute. Much to my surprise, I found that at frequencies as low as 14 MHz most values were completely inadequate.

The resistors were checked on an rf bridge. I found that the R component generally tended to run quite low — at least 10% low in a 1% resistor — except that values in the 25k and higher ranges showed violently reversed characteristics.

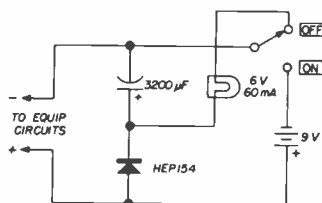


fig. 3. Simple low-drain circuit flashes when the switch is turned off. Current drain is less than 10 μ A.

Resistors from some manufacturers were fairly accurate in this frequency range, but many showed bulk resistance effects. For example, a 25k unit would shoot up to 100k or more!

I haven't completely checked out the results of these tests, but it appears that values of about 300 ohms or so tended to be capacitive at these frequencies, which is opposite to what I would have suspected. Units of 200 ohms or less tended to be slightly inductive. You might bear this in mind the next time you need precision resistors in a frequency-sensitive application.

Bill Wildenhein, W8YFB

homemade heat sinks

Anyone who works with solid-state devices knows that when a large amount of power is applied to a transistor, a heat sink is required.

Many tv, fm and other commercial broadcast stations use tubes such as the 4CX250 and similar types. In most cases these tubes aren't worth rebuilding when they go bad and can be had for the asking.

I found that by cutting the top section from the tube, plus a small amount of filing (and in some cases a little ingenuity), an excellent heat sink can be made for many power transistors. The convenience of having the heat sink around the transistor, rather than spread out along the chassis, can be realized.

Greg Larsen, WAØWOZ

removing IC's

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

I didn't have a vacuum desoldering tool handy, so I used a trick known as wicking. I flattened a short piece of shield braid and held it against the pins of the

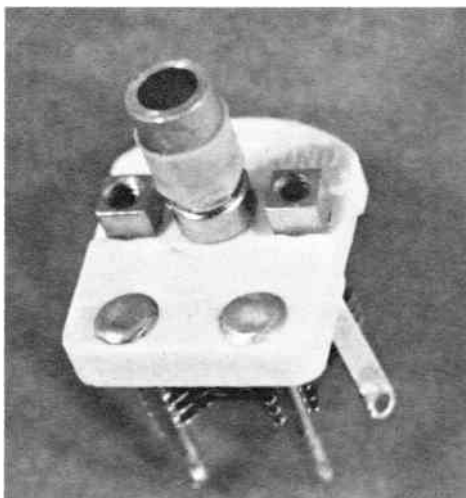


fig. 4. Adding a shaft to APC capacitor with plastic shrink tubing and epoxy.

IC. Applying the soldering iron to the braid causes it to absorb the solder, and the IC comes out easily. Everything must be clean: soldering iron, braid and PC board.

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

Alf Wilson, W6NIF

adding shaft to apc trimmer

Ever faced with that very frustrating situation where the only variable capacitor in the junk box with the required value is an APC trimmer — and you need one with a shaft? Although several solutions have been presented in the past for this problem, the following one is easy, effective, and thus, my favorite.

Slip a piece of shrinkable tubing over the hexagonal rotor extension, and trim to extend at least 1/8-inch. Insert a suitable length of 1/4-inch diameter metal tubing into the open end; apply sufficient heat to shrink the tubing. Now, gently slide the shrinkable tubing and shaft

extension back off the hexagonal rotor extension. Note that it comes off easily but retains the hex shape. Put epoxy on the exposed inner surface of the shrinkable tubing and replace on the rotor extension. Now pour epoxy through the center of the metal tubing, making sure it flows into the original screwdriver slot and that no air pockets are left (see fig. 4).

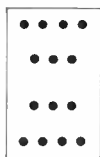
Let it stand for at least twenty-four hours before attempting to turn the shaft. You now have an APC trimmer with a rugged shaft extension; add a knob and you're in business.

V. M. Scott, Jr., W1ETT

surplus IC's

Now that surplus digital integrated circuits are so commonplace, more and more experimenters will be using them in their circuits. Unfortunately, most of these low-cost ICs use 14-lead flat

fig. 5. Full-size drill template for mounting 14-lead integrated circuits on printed-circuit board.



packs — with leads spaced 0.05 inch apart. This small spacing and the natural fragility of the IC makes them extremely hard to solder into a circuit.

However, I have found a simple method that works out very well. First of all, bend leads 2, 4, 6, 9, 11 and 13 down at the case at a 90° angle. Now all the leads are spaced 0.1 inch apart; much better for soldering and making circuit board connections. Bend the remaining leads down, out from the case an 1/8 inch or so.

Use the layout shown in fig. 5 for putting the ICs on a circuit board. I made a small template from a piece of aluminum, and attached it to the circuit board with a piece of Scotch tape and drilled the holes in the board. With this much spacing it's not too difficult to draw circuit trails. You could accomplish approximately the same thing with perfor-

ated board with holes on 0.1-inch centers, but the problem here is connecting wires to the fragile IC leads. It's much easier and quicker with an etched circuit board.

Always check out the IC before you solder it into place. It's next to impossible to remove one of these 14-lead ICs from a printed-circuit board without destroying it. Flip-flops can be tested with the output from your 100-kHz calibrator. Run a lead from the output of the IC to the receiver antenna socket; if the flip-flop is working properly you should pick up calibrate signals at the 50-kHz points on the dial. Integrated-circuit gates can be given a cursory checkout with dc levels. For more involved testing you'll need a square-wave generator and an oscilloscope.

Nat Stinnette, W4AYV

surplus relays

Surplus 24-volt dc relays are often available to the amateur experimenter at attractive prices. Most of these relays are well designed (and originally very expensive), but hams tend to steer clear of them since a 24-volt dc power supply is required. Not so. I've been using these relays, powered off the ac line, for many years; not even a transformer is needed, just a silicon diode, a resistor and a filter capacitor.

Some of the late model hermetically-sealed relays have a relatively high resistance, but most of the open-frame types have a resistance in the range from 150 to 500 ohms. With the circuit shown in fig. 6 the dc output from the diode is 54 volts. If the resistor is chosen with the

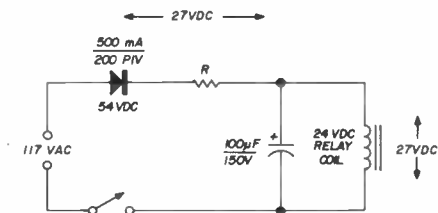


fig. 6. Operating a 24-Vdc relay from the 117-Vac line. Resistor R is a 10-watt wire-wound unit with the same resistance as the relay coil.

same resistance as the relay coil, 27 Vdc will be impressed across the relay. Use a 10-watt wirewound resistor. The filter capacitor only has to be large enough to keep the relay from buzzing; generally 100 μF will be more than enough. An electrolytic rated at 150 volts costs only a few cents more than one with a 50-volt rating, and the higher rating is worthwhile in terms of trouble free operation.

You can obtain a small time delay (that is, the relay will hold in for a small time after the switch is turned off) of 1 to 2 seconds by simply increasing the filter capacitor to 500 to 1000 μF . For this application a 50 working-volt capacitor is suitable. For longer delay periods—up to 30 seconds—use a Globar resistor in series with the coil as shown in fig. 7. I tried two different types of Globar resistors, type FR-9 and FR-50. The FR-9 is used for replacement service in tv sets and is widely available. It works quite well with relays having 150 ohm coil resistance or less. The FR-50 works best with coils with greater than 150 ohms resistance. If you can't find an FR-50, it is interchangeable with type FR-100 and FS-800 (type numbers by Workman Associates, the distributor).

Time delays can be roughly predetermined from fig. 7. If a relay has too high resistance for a particular application, it can be lowered by the simple expedient of putting a composition resistor in parallel with the coil.

The opposite effect—rapid pick-up—can be obtained with the circuit shown in fig. 8. The two no. 313 pilot

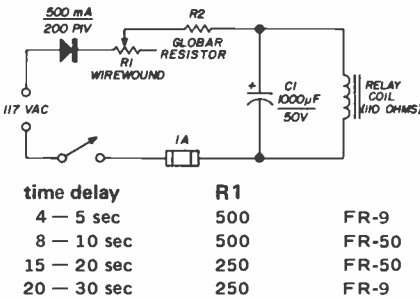


fig. 7. Time delays to expect with different Globar resistors (based on 110-ohm relay coil). Value of C1 can also be varied.

lamps have very low resistance when cold and this permits rapid relay pickup because of the initial over-voltage. Supply voltage is reduced to normal as soon as the lamp heats up and its resistance increases. An ordinary relay in this circuit, in parallel with 60 μF , was sufficiently responsive to follow keying at 20 words per minute.

Neil Johnson, W2OLU

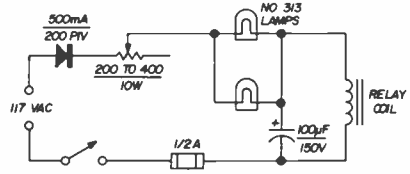


fig. 8. Rapid armature pickup occurs with this circuit because of the low cold-resistance of the lamps, allowing an initial over-voltage.

ferrite beads

Radio-frequency chokes have certain undesirable characteristics that must be recognized and corrected if they are to operate as intended. Rf chokes have "holes" in their frequency-versus-impedance response that can cause resonance in conjunction with stray circuit capacitances.* This phenomenon can result in high circulating currents that will destroy the choke coil.

The flow of rf current can be impeded by using ferrite beads. These beads are made of ferrous particles imbedded in a ceramic material, much like the cores in some rf coils. The physical structure of the beads is somewhat different, however, in that they have a hole through their centers to accept a wire.

characteristics and uses

Running a wire through a ferrite bead greatly increases the inductive reactance of that length of wire. This reactance follows the familiar 6.28fL law, which

*An example is the requirement for modifying the plate-feed choke in high-power tetrode amplifiers when they first became popular. This is explained in George Grammer's "Pi-Network Tank Circuits for High Power," *QST*, October, 1952 and in the 1953 edition of the *ARRL Handbook*, editor.

shows that as the frequency increases, so does the reactance. At 50 MHz, for instance, one inch of wire through a ferrite bead may show an impedance of $50 + j45$ ohms. String on more beads, and the impedance goes up. It increases in a smooth and totally predictable manner — no holes, no peaks.

Now let's see how these beads can be used in ham equipment. For decoupling dc and ac power leads, they're ideal: small, effective, free of dc (or low-frequency ac) resistance, and not susceptible to resonance from associated capacitance.

The increased impedance offered by the beads at higher frequencies suggests another application: that of vhf or uhf parasitic-oscillation suppression. Remember that the impedance consists of a resistive as well as a reactive component. This acts as a potent suppressor of vhf or uhf oscillations, while having very little effect on the primary signal.

This same characteristic can be a temper saver in eliminating rf pickup and rectification of amateur signals by audio equipment. One of the most frustrating jobs you can tackle is that of curing rf pickup (and rectification) in a neighbor's transistorized hi-fi audio equipment. The low impedances encountered in these sets make rf bypassing almost a hopeless task. A ferrite bead or two slipped over the base (or gate) lead of each susceptible transistor can save much hair tearing.

drawbacks

There *must* be some negative factors! For one thing, the beads saturate with too much current through the wire upon which they're strung. This means you can't use them to decouple the filament of a high-powered transmitting tube in the grounded-grid circuit. Also, they're not suitable in place of an rf choke in parallel-feeding a transistor in a powerful transmitter.

One reason why these handy little devices haven't found greater use among radio amateurs is that they are not listed in most supply catalogs. Ferrite beads are manufactured by Stackpole Carbon Company, Electronic Components Division,

St. Marys, Pennsylvania, and by Ferronics, Inc., 66 North Main St., Fairport, New York 14450. You may have difficulty in purchasing small quantities from these sources, however.

In summary, the use of ferrite beads offers a quick, easy, and painless cure for many of the problems confronting the builder of ham equipment. Use them — you'll like the way they work.

Carl C. Drumeller, W5JJ

power, voltage and impedance nomograph

When working with various electronic systems it's often necessary to think in terms of power, voltage, impedance and dBm, and to move rapidly from one unit of measurement to another. Rather than making a calculation each time you change from one unit to another, it is much easier to use the graphs in figs. 9 and 10. These graphs are an analysis of power vs voltage as a function of impedance.

The chart in fig. 9 covers power levels from 1 milliwatt to 1 watt, while the graph in fig. 10 covers the range from 1 watt to 1 kilowatt. The power range in

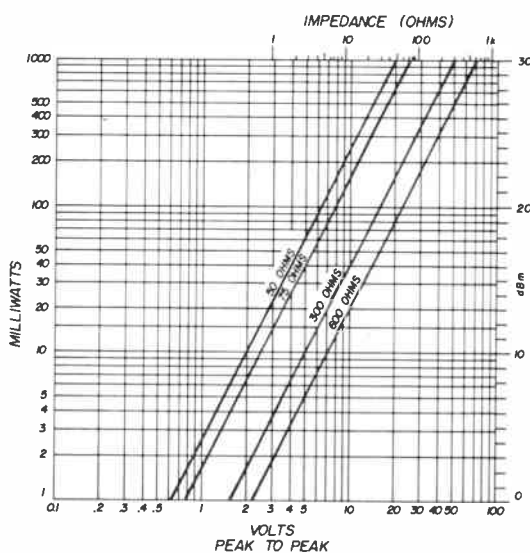


fig. 9. Power, voltage and impedance nomograph for power levels from 1 milliwatt to 1 watt.

dB above a milliwatt, or dBm, is shown on the right hand side of each chart.

These graphs are extremely useful when using an oscilloscope for waveform analysis, and when direct conversion to power level is necessary. They are also useful in the design and calibration of rf voltmeters and wattmeters.

how to use them

When laying out these graphs one of the prime considerations was to make them easy to use. All you have to do is enter the graph with the known quantity and continue to the appropriate impedance line; read the unknown value on the opposite axis of the graph. For example, what power level is represented by 10 volts peak-to-peak across a 50-ohm line? Enter the chart at the 10-volt point on the lower axis, project upward to the 50-ohm impedance scale, then to the left to 250 milliwatts. Note that this corresponds to 24 dBm on the right-hand side of the chart.

Although the impedance curves on the face of the chart are limited to the most common transmission-line impedances, the power and voltage curves are not restricted to these impedances. The im-

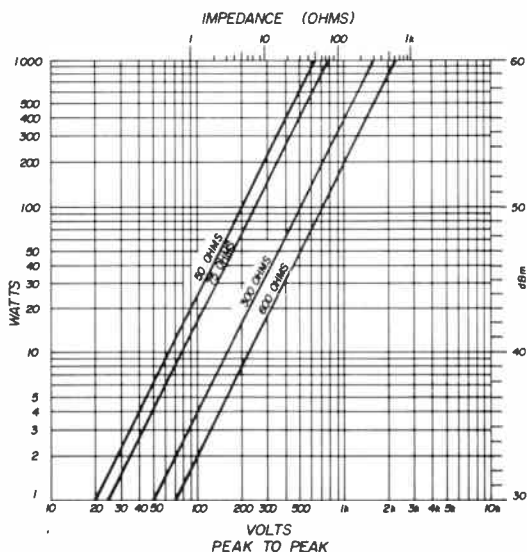


fig. 10. Power, voltage and impedance nomograph for power levels from 1 watt to 1 kilowatt.

pedance ruler at the top of each graph may be used to construct other impedance lines; simply draw a straight line through the appropriate point on the impedance ruler, keeping the new impedance line parallel to those already plotted.

The choice of peak-to-peak volts on the horizontal scale is based on the use of an oscilloscope as the primary measuring tool. If you wish, this axis can be recalibrated in any convenient terms that relate directly to peak-to-peak voltage. For example, for peak voltage, divide by 2; for rms voltage, divide the scale by 2.8.

I would appreciate hearing from any readers who find an unusual application for these graphs.

Walter E. Pfister, Jr., W2TQK

nomograph for reactance problems

Nomographs are aids for quickly solving many electronic circuit problems. A straight-edge placed across appropriate scales allows you to solve equations without using a slide rule or pencil and paper.

The nomograph in fig. 11 can be used to solve reactance problems when one quantity is unknown and two are known. Chart A is used to determine magnitude and decimal-point location. The significant figures are determined from chart B.

practical example

Suppose you're interested in a circuit such as that shown in fig. 12. It's a Q-multiplier that can be added to your i-f strip for increased selectivity. Let's say you have an inductance of fairly high Q whose value is 5 mH. You'd like the circuit to resonate at 1 kHz; what value capacitor should you use in the op amp feedback circuit?

You could determine the capacitor's value by well-known mathematical formulas, but the nomograph of fig. 11 will provide the answer much quicker. Here's how it's done.

In fig. 11A, a line is drawn between the two known values: 5 mH and 1 kHz.

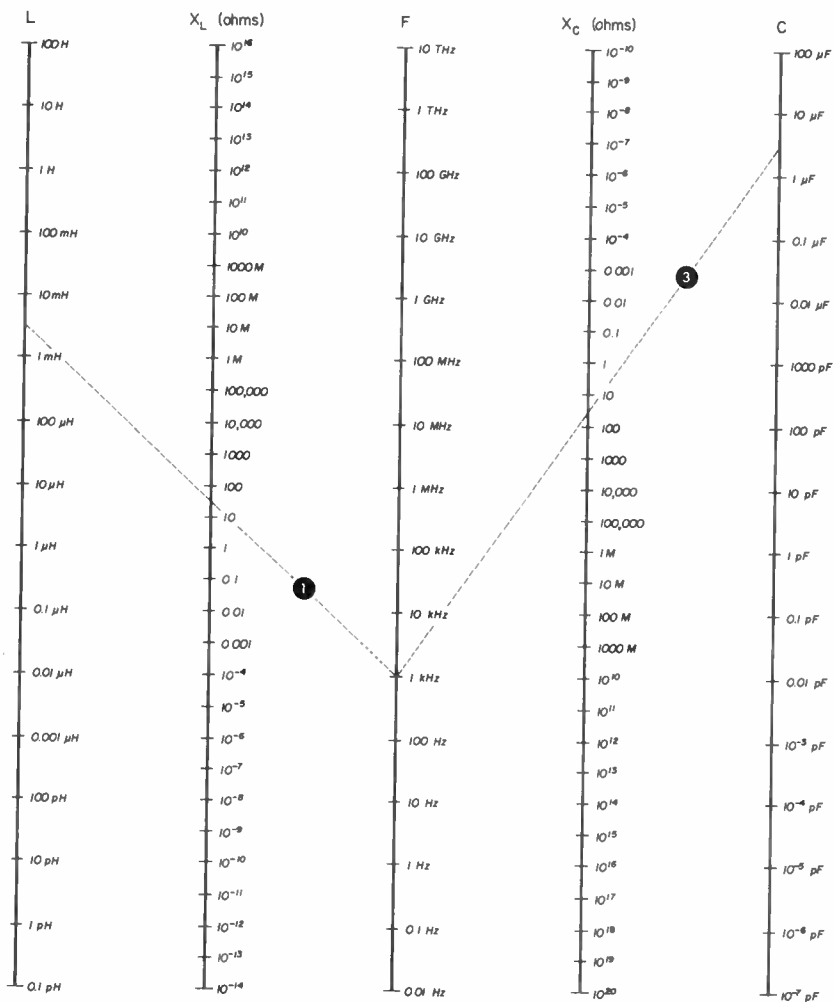


fig. 11A. Reactance nomograph. This chart is used to determine magnitude and decimal location; significant figures are found in fig. 11B.

This is labeled 1 in fig. 11A. The intercept of line 1 on the X_L scale of fig. 11A shows the inductive reactance of this combination to be somewhere between 10 and 100 ohms.

Moving to fig. 11B, a line is drawn between 5 on the L scale and 1 at the top of the F scale. This location of line 2 was purely arbitrary; the line, in this case, could just as easily have been drawn between 5 on the L scale and the bottom of 1 on the F scale.

The intercept of line 2 on the X_L scale of fig. 11B is what's important. It intercepts the X_L scale at approximately 3.2. This is the significant figure, or decimal

multiplier, for the value determined from fig. 11A. The inductive reactance is therefore 10 (from fig. 11A) multiplied by 3.2, or 32 ohms.

determining capacitance

Returning to fig. 11A, a line is shown between 1 kHz on the F scale and 32 ohms on the X_C scale. This is labeled 3 in fig. 11A. Line 3 intercepts the C scale at 5 μF, which is the desired capacitance for C1 of fig. 12.

An inductive reactance of 32 ohms is used to find C1's capacitance, because basic theory says that resonance in a tuned circuit requires that X_C equal X_L .

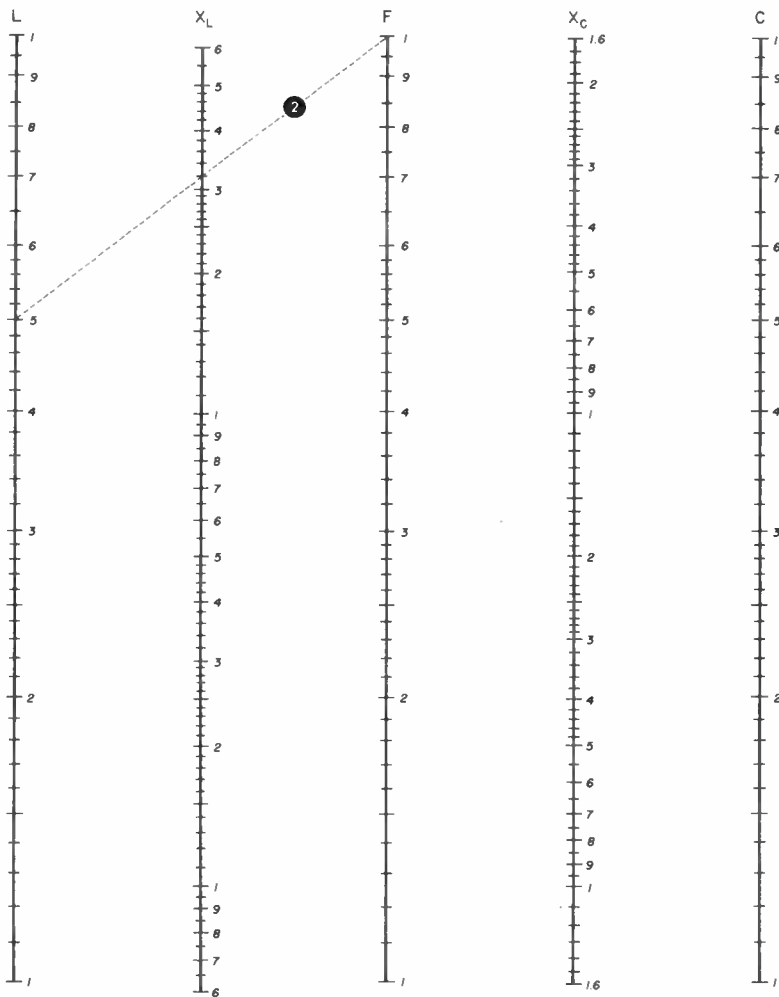


fig. 11B. Reactance nomograph. This chart is used to find significant figures after magnitude and decimal location have been determined from fig. 11A.

This is the principle behind the calculations shown here. The nomograph can be used to solve other reactance problems as well.

useful hints

When working with nomographs, the accuracy of the final result will depend on how accurately you draw the connecting lines between the unknowns. Many nomographs give you a "ballpark" answer. If you wish to refine the result, you'll have to use mathematics. If used with care a nomograph is a great time saver and can provide answers with accuracy sufficient for most problems.

A sharp pencil or draftsman's dividers should be used to locate the end points of the two unknown variables. Lay a straight edge against one leg of the dividers, and rotate the straight edge until you pick up

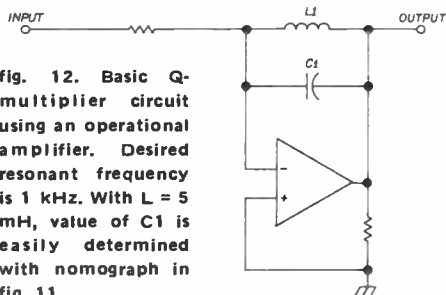
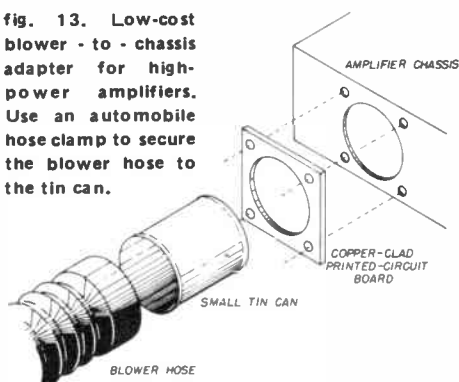


fig. 12. Basic Q-multiplier circuit using an operational amplifier. Desired resonant frequency is 1 kHz. With $L = 5$ mH, value of C_1 is easily determined with nomograph in fig. 11.

the other point. Read the value at the intercept and mark it down on a piece of scratch paper. This will avoid cluttering the nomograph, which can then be used indefinitely as a computational tool.

Alf Wilson, W6NIF

fig. 13. Low-cost blower - to - chassis adapter for high-power amplifiers. Use an automobile hose clamp to secure the blower hose to the tin can.



blower-to-chassis adapter

In many large rf power amplifiers it's not practical to mount the blower directly on the chassis — because of either size or noise. The setup shown in fig. 13 allows the blower to be remotely located. A small tin can with both ends removed is soldered to a piece of printed-circuit board. The printed-circuit board is then mounted to the amplifier chassis with sheet-metal screws. A length of flexible hose is run from the blower to the

adapter. The hose can be secured with a hose clamp or tape.

Bruce Clark, K6JYO

underwriter's knot

I'd bet most hams are pretty adept at replacing ac power plugs and installing new line cords, but how many know how to tie the Underwriter's knot? Most hams use a simple over-hand knot when terminating a line cord in a plug or lamp socket, but this is not only incorrect, it's unsafe. As you can see from the three steps shown in fig. 14, the Underwriter's knot is easy to tie when you know how. Use it next time instead of that half-knot you've been using all these years — it's a lot safer.

Jim Fisk, W1DTY

using undervoltage relays

Although many forms of undervoltage relays are on the market, it sometimes happens (as in my case) that a quick and dirty substitute can be found. Most standard relays exhibit a very wide difference in pickup and dropout current, typically three to one. Therefore, it's often impossible to use a single relay for a specific application.

I needed a relay that had to drop out within approximately 10% of the normal

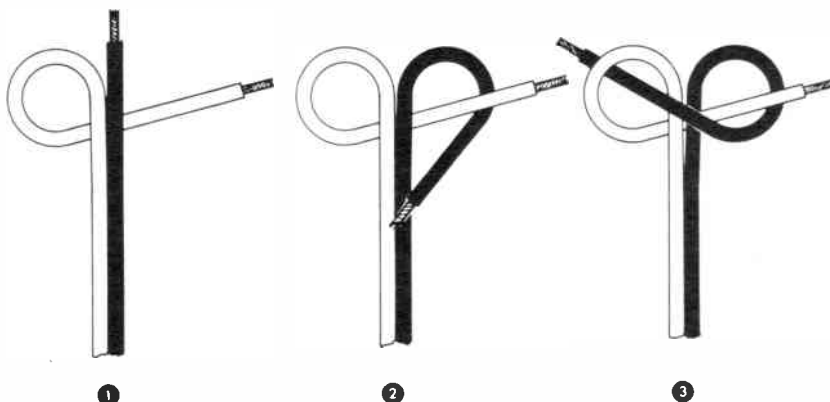


fig. 14. Three steps to tying an Underwriter's knot.

running voltage. Available relays were Allied Control type BO6-D 184VDC. These units have a 13500-ohm coil. Pickup is at approximately 10 mA, but dropout occurs at about 3 mA.

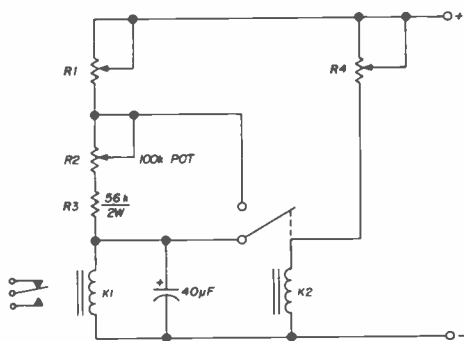


fig. 15. Method for controlling relay drop-out voltage. Resistors can be proportioned to cause pull-in and drop-out over a fairly wide current range.

As shown in fig. 15, R1 is proportioned to energize the relay (K1) at about 20% below the operating voltage. R4 is then set to pull in K2 at a point a little higher — say 10% below the operating voltage. The 100k composition pot is then set to drop out K1 at a point 10% below the operating voltage.

This arrangement was run in a bench setup for eight hours to determine if relay heating would result in any variation of the dropout point. The maximum variation in dropout point was approximately 2% of the operating voltage.

Don't try to use a variation in dropping resistance as shown in fig. 16, or instability will result. Also, don't try to use an extra set of contacts on K1 to perform the function of contacts on K2; i. e., to switch in R2 and R3. Oscillation or instability will result.

A solid-state circuit would have been preferable, except that this particular equipment was to be subjected to a

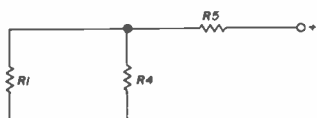


fig. 16. An example of dropping-resistor arrangement that will result in circuit instability.

radioactive environment, which precluded the use of solid-state devices.

Bill Wildenheim, W8YFB

soldering tip

Dig around the lumber yard for a scrap of lumber with a big end-grain knot, the sappier the better. Make a depression in the knot to hold a drop or two of solder. Then, when you want to clean and tin your iron, just rub the tip on the knot. Cover the bottom of the knot to keep it from glopping up your bench.

James T. Lawyer

mounting bnc connectors

Whenever I mount a BNC connector on a panel, I end up scratching the panel as I tighten down the connector's retaining nut. The answer to this problem lies in a simple homemade tool. I cut the tip off an old screwdriver and epoxied a BNC female fitting to the shaft with metal epoxy. Now, when I install a BNC panel connector, I use the modified screwdriver to hold the connector while I tighten the nut — no more frayed tempers or scratched panels.

Elliott Kanter, W9KXJ

trimmers

The components illustrated in fig. 17 have not been ruined. This simple technique of adjustment may be old hat to electronics technicians, but it always seems to surprise hams. It is especially useful when breadboarding with junk-box components.

Resistors may be adjusted upward in resistance as much as 50% with a file, as shown; the cut illustrated raised this resistor by about 20%. With this method it is possible to make your own 1% resistors or to match a set of components. The filing can be done with the resistor connected in a bridge or other active circuit; remember to start with a resistance lower than the desired value. For reasonable changes, the reduction in

results

Results have been excellent as might be expected. Front-to-back ratio is sufficient to knock down most of the W and VE interference. This, with the forward gain, gives a very significant boost to DX signals. While gain measurements are always tricky, it appears that gain in the order of 3 to 5 dB is reasonable, while actual on-the-air reports (compared with a well-tuned dipole) have been so good as to be considered a bit far-fetched. However, signals received with excellent strength on the other side of the world from VE1 can attest to the performance of the array.

George Cousins, VE1TG

dipole center insulator

If you need a dipole center insulator in a pinch and can't get down to your local electronics emporium, try the gadget

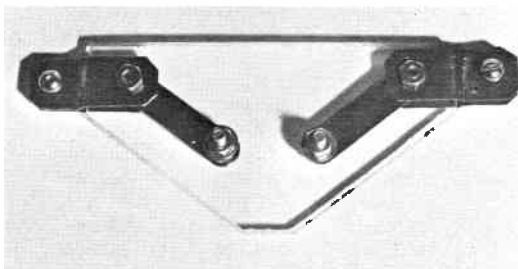


fig. 9. Simple center dipole insulator.

shown in the photo. Just cut out a triangle from $\frac{1}{4}$ -inch plexiglass or other strong plastic, install some copper or brass strips with brass nuts and bolts, hook up your antenna wire and feedline, and you're in business.

Ted Woolner, WA1ABP

repairing broken coax connectors

The insulation in SO-239 coax chassis connectors sometimes breaks and falls out after a period of use. While a new connector costs only 59c, it is often far easier and quicker to repair a broken one

than replace it. Set the piece of gear on the bench so the connector is upright. With a pair of longnose pliers squeeze the four parts of the inner connector sleeve together and fill the inner part with melted candle wax. Epoxy resin is used to replace the original broken and lost insulating material. After the epoxy hardens overnight remove the candlewax with a small screwdriver or with a $\frac{5}{32}$ " twist drill, and the SO-239 will be permanently repaired.

Robert B. Kuehn, WØHKF

a method for remote keying your transmitter

If you're tired of those endless trips up and down the tower to turn the transmitter on and off for swr checks, you'll be interested in this gadget. All you need is a two-conductor phone plug, a spare length of coax long enough to reach from antenna to transmitter (or any other two-conductor cable that's long enough), an spst toggle switch and some electrical tape. Connect the phone plug to one end of the coax, the spst switch to the other end and tape the switch terminals. Put the phone plug into the CW key jack on your transmitter and operate the switch to be sure everything is working properly.

Now load your transmitter into a good dummy load; exchange the dummy load for the transmission line to your antenna,



fig. 10. Remote-keying unit for use while testing antennas.

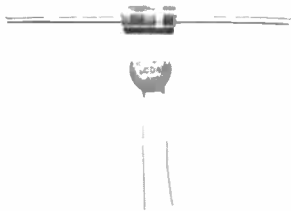


fig. 17. Resistor and capacitor values may be adjusted to precise values by grinding. Resistors increase value; capacitance values decrease with grinding.

power-handling capacity is small.

Ceramic capacitors may also be trimmed in the same manner, with a file or grinder. Practical adjustment range is 50%, and the illustrated capacitor measured about .0025 pF. In-circuit changes are possible.

After reaching the desired value, brush or wash away any dust, and apply a coating of insulating varnish or spray. With careful preparation, these "precision" components are as good as the expensive kind.

Dale E. Coy, W5LHG

using integrated circuits

There are many amateur projects showing up with integrated circuits, but ICs in the dual-inline package (DIP) have two rows of pins spaced 0.1 inch apart. This tight spacing makes casual breadboarding difficult at best, even if you use printed circuits. Here are two simple methods you can use to make the most out of ICs.

Use standard Vector number 169P59/032 perforated board. This board uses a hole arrangement that will fit either the fourteen- or sixteen-lead dual-inline package. Although it is fairly expensive, you can plug your IC in, solder wires to each pin and bring these out to flea-clip. This approach is a natural for casual experiments.

The second method is to turn the IC upside down and secure it to the board with a small dab of glue (be sure to note the alignment dot on top of the IC before

turning it over). This method lends itself to circuits that are going to be used for awhile. When soldering to the IC pins, use a small well-tinned iron and gingerly solder your connections. Normally no heat sink is required if you are careful and treat the device as you would a silicon transistor.

Elliott Kanter, W9KXJ

a dab of paint, a drop of wax

When building gear, there are times when you would like to build, align and forget various assemblies. The perfect example of this is the vfo. The only snag in such wishes is that most modules or assemblies require regular maintenance because, more often than not, the screws holding them together loosen from mechanical vibrations and such. Or perhaps the tuning slugs move a fraction of a turn. Then, the vfo starts to drift, the chirp sounds like an electronic bird-cage and misalignment can lead to out-of-band operation.

Take a hint from the Japanese and put a dab of paint or fingernail polish on each joint. This will prevent the screws from loosening and reduce the number of times you will have to overhaul a unit. Remember, however, that paint is not equivalent to epoxy glue and should not be used as a substitute for lock washers or other hardware. If you ever have to open a "painted" seal, a quick jerk with a pair of pliers will do the trick.

To keep tuning slugs put, use the wax drippings from a lighted candle, preferably a small birthday-cake type. If you use paint on the slugs, the paint may find its way into the threads and prevent the slug from turning — permanently!

D. E. Hausman, VE3BUE

adjustment screwdriver

Many transceivers have internal adjustments for the S-meter, final-amplifier bias, carrier null, etc. Trying to get a screwdriver into these adjustments through the perforations in the cover is

next to impossible. Here's what I use.

I fashioned an ordinary wire coat hanger into an adjustment tool by clipping the hanger about 1½ inches each side of where the Y is formed. Then I straightened this section until it was perpendicular to the hook. Next, I straightened the hook and filed the end until it was flat enough to fit into the transceiver adjustment slots. The finished tool should have the appearance of a T.

The tool is small enough and long enough to go through the holes in the transceiver cover, and it is strong enough to withstand the torque of some of the more stubborn adjustments. It's also handy for tapping those sticky changeover relays that hang up once in awhile.

Dan F. Davis, WAØKGS

soldering fluxes

There may be some confusion regarding the use of "acid-free soldering paste." These paste fluxes are suspensions of zinc-chloride and ammonium chloride, aqueous solutions in grease. They are excellent fluxing agents for most common metals other than stainless steel, but the residue that is left after soldering is highly hygroscopic. Reaction of water with the residue produces hydrochloric acid. Obviously, this is no material to use with electronics equipment. The inorganic residue is not affected by low heat and requires about 1300°F to be boiled off. It is difficult to remove by washing.

Dave Heller, K3HNP

cleaning printed-circuit boards

If you work with printed-circuit boards you probably have had difficulty removing solder from the holes when removing resistors, capacitors or other components. A cutting-torch cleaner — available from a welder's supply house — is a very useful aid. The cutting-torch cleaner comes with 12 different-sized steel rods which can be pushed through

the solder holes. About one-quarter inch back from the smooth tip of each rod there's a machine-cut rasp surface which may be used to enlarge the holes, if necessary. The rods are usually furnished with a small metal case for easy storage.

Felix W. Mullings, W5BVF

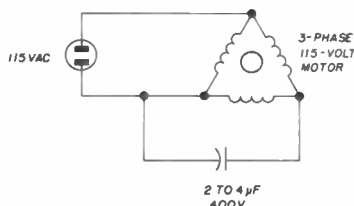


fig. 18. Simple circuit for running a 3-phase motor from single-phase 117-Vac line.

three-phase motors

A three-phase blower motor can be run very easily from single-phase ac by connecting it as shown in fig. 18. It will not develop full power but will run smoothly at somewhat reduced speed. The optimum size for the capacitor should be determined experimentally. If the motor runs backward, connect the capacitor to the other side of the line.

Fred Brown, W6HPH

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

safer suicide cord

Most electronics experimenters have an ac-plug-to-clip-lead adapter (better known as a "suicide cord") on their workbench. Unfortunately, too many experimenters fail to take a few simple safety precautions — precautions that can save much embarrassment or worse!

First of all, always use a fused ac plug such as the El Menco along with 5- to 10-amp fast-blow fuses. For the cord itself use about 6 feet of no. 16 heavy-duty lamp cord. For the clip-lead end, Mueller no. 60 alligator clips are a good bet, but be sure to insulate each alligator clip with a Mueller no. 62 rubber cover.

Bruce Clark, K6JYO

deburring holes

A large drill, at least three times the hole diameter, may be used for rapid deburring. Place the drill in a drill press, and bring each raw hole against it briefly. Be careful to use well-sharpened drill bits when working with brass and plastics.

Jim Ashe, W2DXH

plastic protective material

A self-adhesive plastic sheet product is on the market that is of interest to amateurs. The material I use is called *Plain-Vu* manufactured by Carr Adhesive Products, Inc., Somerville, Massachusetts.

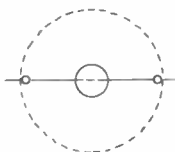
Plain-Vu is a clear plastic contact material, which is good for covering homemade dials, panels, I.D. cards, licenses, QSL cards and the like. It's available in stationery stores in sheets 9 x 11-5/8 inches, two sheets for about a dollar.

Paul White, W6BKX

punching aluminum panels

After breaking my share of Greenlee chassis punches, I have found a way of using them on 1/8-inch aluminum panels. First, drill your starting hole; insert the drive screw; screw on the cutter; and

fig. 19. Punching thick panels is made easier by drilling two small holes where the points of the punch touch the panel.



rotate the cutter by hand to scribe a circle. Then scribe a line through the circle; drill two small holes just *inside* the circle where the points of the cutter touch the panel. These two holes allow the cutter to shear the metal rather than punching through it with brute force.

Joe Kofron, W7DIM

small parts trays

The flat aluminum trays used for TV dinners are divided into compartments and make excellent trays for holding small parts and screws while you're building or repairing ham gear.

Don Farrell, W2GA

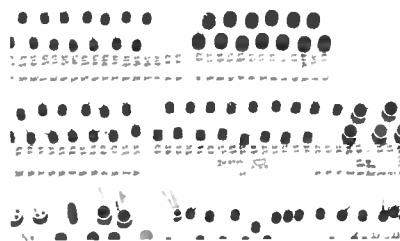


fig. 20. Transistors with measured characteristics are easily stored with double-backed masking tape.

transistor storage

Finding a place to store transistors that have measured characteristics can sometimes be a problem. Try this: lay several pieces of double-backed masking tape across a piece of cardboard as shown in fig. 20. Then, stick the transistors to the tape and write their characteristics just below. You can remove and replace the transistors many times if you use this method.

Tom Lamb, K8ERV

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 vfos 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, VE3BUE



fig. 21. Circuit of electronic temperature indicator. The LM301A is a very stable dc amplifier.

grommet shock mount

When building portable gear or equipment for use in mobile installations, it often becomes necessary to shock-proof certain components, especially relays, since they are bound to rattle, and the contacts become intermittent while going over rough roads. A common rubber grommet of suitable size makes an ideal shock mount. A hole is drilled for the grommet; then the relay, etc. is mounted as shown in fig. 25. The grommet acts as a shock-absorbing washer. The larger the grommet, the greater the shock absorb-

tion. If the inside diameter becomes too large to hold the head of the mounting screw, a flat washer should be added.

D. E. Hausman, VE3BUE

tilt your rig

Modern operating practice for receivers, transmitters and transceivers calls for the cabinets to be slightly tilted. However, most equipment is shipped with four equal-length legs.

To tilt your gear slightly, measure the diameter of the legs at the front of the rig. Then buy some rubber crutch tips from the dime store or your hardware dealer. These will raise the front of the rig sufficiently to allow easier operation. If they're properly fitted to diameter, they slip on and off easily and provide a modern appearance for your operating desk.

George Haymans, WA4NED

the multi-box

Anytime you put a piece of audio gear on the bench there is the problem of mixing and matching the variety of connectors used in current equipment. My little multi-box has proven invaluable when working on such gear. In its present form it consists of three types of jacks mounted in pairs on a $2\frac{1}{4} \times 2\frac{1}{4} \times 5$ mini-box. A terminal strip mounted on the left-hand end of the box accepts pigtail connections. The jacks and terminal strip are all wired in parallel so that any signal fed into one jack is available at any other jack. The three types of jacks on my personal version of the multi-box are standard phone jacks, mini jacks (à la transistor-radio practice) and RCA phono jacks. The RCA phono jacks were mounted with the aid of pop rivets; this is a very neat way to mount this style jack in any gear you may construct.

Al Joffe, W3KBM

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Table 1. Antenna dimensions; length in feet

Frequency (MHz)	length (feet)					
	$5/8\lambda$	$1/2\lambda$	$1/4\lambda$	quad side	inverted vee	radials
3.50	167.143	133.714	66.857	71.714	132.571	68.571
3.55	164.789	131.831	65.915	70.704	130.704	67.606
3.60	162.500	130.000	65.000	69.722	128.889	66.667
3.65	160.274	128.219	64.110	68.767	127.123	65.753
3.70	158.108	126.486	63.243	67.638	125.405	64.865
3.75	146.000	124.800	62.400	66.933	123.733	64.000
3.80	153.947	123.158	61.579	66.053	122.105	63.158
3.85	151.948	121.558	60.779	65.195	120.519	62.338
3.90	150.000	120.000	60.000	64.359	118.974	61.538
3.95	148.101	118.481	59.241	63.544	117.468	60.759
4.00	146.250	117.000	58.500	62.750	116.000	60.000
7.00	83.571	66.857	33.429	35.857	66.286	34.286
7.05	82.979	66.383	33.191	35.603	65.816	34.043
7.10	82.394	65.915	32.958	35.352	65.352	33.803
7.15	81.818	65.455	32.727	35.105	64.895	33.566
7.20	81.250	65.000	32.500	34.861	64.444	33.333
7.25	80.690	64.552	32.276	34.621	64.000	33.103
7.30	80.137	64.110	32.055	34.384	63.562	32.877
14.00	41.786	33.429	16.714	17.929	33.143	17.143
14.05	41.637	33.310	16.655	17.865	33.025	17.082
14.10	41.489	33.191	16.596	17.801	32.908	17.021
14.15	41.343	33.074	16.537	17.735	32.792	16.961
14.20	41.197	32.957	16.479	17.676	32.676	16.901
14.25	41.053	32.842	16.421	17.614	32.561	16.842
14.30	40.909	32.727	16.364	17.552	32.448	16.783
14.35	40.767	32.613	16.307	17.491	32.334	16.725
21.00	27.857	22.286	11.143	11.952	22.095	11.429
21.05	27.791	22.233	11.116	11.924	22.043	11.401
21.10	27.725	22.180	11.090	11.896	21.991	11.374
21.15	27.660	22.128	11.064	11.868	21.939	11.348
21.20	27.594	22.075	11.038	11.840	21.887	11.321
21.25	27.529	22.024	11.012	11.812	21.835	11.294
21.30	27.465	21.972	10.986	11.784	21.784	11.268
21.35	27.400	21.920	10.960	11.756	21.733	11.241
21.40	27.336	21.869	10.934	11.729	21.682	11.215
21.45	27.273	21.818	10.909	11.702	21.632	11.189
28.00	20.893	16.714	8.357	8.964	16.571	8.571
28.10	20.819	16.655	8.327	8.932	16.512	8.541
28.20	20.745	16.595	8.298	8.901	16.454	8.511
28.30	20.671	16.537	8.269	8.869	16.396	8.481
28.40	20.599	16.479	8.239	8.838	16.338	8.451
28.50	20.526	16.421	8.211	8.807	16.281	8.421
28.60	20.444	16.364	8.182	8.776	16.224	8.392
28.70	20.383	16.306	8.153	8.746	16.167	8.362
28.80	20.313	16.250	8.125	8.715	16.111	8.333
28.90	20.242	16.194	8.097	8.685	16.055	8.304
29.00	20.172	16.138	8.069	8.655	16.000	8.276
29.10	20.103	16.082	8.041	8.625	15.945	8.247
29.20	20.034	16.027	8.014	8.596	15.890	8.219
29.30	19.966	15.973	7.986	8.567	15.836	8.191
29.40	19.898	15.918	7.959	8.537	15.782	8.163
29.50	19.830	15.864	7.932	8.508	15.729	8.136
29.60	19.764	15.811	7.905	8.480	15.676	8.108
29.70	19.697	15.758	7.879	8.451	15.623	8.081

table 2. Conversion chart from hundredths of feet to inches. From table 1 an inverted-vee antenna on 3.9 MHz is 118.974 feet long. This is equivalent to 118 feet, 11.64 inches (after rounding off to 118.97 feet).

	0	1	2	3	4	5	6	7	8	9
0	0.00	0.12	0.24	0.36	0.48	0.60	0.72	0.84	0.96	1.08
0.1	1.20	1.32	1.44	1.56	1.68	1.80	1.92	2.04	2.16	2.28
0.2	2.40	2.52	2.64	2.76	2.88	3.00	3.12	3.24	3.36	3.48
0.3	3.60	3.72	3.84	3.96	4.08	4.00	4.32	4.44	4.56	4.68
0.4	4.80	4.92	5.04	5.16	5.28	5.40	5.55	5.64	5.76	5.88
0.5	6.00	6.12	6.24	6.36	6.48	6.60	6.72	6.84	6.96	7.08
0.6	7.20	7.32	7.44	7.56	7.68	7.80	7.92	8.04	8.16	8.28
0.7	8.40	8.52	8.64	8.76	8.88	9.00	9.12	9.24	9.36	9.48
0.8	9.60	9.72	9.84	9.96	10.08	10.20	10.32	10.44	10.56	10.68
0.9	10.80	10.92	11.04	11.16	11.28	11.40	11.52	11.64	11.76	11.88
1.0	12.00	12.12	12.24	12.36	12.48	12.60	12.72	12.84	12.96	13.08

and take the remote-keying line and swr bridge up to the antenna location. This is the procedure I use for keying my transmitter from the roof when tuning up the gamma match on my 6-meter Yagi.

Fred Hock, WA3H DU

antenna dimensions

Here's a chart of antenna dimensions that should be handy if you're getting ready to put up a new sky wire. This is part of a computer listing I ran that covered the spectrum from 3.5 to 30 MHz with readouts for each 50-kHz change in frequency.* The first column of the chart, table 1, is the operating frequency in MHz. The following six columns give the dimensions of 5/8 wavelength, 1/2 wavelength, 1/4 wavelength, the length of one side of a cubical quad, the length of an inverted vee, and the proper radial length for a ground-plane antenna, all lengths given in feet. To convert from hundredths of feet to feet-and-inches use the chart in table 2.

Jim Barcz, WA8JMY

portable all-band antenna

Although the antenna shown in fig. 11 was designed primarily for operation on large wood boats, it looks ideal for any amateur with limited space. Vertical antennas are virtually unmanageable on large non-metal boats and must be constantly retuned, matched and re-resonated; this all-band miniature dipole is relatively non-critical, easy to tune and easily matched. Although it can be rotated to take advantage of any inherent directivity, most users don't go to the trouble. The antenna is tuned up on the lowest band of operation (either 40 or 80 meters) with a grid dipper and antenna-scope. Operation on the higher bands is provided by the two Waters Mobile Band-Adders. The original design of this antenna is attributed to W2INS, and it has been built by a number of boat enthusiasts with excellent success.

*A copy of the complete computer print-out, from 3.5 to 30 MHz in 50-kHz steps, with dimensions to six decimal places, is available from *ham radio* for \$1.00.

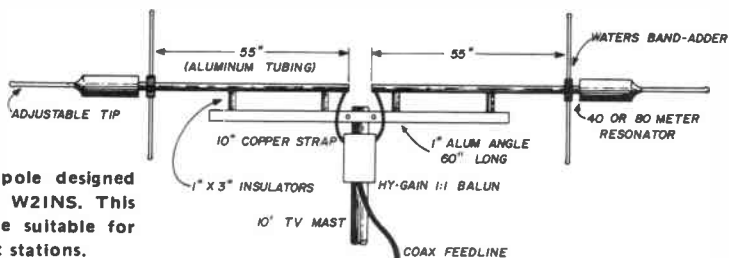


fig. 11. Multiband dipole designed for wooden boats by W2INS. This antenna would also be suitable for portable and apartment stations.

wall-to-wall antenna tuner

Although a well-elevated outdoor antenna is best for radio communications, such an antenna is often impractical or impossible to erect (due, for example, to space restrictions or a grouchy landlord). Fortunately, communications on the ham bands can be effective with only a simple short-wire antenna installed indoors.

I have contacted many stations on the West Coast and in the Midwest using only a horizontal wire strung between two walls. However, the antenna must be properly terminated to be effective. The terminating impedance for most rigs is 50 ohms.

All the equipment needed to terminate the antenna properly is shown in fig. 12. In addition to the rf input and output terminals, the Ameco Model SWB bridge has three leads, which should be connected as shown. The switch selects either forward or reverse power. We want as high a reading as possible in the forward position and as low a reading as possible in the reverse direction. The antenna tuner is adjusted for optimum impedance match between transmitter and antenna.

construction

For convenience, I mounted my bridge in a 4 x 4 x 2-inch aluminum box. The Ameco bridge, which comes with Amphenol SO-329 connectors, is 4½ inches long. It won't fit into a 4-inch-long box, so here's what to do.

Remove the front and rear panels of the box. Then saw one edge of the box so you can spread it apart to accommodate the bridge. Drill holes to allow the Amphenol connectors to protrude. Use a ¾-inch circle cutter.

Mount the bridge on the bottom of the box with screws. Close the box, using a small angle bracket and pk screws to keep it closed tightly.

The 4-inch box also includes the meter (1½-inch square), a miniature toggle switch, and a miniature potentiometer. There's plenty of space for these com-

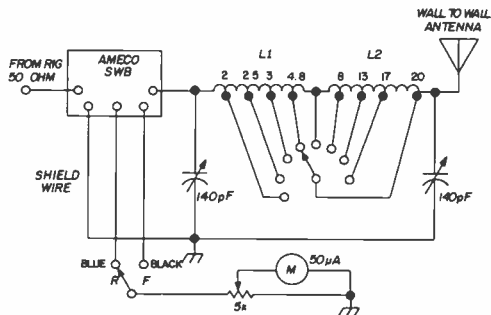


fig. 12 Bridge and tuner for random length antennas. Variable capacitors are Hammarlund type MC140. L1 is 8 turns per inch; L2 is 16 turns per inch. Both coils are 1½-inch diameter. L1 is tapped at 2, 2½, 3 and 4 turns; L2 at 8, 13 and 17 turns.

ponents if they're mounted on the front panel.

the tuner

The tuner is housed in a metal box, 8 x 3 x 2-3/4 inches. The variable capacitors are Hammarlund type MC140. The inductance consists of two separate coils in series, each 1½ inches in diameter. The coils are mounted side-by-side, about one half inch apart.

tune-up

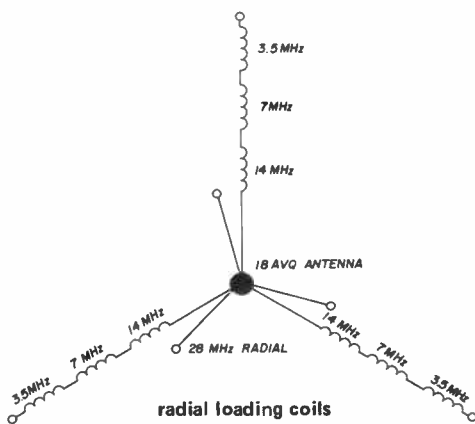
If you have a 50-ohm dummy load, adjust your rig for optimum output. Position the switch to R, and select a tap on the tuner that gives minimum reading on the meter. Adjust the variable capacitors for a minimum meter indication.

I use this system with an indoor antenna on 15 and 20 meters, my favorite bands. The tuner and bridge will probably match any random-length antenna. I've used the circuit of fig. 12 with a vertical whip, 6 feet high, and with an outdoor long-wire antenna 50 feet long. Results have been very good.

I. Queen, W2OUX

multiband ground-plane

Although the popular Hy-gain 18AVQ multiband vertical antenna is designed to

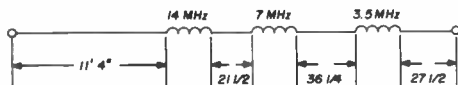


14 MHz	29 turns no. 20, 1.5-inch diameter, 16 turns per inch (Air-Dux 1216T)
7 MHz	46 turns no. 18, 2-inch diameter, 16 turns per inch (Air-Dux 1616T)
3.5 MHz	94 turns no. 18, 2-inch diameter, 16 turns per inch (Air-Dux 1616T)

fig. 13. The Hy-Gain 18AVQ multiband vertical can be used as a ground-plane antenna by using a system of inductance-loaded radials.

be mounted on the ground, it can be used as an elevated ground plane. All that is required is a suitable arrangement of inductance-loaded radials that provide resonance on each of the amateur bands.

JA1QIY reports excellent results with the system shown in fig. 13. The radials shown in fig. 14 provide high performance on 3.5, 7, 14 and 21 MHz; a separate set of radials is used for ten meters. The radials permit the antenna to be put above surrounding objects where it can do the most good. If the radials are allowed to slope away from the antenna at the 18AVQ they can also be used as guys. With the radials sloping away from the antenna at



NOTE: SEPARATE 28-MHz RADIAL IS 8' 4" LONG

fig. 14. Construction details for the loaded radials.

about 45°, the antenna provides a relatively good match to 50-ohm coaxial cable. The dimensions in fig. 14 are for the cw end of the band, but with a little cut and try, equal performance can be obtained on the phone bands. The multi-band swr of the JA1QIY antenna is shown in fig. 15.

JA1QIY has reported excellent DX performance with this antenna, particularly on 80 and 40 meters. On 80 he has worked Soviet Russia, Korea, Okinawa and the Philippine Islands; on 40 he has been able to work into the United States and Canada, no small feat with relatively low power.

In the original version of this antenna each of the radial loading coils is mounted around a length of phenolic rod. However, ceramic strain insulators could be used for more strength and better performance in wet weather. The 28-MHz radials are spaced a few inches away from the low-frequency radials.

Jim Fisk, W1DTY

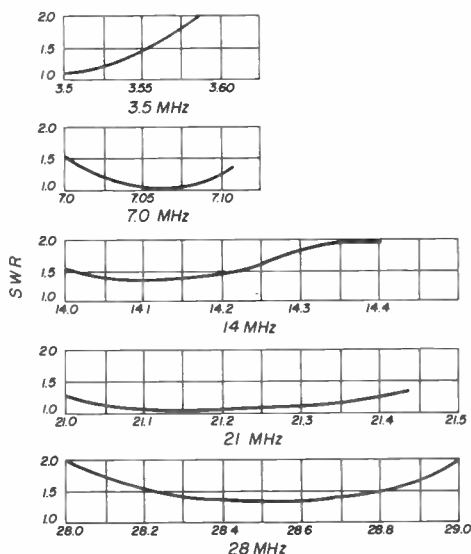


fig. 15. Swr performance of the multiband ground plane. Loaded radials were designed for the CW end of each band.

vhf fm

fm deviation meter

With fm becoming more and more popular, a deviation meter is a nice piece of equipment to have in your station. The method described can be used for most frequencies, and on 2 meters it works fine.

features

The deviation meter allows you to check deviation of on-the-air signals as well as your own. The use of an oscilloscope is ideal, and peak deviation can be monitored.

A two-channel fm receiver is used, with channel A tuned to the desired frequency and channel B tuned either 15 kHz above or below channel A. The setup is shown in fig. 1. Discriminator output is fed to the vertical input of a scope. Internal sweep is used for the horizontal plates.

With no signal, noise will be displayed on the scope. When an unmodulated signal is received, a straight line will appear, possibly with some noise, depending on signal strength.

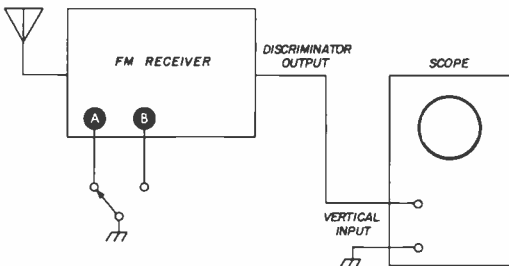


fig. 1. Fm deviation meter using 2-channel receiver and oscilloscope. Method allows monitoring of peak deviation of received signal or your own.

operation

Switch the receiver to the desired signal.

Adjust vertical centering so that the line appears in the center of the display. Now switch to channel B. The line will swing either up or down. Calibrate the scope vertical gain for a reasonable display.* Switch back to channel A, and modulate the transmitter to be tested. If voice peaks hit the calibration points, the signal is deviating ± 15 kHz.

A more elaborate unit could be made using a three-channel receiver, which would allow calibration above and below the received frequency. Two channels should suffice, however, since the signal deviates equally well both ways.

Vern Epp, VE7ABK

fm deviation measurement

Accurate fm deviation measurements are much more involved than they may seem. The i-f and discriminator must be extremely broad. Pulling out the i-f filter and reducing the size of the discriminator load resistors by about 3:1 is a good start. It's also a good idea to tweak the discriminator for best linearity — it's probably not as good as you might think.

In reference to the fm deviation meter described above, a somewhat less expensive calibration procedure involves switching in a 455-kHz crystal first, then a 440-kHz or 470-kHz crystal in place of the second conversion oscillator crystal.

*Or use a grease pencil to mark deviation limits on the face of the scope tube. editor.

If you don't have a dc-coupled oscilloscope, you can use the same technique used at the Motorola factory: use an audio-frequency multivibrator to switch between a 440-kHz and 470-kHz oscillator, and couple the output into the receiver i-f strip.

If the whole thing sounds a bit cumbersome, you can set your deviation quite adequately for amateur purposes by bellying into the microphone, and adjusting the deviation control to just below the point where first limiter current starts to kick downward with modulation. This may not sound very sophisticated, but it's probably at least as accurate as most homemade deviation meters.

J. A. Murphy, K5ZBA

fm repeater receiver isolation

If you are interested in obtaining channel isolation in your fm-repeater receiver, the Motrac units made by Motorola are worth considering. These units have a five-cavity front end and are readily available from surplus sources dealing in fm gear.

motrac receivers

It should be noted that there are two basic types of Motrac receivers. The type-L receiver uses a bipolar rf amplifier stage and is a poor choice for just about any fixed-station application. Its desensitizing characteristics are 5 to 10 dB inferior to the tube-type G receiver, and 10 to 15 dB worse than the old type-A receiver. The intermod characteristics are even worse: 10 to 15 dB more than the G, and 20 to 25 dB more than the A. The later type-M receivers use a fet mixer in the front end; it is about equal to the old type-A in desensitization and far, far better in the intermod department.

The order of preference for Motorola receivers is the type-M (new Motrac), type-A (wide-chassis tube type), type-G (narrow chassis tube type), and last, the type-L (early Motrac).

J. A. Murphy, K5ZBA

continuous tuning for fm converters

The Collins 75A-3 and 75A-4 receivers, with their calibrated-tuning capability, can be used with most crystal-controlled converters to provide continuous coverage of the two-meter fm band.

The appropriate converter is one having an output between 26-30 MHz. The 75A-3 receiver will tune this range directly; the 75A-4 will cover most of this range, and an appropriate crystal can be selected to cover the frequency portion desired.

On all later production models of the 75A-3 (and on all 75A-4 receivers) a broadband i-f output jack is provided, which was originally intended for use with a panadapter. This output can be coupled via 52-ohm coax to the i-f strip of any of the surplus Motorola receivers, most of which use the same 455-kHz i-f.

For early 75A-3 receivers that don't have the panadapter output, the instruction manual gives complete details for installing the output jack, which is a very simple job.

If you have a 75A receiver, this is an excellent way to become acquainted with the two-meter band and discover which spot frequencies are active. Then you can decide which crystals to purchase for the fixed-tuned equipment you would eventually use. You can't beat fm for communications reliability and noise-free operation.

Skip Tenney, W1NLB

three channels from two

Many of the common fm mobiles are equipped for two-channel operation. However, with the different frequency combinations in repeaters using simplex, it might be necessary to transmit on channel B and receive on channel A (or vice versa). This can be accomplished easily with the addition of steering diodes (fig. 2). All you'll need are four diodes

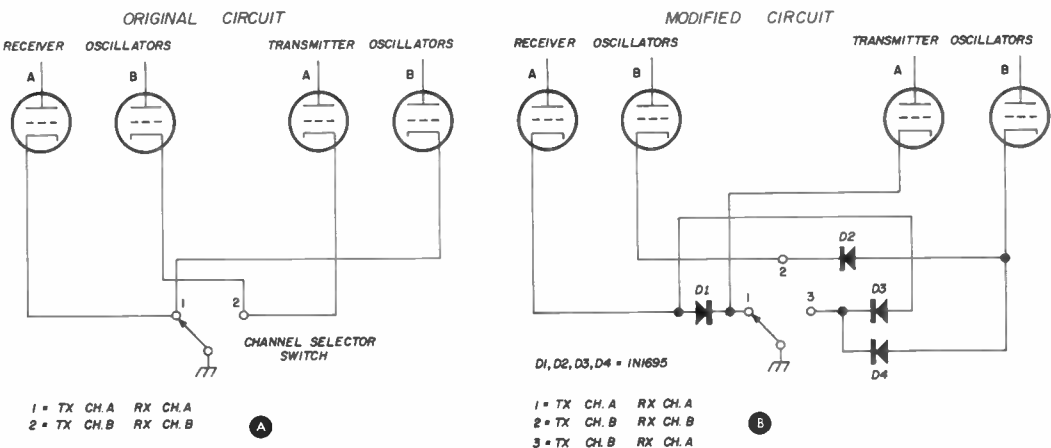


fig. 2, Method for increasing channel coverage in a typical fm oscillator circuit. Steering diodes, B, allow an extra transmit-receive channel.

capable of carrying the current of the circuit and a single-pole, triple-throw switch. This system has been used in commercial applications for a private mobile system in this area.

The circuit shown in fig. 2B provides two simplex channels with a modified two-channel set. The idea for this application is from VE7BDY, to whom I'd like to express my thanks.

Vern Epp, VE7ABK

tone encoder and secondary frequency oscillator

While working on various RTTY and fm selective-call projects, the need for an accurate stable tone source became apparent. The device shown here is presently programmed for eight frequencies, provides a good sine-wave output and is quite stable over a wide range of temperature, voltage and loading. The oscillator has been duplicated by a number of amateurs and used as a digital selective-call encoder on many of the West Coast repeaters. Its reliability and stability have been proven under all types of field conditions. Further, the unit can be easily modified to provide tone-burst operation.

It should be obvious that to obtain frequency accuracy closer than 1%, a

frequency counter must be used. However, close to 1% accuracy can be expected for the frequencies listed with combinations of good quality Mylar capacitors.

For burst use, return the negative common to the push-to-talk line instead of ground; add a zener regulator for power or use a dry battery. Connect the two diodes and the RC network as shown in the box in fig. 3. For mobile operation, be sure *not* to ground or common *anything* except through the audio return line to the transceiver.

Digital encoding can be easily obtained

table 1. Frequencies and tuning capacitors used in the tone encoder built by WB6YOC.

Channel	Design Frequency (Hz)	Measured Frequency (Hz)	Capacitance (μF)	Use
1	1275	1271.5	0.150	RTTY space frequency*
2	1700	1699.6	0.068	RTTY mid-band*
3	1800	1803.0	0.0568	selective-call tone
4	2125	2125.8	0.0340	RTTY mark frequency*
5	2400	2401.4	0.0181	selective-call tone
6	2550	2551.4	0.0106	RTTY mid-band
7	2700	2704.1	0.0071	selective-call tone
8	2975	2973.3	0.0011	RTTY space frequency

*wide shift fsk/ssb transmitter.

by inserting a set of dial-pulsing contacts in series with the 3.3k oscillator base-bias resistor, or in series with the 0.1 μ F oscillator output coupling capacitor. The unloaded output voltage is in excess of 2 V rms above 1700 Hz. With a 100-ohm load, the output drops to 300 mV rms, still enough to provide full deviation in most fm rigs.

The frequencies I used are shown in table 1 but other tones can be easily substituted for your particular requirements. Printed circuit boards for the basic oscillator circuit are available from the treasurer of the Grizzly Peak VHF Radio Club, Post Office Box 1333, Richmond, California 94802 for \$1.50 each. I would like to extend my thanks to WA6UFW who did the initial design work with the basic oscillator.

David M. Stahley, K8AUH

vhf fm modulation standards

It is not widely realized that there are certain variables in fm that must be

defined and standardized before full compatibility is obtained between transmitting and receiving equipment. I will attempt to identify these variables, point out current standard practice and discuss how these standards affect transmitter and receiver circuitry.

In amplitude-modulation systems the modulation limit is related to carrier level. This limit is called 100 percent modulation. There is no such inherent limitation for fm systems. Any modulation level, or deviation, may be transmitted as long as the receiver bandwidth will accept it.

Two standard receiver bandwidths are currently found in amateur practice. These bandwidths, as well as most of the other standards which we will discuss, stem from commercial practice — and the large amount of commercial fm equipment used by amateurs. The most common bandwidth permits a deviation of ± 15 kHz; this is referred to as wideband. Newer commercial equipment permits a deviation of only ± 5 kHz; this is referred

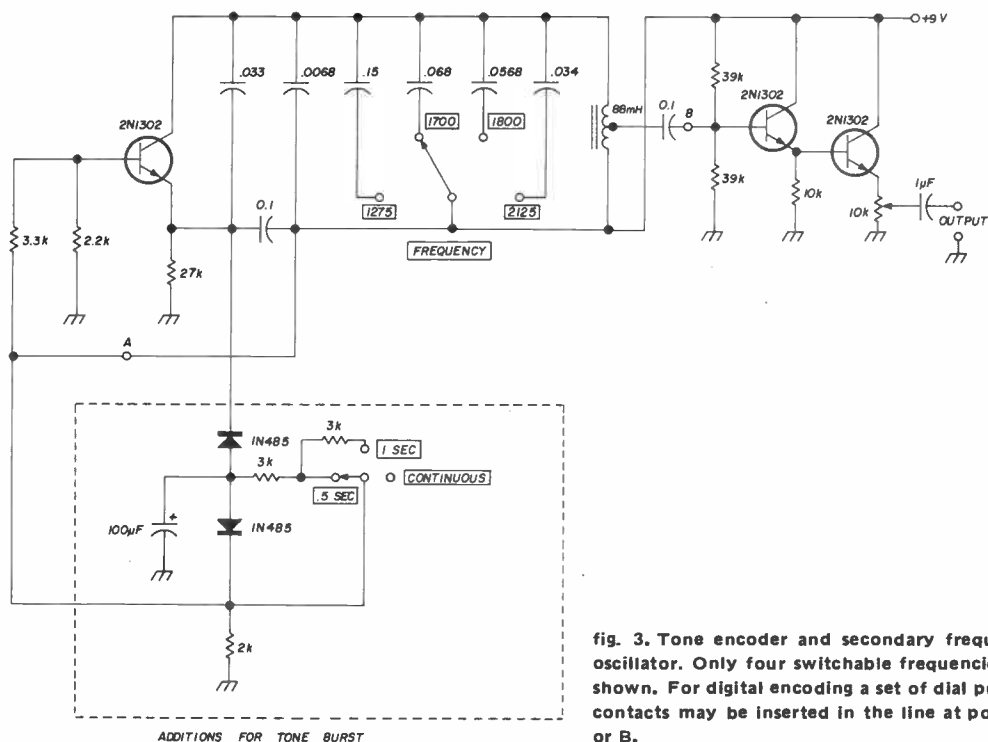


fig. 3. Tone encoder and secondary frequency oscillator. Only four switchable frequencies are shown. For digital encoding a set of dial pulsing contacts may be inserted in the line at point or B.

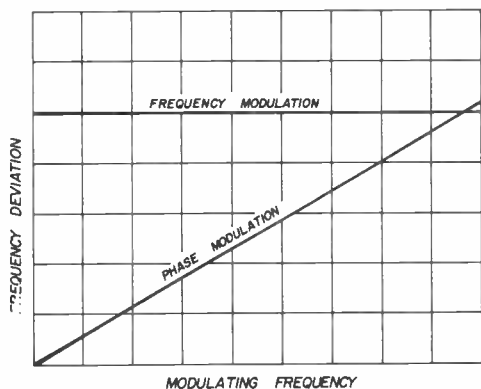


fig. 4. Modulating frequency dependence of fm and pm with constant audio input level.

to as narrowband. (Narrowband should not be confused with the nbfm permitted on the amateur bands below 30 MHz; nbfm is limited by regulation to $\pm 3\text{kHz}$.)

Narrowband may be copied on a wideband receiver with only a slight loss of audio, but wideband is not copyable on a narrowband receiver because of modulation excursions out of the receiver passband. When both types of equipment are in use, modulation levels are set for the narrower receivers.

audio response characteristics

Through use, "frequency modulation" has come to refer to any angular modulation system, either true fm or pm (phase modulation). Although the difference

A constant audio level applied to a frequency modulator will result in a certain frequency deviation which does not change with the modulating frequency. However, a constant audio level applied to a phase modulator will only result in a constant peak phase shift. The frequency deviation depends on how rapidly the phase shifts. Since the phase shift becomes more rapid as the modulating frequency is increased, the frequency deviation of a phase-modulated transmitter is directly proportional to the modulating frequency as shown in fig. 4.

The result is that a pm signal detected in an fm discriminator will have a 6 dB per octave rising audio characteristic. This can be overcome in one of two ways. If an RC network that will cause a 6-dB-per-octave *roll-off* across the entire audio range is placed in the transmitter audio (before the phase modulator) the transmitted signal will be identical to a true fm signal (fig. 5B). The alternative is to place the same RC circuit after the fm discriminator in the receiver (fig. 6A). In this case the system audio will still have a net flat response but the transmitted signal will be pm.

It is pm which is standard for commercial^{1,2} and amateur use. For this reason, when a frequency modulator is used an RC network with a 6-dB-per-octave *rising* characteristic is placed in the transmit audio circuit prior to the modu-

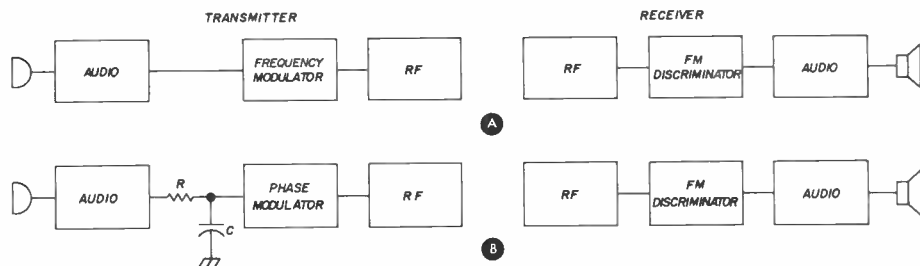


fig. 5. Frequency modulation and fm-equivalent system.

between an fm and a pm modulator is known, it is not widely realized that the two systems result in an inherent difference in audio-response characteristics.

lator (fig. 6B). If steps are not taken to assure standardized audio response different equipment combinations can result in either high-or low-pitched received

audio with accompanying loss in intelligibility.

The RC rolloff network used in the above examples should have a time constant of $RC=530$ microseconds for a low-frequency limit of 300 Hz. The rising response RC network for use with a frequency modulator should have a time constant of $R'C'=53$ microseconds for a high-frequency limit of 3 kHz (R in ohms, C in farads). The closest standard component values may be used.

An improvement in reception may be gained when slope detecting pm on an a-m receiver if audio rolloff is added as with the fm discriminator. Not only will the unnatural high pitched quality be eliminated, some noise reduction will result. A shunt capacitor may be selected for the proper time constant (530 microseconds may be used) in conjunction with an existing plate load resistor (see fig. 7). Provision should be made to switch the capacitor out for a-m reception. This arrangement is recommended for monitoring purposes only because of the inferior reception provided by slope detection. Also, tunable receivers are discouraged for fm communications because they encourage poor operating practices.

speech clipping

Speech clipping is a useful method of maintaining high average deviation levels without going beyond the receiver band-

it's obvious that fixed amplitude clipping by itself will not work unless it is made frequency dependent. This is normally done as shown in fig. 8 by preceding the clipper with a network with a 6-dB-per-octave *rising* characteristic. This enables the clipper to take a bigger bite of the higher frequencies. The clipper is followed by a 6-dB-per-octave *rolloff* net-

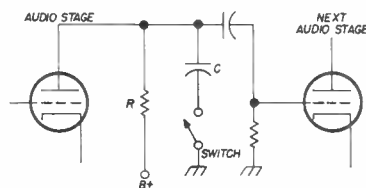


fig. 7. Modifying an a-m receiver to slope detect pm. Capacitor C and switch are added. Shunt circuit impedances are assumed to be high relative to R and are ignored in computing RC.

work that restores the unclipped audio to a flat response as shown in fig. 9. The net result is a pm signal clipped to a constant maximum frequency deviation.

When the audio clipper is used with a frequency modulator rather than a phase modulator, network RC is left out but $R'C'$ is left in. The resulting signal is the same as pm limited to a constant maximum frequency deviation.

It should be noted that excessive clipping with this method will cause a noticeable loss of high audio frequencies. However, at normal clipping levels the

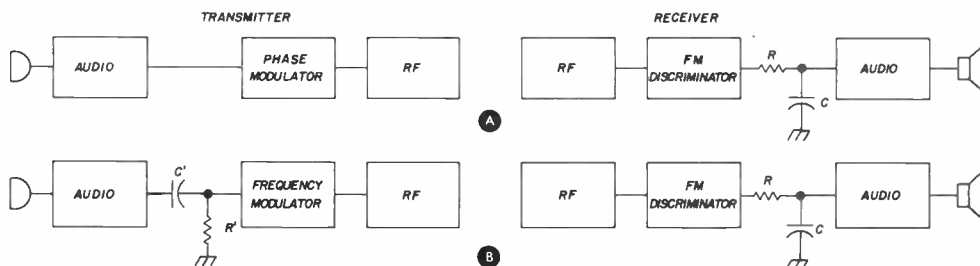


fig. 6. Phase modulation and pm-equivalent system.

pass. It has previously been established that the system in use is phase modulation; since pm exhibits a different deviation level for each modulating frequency

spectral distribution of speech is such that little high-frequency clipping takes place, and the highs appear normal. This loss effect has been noted on many

improperly adjusted repeaters around the country where the receiver is overdriving the clipper. Not only is excessive distortion created by too much clipping, but further degradation of intelligibility is

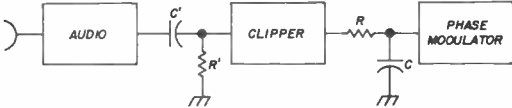


fig. 8. Speech clipping for constant maximum frequency deviation with phase modulation. $R'C' = 530$ microseconds and $RC = 53$ microseconds for 3-dB points at 300 and 3000 Hz.

caused by the muffled highs.

Despite the fact that fm is the general term applied to angular-modulated vhf and uhf work, the truth is that pm is the system in use from the point of view of

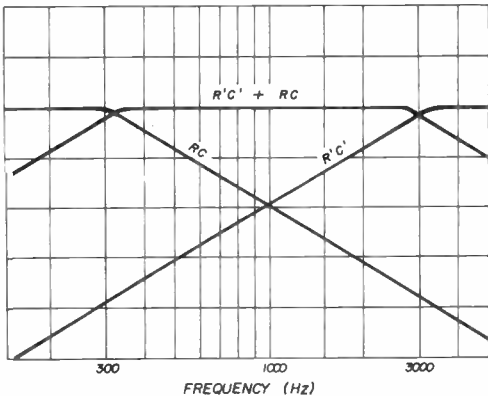


fig. 9. Normalized response for the circuit of fig. 8.

system audio response. Audio compensation must be used with fm modulators and detectors to maintain correct audio recovery for maximum intelligibility.

Modulation levels are restricted only by receiver bandwidths (except on those lower frequencies where the FCC specifies maximum bandwidths). Speech clipping is almost universally used but special audio frequency processing is necessary in the transmitter to limit a pm signal to a

constant maximum frequency deviation. Standard modulation levels are wideband (15-kHz deviation) and narrowband (5-kHz deviation).

references

1. EIA Standard RS-52A, "Minimum Standard for Land-Mobile Communications FM or PM Transmitters 25-470 Mc.," Electronic Industries Association, 1959, Section 6.
2. EIA Standard RS-204, "Minimum Standards for Land-Mobile Communications FM or PM Receivers," Electronic Industries Association, 1958, Section 11.

Les Cobb, W6TEE

isolating receivers in vhf fm repeaters

The popularity of vhf fm repeaters is steadily increasing in the ranks of amateur radio despite several problems inherent in the design and operation of

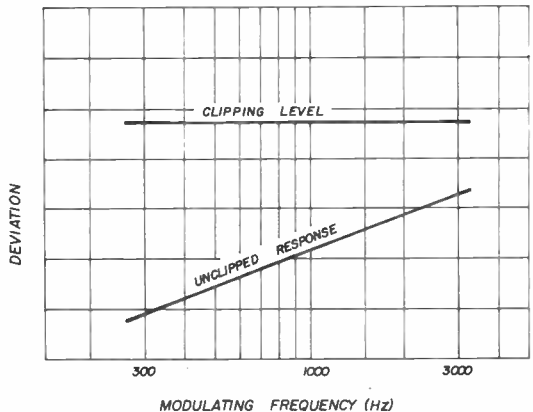


fig. 10. Deviation characteristics of the circuit in fig. 8.

such stations. Perhaps the most serious and least understood is receiver blocking, or more properly, receiver desensitization.

This problem appears in two ways. First, it can severely limit the effective talk-in range of the repeater system. It can also show up as repeater chatter which is cyclic keying of the station. First a signal breaks the receiver squelch and keys the transmitter. With the transmitter

on, receiver sensitivity falls off, and the squelch closes. When the squelch closes, the transmitter shuts off, and the receiver sensitivity returns to normal; if the signal is still present, the squelch opens and the cycle repeats.

noise

A transmitter will affect a nearby receiver's sensitivity in two ways. First, it can significantly reduce the receiver's front-end gain. This occurs when the transmitter carrier is present at the front end at a sufficient level to cause rectification in the amplifier or mixer input circuit. The resultant change in bias reduces the stage gain, which in turn reduces the noise input to the limiters, and the limiter current. The effect of all this is ultimately to reduce the receiver's

be many times greater than front-end or antenna noise. This increased noise input to the limiters produces an increase in limiter current, which in turn reduces receiver sensitivity.

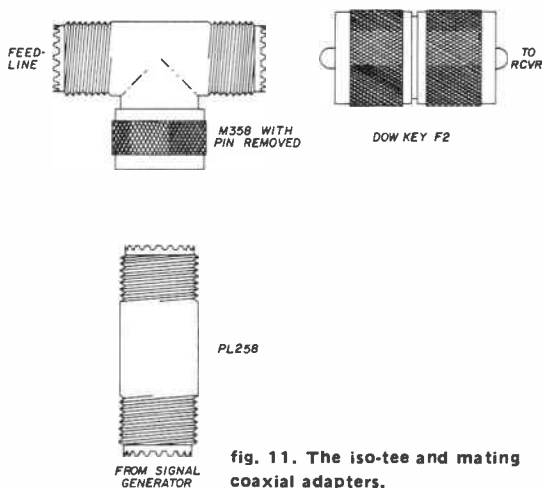
While rectification decreases limiter current, transmitter noise increases it. Thus, since both effects can occur simultaneously, it's not uncommon to find them canceling each other in regard to changing limiter current. Therefore, there may be severe receiver degradation with no apparent change in limiter current readings.

Before proceeding to a description of techniques for curing the problem, I'll develop a method for measuring the effects so corrective steps can be properly evaluated. The problem is basically how to measure receiver sensitivity, or to be more precise, the change in sensitivity. The term sensitivity refers not only to the receiver itself, but to the entire receiving system including antenna and feedline.

The basic piece of test gear required is a controlled signal source. A calibrated signal generator is ideal; however, most of us don't have one. Since we are primarily interested in changes of sensitivity, rather than the actual measurement of it to a hundredth microvolt, simple equipment can be used with great success. You'll require a well-shielded signal source, preferably crystal controlled, and a step attenuator. Both are within the pocket-book range of most hams, supplemented by home-brewing.

Also required is an *iso-tee*. This is simply a coaxial tee adapter, such as the M-358, with the center pin of the male termination removed (fig. 11). The female terminations are connected to the receiver and the antenna feedline. The output of the signal source is connected through the step attenuator to the male termination of the tee. The tee provides very loose coupling of the signal source to the receiving system.

To measure the receiver degradation produced by the transmitter, connect an ac voltmeter across the receiver speaker terminals (see fig. 12). Open the receiver squelch and increase the output of the



sensitivity.

The second effect is produced by the transmitter's noise spectrum. It is a sad but true fact that all transmitters produce not only a carrier and modulation sidebands, but noise sidebands as well. These sidebands may extend several hundred kilohertz on either side of the carrier. If the transmitter and receiver frequencies are only a few hundred kilohertz apart, which is usually the case in amateur repeaters, the transmitter noise output that lands on the receiver frequency can

signal source until the receiver output decreases by 20 dB, just as when making a standard 20-dB quieting sensitivity check. It will probably require several hundred microvolts of rf output from the attenuator to produce this quieting, as the isolation of the tee fitting is substantial.

The next step is to key the transmitter and repeat the measurement. Ideally, it should take the same amount of signal to quiet the receiver with the transmitter on as with it off. More likely, it will require considerably more signal. If, for instance, you must inject 30 dB more signal to produce 20 dB quieting with the transmitter on, the transmitter is degrading the receiver sensitivity by 30 dB! Now the question is, what to do about it?

the cure

If the transmitter and receiver are located at the same site, the first step is to shield the two units. This includes filtering all power, control and audio leads entering or leaving the shielded enclosure. Shield kits for this purpose are available from the manufacturers, or you can brew your own. The effectiveness of the shielding can be tested by the same technique previously outlined. To test the transmitter shielding, replace the transmitter antenna with a dummy load and make the quieting measurement. To test the receiver shielding, replace the receiving antenna with a dummy load and perform the measurements.

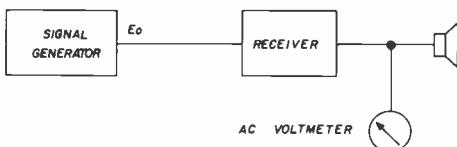
If the shielding is effective, no degradation will be apparent during either test. Of course, this presumes no radiation from the dummy load, so forget that light bulb nonsense. Once the units are properly shielded, reconnect the antennas and check for degradation. If it is still severe or objectionable, the next step is antenna spacing.

The transmit and receive antennas should be as far apart as possible. If they are mounted on the same tower, they should be at least 100 or more feet apart. Unfortunately, few hams can realize the benefits of 200-foot towers or separate transmitter and receiver sites, so a more

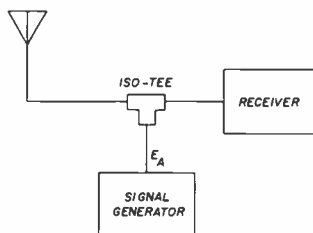
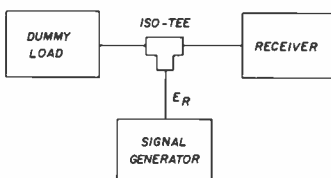
practical solution is the use of tuned cavities.

tuned cavities

A tuned cavity is essentially a very high-Q tuned circuit. A cavity placed in the receiver feedline and tuned to the receiver frequency will pass signals on the



BASIC SENSITIVITY = E_0 = VOLTAGE FOR 20 dB QUIETING



$$\text{EFFECTIVE SENSITIVITY} = E_0 \left[\frac{E_A}{E_R} \right]$$

$$\text{DEGRADATION DUE TO ANTENNA NOISE (dB)} = 20 \log \frac{E_A}{E_R}$$

$$\text{DEGRADATION DUE TO TRANSMITTER (dB)} = 20 \log \frac{E_A \text{ XMTR "ON"}}{E_A \text{ XMTR "OFF"}}$$

fig. 12. Test equipment set up for measuring receiving-system degradation.

receive frequency, while rejecting all other frequencies. A cavity placed in the transmitter feedline and tuned to the Transmitter frequency will pass the transmitter signal, while rejecting noise on the receiver frequency.

The criteria for selecting a cavity, either commercial or homebrew, are physical size and power rating. As a rule of thumb, the bigger they are, the better.

Many commercial units are available with a choice of coupling loops, thus providing a choice of different amounts of selectivity. There is a tradeoff involved here as the higher the selectivity, the greater the insertion loss and the lower the power rating.

To determine the appropriate location for the cavity, watch the limiter current as the transmitter is keyed. An increase indicates degradation due to transmitter noise, and the cavity should be placed in the transmitter feedline. A decrease indicates desensitization due to rectification, and the cavity should be placed in the receiver feedline.

If the transmitter noise is the culprit and a separate exciter and final are used, try placing the cavity between the exciter and the final. Since most of the transmitter noise is generated in multiplication stages, a cavity at the output of the exciter will reduce the noise output of the amplifier. An added benefit of this location is the fact that, since you are dealing with a low-level signal, you can use a cavity with a higher selectivity (remember the tradeoff) thus improving the noise reduction. Also, due to the reserve gain of the final amplifier, insertion at this point will result in less loss of radiated power than if the same filter were placed after the final.

After the first cavity is installed, a noticeable improvement should be observed. If some degradation still exists, it may be necessary to resort to additional cavities. Complete elimination of the problem may require several cavities in both the transmitter and receiver feedlines.

Many hams will object to the use of cavities because of their insertion loss. However, from a system viewpoint, they are the least of several ills. A repeater is useful only to stations who can both hear and be heard by it. If your repeater can be heard across the state, but it can only hear across town, then the repeater's effective range is just across town. Stations outside the talk-in range can't use the repeater and so are probably not very interested in hearing it. As far as

they are concerned, it just ties up the frequency.

If you install a cavity or two on the transmitter, you may reduce your talk-out range to halfway across the state, but you may have reduced receiver desensitization to the point where you can hear halfway across the state. Thus, you've increased your effective range to halfway across the state, insertion loss and all.

Similarly, in the case of the receiver, what really counts is not just receiver sensitivity, but effective sensitivity when the transmitter is keyed. If adding cavities to the receiver feedline reduces the degradation produced by the transmitter, then you are improving performance and increasing operational range even at the expense of additional insertion losses.

receiver sensitivity

An interesting point regarding receivers arises at this point. If a receiver is capable of hearing external noise from the antenna, this noise is the limiting factor in weak signal detection. Low noise preamps and high gain antennas won't help matters, and they may produce additional problems. With an *a-m* receiver, the most simple test for sensitivity is to replace the antenna with a dummy load and see if the noise output of the speaker decreases. If it does, the receiver is hearing external noise. This test won't work with fm receivers; and if it does, you'd better start replacing tubes.

An equivalent test for fm receivers can be made by watching limiter current, but a much more accurate method is to perform the same test as used for measuring degradation due to transmitter noise. First, make the quieting check with a dummy load on the receiver. Then replace the normal antenna, and again check the quieting level. The difference in levels required to produce 20 dB of quieting is the amount of external degradation.

If the receiver sensitivity is being degraded by, say, 8 dB of antenna noise, then up to 8 dB of additional loss, either in the form of cavity insertion loss or attenuators, can be inserted in the feed-

line without affecting the effective sensitivity by more than 2 or 3 dB. The attenuator, by the way, will provide some improvement in desensing characteristics and considerable improvement in inter-modulation protection.

If the receiver is not being degraded at all by antenna noise, then the limiting factor is the receiver front-end noise, and you may want to add a low-noise preamp or switch to a higher-gain antenna. If you decide on a preamp, be careful. Many preamps, especially those using bipolar transistors, are extremely susceptible to desensing problems. The same unit that works wonders in a mobile or base station may prove disastrous in a repeater.

summary

The steps toward elimination of receiving system degradation in repeater applications are transmitter and receiver shielding, spacing between antennas and the proper use of tuned cavities. With enough cavities, it's even practical to use the same antenna simultaneously for transmitting and receiving.

J. A. Murphy, K5ZBA

improving the fm repeater transmitter

A large number of amateur 144-MHz repeaters and remote-base stations make use of the Motorola PA-8491 60-watt highband transmitter. This transmitter is ideally suited for repeater use but a few simple modifications will greatly improve its operation.

This transmitter, similar to units produced by Motorola and others, uses relatively low-Q tuned circuits. The low-Q circuitry permits moving the transmitter frequency a megahertz or more without retuning, and also permits replacement of oscillator and multiplier tubes with essentially no retuning required. In amateur repeater service these advantages, however, are far overshadowed by the problems generated by the broad spectrum of noise that the low-Q circuits permit the transmitter to emit.

The "white noise" is the cause of most receiver desensitization in amateur repeaters, especially in close-spaced systems that receive on 146.34 MHz and transmit on 146.76 or 146.94 MHz. The broad

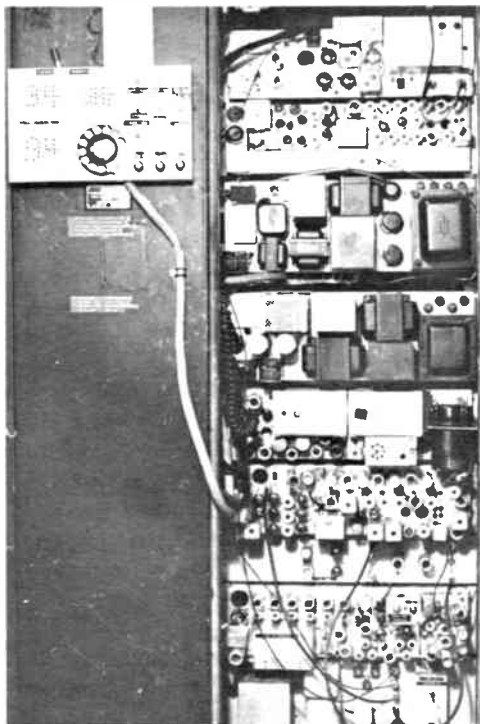


fig. 13. Grass Valley repeater station, Wolf Mountain, California. This repeater covers most of the Sacramento Valley.

noise spectrum from the transmitter covers weak signals on the receive frequency, causing the repeater receiver to be less sensitive with the repeater transmitter on than it is when the transmitter is off. This problem was covered very well by K5ZBA on page 28.

This desensitization usually results in repeater "pumping." This occurs when the signal received at the repeater receiver is strong enough to actuate the repeater when the repeater transmitter is off, but is not strong enough to keep the repeater on when the repeater transmitter (and its white noise) is on. Under these conditions the repeater transmitter comes on, gener-

ates white noise that causes the repeater receiver to lose the incoming signal, and the repeater transmitter turns off. As soon as the transmitter is off, the receiver again detects the incoming signal and the cycle repeats itself.

desensitization measurements

While preparing a repeater remote station package for mountaintop installation, desensitization tests were made under laboratory conditions. The 146.94

the zero-to-50-dB attenuator and an unmodified Motorola Sensicon A receiver.

Receiver sensitivity was measured through the probe with the transmitter off; then the transmitter was keyed, and the zero-to-50-dB pad was adjusted to produce a loss in sensitivity of the receiver. This loss in sensitivity (desensitization) was measured as the number of dB of signal power increase required to produce the same amount of quieting (20 dB was used for these tests) as when the transmitter was off.

This measurement was performed with various amounts of transmitter power fed through the zero-to-50-dB attenuator into the receiver, thus simulating different amounts of antenna coupling loss; this loss is equal to the zero-to-50-dB attenuator setting plus the 30-dB probe loss.

It was initially determined that before we modified the repeater transmitter, 79 dB of coupling loss produced 3 dB of receiver desensitization. This is equivalent to more than 60 feet of vertical separation between dipole antennas for receive and transmit. After modifying the transmitter, 41 dB of coupling loss produced 3 dB of desensitization; this can be obtained with 15 feet vertical separation between dipoles, and represents 38 dB improvement.

Note that for these tests coupling from transmitter to receiver must be only through the zero-to-50-dB attenuator. Direct radiation coupling must be incidental. This can be accomplished by separating the transmitter from the receiver in the rack, and placing other equipment in between. Shielding was not found to be necessary.

Transmitter modification consists of raising the circuit Q of the intermediate tuned circuits. This is done by adding capacitance across the tuned circuits in question, and in one case by reducing the coupling between tuned circuits. It is possible to obtain a couple of dB of additional benefit by critical slight detuning of the transmitter tuned circuits; however, this added improvement is not generally worth the added difficulty of adjustment.

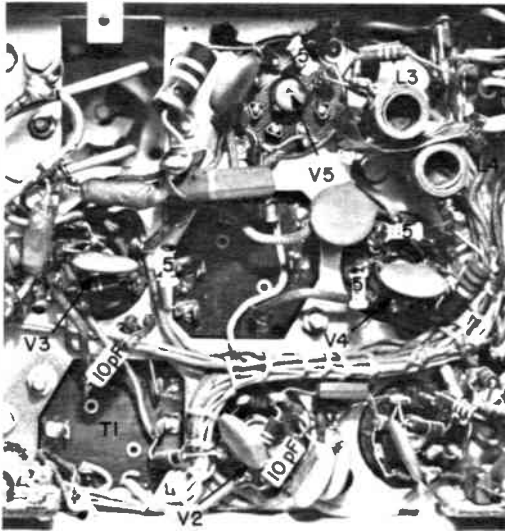


fig. 14. Bottom view of chassis showing added components.

MHz transmitter was fed into a dummy load through a Bird Thru-line wattmeter and adjusted for 60 watts output. A probe (such as the *iso-tee*) was coupled to the dummy load, through a 10-dB 50-ohm attenuator to a milliwatt meter; the probe was adjusted for 60 mW indicated on the milliwattmeter (this represents a 30 dB loss from the transmitter to the milliwatt meter).

The connection from the 10-dB attenuator to the milliwatt meter was then disconnected and connected instead to the receiver input through a precision zero-to-50-dB 50-ohm attenuator. A signal generator was coupled through a 40-dB-loss probe into the line between

transmitter modifications

The necessary transmitter modifications are quite simple, and are shown in the accompanying photographs.

1. Install two high quality 10-pF tubular ceramic capacitors across transformer T1: one capacitor from the plate of V2 to ground, the other from the grid of V3 to ground. These new capacitors are shown in fig. 14.

2. Install high-quality 5-pF tubular ceramic capacitors across T2, L3 and L4 as follows: from the plate of V3 to ground, from the grid of V4 to ground, from the plate of V4 to ground, and from the grid of V5 to ground.

3. Install a 1-to-8-pF variable capacitor across the grids of the final amplifier. The E. F. Johnson 160-104 capacitor shown in fig. 15 does an excellent job but any similar air variable should work as well. This modification increases grid drive, improves efficiency, and raises the power output of the 829B final. An insulated tool must be used for adjusting this capacitor.

4. The coils at the plate of V4 and the grid of V4 (L3 and L4) are overcoupled, thus lowering the effective Q of this interstage network. Decouple these coils by mounting a small copper tab between them (fig. 16). Adjust the size of the tab so that after retuning, the driver grid drive is slightly less than before; an indicated drop of one microampere as measured at test position 5 should be about optimum. You will notice that the coils tune sharper as you decouple them; this is an indication of increased circuit Q.

results

At our installation, we measure 64 dB of coupling loss between two three-element Yagis spaced 18 feet vertically on our tower. This is an acceptable figure for the modified transmitter. The top antenna is used for transmit on 146.94, and the bottom antenna for receive on 146.34 MHz. A 146.94 MHz receiver is also

connected through the transmitter antenna relay to the upper antenna. Both 146.94 and 146.34 are retransmitted as received to the control stations on the 440 MHz "down channel." No cavities are used in either the transmitter or receiver feedlines. With a weak signal (15 dB of quieting — about 0.3 microvolt) being received on 146.34, no change in limiter current or in quieting is noted by switching the 146.94 MHz transmitter on and off.

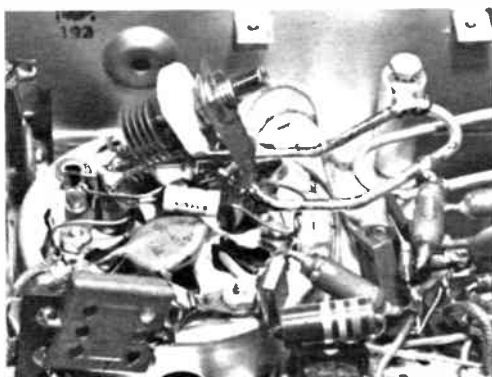


fig. 15. This photo shows the 8-pF air variable which is installed between the grids of the final amplifier.

One note of caution, however; the swr must be low on both antenna feedlines. When the swr is high, the feedlines become part of the antenna, and since the feedlines usually run parallel to each other, the coupling loss between antennas is lowered.

It is a good idea to measure the coupling loss between antennas to determine if it is adequate. The coupling loss between the two antennas may be measured by first connecting the signal generator (on 146.34 MHz) through a 6- to 10-dB attenuator to the 146.94-MHz antenna. Connect the 146.34-MHz antenna through a 6- to 10-dB attenuator to the 146.34-MHz receiver. Adjust the signal generator for a reference limiter current reading in the receiver. Now disconnect the attenuators from the antennas, and connect the attenuators to each other.

The signal generator is then connected through two attenuators to the receiver. Reduce the generator attenuator loss until the reference limiter reading is again obtained. The difference in attenuator settings on the signal generator is the coupling loss.

J. Jay O'Brien, W6GDO

improving the voice commander

Many requests have come to my attention for help with the General Electric Voice Commander. Many hams aren't happy with these sets because of receiver instability. This article offers a means for resolving the receiver instability problem and provides information for converting narrowband receivers to wideband operation. A couple of minor but useful transmitter modifications are also described.

The Voice Commander was manufactured in three models: Voice Commander I, II, and III. Model I used subminiature tubes in the transmitter, but the receiver is essentially the same as in the other models. Components and physical layout for the receivers were changed only slightly throughout the entire model

series. Models II and III have nearly identical receivers.

An rf amplifier was added to the receiver of the Model III. This circuit is on a separate chassis hidden under the transmitter board. Also included is a ptt relay for an external microphone. In this article, all references are to the Model III; however, they apply to Model II except for the part about the receiver rf amplifier.

The Voice Commanders were manufactured in two production splits. Split 1 covered 132-150 MHz; split 2 covered 150-174 MHz. Most high-split units will tune to 146-147 MHz with no trouble. Information is included here on appropriate padding to accomplish this tuning range.

receiver rf amplifier

The receiver rf amplifier is model 4EA19A10 (low split), or 4EA19A11 (high split). The original unit used G. E. part no. 19C300037-2 transistor. It was replaced in revision A with part no. 19A115413-1 (2N2996). The higher gain of this transistor produced oscillation in some receivers. This led to revision B, in which the collector was moved to a tap on the output coil, and a 10k ¼-watt resistor was added across the coil.

If your receiver is revision A, the 10k resistor is sufficient to tame it. To pad from high to low split, which is normally not necessary, add a 3-pF capacitor across input and output coils.

The basic complaint of receiver instability can be corrected by following the steps outlined below, in which dc voltage distribution is rearranged to eliminate a tendency toward regeneration and oscillation. These changes were incorporated by G. E. into later versions of the Voice Commander. A manual for the Voice Commander III is available from G. E.*

The first modification step is to identify the chassis. Inside the square can are three small chassis with a lid on each.

*General Electric Co., Box 4197, A & SP, Lynchburg, Virginia 24503. Enclose check or money order for \$1.00.



fig. 16. The decoupling tab between coils L3 and L4 increases circuit Q.

Remove the lids. The chassis are identified as follows, although the numbers are sometimes hard to see.

1. 4EL13A10 (low i-f gain and discriminator board).

2. 4EF29A10 (narrowband) or 4EF-29B10 (wideband) high i-f gain second-oscillator/mixer and low i-f filter board.

3. 4EF14A10 (low split) or 4EF14A11 (high split) front-end board.

Begin with the low i-f gain and discriminator board. At each end of the board is a black lead running to the high i-f gain and low i-f filter board. Unsolder the black lead at hole 14 (near the discriminator transformer) and let it hang loose for now. Move the other black lead from hole 13 to the top of R3 (2.2k, $\frac{1}{4}$ W) next to hole 13. You're now temporarily finished with the discriminator board.

Remove the high i-f gain and low i-f filter board (center chassis) and turn it over to the solder side. You'll see a wire in sleeving connected from a point in the center of the board (hole 4) to one end. The end connection is at hole 9; remember its location and unsolder the end of the wire from this hole. Move the end of the wire to the opposite end of the board. The black wire hanging there is in hole 14. (You disconnected the other end from the discriminator board.) Remove this black wire and discard. Connect the wire previously removed from hole 9 to hole 14. Install a 2K, $\frac{1}{4}$ -watt resistor between hole 4 and hole 9. Use sleeving and dress the leads carefully.

Now turn the board over and locate R2, a 6200-ohm, $\frac{1}{4}$ -watt resistor next to hole 13. Install a 0.047- μ F capacitor from the top of R2 to hole 13. Unless you have very small components, I recommend that you order the capacitor from a G. E. service center. The part number is 5492638-P6.

Unsolder the black wire from hole 13 and let it hang. This wire connects to the 4EF14 front-end board.

If your Voice Commander is a narrow-band unit and you wish to convert it for wideband use, you'll need a 47k, $\frac{1}{4}$ -watt

resistor, a miniature 1000-pF capacitor (G. E. part no. 5491500-P7), a miniature 1200-pF capacitor (part no. 5491500-P8), and a miniature 12-pF capacitor (part no. 5495334-P42).

Looking at the top of the high i-f gain and low i-f filter board, locate Q2, which is near the center of the long side and next to one of the four slug-tuned coils. (Q2 is a TO-18-size transistor.) Locate the two small holes for the 47k resistor between Q2 and the slug-tuned coil. The 47k, $\frac{1}{4}$ -watt resistor will fit into this space, although it's a squeeze. In the area bounded by the four coils are seven capacitors. Four of these are 110 pF. Leave these alone. Replace the 3300-pF, 2.7-pF, and 4700-pF capacitors with 1000-pF, 12-pF, and 1200-pF capacitors respectively.

If your unit is one of the high-split sets and you wish to convert it to low-split operation, remove front-end board 4EF14A11. Locate the three air-wound coils, and pad each coil with a 4-pF capacitor. Turn the board over, and you'll see a short, black wire hanging from J2. Remove and discard this wire. Install a 5½-inch-long length of black no. 24 wire in J2.

The first-oscillator crystal is interchangeable between Voice Commanders I, II, III, the Progress Line portable, the transistorized portable (TPL), and Voice Director receivers — in case you want to borrow one to check it out.

reassembly and alignment

At this time, replace the three chassis into the square can. Use care, especially with the center board with the added 2k resistor on the bottom. Dress the black wire hanging from J2 on the front-end board through the same slot containing the coax cable. Run the black wire along the edge to the audio-squelch board. (This is a square board with a round hole in the center.) This board will be identified as 4EA18A10 (narrowband) or 4EA18A11 (wideband). Only the purist will worry about the difference here.

Connect the black wire from the

front-end board to hole 8, the location on the audio-squelch board already having a black wire going to the low i-f discriminator board. Do not disconnect the wire already attached to hole 8.

The receiver is now ready for alignment. The low i-f is 290 kHz; the high i-f is 8.7 MHz. Test point 1 is the limiter metering, and test point 2 is the discriminator secondary metering. Be careful of the four slug-tuned coils in the low i-f filter, as a strand of rubber is usually

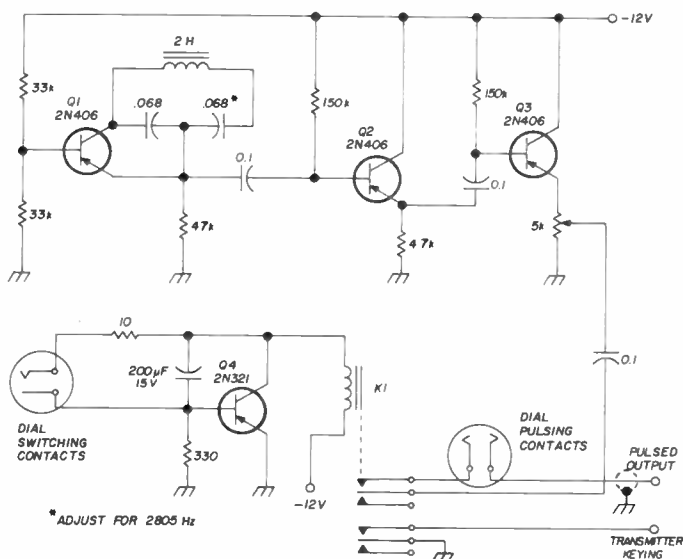
With the modifications described in this article installed in your Voice Commander you will have a unit that is a pleasure to use.

Don Chase, WØDKU

fm sequential encoder

A stable pulsed encoder for controlling fm repeaters is a relatively simple device requiring only a few components and a telephone dial. Since more and more fm

fig. 17. Circuit for the portable sequential encoder. Relay K1 is a 12-Vdc relay with dc resistance between 240 and 560 ohms.



placed inside as a friction device and it has a strange feel on the tuning tool.

transmitter

The Voice Commander II and III transmitters are simple and straightforward. On some models, the crystal plugs in; on others it's soldered in. A tuning chart is mounted inside the back cover of all units. The only coils needing alteration are those in the driver and final, and a simple squeeze will suffice.

The audio system of the Voice Commander transmitter is pretty hot. If you wish to speak close to the microphone, reduce the sensitivity by shunting a 390 or 470-ohm resistor across the microphone cartridge.

repeaters are using pulse-type decoders and additional control it is convenient to have a small pulse encoder on hand whenever you need it.

The pulsed encoder in fig. 17 uses low cost parts, is easy to build, and provides an excellent sequential encoder for mobile fm operation.

All components except the toroid oscillator coil should be available at your local parts store. Although I used a 2-Henry toroid in the circuit in fig. 17 the surplus 88-mH toroids used by RTTY enthusiasts will do an excellent job if you juggle the values of the tone determining capacitors.

The circuit in fig. 17 consists of a 2805-Hz oscillator (Q1), buffer stage

(Q2) and emitter follower (Q3). The oscillator tone is set to the desired frequency by adjusting the .068- μ F capacitor marked with an asterisk. In some cases it may be necessary to put two capacitors in parallel to obtain the desired tone.

The 3-second hold circuit (Q4) provides a 3-second carrier and tone at the end of the pulsing sequence. This allows for selective calling of a control pulse with "hands off" capability. I used an Allied 12-Vdc dpdt relay with a coil resistance of approximately 560 ohms at K1. This is not critical however, and almost any 12-Vdc relay will work. Transistor Q4 heats up a little, but since control is not continuous it is not necessary to use a heavy-duty transistor.

If you use this circuit without a separate power supply (a 9-volt transistor radio battery will work), you must bypass circuit ground or *tone low* through a 0.1 μ F capacitor to isolate the positive voltage from system ground. Since the circuit is open (no tone) until it is dialed and the relay picks up, the unit can be connected directly across the microphone input of your transmitter. The output matches the inputs of most fm transceivers (except those with carbon microphones).

John S. Hollar, W3JJU

motorola fm receiver mods

During the past year I have retuned several 43G and 43GGV Motorola Sensicon transceivers for amateur use on two meters. In some of the units I discovered that the squelch control had to be advanced almost fully clockwise to quiet the channel. After checking tubes, components and voltages and a careful alignment I still could not find the problem.

Checking through a number of manuals, I found some resistor changes in the schematic for a late T-Power receiver strip. When these resistor values were incorporated in an otherwise identical older version of the receiver the result was vastly improved squelch-control

table 2. New resistor values for the Motorola Sensicon G fm receivers.

resistor	old value	new value	
		wide band	narrow band
R139	—	56k	180k
R140	—	56k	180k
R145	150k	330k	330k
R146	—	680	220
R152	560k	820k	1.5M

action. The new resistor values are shown in table2.

To provide a measure of protection to the 6AQ5 audio output stage I installed a 120-ohm, 1-watt resistor between the cathode (pin 2) and ground, bypassed with a 50- μ F, 25-V capacitor. Gain of the stage is not greatly affected, and the tube will be protected by self-bias.

Table 3 lists a number of dc voltage readings at various points in the squelch and audio stages. These should be helpful in tracking down circuit ailments.

Murray Ronald, VE4RE

phase-locked loop fm detector

The circuit in fig. 18 shows how the Signetics NE561B may be used as an fm detector. The fm input signal is connected to pin 12 and 13; audio output appears across the 15k load resistor connected to pin 9. The capacitor between pin 10 and ground establishes the proper de-emphasis.

table 3. Squelch and audio-voltage readings are helpful in troubleshooting Motorola Sensicon G fm receivers. (Vrms readings are noise voltage measurements made with ac vtvm.)

test point	unsquelched	squelched
V112, pin 1	3 V rms	3 V rms
V112, pin 2	7 Vdc	2 Vdc
V112, pin 5	50 V rms	50 V rms
	190 Vdc	100 Vdc
V112, pin 6	100 Vdc	60 Vdc
V113, pin 6	12 V rms	12 V rms
V114, pin 1	45 Vdc	25 Vdc
V114, pin 2	-8 Vdc	2 Vdc
V114, pin 3	3 Vdc	3 Vdc
V114, pin 6	180 Vdc	190 Vdc
V114, pin 8	50 Vdc	50 Vdc

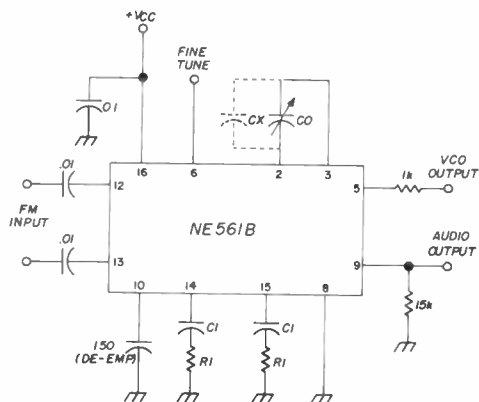


fig. 18. Using the phase-locked loop IC as an fm detector.

The capacitor between pins 2 and 3 determines the frequency of the voltage-controlled oscillator. Typical capacitor values for intermediate frequencies of 4.5

and 10.7 MHz are given in fig. 18. For fine tuning a small capacitor may be placed in parallel with C_o (C_x in fig. 18) or a small variable voltage may be applied to pin 6. This can be in the form of a potentiometer connected across the power supply. The low-pass loop filter components are connected between pins 14 and 15. Component values, which depend on the i-f frequency, are shown.

The frequency range over which the vco will track the signal frequency can be altered by applying appropriate dc control voltages to pin 7 from an external and adjustable supply voltage source. In some specialized application it may be necessary to offset an internal dc component. This can be done with an external voltage source applied to pin 11.

Ed Noll, W3FQJ

keying and control

contest keyer

I built this gadget, the Scratch-1 programmable keyer, for the 1969 ARRL Field Day Contest. It was designed to replace an old keyer that used the rectified output of a tape recorder. However, the Scratch-1 offers several advantages: small size, no power connections, variable speed and access to different parts of the message.

Although the basic concept is quite simple, a number of prototypes were built to find an easy and durable keying system. The unit shown in the photo uses Scotch no. 49 aluminum tape for the keying element. As a preventative measure, a Q-Tip dipped in NO-OX (contact cleaner) was run over the keyer contact surface just before the field day contest. The keyer performed perfectly and shows little wear after completing over 300 contacts.

Since the code characters are fairly small, the smoothness of keying is critically dependent on how smoothly the

operator can scratch the stylus down the base. I wanted a compact unit so I used quarter-inch dashes, but I would recommend making them a little longer to facilitate smooth operation. This keyer was designed for low-power applications such as grid-block keying — it probably won't take the rigors of cathode keying, particularly in medium to high-powered transmitters.

Martin Davidoff, K2UBC

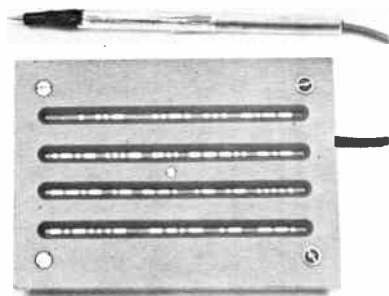
simple grid-block keying

The merits of a differential grid-block keying system are discussed at length in many of the standard handbooks. However, most of the circuits described in the literature involve one or more vacuum tubes, many parts, and in general cause the builder to shy away from all that hardware and stick to a simpler, *poorer* system.

The system shown in fig. 1 is very simple, involves no active elements, and does a nice job of providing true differential keying at minimum cost. It is simply a voltage divider, with a zener diode providing the differential function normally provided by a voltage-regulator tube.

To design the system, determine the proper operating grid voltage and current to place your final amplifier in class-C operation. Resistors R1 and R2 are calculated as follows:

$$R1 = \frac{E_{gg}}{I_{g1}}$$



The Scratch-1 programmable contest keyer.

$$R2 = \frac{R1 \times E_{g1}}{2E_{gg} - E_{g1}}$$

where E_{gg} is the negative bias voltage, E_{g1} is the final grid bias and I_{g1} is the final grid current.

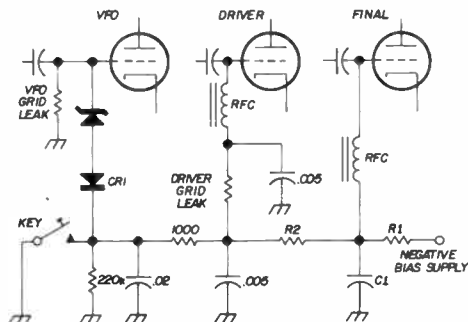


fig. 1. Simple grid-block keying system.

The zener voltage is chosen by subtracting the vfo bias (key down) from the supply bias and choosing a zener diode with a breakdown voltage smaller than this value. Several values should be tried to get optimum keying. C1 is chosen by trial and error to get the proper keying waveform, according to your particular taste. CR1 is any general-purpose diode.

In operation, power-amplifier bias falls exponentially as the key is closed. However, at a trigger point well before the driver and final conduct, the zener stops conducting and the vfo is turned full on, while the other stages become operative later (depending on capacitor C1). When the key is released the final and driver gradually turn off, but the vfo remains on until the bias is nearly at its peak value, at which time the zener conducts, placing bias on the vfo to keep it from oscillating.

The circuit provides a minimum bias on the final of one-half the negative bias supply. This protects the final amplifier stage in case excitation is lost. Be sure that the grid leak in the vfo is at least 4700 ohms.

This circuit is a variation of the design used in the Heath HW-16 transceiver. Existing grid-block keying systems can be

made differential by simply adding the two diodes to the vfo grid circuit.

Dick Blasco, WA4DHU

Independent keying of Heath HG-10B vfo

For owners of the Heathkit HG-10B vfo who might wish to make it independent of other equipment, the following idea will be helpful. My vfo is keyed with a Heathkit HD-10 keyer. The problem was how to use this model keyer with the vfo. The instruction manual for the keyer warns against using a voltage above -105 volts.

The circuit of fig. 2 shows how I solved the problem. All components except the transformer and fuse holder were mounted on a small piece of Vector board. The existing 4-conductor cable was removed and replaced with a length of ac cable. The fuse holder was mounted between the key jack and ac cable. The transformer, an inexpensive Japanese unit, was mounted vertically on the underside of the chassis next to the mode-switch wafer. The Vector board will fit nicely between the transformer

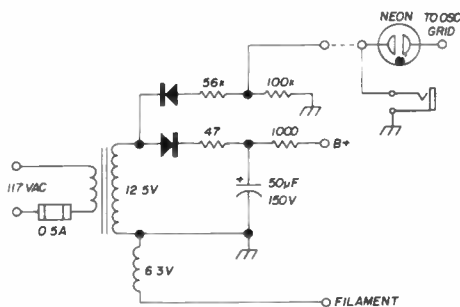


fig. 2. Isolation circuit for use with Heath HG-10B vfo and HD-10 keyer.

and the rear chassis wall next to the terminal strip holding the neon lamp.

The connection from the key jack was removed from the cathode circuit and connected as shown. No switch was used, as I intended to leave the vfo on constantly.

James H. Crouch, K4RR

transistor replaces relay

About five years ago I built a TO-type keyer based on Jim Ricks' circuit. The relay that I used originally was not the recommended (and expensive) type and it lasted about three months. I replaced it with another one that was hermetically sealed; a surplus unit that originally cost someone about \$55. It lasted about nine months.

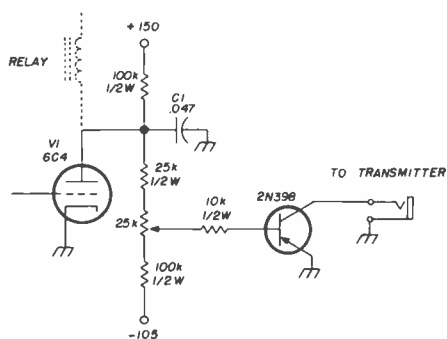


fig. 3. Solid-state keyer for the W9TO keyer. C1 eliminates any transients generated by the multivibrator.

So I tried a cheapie — an open-frame relay that I picked up at a surplus store for 29c. Surprisingly, it lasted four years and was still going when I decided to replace it with a transistor.

The idea was to do a minimum of reworking on the TO circuitry. Fig. 3 shows the changes. V1 is the relay driver tube; the relay coil is disconnected from the plate, and the plate is tied to the resistor network shown in the diagram. The 25k potentiometer can be a miniature unit put in some inconspicuous but accessible place; once it's set, it can be forgotten.

The output to the transmitter is the transistor collector. This type of keying can only be applied to a system using grid-block keying where the keyer shorts a negative voltage to ground through a relatively high resistance.

After completing the modification,

turn the potentiometer fully counter-clockwise before turning on the power. Recheck your wiring to be sure. Turn on the transmitter and the keyer. When warmed up, the transmitter will come on as though you were holding down the key. Press and hold the dot paddle and, while doing so, slowly turn the pot clockwise until you see that the transmitter is being keyed by the dots. To acquire a feel for the marginal area, turn the pot back and forth and notice the transmitter reaction. Turn the potentiometer clockwise a little past the marginal area — that's it.

The purpose of C1 is to eliminate any transients generated by the primary free-running flip-flop. The transient occurs during the dash period when the free-running multivibrator goes negative. When the relay was used, the transients were not apparent because of the relay's inertia and the short time of the transient. But with a fast transistor, the transient could (and did) appear on the carrier. C1 was selected to provide adequate filtering without materially affecting the dot-space ratio. It did have a slight effect which was compensated by readjusting the keyer weight control.

It's a great improvement. No more relay chatter (which the wife appreciates). No more contact bounce, no relay

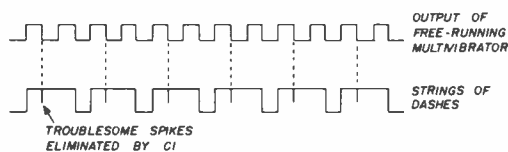


fig. 4. Wavetrain when the dash paddle is keyed. The multivibrator transients weren't a problem with the original keying relay because of the inertia of the armature.

replacements to make, and it's really fast!

A note on the 2N398: The technical data shows that it can hold off 105 volts. The voltage that it's holding off in my Heath Apache is 105 volts. This is really riding it close. However, it's been on the air every night for several months and is still going strong.

Frank Case, W3NK

painless vox keying

The convenience and speed of vox keying is well known by those who have this feature in their transmitter or transceiver. With this method, changing from receive to transmit mode occurs at the first touch of the key, which activates the vox circuit normally used on phone. A recently purchased transceiver, while excellent in many other respects, didn't have this feature. So I incorporated the keying method shown in fig. 5 to provide vox keying. It also has other possible uses.

Before using this method with your equipment, you should determine what additional modifications are necessary. The following conditions must be met.

audio to the microphone input, place the operate switch in vox position, and go ahead.

Grid block keying was used in the original circuit. A -180 V line was used to cut off the driver stage during standby. The -180 V line was shorted directly to ground by the key. This made it difficult to use the same contacts to operate other circuits. In fig. 5, a silicon-controlled rectifier replaces the key contacts, and a positive-gate turn-on voltage is keyed. This removes the shock hazard and provides a keyed source of low voltage to turn on other circuits.

The audio oscillator module is a small code-practice type available at most amateur equipment dealers. The speaker

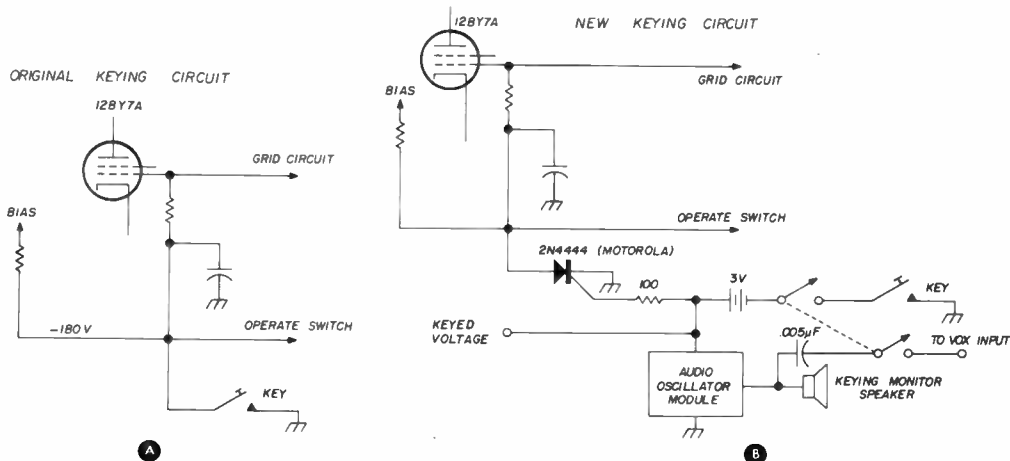


fig. 5. Original vox keying circuit, A, and new circuit, B.

1. The speech amplifier must be disconnected from the vox input or its power supply when the transmitter is placed in the CW position.
2. The vox circuit must remain in operation with the transmitter in the CW position.
3. The vox relay must control the same circuits as the CW operate switch.

If all these conditions are met, all that's necessary is to connect the keyed

is a tiny 8-ohm transistor radio replacement. Your junk box might provide a phase-shift or unijunction oscillator. However, at the price of the code-practice module, constructing one would hardly be worth the effort.

R. J. Beck, VE7IG

break-in control system

For really fast and effective CW break-in, many amateurs use separate transmitting and receiving antennas. With such

solid-state audio oscillator-monitor

Prompted by a recent article in *ham radio*, I fulfilled a long-time desire to design an audio oscillator that works with digital integrated circuit supply voltages, produces a clean sine wave at 1 kHz, drives a speaker, and is inexpensive and

I use a 3.2-ohm speaker in series with a 22-ohm resistor as a load. Any speaker will work, even the 40-ohm transistor-radio types. Use a scope to ensure a good waveform across the whole load. Q4 can be prevented from heating up by adding more resistance in series with the speaker. I don't use a volume control, since the level of my unit is perfect for my

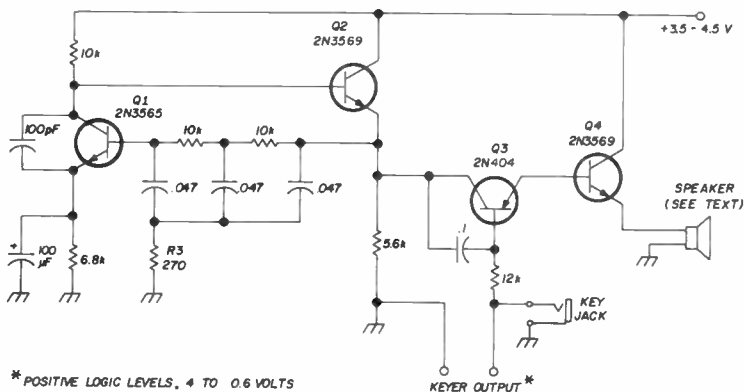


fig. 8. Schematic of the 1-kHz audio oscillator. Output frequency is determined by integrator circuit between Q1 and Q2.

easy to build. The result is described in the following paragraphs.

circuit description

A minimum number of components is used in an efficient sine-wave oscillator circuit (fig. 8). Transistors Q1 and Q2 form a high-input and low-output impedance amplifier, a feature of operational amplifiers. A "basic isolated integrator network" is inserted between input and output. The circuit can be made to oscillate at the frequency determined by the network constants by omitting a resistor between the base of Q1 and ground. Slight adjustment in frequency can be made by changing the value of R3; however, don't make too big a change since stability may be affected. Q3 operates as a series switch and is turned on by grounding its base resistor. The transistors specified for Q3 are germanium types because of their low saturation voltage, which promotes efficient switching. Q4 is an audio power amplifier and drives a speaker directly.

requirements. However, if one is desired, see fig. 7.

power supply

With a 4-volt supply and key down, the total power drain is 35 mA. For a power supply you can use three D cells in series or obtain 3.6 to 5 volts from your electronic keyer. A simple power supply is shown in fig. 9.

Logic levels of 4 to 0.6 volts are commonly encountered with keyers. I would be interested in hearing how this circuit can be tied into the many keyers described in the amateur literature.

N. J. Nicosia, WA1JSM

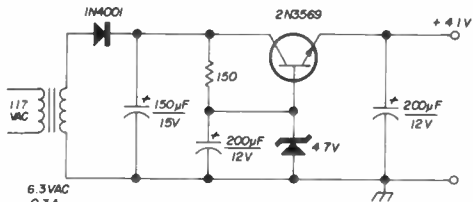


fig. 9. Suggested regulated power supply.

measurements and test equipment

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

transformer shorts

Transformer shorts occur in many ways. They can be caused by moisture absorption, hydroscopic leakage on the insulators, damp cotton-covered output wires, overload, poor or old paper insulation, just to name a few. The phenomenon of a shorted transformer can be very puzzling because the line fuse will blow when nothing seems to be wrong, especially when the transformer has been disconnected from the power supply.

Here is the way some of the old timers check a transformer. First, disconnect the secondary winding leads. Next, put an ordinary 50- to 100-watt light bulb in series with the primary winding. If the transformer is good the lamp will glow dimly when it is plugged into the 120 volt ac line because of the resistance of the winding. If the transformer has a shorted turn, the lamp will burn brightly. This will also happen if there is a short in either the primary or secondary winding.

You can prove this to yourself by shorting the secondary winding with a clip lead; the lamp will burn brighter. Unless you tie the clip leads to a long stick however, this experiment should only be used with low-voltage transformers. Pull the line cord while shorting, and keep your hands off while the power is on — we don't want you at the bottom of a hole. High-voltage ac has a propensity of going through insulation to your fingers, so please, put the clip leads on a long stick!

Ed Marriner, W6BLZ

unmarked filter chokes

Have you ever picked an old power choke out of your junk box and wondered how you could find out its induc-

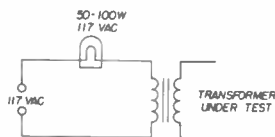


fig. 1. Simple circuit for checking for short circuits in high-voltage transformers.

tance? You probably checked the dc resistance, scratched your head, and put it back in the junkbox as another of life's insoluble problems.

Well, there is an easy and painless way to figure out the inductance of any power choke. The only equipment you need is a filament transformer, a vtm with reasonable ac accuracy and a composition

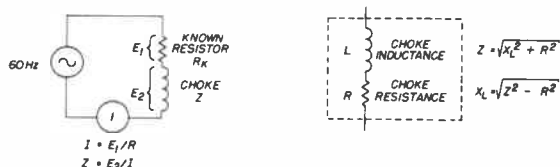


fig. 2. Here's the circuit used for determining choke inductance. The principles are explained in the text.

resistor of known resistance ($\pm 5\%$ is good enough). There is only one limitation — you obtain the zero-dc-current inductance. If you're looking for the inductance of an audio filter, this is fine. For power filtering purposes with dc flowing in the choke, the inductance may be perhaps 30 or 40 percent less, depending on the characteristics of the core and the amount of current.

Let's look at a thumbnail sketch of the process first and then run through a sample calculation. The process is based on Ohm's law for ac circuits where impedance (Z) is substituted for resistance (R). The formulas are written the same way as the dc version of Ohm's law (e.g., $E = IZ$, $I = E/Z$, $Z = E/I$).

One further point: if the voltage is measured as rms, the current will also be rms. The same applies to peak-to-peak measurements. As long as the system of measurement is consistent, only the values are important. The second formula you need is a variation on $Z = \sqrt{X_L^2 + R^2}$. If you know Z — more on this in a moment — and R , the choke resistance, you can find the inductive reactance of the choke from $X_L = \sqrt{Z^2 - R^2}$. By now you may see where we're heading.

The first step in the process is to

measure the resistance of the unknown choke. Your ohmmeter is accurate enough for this. Now connect the choke in series with a resistor of known value. Apply 5 to 10 Vac from a filament transformer to the ends of this RL network. Use the vtm on its ac voltage ranges to measure the ac voltage across the known resistor.

With this information, you can calculate the ac current flowing in the circuit. Now measure the ac voltage across the choke. When you know the ac voltage across the choke and the ac current flowing through it, the impedance of the choke at 60 hertz may be calculated from $Z = EI$.

You already know the dc resistance of the choke, so by using Z and the formula $X_L = \sqrt{Z^2 - R^2}$, you solve for inductive reactance. As mentioned earlier, this will be the reactance of the no-dc-current inductance. While you're at it, you can also determine the Q of the choke from $Q = X_L/R$. This may be important if you are building an audio filter or other unit requiring a resonant audio circuit.

example

Assume a choke with 320 ohms dc resistance. You measure 4.7 Vac across a 1000-ohm resistor; this gives an ac current of .0047 amperes. You measure 7.5 Vac across the choke. Dividing 7.5 by .0047, you find the choke impedance at 60 hertz is 1600 ohms. From $X_L = \sqrt{Z^2 - R^2}$, where R is the choke resistance, you find X_L is 1570 ohms. To find the inductance, divide 1570 by 377 ($2\pi f$, where f is 60 Hz); the inductance is 4.17 henrys.

table 1. summary of process

1. Measure the ac voltage across the series resistor.
2. Compute I from E/R_1 where R_1 is the series resistor.
3. Measure the ac voltage across the inductor.
4. Compute Z from E/I .
5. Compute X_L from $\sqrt{Z^2 - R_2^2}$ where R_2 is the internal resistance of the choke.
6. Compute L from $Z/2\pi f$ where $2\pi f = 377$ for 60 hertz.

Now that you see how it's done, there shouldn't be a single unknown choke in your junkbox.

Bob Tellefsen, WØMKF

rf generator clip

Connecting an rf generator to a piece of equipment for test purposes can sometimes pose a problem. Many times a simple expedient is used: a couple of test leads with alligator clips. In most cases this is not too satisfactory. First, there is no isolation between stages, because the leads are usually quite long and radiate heavily. Besides, it is pretty unhandy.

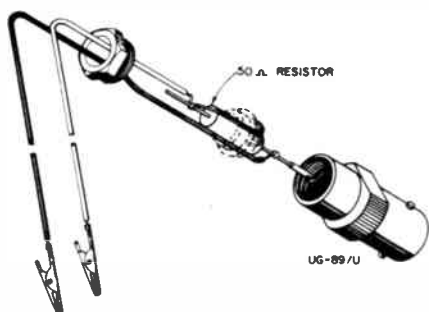


fig. 3. Exploded view of the rf generator clip. The 50-ohm resistor is mounted inside the coaxial connector. If a 1/4-watt resistor is used, there's space left for a series coupling capacitor.

The generator clip described here is based on a BNC connector, because all the test gear in my shop uses them. Although a UG-89B/U was used in this case, a UG-23/U, PL-259 or an Amphenol audio connector could be adapted in just about the same way. To insure that the generator is terminated in a 50-ohm load (or thereabouts), a 1/2-watt, 5%, 51-ohm resistor is mounted inside the coaxial connector. One lead of this load resistor is soldered to the female pin of the connector. A short length of flexible test-prod wire with an alligator clip is also soldered at this point.

The other end of the resistor is connected to the ground side of the test circuit. This is done by stripping the insulation off another piece of test-prod wire about one inch back from the end.

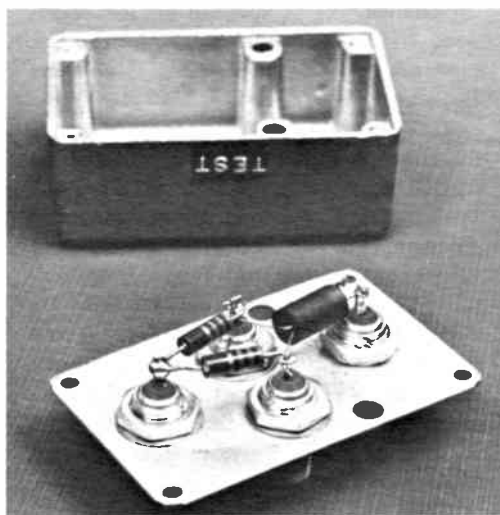
The resistor is soldered to the wire up next to the insulation, and the wire strands are brought down around the body of the resistor. The braid-retaining washer from the coaxial connector is then put over these wire strands and the body of the resistor; the strands are bent over the shoulder of this washer. When the cable nut is tightened into the connector body, these strands will ground the test lead and hold the test wires and resistor in the connector.

To protect the assembly and to prevent the test leads from breaking at the solder points, a short length of heat-shrinkable tubing is placed over the unit and shrunk into place. After it's all put together, check with an ohmmeter to make sure the two test leads are not shorted; they should exhibit 50-ohms resistance.

Some amateurs might like to place a coupling capacitor within the connector. There is room for a small 50- or 100-pF ceramic tubular capacitor if a 1/4-watt load resistor is used. In some cases you might want to delete the terminating resistor. In any event, this little clip provides a very simple and convenient way of getting rf energy from the generator into a circuit.

Jim Fisk W1DTY

fig. 4. Construction of the impedance bridge. Unit must be housed in a metal box.



impedance bridge

Here is a useful little bridge that can be used to compare the impedance of an unknown antenna (or network) to a known termination over the range from 2 MHz to 900 MHz. By changing the value of the termination, impedance can be checked in the range from 25 to 500 ohms; to check the match of a 50-ohm antenna for example, a 50-ohm load would be placed across the *termination* terminals. When the impedance across the *test* terminals is equal to the impedance across the *termination* terminals, there is no output.

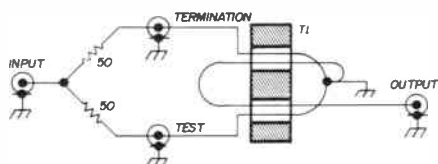


fig. 5. Impedance bridge compares "termination" impedance to the "test" impedance. Transformer T1 has two windings; each winding is one turn through a small double-hole toroid, the windings connected as above.

This bridge may be used with a sweep generator and oscilloscope to display a circuit's resistance characteristic over the desired bandpass. To determine the value of an unknown impedance connected at the *test* socket, a calibrated variable resistor may be used at the termination terminals. When building this simple instrument, make sure all lead lengths are equal. It is not necessary for the impedance of the measuring equipment connected across the *input* and output terminals to have the same value as the unit under test.

Oliver W. Swan, W6KZK

harmonic generator

The circuit shown in fig. 6 will produce 50- μ V harmonics through 1296 MHz with an input of 0.15 to 1 V from a 100- or 1000-kHz crystal oscillator. With a germanium diode instead of a tunnel diode, harmonics can be heard up to

about 147 MHz.

A 2- to 5-ohm bias resistor can be inserted between emitter and ground to prevent possible thermal runaway. As the collector current increases with temperature, the voltage developed across the resistor alters the base current, through

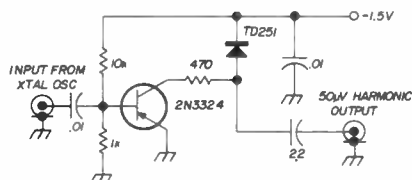


fig. 6. Basic harmonic generator using a tunnel diode.

dc feedback, to stabilize the collector current. I am not using an emitter resistor, however, and have had no trouble. A larger resistor, up to about 470 ohms, can be used. This will affect the output of the circuit. The emitter resistor should be bypassed to prevent gain loss due to degenerative feedback.

A complete circuit is shown in fig. 7 for those who don't have a 100- or 1000-kHz oscillator.

Chuck Spurgeon, W5GDQ

new uses for a grid-dip meter

Every once in a while I get interested in playing around with antennas from the standpoint of home design. There are many books available, and all describe characteristics of beam antennas in terms of electrical length. Since the electrical length varies with tapering rod diameter

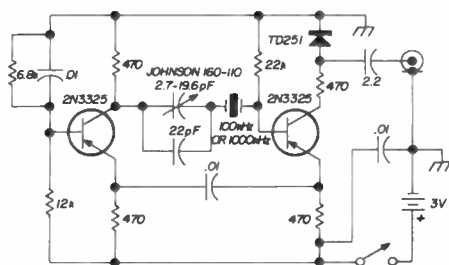


fig. 7. Complete harmonic generator circuit including reference oscillator.

and type of antenna (folded vs normal dipole), it is extremely hard to measure in terms of linear – and conventional – dimensions.

The grid dipper is one of the most versatile instruments available to the ham and is next in importance, perhaps, to the multimeter. Ken Lockhart, K2HAK, has been able to resolve the antenna situation by proving out the theory and permitting us to publish his findings.

As is well known, the coil of the gdo, when coupled to a rod, shows the rod's resonant frequency. This corresponds to a half-wave-length of the rod – with such illusive parameters as length-to-diameter ratio and propagation velocity thrown in. Measurements made in this way are exact and provide the practical answer we're seeking.

However, one practical limitation not accounted for by manufacturers of test instruments is this: A small coil produces a rather constricted field pattern; while this field will link a small rod, it will not link a rod that is comparable in diameter to that of the coil itself (see fig. 8A). This condition is easily remedied by the rather obvious technique of building a larger

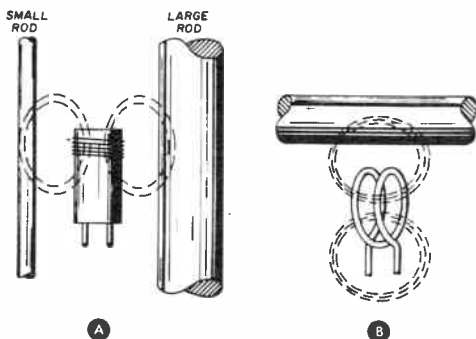


fig. 8. It is difficult to couple a grid-dip meter to large antenna elements because of the small diameter of the coil (A). A larger coil, shown in B will do the job.

diameter coil whose larger field pattern can be coupled to larger diameter rods (see fig. 8B).

antenna gdo

One method of solving the problem is shown in fig. 9. The unit is calibrated by

comparing the radiated energy to various TV stations. If the gdo is tuned to channel 2, for example, the picture will tend to roll out of sync; if tuned to ± 4.5 MHz from this frequency the sound will be garbled or blank completely. If the TV

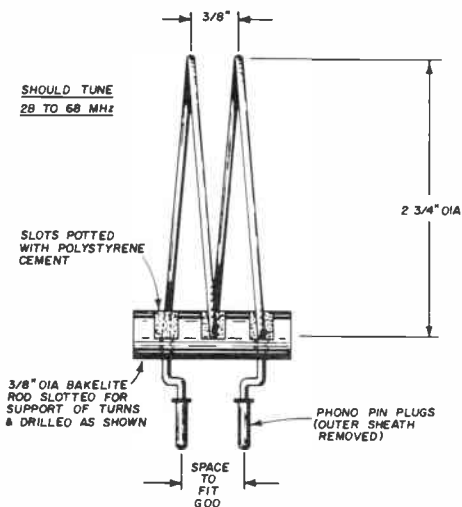


fig. 9. Method of building a large coil for your grid-dipper.

signals are strong, it may be difficult to blank the sound with the oscillator tuned to the picture (51.25 MHz), but this can be overcome by disconnecting the TV antenna.

You can calibrate the grid dipper above 6 meters with TV signals. This won't work below 6 meters; this can be accomplished by beating the second harmonic of the oscillator with the stations shown in table 2. If an fm receiver is handy, excellent checkpoints may be obtained by using the second harmonic of the gdo. If a smooth curve is plotted in this manner, extrapolated accuracy is good enough for the most demanding purposes.

constructing an antenna by gdo

Let's try an example. A good three-element beam is the one with 0.15-wave-length spacing from driven element to reflector; the reflector is approximately 1.06 times the length of the driven element. The director is 0.943 the length of the driven element and spaced 0.1

table 2. Television and fm frequencies which can be used for calibrating a grid-dip meter.

channel	video	sound	video -4.5 MHz	video second har- monic	sound second har- monic
2	55.25	59.75	50.75	27.6	29.9
4	67.25	71.75	62.75	33.6	35.9
5	77.25	81.75	72.75	38.6	40.9
fm	88-108	—	—	44 to 54 MHz	—

wavelength away.

Since boom length is 0.25 wavelength at 50 MHz, its resonant frequency will be 100 MHz; the length is cut until gdo resonance is indicated. The driven element is shorted out (either with a direct short or a terminating resistor) and cut until it resonates at 50 MHz. The director is cut to resonance at 50×1.06 or 53 MHz and the reflector at 50×0.943 or 47 MHz. Now we have the complete antenna specified in terms of the resonant frequencies we can measure with a grid dip meter:

boom:	100 MHz
driven element:	50 MHz
reflector:	47 MHz
director:	53 MHz

Once this method has been mastered, it's likely that you'll never refer to antenna elements in feet or inches but in MHz. I have now built several antennas using this technique (two trapped jobs for 6, 10 and 15 meters; one for 6 and 10 meters, and one for 220 MHz) and have had excellent results.

If you should decide to plot the radiation pattern of your new antenna, the coil and your gdo will work nicely as a low-power source of rf. The coil should be horizontally polarized and in the horizontal field of your beam — so you'll have to get it up as high as humanly possible.

Bob Brown, K2ZSQ

versatile resistor decades

Accurate decade standards may be

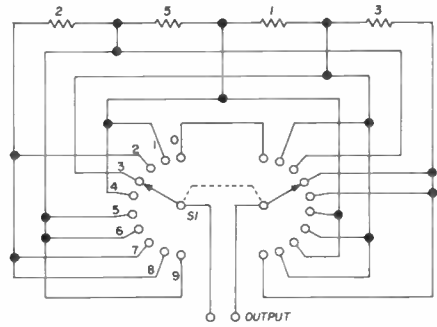


fig. 10. Zero through 9 ohms are obtained by switching various combinations of four standard resistors. For higher resistance values the resistors are increased by factors of ten. The switch is a 2-pole 10-position rotary.

assembled at considerably less than their usual cost by using one of the switching arrangements shown in fig. 10 and 11. The precision components represent the greatest cost of a decade standard; therefore, any reduction in the number of components directly reduces cost.

Only four precision resistors or capacitors are required in this design, instead of the usual ten. This savings may be used to simply reduce cost, or if you wish, to buy closer tolerance components for a more accurate decade.

Several construction precautions should be observed. Low-resistance decades require very low contact resistance switches to preserve accuracy. This requirement may be met by parallel wiring two decks of a conventional rotary switch, instead of single decks shown in fig. 10.

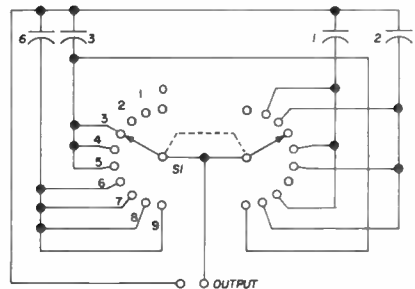


fig. 11. Nine individual capacitance values are provided by switching combinations of four capacitors. Switch is a 2 pole 10-position rotary.

The capacitance decade's low-value capacitors must be well spaced from all metallic components to reduce errors due to stray capacitance effects. Also, wide spacing between capacitor terminals and jumpers is required. In the low-capacitance and high-resistance decades, rotary switches with ceramic insulation are recommended.

The economical decade switching systems shown here may be extended to the design of other useful decades including inductance, zener diodes and back-to-back diode clippers.

Gene Brizendine, W4ATE

economical decade standards

Resistor decades are convenient for experimental use but are bulky and expensive, especially the higher-accuracy units. A more convenient and compact precision, direct-reading variable resistance may be easily assembled using the ten-turn potentiometers and dials now available on the surplus market. Pots with accuracies of 1% and 0.1% linearity are available – which is more than adequate for most applications.

Potentiometer values in multiples of ten are used so that the selected resistance value may be read directly from the 10-turn dial by adding the proper number of zeros. Several potentiometer/dial units may be assembled in the form of the usual decade box.

The versatility of this arrangement may be expanded by bringing all three potentiometer terminals to binding posts on the front panel. With three binding

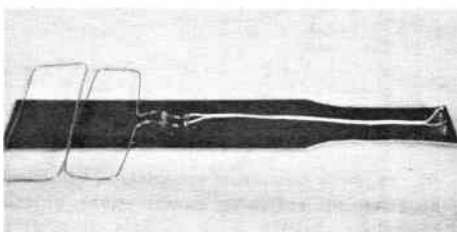


fig. 12. Rf current probe.

posts continuously variable voltage dividers of accurately known ratio are conveniently available to the experimenter.

Gene Brizendine, W4ATE

rf current probe

When working with antennas, you often need a way of checking relative current magnitudes in individual elements. Here is a gadget that will do this without breaking the element. It is essentially a loop connected to an rf

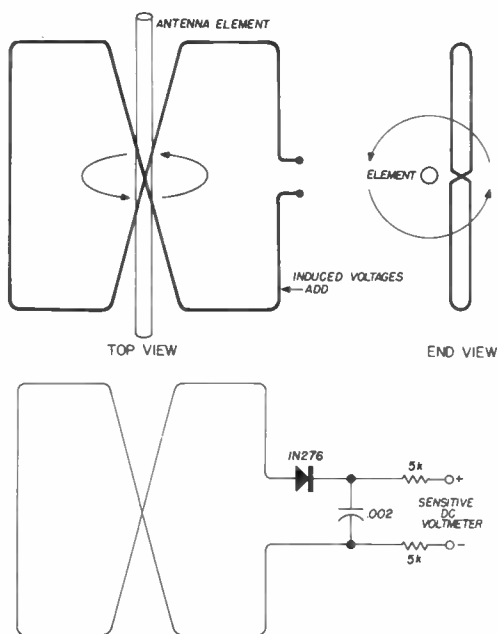


fig. 13. Construction of the rf current probe is not critical although the two loops should have the same area.

voltmeter which permits inductive coupling to the antenna element. The voltmeter gives a reading which is proportional to the current in the conductor.

The loop is bent into a figure-eight shape for a special reason. Any *uniform* rf field will induce equal voltages in each half of the figure-eight. However, since one loop is wound clockwise, and the other counterclockwise, the two induced voltages will cancel, and the meter will

not respond to this type of field. But if the probe is held against an rf current-carrying conductor as shown in the diagram, the magnetic field surrounding the conductor will thread one loop in one direction and the other loop in the opposite direction — instead of canceling, the induced voltages will add. Consequently, the meter will only respond to the current in this conductor and not to radiated fields from other elements.

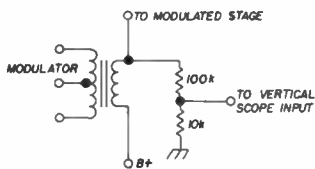


fig. 14. Resistive divider is used to pick off signal for a-m modulation monitor.

Construction details should be obvious from the photograph. This model was built on a 2-inch wide strip of masonite. No dimensions are critical except that the two loops should have approximately the same area.

Fred Brown, W6HPH

vhf a-m modulation monitor

If you have a dc-coupled oscilloscope in the shack, such as the Eico 460, here is a low-cost and simple method of keeping track of your percentage of modulation. At vhf the usual techniques of envelope patterns and trapezoidal patterns become very difficult, if not impossible. The method shown in fig. 14 is based on the fact that the plate voltage applied to an amplitude-modulated stage varies between zero and twice the dc supply voltage for 100% modulation of a properly adjusted a-m transmitter.

Connect the vertical scope input to the modulated dc which feeds the modulated rf stage. A resistive voltage divider (no capacitors) is necessary if the voltage is

high enough to be dangerous. Select a convenient sweep rate of about 100 Hz or so. Position the horizontal trace so it coincides with one of the lower graticule lines. Apply plate voltage to the modulated stage (with no modulation) and adjust the vertical gain control to move the horizontal trace to the center graticule line. The scope is now set up to show modulation amplitudes varying around the dc supply voltage.

Audio modulation gain should be adjusted to keep the scope pattern between the limits of the lower (zero volts dc) graticule line and the corresponding graticule line above the center. Fig. 15 shows the relationships. For my equipment, with a 750-volt dc power supply, I chose a resistive voltage divider of 100k and 10k. A suitable ratio permits convenient measurement of the final dc supply voltage without exposure to lethal high voltages.

Harry Ferguson, K7UNL

switching counter readouts

One of the more expensive parts of a frequency counter, or counter dial, is the indicating or readout equipment. As a result many applications provide the readout for only three or four digits of the count. In some cases you might want to read kHz but not individual Hz; in other cases you may want to read kHz or Hz but not MHz.

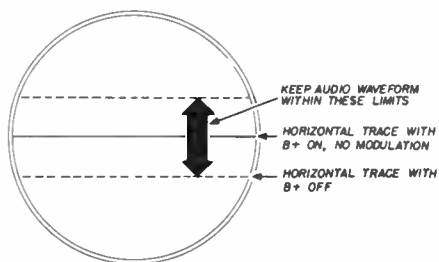


fig. 15. Modulation percentage is less than 100% if waveform is between limits shown here. Equipment setup is described in text.

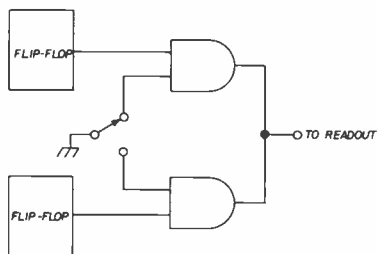


fig. 16. Using dual-input gates a readout can be switched to the Q output of either flip-flop to reduce cost. If the gates do not permit tying the outputs together, use steering diodes or a third gate.

One way to obtain a readout when needed, without providing more Nixie tubes and associated storing and decoding circuitry, is to switch the indicating system from one part of the counter to another so all digits can be read when required. This can be accomplished by switching half as many indicators as the available total of digits counted, or by having just one digital indicator which is switched from one digit to another. A rotary switch, with four double-throw contacts per decade, will do the job.

Digital logic handbooks often show a "data selector" circuit using a gate in each data stream with provision for applying a signal to select which data stream appears at the output. Fig. 16 shows one way to do this when the gate design permits output ORing. It uses a dual-input gate for each flip-flop output, or four gates per decade. Four more gates

may be needed for the alternate decade which might be switched to the readout. A single toggle switch is sufficient for all gates.

Since no operation is required of the later flip-flops when the earlier ones are connected to the readout equipment, half the gates can be eliminated. This can be done by providing *preset* or *clear* signals or removing the B+ from the later flip-flops which are not to operate the digital readout indicator. If direct interconnection of flip-flop outputs causes a problem, steering diodes can be inserted.

Surplus diodes and a few resistors can perform the gate functions if a steering diode is included between the gate and the digital readout. This diode prevents the two interconnected gates from operating as a single gate. Again, a single toggle switch is sufficient for all gates involved in the readout switching (see fig. 17).

In addition to other types of diode switching that probably can be worked out to do the job at little cost, some other simplifications appear possible. One possibility is the elimination of one gate on the output of the tail-end flip-flops, turning these off by other means.

RTL and DTL flip-flops usually have their Q outputs connected to transistor collectors; the collectors are connected to $+V_{CC}$ through a resistor. Shorting the output of these circuits may overheat the resistor (although it *can* be replaced externally to the IC). Putting $+V_{CC}$ on the Q outputs can result in too much collector voltage on the transistors.

One simplification which I have not tried is shown in fig. 18. Diode CR1 forms a *one-input gate* when FF2, in an earlier position in the counter, is connected to the readout by the switch placing $+V_{CC}$ on the resistor. Whenever FF2 is *high* there is little voltage drop across R, so the *high* passes through steering diode CR3 to the readout equipment. When FF2 is *low* the voltage from R goes into the flip-flop, creating a *low* input to the readout equipment. The other switch position puts $+V_{CC}$ on FF1 and other tail-end

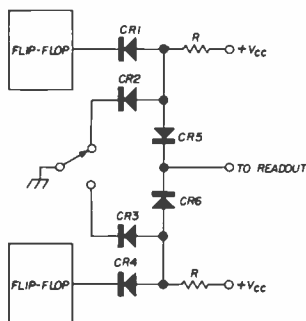


fig. 17. AND gates consisting of two diodes and a resistor can perform the switching between Q outputs, but require steering diodes CR5 and CR6.

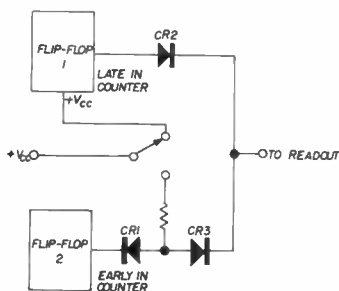


fig. 18. Simplified diode switching of readout when flip-flop 2 is earlier in the counter string than flip-flop 1.

flip-flops to be read out. This allows these flip-flops to operate, feeding the FF1 Q output through steering diode CR2 to the readout equipment. Diode CR2 prevents the voltage across the resistors from feeding back into the Q output of FF1 when FF1 is turned off by the switch. CR2 may not be necessary if the voltage on the Q output is not harmful when FF1 has no $+V_{CC}$ supply.

Bill Conklin, K6KA

multitester

Although most ham shacks have at least one volt-ohmmeter or vtvm, a low-cost utility multitester can fill in when you have to make simultaneous voltage and current measurements. The multi-

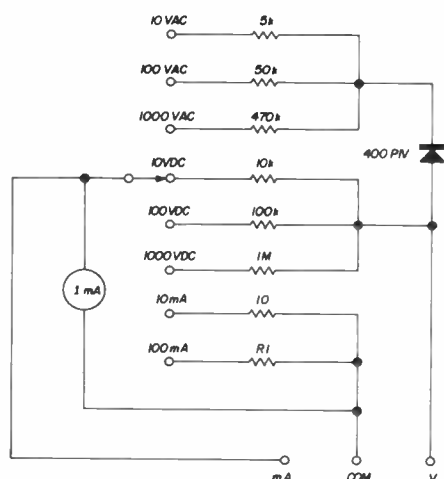


fig. 19. Simple utility meter. R1 is 5 feet no. 36 wire wound around high-value resistor.

tester in fig. 19 covers the most useful voltage and current ranges and uses a low-cost 1-mA meter movement. The multiplier and shunt resistors, and diode are mounted on the selector switch, a Centralab PA1001. All resistors can be 5% carbon composition types, although for higher accuracy 1% precision resistors should be considered. However, to keep the cost down, carbon composition resistors provide acceptable accuracy for most purposes. R1, the 100-mA shunt, consists of 5-feet no. 36 wire wound around a high-value resistor. The 1-mA meter is an imported unit, such as available from Lafayette.

Jim Fisk, W1DTY

simple resistance standard

As many amateurs know, ohmmeter readings taken with a vom can be notoriously inaccurate. However, there are times when you need an accurate resistance measurement of a higher order than that obtainable with an ordinary multitester. Higher accuracy generally requires a Wheatstone bridge or similar laboratory-type equipment. For occasional use such expense cannot be justified.

The inexpensive "resistance checker" shown in fig. 20 is a simple and direct solution to this problem. Four 1% resistors are wired as shown; it is apparent that there will be resistance check points at 10k, 20k, 30k and 40k ohms. If your volt-ohmmeter has not been grossly abused, you can set the meter to 10k ohms when connected across one leg of the standard — merely adjust the "zero ohms" knob so the meter coincides with the selected resistance value. Thereafter you may read resistance values with good accuracy up to 10 or 20 percent removed from the check point.

In addition to the multiple 10k check points, parallel connections will result in resistor standards at 2,500, 3,333, and 5,000 ohms (see table 3). Other standard

values may be obtained through suitable interconnection of the various resistors. The chart is provided since most of us don't want to get out the slide rule when a quick resistance check is needed. Obviously other decade values may be selected but the above setup provides resistance checkpoints at more than one dozen intervals throughout the range from 2.5k to 40k ohms.

Neil Johnson, W20LU

uhf tuner tester for tv sets

Most amateurs have tv sets, so I thought this idea would appeal to *ham radio* readers. It's an adaptation of a miniature solid-state GE tuner for use as a piece of test equipment. With this tester, you can check the uhf front end of your tv set without a signal generator.

A number of surplus-type uhf tuners are available that will work nicely as signal generators. I purchased mine for \$5.20 postpaid from Arcturus Electronics.

Remove the tuner from its cover by unhinging the securing wire. Note the junction of two resistors feeding the stator of the oscillator section of the tuning capacitor. Attach a one-half-inch piece of insulated wire to this junction. Drill a hole in the cover to accommodate the wire when the cover is replaced. Leave about one-eighth inch of bare wire at the other end. Replace the cover.

Connect a half-watt, 330-ohm resistor to the insulated stud. (This is the spot where the tuner instructions say to install

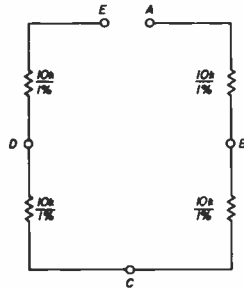
table 3. Parallel connections required for resistance values from 2500 to 40 kilohms.

desired resistance value	connect together	read ohms between
2500	A-C B-D C-E	A-B
3333	A-C B-D	A-D
5000	A-C	A-B
6667	A-D	A-C
7500	A-E	A-D
10,000	—	A-B
13,333	A-C B-D	A-E
15,000	A-C	B-D
16,667	A-D	C-E
20,000	—	A-C
25,000	A-C	B-E
30,000	—	A-D
40,000	—	A-E

the 15 kilohm resistor furnished.) Attach a few inches of wire to the other end of the resistor. Tape wire and resistor to the tuner case. Install the mounting bracket and nut according to tuner instructions, as well as the fine and coarse tuning knobs. Connect a 6-volt battery charger (must be unfiltered), with negative lead to the case and positive lead to the wire connected to the 330-ohm resistor.

Tune the tv set to a known uhf channel. Adjust the tuner slightly so the picture just disappears. Run a jumper from the wire emerging from the oscillator section of the tester to the uhf antenna terminals of the tv set. Apply power to the tuner, and tune until two black and two white bars appear on the tv screen. (This assumes a full wave charger or rectifier circuit.) Set the tester dial to read the same as the uhf dial on the tv set

fig. 20. Simple resistance standard provides accurate checkpoints from 2500 to 40,000 ohms. Connections are shown in table 1.



minus the offset to make the picture disappear.

This completes the tune-up and tester calibration. Other modulated 6-12 volt sources may be used, providing the transistor in the tester isn't overdriven.

John R. Schuler

testing unknown meters

A typical 10-megohm or higher vtvm is an excellent tool for testing unknown meters to determine their approximate current ranges without the danger of pinning the needles or burning out the windings.

Set the vtvm to OHMS and the range dial to $R \times 1$ MEG. Connect the test leads to the unknown meter. Move the range dial, one step at a time, through $R \times 100k$, $R \times 10k$, etc., until the meter under test reads upscale.

The maximum current through the meter from a typical 10-megohm vtvm (Lafayette KT-174, for example) is as follows:

range	I max	vtvm 1% series res.
$R \times 1$ meg	0.156 μA	10 meg
100k	1.56 μA	1 meg
10k	15.6 μA	100k
1k	156.0 μA	10k
100	1.56 mA	1k
10	15.6 mA	99.1*
1	170.0 mA	9.1*

*0.9 ohm is allowed for the internal resistance of the D cell and test lead.

The effect of the meter resistance on the first four ranges of the vtvm is usually negligible. Don't try this idea with a vom. The currents will almost surely be much greater than those shown above, and the meter being tested may be destroyed.

Harold E. Brown, W1ONC

measurement of electrolytic capacitors

Recently I had occasion to measure the value of an electrolytic capacitor. Since test equipment was not available, I

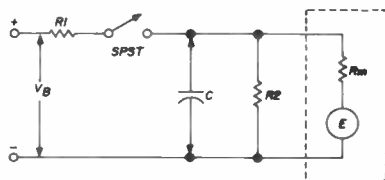


fig. 21. Circuit for determining capacitance of electrolytic capacitors. After measuring discharge time as discussed in text, capacitance can be found from nomograph in fig. 22.

devised a simple method using a resistor, a voltmeter and a stop watch. The capacitor was connected as shown in fig. 21.

The switch should remain closed for several minutes before making the measurement. This permits the electrolytic to form and stabilize. This is particularly important if the capacitor has been out of use for a long time. Resistor $R1$ is included in the circuit to limit the initial surge in charging current; its value can be on the order of 1000 to 5000 ohms.

When the switch is closed, read the voltmeter. Make a note of this reading (V_B). Now, mark the voltmeter at half this reading. This can be done by simply laying a strip of paper over the voltmeter scale at the $0.5V_B$ point. With a stop watch or watch with a sweep second hand, starting with the instant the switch is opened, measure the time required for the voltage to drop to the $0.5V_B$ point. Resistor $R2$ should be adjusted to produce a discharge time in excess of 30 seconds; this increases accuracy since it permits easier timing.

The value of the capacitor can be calculated from the following equation

$$C = 1.44 (t/R) \quad (1)$$

where t is in seconds, R in megohms, and C in microfarads (see fig. 22). The value of R in eq. 1 is the parallel equivalent of $R2$ and R_M , the internal resistance of the meter (eq. 2).

$$R = \frac{R2 R_M}{R2 + R_M} \quad (2)$$

The resistance of a voltmeter is usually

given in terms of "ohms per volt." If, for example, your voltmeter is rated at 20,000 ohms per volt, when set on the 100-volt scale R_M is 2 megohms (100 volts $\times$ 20,000 ohms = 2 megohms). Since R_2 may be on the order of 1 megohm, voltmeter resistance cannot be neglected if you want accurate results.

Although this method of measuring capacitance can be used with any type of capacitor, on small-value units it's difficult to measure discharge time accurately enough to obtain meaningful results.

Edwin L. Clark, W2NA

grid dipping transmission lines

Most amateurs are familiar with the many uses of the grid-dip oscillator. If a gdo is coupled to a tuned circuit made of lumped constants (fig. 23A), the meter will show a dip at resonance. Similarly, a

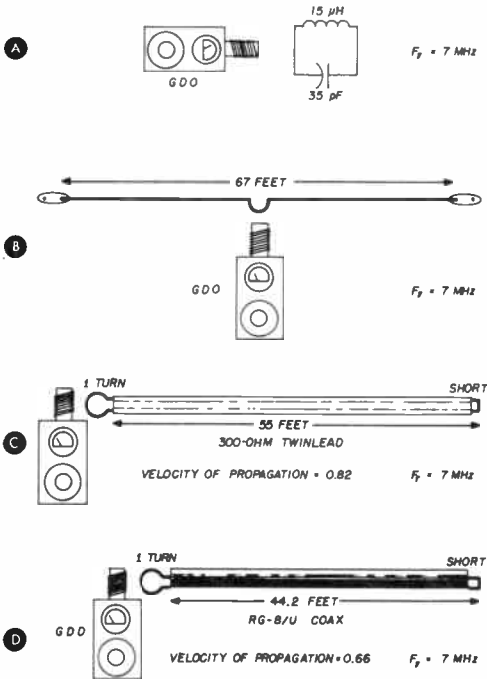


fig. 23. Illustrating the use of the grid-dip oscillator for determining resonant frequency of circuits with lumped and distributed LC constants.

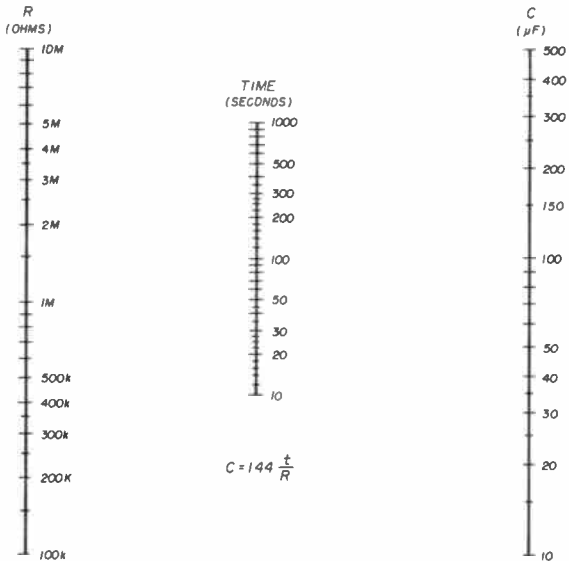


fig. 22. Nomograph for determining capacitance of electrolytic capacitors using method discussed in text.

gdo coupled to a circuit composed of distributed constants, such as a half-wave antenna (fig. 23B) or a half-wave section of transmission line (fig. 23C), the meter will dip at the resonant frequency of the antenna or transmission line.

The lowest frequency at which the gdo gives an indication of resonance is at a half-wavelength in the case of a Hertz or dipole antenna. When grid-dipping transmission lines, a propagation-velocity factor must be used. This factor is 0.82 for typical tv twinlead and 0.66 for commonly used coax cable (fig. 23C and 23D).

When grid-dipping antennas or sections of transmission lines, either of two methods may be used. The usual method is to determine the lowest frequency of resonance. If this isn't possible, you can grid-dip the line at several frequencies; i. e., at 4, 8, 12, and 16 MHz. A halfwave antenna has a "harmonic family" of resonances, roughly speaking, so you can expect similar behavior from a similar section of transmission line.

coiled transmission lines

Some hams roll up sections of trans-

mission line for compactness. This is where problems can occur. It's practicable to coil sections of coaxial cable, but this procedure isn't recommended for balanced lines, such as twinlead.

I experienced some frustrating results with a gdo coupled to coiled twinlead. Normal response was obtained when the twinlead was unrolled from the spool; but with the twinlead rolled up, gdo response was baffling. I obtained a series of minor and major dips, none of which seemed to produce a sensible pattern. When uncoiled, the twinlead behaved in normal fashion.

This anomaly may help to explain some rather odd reactions by those who have used twinlead for matching sections and, when coiling the line, have run into strange and unexpected results.

Neil Johnson, W2OLU

low-voltage zener tester

With the advent of transistors and ICs, it's desirable to have low-voltage power supplies and transistor testers for use on the workbench. The tester described here is not the ultimate in zener testers, nor do I make any claim for its originality. But it does work, is very accurate, and the cost is within the reach of most serious hams and experimenters. It will measure zener voltage up to 10 volts at the desired current from 15 mA to 40 mA (see fig. 24).

I must point out that you can't see the response curve of the device (a curve

tracer is needed for this); but if you're certain the device is a zener, this tester will measure its breakdown voltage and allow selection of precise voltages. You can also check voltages of unmarked and surplus devices.

For testing low-voltage zeners, I feel it's desirable to use a constant-current source. A standard parameter used by many companies for 400 mW, 2.2- to 10-V zeners is 20 mA.

The voltmeter consists of the 100- μ A meter and its 100k multiplier. The accuracy of your tester will be determined by the accuracy of these two components.

I used a 100 μ A API meter with 0.5 percent accuracy. Less expensive parts can be used if you're not too concerned with accuracy. The meter should be no larger than 100 μ A, though, because it loads the circuit.

To set up the tester, merely plug it in and short circuit point A to point B through a 50-mA meter. (Yes short circuit — this is a constant current source.) Adjust the 200-ohm pot for the desired test current. (I use 20 mA for 400 mA zeners.) Don't be concerned when the meter reads off scale without a zener across terminals A and B. The maximum output to the meter is only about 12 V dc, which means it's only about 20 percent overscale. This won't damage the meter.

After this adjustment, place the zener to be tested across test points A and B. The voltmeter will drop to the zener voltage at the set current.

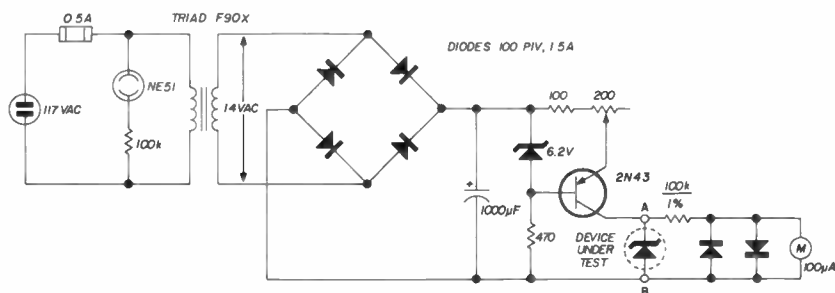


fig. 24. Zener-diode tester uses constant-current source that is adjustable from about 15 mA to 40 mA. Zener voltage is read on the voltmeter consisting of the 100-microampere meter and 100k multiplier.

If a unit should be placed backwards across points A and B, the voltmeter will drop to about 0.8 volt.

The layout of components isn't critical. I built my tester on vectorboard and placed it in a Bud 4 x 5 x 6-inch minibox.

M. Weinschenker, K3DPJ

vtvm modification

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

By adding a few parts and making a few circuit modifications, it is possible to do away with the flashlight cell altogether. The diagrams show the modifications I made to my Heathkit V-7A. This is typical of all inexpensive dc vtms, whether made by Heath, Allied, Eico or RCA.

As shown in fig. 25, the 6.3-volt filament voltage is rectified by a conventional bridge rectifier; the resulting dc is dropped by the 18-ohm resistor in conjunction with the regulating diodes, giving the required 1.5 volts dc.

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

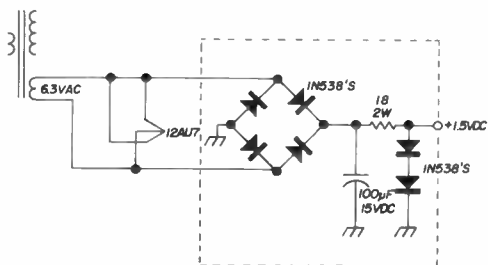


fig. 25. Deriving a 1.5-volt dc supply for the vtm from the 6.3 Vac filament transformer. Added circuitry is enclosed by the dotted line.

Two 1N538s in series act as a zener to regulate the output voltage. The barrier potential for a silicon diode is about 0.75 volts; two in series regulate the output nicely at 1.5 volts.

The original flashlight cell was connected to a resistor string tapped by the

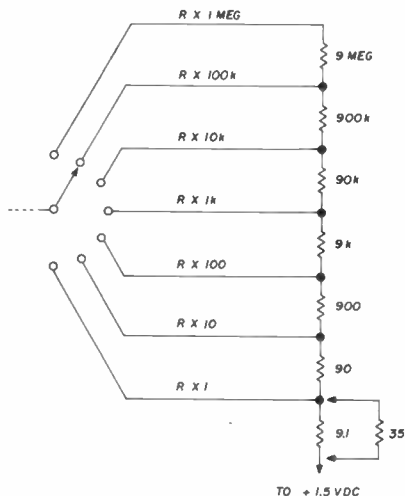


fig. 26. Modifying the ohmmeter circuitry for the 1.5 Vdc supply.

ohmmeter range switch as shown in fig. 26. Notice that the total resistance switched into the circuit is always 10 times the range switch position; this assumes cell internal resistance of 0.9 ohms. The internal resistance of this power supply is higher than 0.9 ohms, so it's necessary to lower the 9.1-ohm resistance by shunting it with 35 ohms as shown. The two in parallel then make 6.75 ohms. When this is added to the power supply internal resistance of 3.25 ohms, it makes up the required 10.0 ohms.

In my vtm I also replaced the 6AL5 "ac volts" rectifier with a pair of high back resistance, 600-PIV silicon diodes. This saves about as much heater current as consumed by the 1.5-volt power supply, so the transformer is still running cool.

The diodes that replace the 6AL5 should be forward biased slightly to overcome their barrier potential. This is

Fred Brown, W6HPH

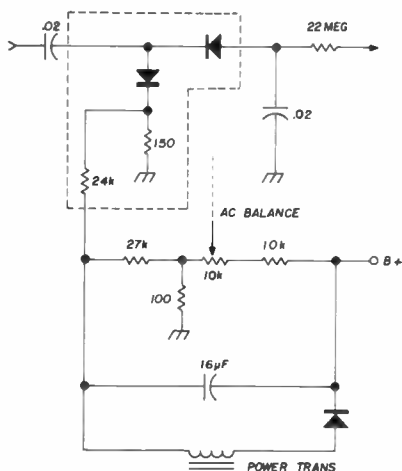


fig. 27. How to replace the 6AL5 with forward-biased silicon diodes.

simple scope calibrator

The calibrator shown in fig 28 was built in one afternoon and added to an Eico 460 oscilloscope. The same circuit could be added to other inexpensive scopes or built as an output accessory in a minibox. It's simple, the cost is low and the added current drain on the scope power supply is negligible.

The calibrator voltage is derived from a 22-volt zener diode. It is chopped at 60 Hz by the transistor, and the resulting 60-Hz square wave is trimmed to exactly 20 volts and applied across a precision resistor voltage divider. The calibrator can

To install the calibrator in the Eico 460 scope, remove the pilot light, X1, and resistors R5 and R44. The pilot light is a filament tie point, so the new wiring must take this into account. The best procedure is to run new wires from TB2 along the path of the old leads. Wiring to the phasing control (R45) is checked for continuity, and the scale dimmer (R80) is wired directly to TB2. The pilot light mounting hole is used for the calibrator rotary switch, and the "60-Hz" binding post is used for calibrator output.

The rotary calibrator switch is completely assembled before it's put into the scope; 1% resistors are used in the divider for maximum accuracy. After mounting the rotary switch, the zener and chopper transistor are mounted on an additional two-terminal tie strip mounted over TB4 and near XV6. The calibrate potentiometer (R6) is rotated 180 degrees in its hole and wired into the circuit as a voltage trimmer. The 820-ohm resistor, R5, is re-used in the base circuit of the transistor.

The transistor may generate voltage

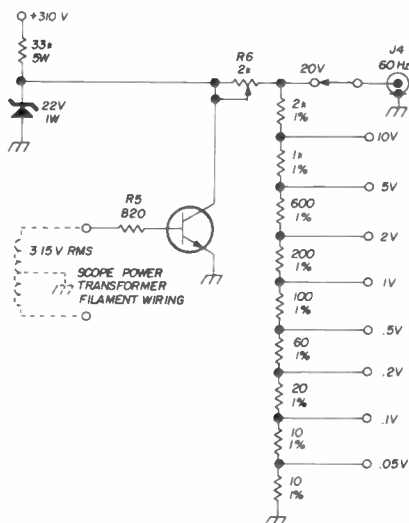


fig. 28. Simple scope calibrator circuit.

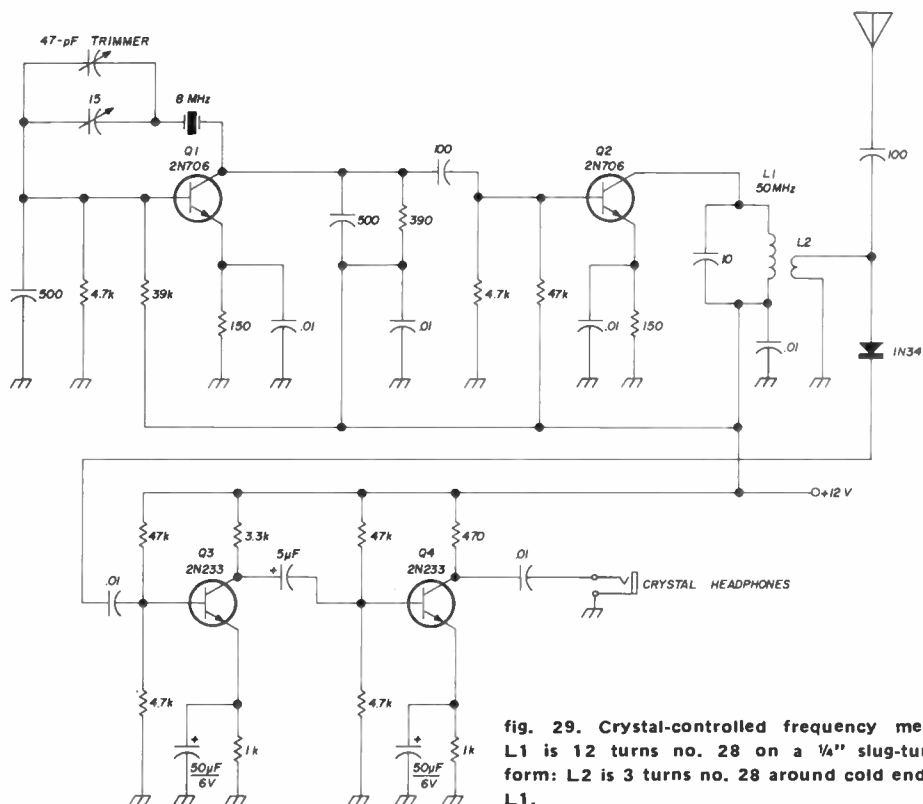


fig. 29. Crystal-controlled frequency meter. L1 is 12 turns no. 28 on a 1/4" slug-tuned form; L2 is 3 turns no. 28 around cold end of L1.

spikes that will crosstalk into the scope input unless the new wiring is dressed close to the panel where it's connected to the 60-Hz post. Also, the divider ground lead should not be connected to the scope input ground post — return it to the point where the transistor emitter is grounded near XV6. Nearly any npn transistor will work as the chopper; I tried surplus 2N706s and 2N1302s among others.

Bert Kelley, K4EEU

crystal-controlled frequency meter

For the past couple of years amateur radio magazines have been running articles about sophisticated frequency measuring devices, mostly using integrated circuits and other solid-state components. The crystal-controlled frequency meter shown here is as simple as can be constructed and is based on the old

method of zero beating one frequency against another.

The oscillator in fig. 29 is an untuned crystal-controlled oscillator. The crystal frequency can be "pulled" from 2 to 4 kHz, depending on the crystal. I use an 8 MHz crystal and multiply it to 50 MHz; the 50 MHz signal is rectified and fed to an audio amplifier. With this carrier turned on, every time another carrier of the same frequency is fed to the amplifier a zero beat is heard. It even radiates enough to check receiver frequency.

The stability of the oscillator was checked with a digital frequency counter and only ± 10 Hz drift was observed at 8 MHz during a two-hour period. The frequency meter uses a 12 volt battery supply; current drain is 8 mA — due chiefly to the low-level of amplification used in the circuit.

Construction is not critical. It can be hand wired, or everything can be put on printed circuit boards; I prefer printed

circuit boards, because it is quicker as well as compact and neat. Transistors are the four-for-\$1 type and crystals were 50c each. The transistors in the amplifier should be npn germanium with a beta of

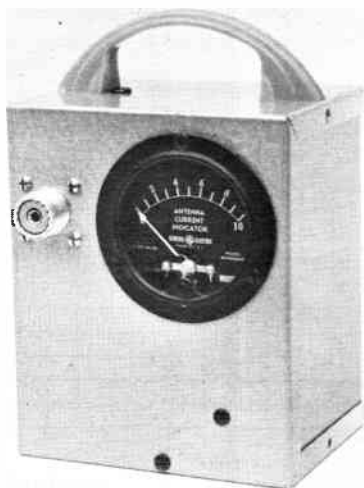


fig. 30. Low-cost dummy load/rf wattmeter.

at least 90; the 2N233 or 2N170 will do the job fine.

Crystals from 5 to 13 MHz will work in the oscillator. Adjust the 47-pF trimmer to zero a certain frequency with the proper crystal; make sure the 15-pF capacitor is half open. You should be able to deviate the crystal frequency ± 1.5 kHz. To set a transmitter vfo, set the 15-pF capacitor to the desired frequency, plug in a crystal headphone, load the transmitter into a dummy load and vary transmitter vfo to zero beat.

John J. Sury, W5JSN

dummy load/wattmeter

The advantages of using a dummy load for rf tests are many. No interference is created, and station identification isn't required for prolonged test periods. If the dummy load contains a calibrated meter, you can measure rf power with good accuracy.

Most hams have dummy loads that will handle a kilowatt. While these are fine for

their intended purpose, most won't provide accurate readings at power levels of 25 watts or less.

The unit described in this article is a highly accurate dummy load and rf watt-meter that can be used for testing low-power transmitters. Two versions are described: one is limited to 11 watts, and the other is good for 25 watts. Essential components cost less than two dollars. The enclosure, connector, and handle bring the total cost to less than four dollars. Sound interesting?

11-watt load

My original load used eleven 2-watt resistors in parallel. To save a buck, I used ten composition resistors, each 560 ohms, and one 680-ohm resistor. These are rated at 2 watts, 10 percent tolerance. I chose them because they're widely available and cost less in quantities of ten units.

The resistance of this dummy load is 51.73 ohms, which provides a good match for popular coaxial cables. Power capacity is limited to 11 watts. This is because composition resistors should not be used at more than 50 percent of their power rating; otherwise their resistance will change, which will degrade the accuracy of the wattmeter.

I mention this for those who might wish to build a very low-power unit with readily available resistors. The schematic of fig. 31 can be used, substituting the composition resistor bank for R (see photo).

25-watt load

Better resistors will allow rf power measurements up to 25 watts. The better units turned out to be another of those

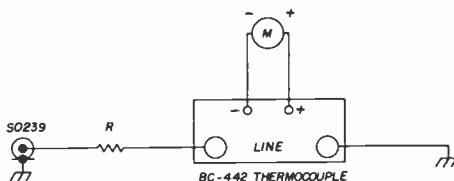


fig. 31. Schematic of the 25-watt dummy load and rf wattmeter. Resistor is two 100-ohm resistors in parallel; each resistor 13 watts, 1% non-inductive type.

table 4. Data for the CGW 25-watt resistor.

tolerance	2.5 or 10%
stability	less than 1% permanent resistance change*
operating temp	25° C
manufacturer	Corning Glass Works, Raleigh, N. C.
marking	CGW R35 25W 50 ohms

*When operated at 10 times rated power for 5 seconds.

surplus bargains that appear from time-to-time. I ordered two of these and connected them in parallel to obtain 50 ohms.

In an earlier project I used the 25-watt version of these resistors in a dummy load. Four resistors, each 50 ohms, were connected in series-parallel. The CGW resistors can be extremely overloaded and will retain their accuracy to a remarkable degree. For those who may have some, the data provided in table 4 should be helpful. I believe the overload characteristics with respect to permanent resistance change also apply to the 13-watt units.

thermocouple and meter

These components are another "ham special" value. Both units came from the BC-442 antenna tuning unit. They are first-rate, rugged, accurate components originally designed for military use. The dc meter has a nonlinear scale, a relatively expensive type of meter construction not generally associated with amateur equipment. This gives an expanded scale at the low end, and rf output as low as 2 watts

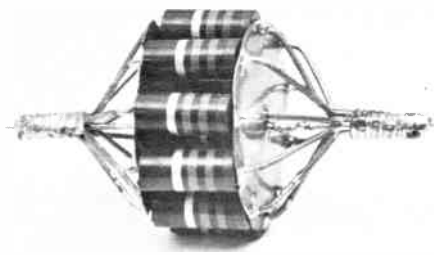


fig. 33. Composition resistor bank used in the 11-watt unit.

can be measured.

My experience with this meter and thermocouple has shown them to be highly accurate and dependable. This comes as no surprise when it's realized that these units had to withstand constant vibration in flight, hour after hour, plus the many shocks in taking off and landing at some rugged air strips.

construction

Construction details are shown in the

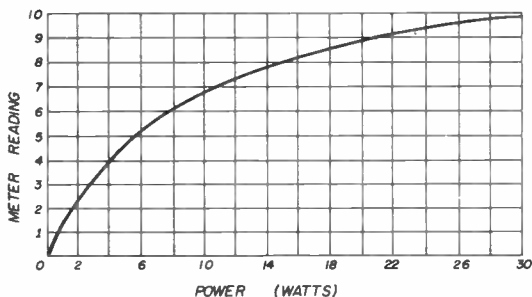


fig. 32. Calibration curve for the low-power wattmeter.

photos. The usual procedure of keeping rf leads short and direct should be followed in wiring this simple instrument. This will ensure a vswr of 1.1 or better throughout most of the amateur high-frequency bands.

calibration

The calibration data shown in table 5 and fig. 32 were obtained with the setup shown in fig. 34. The wattmeter was calibrated at 60 Hz, using mirror-scale laboratory instruments: a Weston Model 433 ac ammeter and a Simpson Model 261 voltmeter.

The voltmeter takes power, so I checked it against the Weston 433 before starting the calibration procedure. Readings were checked going up and down the scale. Then a second check was made to ensure repeatability of the data.

Note that the meter reading at 10 (table 5) occurs when the resistors are operated beyond their maximum power

†Fair Radio Sales, P. O. Box 1105, Lima, Ohio 45802; \$1.25 for both thermocouple and meter.

table 5. Calibration data.

meter reading	True rf amperes	rf watts into 51 ohms
1	—	0.9
2	—	1.8
3	0.23	2.7
4	0.28	4.0
5	0.335	5.6
6	0.39	7.9
7	0.46	11.0
8	0.54	15.2
9	0.64	21.4
10	0.78	31.4

rating. No harm is done to either meter or resistors. A temporary effect was that the resistors changed value to 50.5 ohms, and the entire dummy-load resistance increased to 51.5 ohms. The resistors returned to their original value upon removal of rf power.

The 51-ohm resistance of the unit is due to the series combination of the resistors (50 ohms) and the thermocouple, approximately 1 ohm. Note that readings below 0.23 ampere are not valid.

If you intend to use the instrument at levels above 15 watts for extended periods, some means of auxiliary cooling should be used. Removing the cover will also help.

conclusion

The amount of abuse the CGW resistors will take staggers the imagination. For example, my old 200-watt load withstood a kilowatt for 5 seconds maximum without resistor values changing more than 1 percent (permanent change after cooling). By the same token, the smaller resistors, which have a nominal rating of

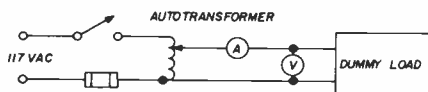


fig. 34. Circuit used for calibration. A is a Weston model 433 ac ammeter; V is a Simpson model 261 ac voltmeter (see text).



fig. 35. Inside the wattmeter, showing the CGW "R Series" resistors.

26 watts per pair, could be run up to 260 watts.

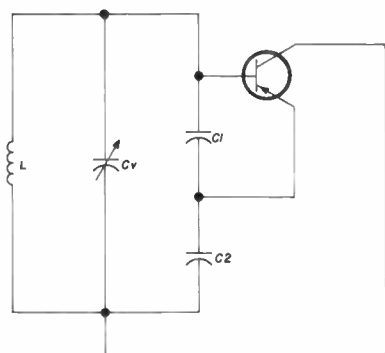
A high-quality rf wattmeter, good for any power up to 25 watts, is not a bad investment for \$4 and a pleasant hour's work.

Neil Johnson, W2OLU

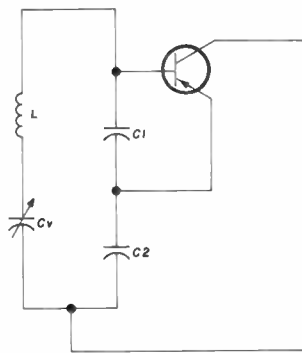
checking coaxial cable

I wonder if any other amateurs have stumbled onto a very useful secondary function of my QRP wattmeter, *coaxial cable quality checking*. You simply feed the output of your QRP rig or exciter into the rf wattmeter. This will give a reading (let's say) of 7 on the meter. Next, the cable to be tested is inserted between the exciter and the wattmeter. If the cable has a nominal 50 ohms impedance the wattmeter reading should be reasonably close to the earlier reading. If not, the watts lost in a given length of 50-ohm cable can be directly calculated by subtracting Watts_{far} from Watts_{near} as read directly from the QRP wattmeter.

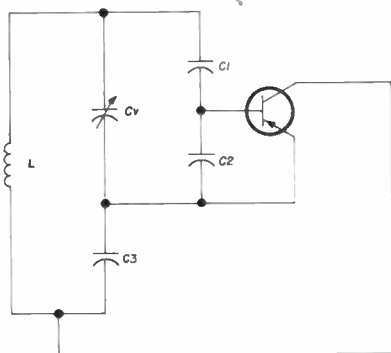
Neil Johnson, W2OLU



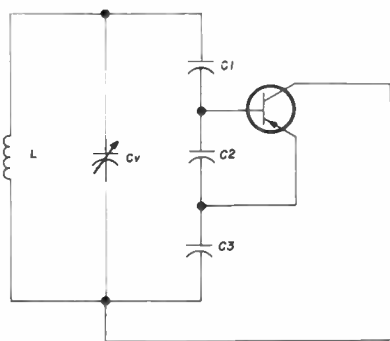
COLPITTS



CLAPP



VACKAR



SEILER

fig. 2. Circuit configurations of the Colpitts, Clapp, Seiler and Vackar oscillators. The Clapp, Seiler and Vackar circuits are derivations of the basic Colpitts circuit.

variable capacitor (Cv) in the Seiler oscillator, the frequency of oscillation may be simplified to:

$$f_{osc} = 2\pi \sqrt{L(C1 + C_v)}$$

Several vacuum-tube versions of the Seiler oscillator have appeared in the amateur-radio magazines, but in most cases the designers weren't aware that their circuit was an extension of W8PK's original design. In at least one case, the author called his circuit a, "groundplate Colpitts type."

To my knowledge, the first transistorized version of the Seiler oscillator was W3JHR's "synthetic rock" which was published in *CQ* in 1963. This circuit was extremely popular and subsequently appeared in amateur magazines in England, Germany and South America. W3JHR used an old ARC-5 transmitter as

the basis for his vfo; he cut the unit down and used the original variable capacitor and tuning coil to cover the frequency range from 4.9 to 6.1 MHz. Although only the oscillator stage is shown in fig. 2, he included a 2N384 emitter-follower buffer for isolation from the next stage.

K9ALD described another transistorized Seiler oscillator for ssb in 1964 and claimed exceptionally stable results. His oscillator, designed to cover the range from 4.95 to 5.6 MHz, is shown in fig. 3. Because of the relatively low-capacitance characteristics of the 2N2219, the feedback capacitors from base to emitter and from emitter to ground are smaller than those which are usually used in the Seiler oscillator. However, drift was negligible — about 25 Hz after warmup, and that was measured with a digital counter!

Don't let that 200-pF capacitor in series with the variable capacitor confuse you. It was used to set up the bandwidth range of the variable capacitor.

Another transistorized Seiler oscillator was described by G3BIK, although he mistakenly identified it as a Vackar. This oscillator used a 2N706 and covered the range from 1.8 to 2 MHz (fig. 4). G3BIK

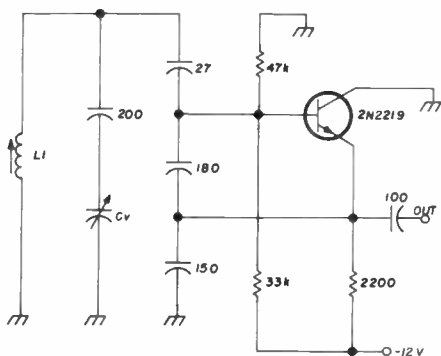


fig. 3. This Seiler oscillator, designed by K9ALD, tunes from 4.95 to 5.6 MHz. Total drift is reported to be 25 Hz. L1 is 2½ turns number 16 on a 1¼" ceramic form. Variable capacitor C_v is a 100-pF variable in parallel with an 82-pF silver mica.

reported exceptional stability with this circuit — a change in voltage from 12 to 6 volts results in a 100-Hz change in frequency. He did experience some difficulty with temperature drift, but cured it by using a high-Q coil and silver-mica capacitors and by putting the complete circuit in an enclosed metal box. This doesn't reflect on the Seiler oscillator though — it's good construction practice with *any* vfo!

Since all the amateurs who have built transistorized Seiler vfos have claimed such extraordinary results, I thought that an fet would make a good thing better. I was right; the results with the circuit shown in fig. 5 were nothing short of remarkable! When the circuit was breadboarded on a piece of Vector board, drift was unmeasurable, even with a fresh spring breeze blowing through the window. When the supply voltage was varied from 22 to 9 volts, total drift was

less than 1 kHz. This could be cured quite easily by putting a zener diode in the circuit.

The total current drain of this circuit is a little over 4 mA, so a couple of 9-volt transistor-radio batteries would power it for many months of operation. The output is constant within 2 dB over the complete tuning range, 3.49 to 4.01 MHz, so it makes an ideal rf driving source. When it's keyed, there is no chirp or drift; it sounds as though it's crystal controlled. It far surpasses any vfo circuit I've ever built, transistor or vacuum tube.

Seiler design

The design of the Seiler circuit closely parallels the design procedure used for the basic Colpitts oscillator. First of all, choose a transistor that has an f_T several times greater than the frequency you're interested in. Then design a bias network which will put the transistor in the linear operating range. Choose a value of tank tuning capacitance (C_T) from the following formula:

$$C_T = Q/6.28fZ$$

Where C_T is the sum of C_v and C_1 (fig. 2); f is the center of the desired frequency range; Z is the impedance of the tank circuit at resonance; and Q is the tuned-circuit Q .

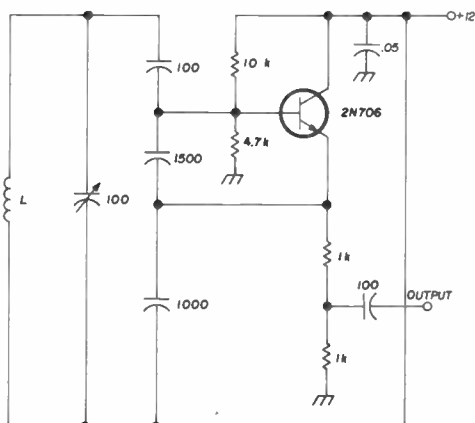


fig. 4. G3BIK's Seiler oscillator covers the frequency range from 1.8 to 2.0 MHz. L1 is 65 turns of number 30 on a 5/8" diameter form.

For maximum power transfer from the transistor, the tuned-circuit impedance should equal the transistor output impedance and may be approximated from:

$$Z = V_{CE} / I_C$$

Where V_{CE} is the voltage between the collector and emitter of the transistor and I_C is the collector current.

Choose a value of Q as high as possible, because oscillator stability is very closely related to tank-circuit Q . For all practical purposes, the Q of the tank will be determined by the inductance you select, so use the best coil you can. If you have lots of room, airwound coils are very good; if you're interested in miniaturization, try a ferrite toroid. In any event, when you're calculating for tuned-circuit capacitance, use a value of Q that is attainable in practice.

After you've calculated the total equivalent tank-circuit capacitance that you need, you can choose the coil to resonate in the center of the desired tuning range.

The values of the two large capacitors in the capacitor divider network, C_2 and C_3 (fig. 2), are not critical. However, they should be quite a bit larger than C_1 or the variable capacitor. Typical values range from 150 pF up to several thousand picofarads, depending on the frequency of interest and the gain of the transistor. The rule of thumb to follow here is to use the largest capacitors that will still result

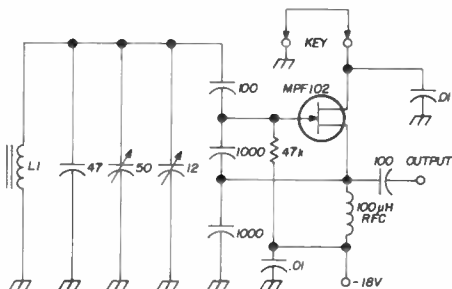


fig. 5. Stable Seiler oscillator using an fet. The tuning range of this circuit is 3.49 to 4.01 MHz. L_1 consists of 44 turns number 30 on a $\frac{1}{2}$ " ferrite core (Amidon T-50-2*).

in oscillation. If a high-gain transistor is used, these two capacitors are usually equal. If a relatively low-gain device is used, it may be necessary to set the ratio of C_3 to C_2 less than the current gain of the transistor.

The variable capacitor, C_v , usually consists of a variable in parallel with a padder. The padder can be adjusted so that the variable will cover the desired

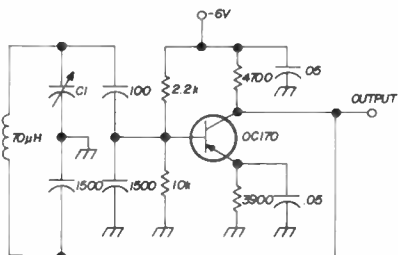


fig. 6. Transistorized Vackar oscillator designed by L. Williams, a British swt. C_1 is a 30-pF trimmer in parallel with a 75-pF air variable.

frequency range. Capacitor C_1 determines the amount of drive to the transistor and is relatively small. The best approach here is to start off with about 100 pF at C_1 and reduce it until the oscillator ceases to function. Add about 50% to this value as a safety factor for the final value of C_1 .

This design method will put you in the right ball park with a working oscillator. All that is left is to set the tuning range of the variable capacitor. This is best accomplished on the bench. First, put in a variable that you think will do the job and measure the frequency with your grid dipper. If the circuit covers the frequency range you want, but the tuning range is too broad, reduce the size of the variable and put in some padding capacitors. If the range is about right, but the center frequency is off, change the size of the inductor. With a few minutes work on the bench, you'll have a stable vfo that tunes just where you want it to.

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607 (formerly Ami-Tron Associates). T-50-2 ferrite cores are 45c each; minimum order, \$1.00. Add 25c for packing and shipping.

the Vackar oscillator

The Vackar circuit was another solution to the same problem — to reduce the load on the tuned circuit. In the Vackar, the transistor is again connected across a relatively low impedance and is very loosely coupled to the tuned circuit. This

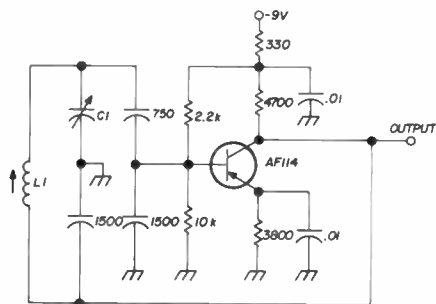


fig. 7. Vackar oscillator design by G5BB for use on 21 MHz. L1 is 19 turns number 22 on a 1/4" form. C1 is a 35-pF air variable in parallel with a 30-pF trimmer.

oscillator will tune over a frequency range of at least 2.5:1; the output can be made absolutely constant, and, according to Jordan, it has the greatest inherent stability of any known oscillator configuration except for a design with independent external load feedback. Those are pretty strong words!

Although the Vackar circuit was originally described in 1949, and publicized, at least in this country, by Clapp in 1954, it has remained virtually unused. W91K described a vacuum-tube Vackar oscillator built by W9TO, and a design appeared in *Radio and TV News*, but that was over ten years ago.

The Vackar oscillator was resurrected when the first transistorized version was published in the *RSGB Bulletin* in July, 1966. This circuit, shown in fig. 6, tunes over the frequency range of 2 to 2.5 MHz. The designer reported the prototype "will stay zero beat with a crystal frequency standard for hours."

This article aroused considerable interest in the Vackar oscillator. G3RAE reported that he modified the circuit

shown in fig. 6 for use as a 465-kHz bfo. He increased the inductance to 460 microhenries and changed the tuning capacitor to 100-pF in parallel with a 270-pF fixed capacitor. All other values were the same as shown in fig. 6.

Shortly thereafter, G5BB described another transistorized Vackar, this one designed for 21 MHz (fig. 7). He experienced some difficulties with temperature drift, but felt they could be cured by putting the circuit in a diecast box. I suspect that replacing the slug-tuned coil with an air-wound inductor would also help.

The latest transistorized Vackar oscillator was described by G. B. Jordan in the February, 1968 issue of *The Electronic Engineer*. He has done a lot of experimental work with the Vackar oscillator and found it to be an extremely stable circuit.

Jordan's circuit, shown in fig. 8, is particularly interesting since it was designed to tune from 26.9 to 34.7 MHz, both the CB and 10-meter bands. The output amplitude varied 1.5 dB over the frequency range, and the temperature drift was linear from +20 to +100°F. When he compensated the circuit with N750 capacitors at C2 and C3, temperature drift dropped to 10 Hz per degree F. Further compensation would reduce drift to negligible amounts.

Since I had such good luck with the fet version of the Seiler oscillator, I tried the same thing with the Vackar (fig.

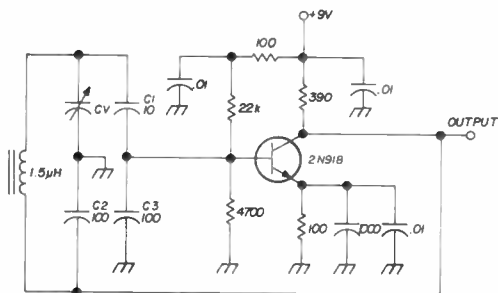


fig. 8. This Vackar oscillator designed by G. B. Jordan tunes from 26.9 to 34.7 MHz. CV is a 12-pF air variable in parallel with a 6.2-pF silver mica.

9) — again, the results were fantastic. Stability was at least as good as the Seiler; drift was negligible, and the keyed note was crystal clear. I went on to add an fet buffer stage, a 2N706 driver and 1-watt 2N697 final. Still no chirps or drift.

Although this circuit was designed to cover the range from 3.5 to 4.0 MHz, by reducing the number of turns on L1, the same basic design could be used as a remote 5-MHz ssb vfo or 8-MHz vfo for vhf use.

Except for output amplitude stability, I could detect *no* difference between the Seiler and Vackar circuits. Perhaps with a counter and a controlled temperature environment, different drift characteristics would be apparent, but in the typical amateur environment, there doesn't seem to be any detectable difference. As far as amplitude stability goes, with the Vackar circuit, the output level changed less than 1 dB over the range from 3.5 to 4 MHz; the Seiler output varied slightly less than 2 dB. This is a pretty small difference.

Vackar oscillator design

As with the Seiler circuit, design of the Vackar is very closely akin to Colpitts design. Since the frequency of oscillation is determined essentially by the value of the variable capacitor and C1, these variable capacitors may be taken as the total tank-tuning capacitance. With this in mind, the tank-tuning capacitance and inductor are chosen by the same method we used for the Seiler circuit. Capacitors C2 and C3 are found from the following formula:

$$C2 = C3 = 3000/f(\text{MHz})$$

According to Jordan, this formula yields about optimum oscillator stability compatible with other requirements. Capacitor C1 is adjusted so that the transistor operates essentially class A and is not driven into cutoff or saturation. In the circuit in fig. 8, with 10 pF at C1, the peak-to-peak voltage at the junction of the variable capacitor and the inductor was 1½ times the B+ supply. This is a

good rule of thumb to go by when you're designing an oscillator of this type.

Most of the authors who have described transistorized Vackar and Seiler vfos have noticed a tendency for these circuits to oscillate at audio frequencies. Since the feedback loop from the collector to the base of the transistor is through the power supply, the base-bias resistors should be decoupled from the collector resistor by a bypassed resistor as shown in fig. 8. Another precaution used

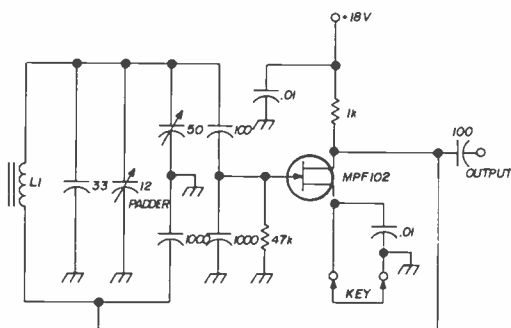


fig. 9. fet version of the Vackar oscillator is extremely stable. L1 is 48 turns number 30 on a ½" ferrite core (Amidon T-50-2).

by Jordan was to bypass the emitter for both audio and rf, although this may not be necessary.

summary

Both the Seiler and Vackar circuits are similar in design and, from my experiences with the fet versions, similar in stability and output. The original tube-type Vackar circuit used high-C tuning whereas Seiler designed for low-C tuning; the high-C was provided by a large trimmer across the main tuning capacitor. There *may* be some advantages to the Vackar circuit for very wide tuning ranges and some advantages to the Seiler when the low-C approach is used, but for amateur vfos I doubt if there is any significant advantage with either circuit. With both of these circuits, stability is independent of the LC ratio, and not very dependent upon the transistor used.

All of the designers of the circuits

Jim Fisk, W1DTY

cure for cranky oscillators

Bill Wildenhein, W8YFB

vfo transistors

When choosing a transistor for a vfo, several characteristics should be considered: power gain, thermal resistance

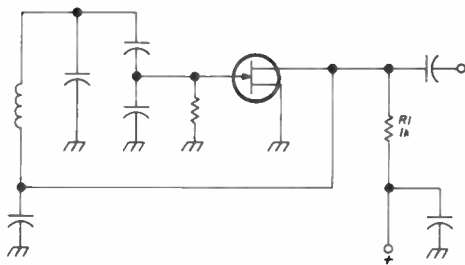


fig. 10. Typical Vackar oscillator using the MPF102 fet. Replacing R1 with a small rf choke ensures oscillation with low drain voltage.

and capacitance. Transistors with high power gain allow loose coupling to the resonant circuit, so transistor changes have less effect on frequency; high f_T ratings imply high power gain.

Thermal resistance characteristics (degrees rise per milliwatt) are important because transistor characteristics and capacitances drift with temperature; less temperature rise means less drift. Lower thermal resistances are found when the collector is tied to the case, so this is often an important consideration.

Some applications require appreciable power output from the vfo, but if you don't need it, low-power operation is preferred. Pick a transistor that has high

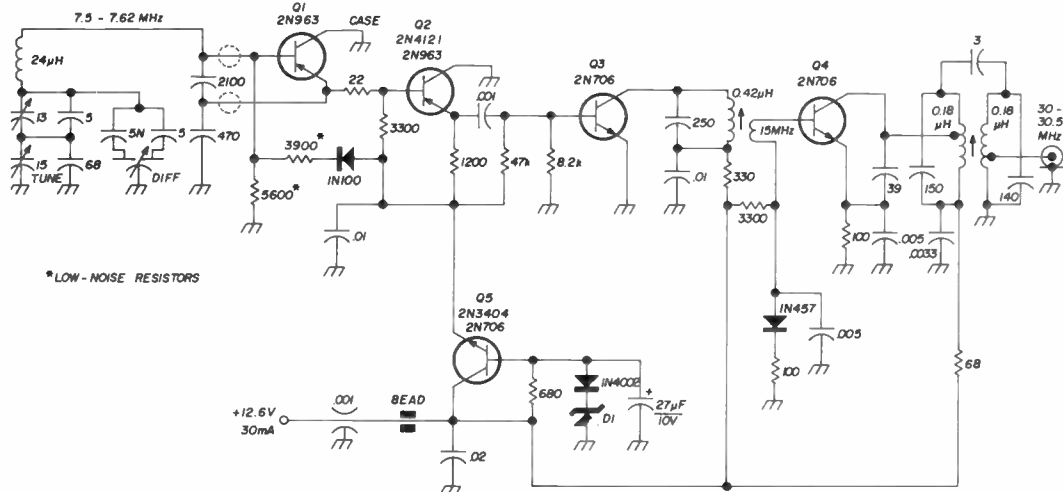


fig. 11. Vfo used by W1COP as part of his two-meter system. Diode D1 is an 8.4 to 9.3 volt reference diode such as the 1N935A or 1N3154; temperature coefficient is important. According to calculations, the power dissipated in the oscillator tuned circuit is less than 3 mW.

table 1. Oscillator transistors listed by figure of merit.

JEDEC Number	Type		f_T (MHz)	P_{in} (mW)	C_{ob} (pF)	Thermal Resistance $^{\circ}/W$	Figure of Merit	Remarks
2N1141	pnp	ge	300	5	2	100	400	class C
plastic 2N918	npn	si	400	10	1.7	200	154	slow heating
2N963	npn	si	300	5	2.5	250	150	low-cost switch
2N700A	pnp	ge	400	5	1.4	600	110*	vhf
2N2368 40405	npn	si	300	10	3.5	150**	79	saturated switch
2N4121	pnp	si	400	10	4.5	250	76	slow drift
2N3866	npn	si	600	50	4	34***	60	class C
2N918	npn	si	400	10	1.7	580	53	vhf
2N706 40404	npn	si	200	10	5	150**	45	saturated switch, some noisy
2N1742	pnp	ge	500	12	2.5	1000	27	slow heating
2N3250	pnp	si	300	50	6	150	16	general purpose

* When biased at 5 V, 1 mA; figure of merit drops to 70 when device is biased to 5 V, 1 mA.

** Use thermal resistance of 200 $^{\circ}/W$ because of polarity.

***Use thermal resistance of 100 $^{\circ}/W$ because of polarity.

gain at low voltage and current; some types such as "forward-agc i-f and rf amplifiers" are poor performers at low voltage. Most transistors that are characterized for class-C operation are good. Computer types are usually good, but silicon "saturated-switch" types may be noisy at low frequencies. Power level for crystal-controlled oscillators is limited by crystal heating; high f_T minimizes this.

Other important considerations in vfo design are transistor size and polarity and supply voltage. Smaller transistor packages are preferred; as far as polarity goes, pnp transistors are preferred for grounded-collector operation in mobile power supplies in American cars. Mobile operation is also an important consideration when discussing power supplies: automotive battery voltage means a maximum of nine volts regulated — well regulated — to the oscillator.

The best way of choosing a semiconductor is to establish a figure of merit for

oscillator transistors. This figure of merit should be proportional to f_T , and inverse as the square root of the output capacitance at the operating voltage, inverse as the thermal resistance and inverse as the dc power input at the operating point. The transistors in table 1 are listed according to this figure of merit.

The vfo transistor in the circuit of fig. 11 is a 2N963, an inexpensive switching transistor with an extremely good oscillator figure of merit. In this circuit I built, emitter voltage on the 2N963 is 5.3 volts and emitter current is 1.1 mA; total input power is 5.8 mW. Since the thermal resistance (junction to case) is 0.25 $^{\circ}C/mW$, the temperature rise of the semiconductor junction in this circuit is 1.46 $^{\circ}C$ maximum. Power dissipated in the tuned circuit is less than 3 mW.

The basic 7.2 to 7.5 MHz output of the vfo drives the doubler stage Q3; the output from Q2 drives another 2N706 doubler, Q4, to 30 MHz. The 30 to 30.5

MHz output is used with a 114-MHz crystal controlled source in a conversion scheme to provide vfo control on two meters.

Hank Cross, W1OOP

crystal-controlled multivibrator

The crystal-controlled oscillator circuit shown in fig. 12 uses no tuned circuits. Since there is no reactance in the circuit

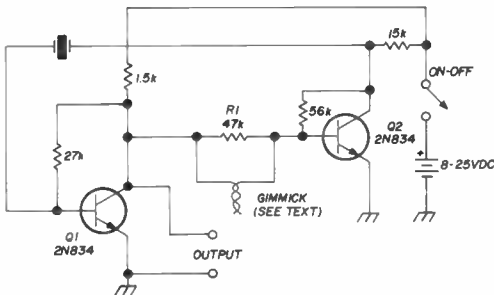


fig. 12. Crystal-controlled multivibrator uses no tuned circuits but will oscillate at any frequency between 2.5 kHz and 5 MHz by simply changing the crystal.

except for the small gimmick capacitor, the circuit will oscillate at any frequency from 2.5 kHz to 15 MHz by simply changing the crystal. Rf output voltage is approximately the same as the supply voltage, which can be from 6 to 25 volts.

I have used this circuit as a wideband frequency spotter, as a marker generator (output is rich in harmonics), and as a crystal activity and stability tester.

When building the oscillator, keep all interconnecting wires as short as possible. Although it is not required in every case, you may need the gimmick capacitor to obtain oscillation. The gimmick is made by twisting together two 2-inch pieces of insulated hookup wire; make sure the two wires don't short together.

Mike Centore, WN2MQY

crystal oscillators

In a previous article, I discussed the general nature of crystal oscillators using

tubes, transistors and field-effect transistors (fets). Since 1966, when that article was written, a great deal has happened in the semiconductor industry that might affect your choice of an active device for use in a crystal oscillator. In the following paragraphs, I've considered all the solid-state devices available at reasonable cost that can be used in crystal oscillator service.

A brief review of their application in conventional circuits is first presented. This is followed by an extensive treatment of these units as used in modified versions of the basic circuits. I've also given some recommendations for certain precautions and design considerations that should be used. If you're solid-state oriented, this article will be invaluable as a reference source the next time you consider a crystal oscillator design.

Prices have dropped, and performance has increased significantly on n-channel junction fets, insulated gate fets, digital integrated circuits and linear integrated circuits. The price reductions appear to

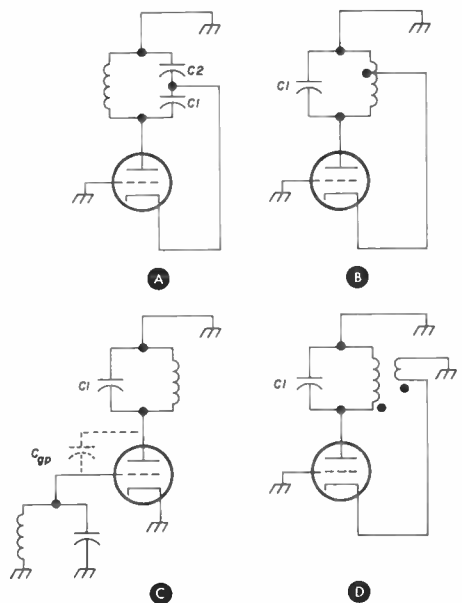


fig. 13. The four basic oscillator circuits: Colpitts (A), Hartley (B), tuned plate-tuned grid (C) and tickler-feedback (D).

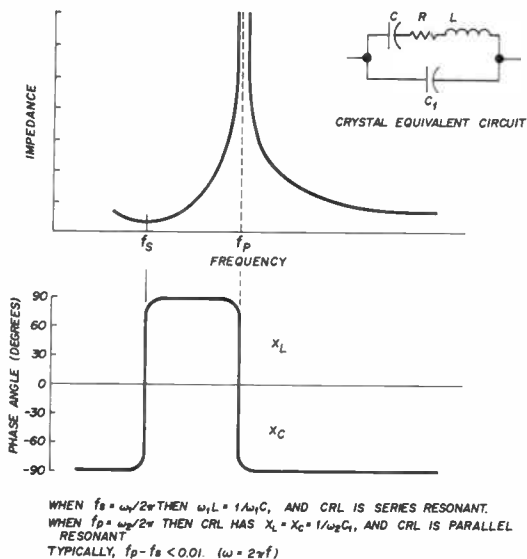


fig. 15. Series and parallel resonance impedance characteristics of a crystal.

be due mainly to lower-cost plastic packaging and the acceptance of these devices by large commercial makers of computers and television sets. The lower costs and the increased performance are just what the doctor ordered for amateur applications.

basic circuits

Four forms of the vacuum-tube oscillator are generally used in amateur

designs. These are the Colpitts, Hartley, tuned-plate-tuned-grid and tickler-feed-back circuits, as shown in fig. 13. They can all be modified to incorporate a crystal, as shown in fig. 14. Note that some of these circuits use the crystal as a series-resonant circuit, and others as an inductance. A single crystal can exhibit either of these characteristics, as shown in fig. 15; but you must account for the fact that *each occurs at a different frequency*. This is one reason for the various "CR" specifications; some crystals are cut for series-resonant and some for near parallel-resonant operation. A listing of military CR specifications is given in table 2.

modified forms

One of the most successful of the

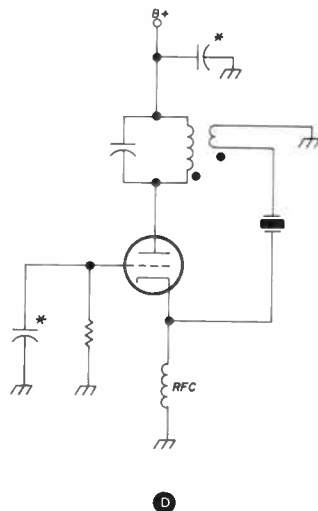
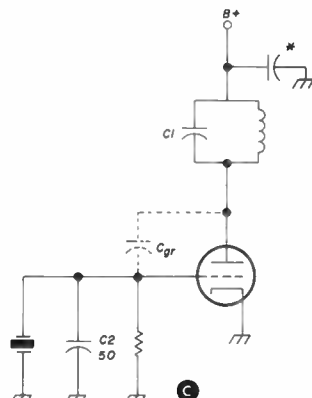
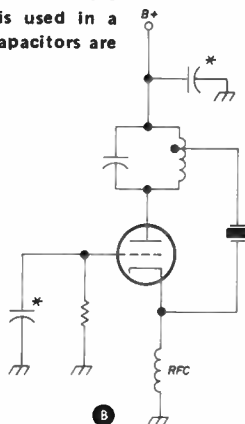
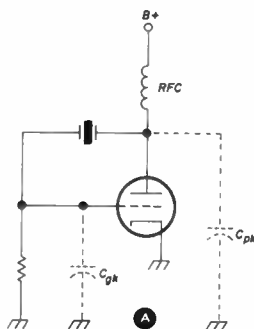


fig. 14. Descendants of the oscillators in fig. 13 using crystals for frequency control. In the Pierce (A) Cpk and Cgk replace C1 and C2 of fig. 13(A). Series-mode crystal is used in the Hartley circuit (B). The Miller oscillator (C) is a version of the tptg. In (D) a series-mode crystal is used in a tickler-feed-back oscillator. Blocking capacitors are denoted by an asterisk.



crystal unit, (military)	military holders	nominal frequency range (kHz)	frequency tolerance (± percent)	resonance	load capacitance (pF)	mode of operation
CR-1(A)/AR	HC-11 or 12/U	2,000-15,000	0.02	Parallel	35.0±0.5	Fundamental
CR-2/U	FT-241A, HC-17/U	200	0.009	Parallel	125.0	Fundamental
	FT-241A, HC-17/U	500	0.010	Parallel	64.0m	Fundamental
CR-3/U	FT-241A, HC-17/U	300-600	0.02	Parallel	—	Fundamental
CR-4/U	FT-241A, HC-17/U	500-1,200	0.02	Parallel	—	Fundamental
CR-5/U	FT-241	2,000-10,000	0.02	Parallel	25.0±0.5	Fundamental
CR-6/U	FT-243	2,000-10,000	0.02	Parallel	12.0	Fundamental
CR-7/U	HC-14/U	3,750-10,000	0.004	Parallel	28.0±0.5	Fundamental
CR-8/U	FT-243	1,000-10,000	0.02	Series	—	Fundamental
CR-9/U	HC-10/U	15,000-50,000	0.01	Series	—	Overtone
CR-10/U	FT-243	5,000	0.005	Parallel	25.0	Fundamental
CR-12/U	FT-243	2,000-10,000	0.02	Parallel	32.0	Fundamental
CR-13	FT-243	455	0.02	Series	—	Fundamental
CR-13/U	FT-243	5,250	0.02	Series	—	Fundamental
CR-14/U	FT-243	2,000-10,000	0.01	Parallel	32.0	Fundamental
CR-15/U	HC-5/U	80-200	0.01	Parallel	32.0±0.5	Fundamental
CR-16/U	HC-5/U	80-200	0.01	Series	—	Fundamental
CR-17/U	HC-10/U	15,000-50,000	0.005	Series	—	Overtone
CR-18/U	HC-6/U	800-16,000	0.005	Parallel	32.0±0.5	Fundamental
CR-19/U	HC-6/U	800-20,000	0.005	Series	—	Fundamental
CR-23/U	HC-6/U	10,000-75,000	0.005	Series	—	Overtone
CR-24/U	HC-10/U	15,000-50,000	0.005	Series	—	Overtone
CR-25/U	HC-6/U	200-500	0.01	Series	—	Fundamental
CR-26/U	HC-6/U	200-500	0.002	Series	—	Fundamental
CR-27/U	HC-6/U	800-15,000	0.002	Parallel	32.0±0.5	Fundamental
CR-28/U	HC-6/U	800-20,000	0.002	Series	—	Fundamental
CR-29/U	HC-5/U	80-200	0.002	Parallel	32.0±0.5	Fundamental
CR-30/U	HC-5/U	80-200	0.002	Series	—	Fundamental
CR-31/U	HC-6/U	1,000-10,000	0.005	Parallel	12.0	Fundamental
CR-32/U	HC-6/U	10,000-75,000	0.002	Series	—	Overtone

table 2. Operating specifications for various military crystals.

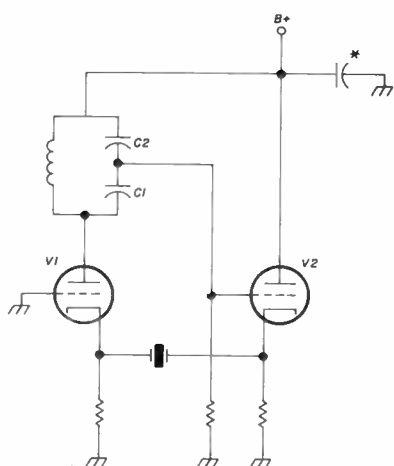


fig. 16. The Butler oscillator — a modified Colpitts for use with series-mode crystals.

vacuum-tube circuits, for use with series-resonant crystals, is shown in fig. 16, the Butler oscillator. It's similar to that in fig. 13A, except that an impedance-lowering cathode-follower has been added, so that the low impedances of the series-resonant crystal and cathode of V_1 can be more easily driven.

Two other types of crystal oscillators somewhat familiar to hams are the negative-resistance oscillator, as exemplified by the transitron and dynatron oscillators using tubes, and the tunnel-diode oscillator of the solid-state world. The general negative-resistance crystal oscillator is shown in fig. 17, and the transitron and dynatron in fig. 18. Two tunnel-diode crystal oscillators are shown in fig. 19.

From fig 17 it might seem that the

crystal unit, (military)	military holders	nominal frequency range (kHz)	frequency tolerance ($\pm$ percent)	resonance	load capacitance (pF)	mode of operation
CR-33/U	HC-6/U	10,000-20,000	0.005	Parallel	32.0 ± 0.5	Overtone
CR-35/U	HC-6/U	800-25,000	0.002	Series	—	Fundamental
CR-36/U	HC-6/U	800-15,000	0.002	Parallel	32.0 ± 0.5	Fundamental
CR-37/U	HC-13/U	90-250	0.02	Parallel	20.0 ± 0.5	Fundamental
CR-38/U	HC-13/U	16-100	0.012	Parallel	20.0 ± 0.5	Fundamental
CR-39/U	HC-15/U	160-330	0.004	Series	—	Fundamental
CR-40/U	HC-15/U	160-330	0.003	Series	—	Fundamental
CR-42/U	HC-13/U	90-250	0.003	Parallel	32.0 ± 0.5	Fundamental
CR-43/U	HC-16/U	80-860	0.035	Parallel	45.0 ± 1.0	Fundamental
CR-44/U	HC-6/U	15,000-20,000	0.002	Parallel	32.0 ± 0.5	Fundamental
CR-45/U	HC-6/U	455	0.02	Series	—	Fundamental
CR-46/U	HC-6/U	200-500	0.01	Parallel	20.0 ± 0.5	Fundamental
CR-47/U	HC-6/U	200-500	0.002	Parallel	20.0 ± 0.5	Fundamental
CR-48/U	HC-6/U	800-3,000	0.0075	Parallel	32.0 ± 0.5	Fundamental
CR-49/U	HC-6/U	800-3,000	0.0075	Parallel	32.0 ± 0.5	Fundamental
CR-50/U	HC-13/U	16-100	0.012	Series	—	Fundamental
CR-51/U	HC-6/U	10,000-61,000	0.005	Series	—	Fvertone (Pressure)
CR-52/U	HC-6/U	10,000-61,000	0.005	Series	—	Overtone (Plated)
CR-53/U	HC-6/U	50,000-87,000	0.005	Series	—	Overtone (Pressure)
CR-54/U	HC-6/U	50,000-87,000	0.005	Series	—	Overtone (Plated)
CR-55/U	HC-18/U	17,000-61,000	0.005	Series	—	Overtone
CR-56/U	HC-18/U	50,000-87,000	0.005	Series	—	Overtone
CR-57/U	HC-6/U	500	0.001	Parallel	32.0 ± 0.5	Fundamental
CR-58/U	HC-17/U	3,000-20,000	0.005	Parallel	32.0 ± 0.5	Fundamental
CR-59/U	HC-18/U	50,000-91,000	0.002	Series	—	Overtone
CR-60/U	HC-18/U	7,000-20,000	0.005	Series	—	Overtone
CR-61/U	HC-18/U	17,000-61,000	0.002	Series	—	Overtone

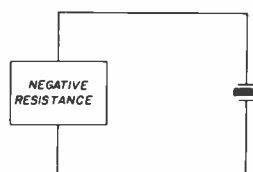
obvious way to build a tunnel-diode crystal oscillator is as shown in fig. 19A. While it may be possible to make such a circuit work, if just the right rf choke can be found, I had no luck with it. The circuit oscillates, but the crystal has no effect other than that of its holder capacity on the frequency. Instead of oscillating at the crystal frequency, the circuit oscillates at the series-resonant frequency of the rf choke and the diode equivalent capacitance.

A more satisfactory crystal oscillator can be built as shown in fig. 19B. At the crystal's series resonant point, the two resistors are effectively in parallel (because the crystal is a "short" between them). The two resistors must be smaller than the absolute magnitude of the tun-

nel-diode negative resistance yet large compared to the equivalent series resistance (at series resonance) of the crystal. Also, L and C are the same impedance as the resistors. That is, at 8775 kHz, L equals $+j51$ ohms and C equals $-j51$ ohms.

Since the equivalent series resistance at series resonance is so important to tunnel-diode crystal oscillators (and to many others as well), the resistance of different

fig. 17. Negative-resistance, or two-terminal oscillator.



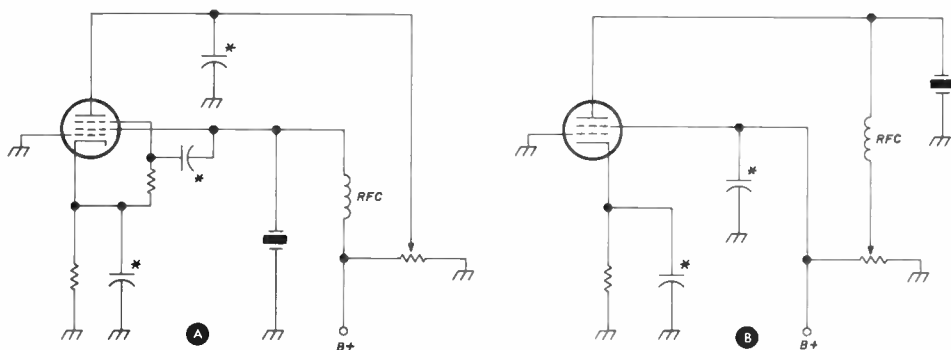


fig. 18. Transistron (A) and dynatron (B) crystal oscillators. Blocking capacitors denoted by an asterisk.

crystals for various frequencies must be known for good design. Such a listing is shown in table 3. Note that these figures were taken from one manufacturer's data sheets; resistances of other manufacturer's crystals will vary from these somewhat. Nevertheless, the table provides a "ball-park" figure upon which design can be started.

the multivibrator

The multivibrator crystal oscillator didn't really come into its own until the advent of bipolar transistors, although it has been built using tubes. It is simply an astable multivibrator with one of its coupling capacitors replaced by a crystal (or with a crystal and series-tuning capacitor). Fig. 20 is representative of such an oscillator. This circuit was used as a one-megahertz crystal calibrator in vhf receivers.

The multivibrator crystal oscillator isn't particularly noteworthy as shown (built with discrete components), but it's the basis for most of the crystal oscillators using digital integrated circuits.

The main use of multivibrator-type crystal oscillators has been for frequencies below 100 kHz, where it's impractical to provide large inductors for conventional oscillators.

using bipolar transistors

The translation of tube-type crystal oscillators to solid-state circuits must be approached with some caution. Too often

a bipolar transistor symbol is just drawn in place of a triode tube symbol. This sort of "engineering" has resulted in circuits like that of fig. 21, which *sometimes* oscillate.

A bipolar transistor is quite unlike a vacuum tube; a *current* into the base controls the collector current in the transistor. In a tube, a *voltage* on the grid controls the plate current. As a result of these different behaviors, the base of a common-emitter bipolar transistor has a much lower input impedance than the grid of a grounded-cathode vacuum tube.

The lower impedances presented by the bipolar transistor make it difficult to

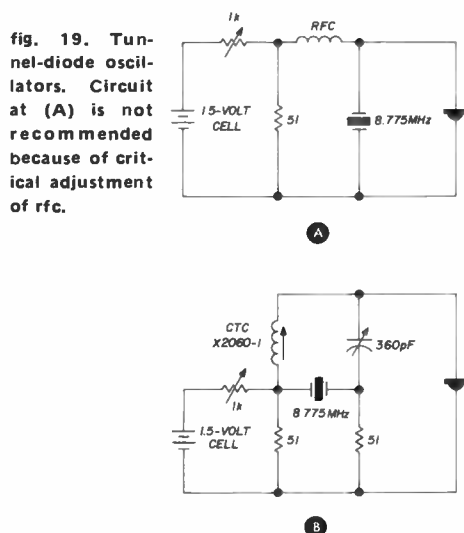


fig. 19. Tunnel-diode oscillators. Circuit at (A) is not recommended because of critical adjustment of rfc.

table 3. Series-resonant resistances of crystals for different frequencies.

Reeves-Hoffman type	Frequency (kHz)	Series resonant Resistance (ohms)	Remarks
J element	1	150,000	+5° X cut
J element	2	75,000	+5° X cut
J element	4	36,000	+5° X cut
J element	8	30,000	+5° X cut
J element	12	15,000	+5° X cut
N element	20	9,000	NU cut
N element	50	10,000	NT cut
E element	100	1,800	+5° X cut
D element	200	950	Y cut
C element	500	1,000	Y cut
A element	1×10^3	300	AT cut (fundamental)
A element	3×10^3	65	AT cut (fundamental)
A element	6.5×10^5	15	AT cut (fundamental)
A element	1.15×10^4	12	AT cut (3rd overtone)
A element	2.08×10^4	8	dT cut (3rd overtone)
A element	4.04×10^4	35	AT cut (3rd overtone)

use in crystal oscillators that use parallel-resonant crystals. However, for use with series-resonant crystals, the bipolar transistor is just great. Fig. 22 shows how series-mode crystals and bipolar transistors can be used to take advantage of the compatible impedances of each.

Note also in the circuits of fig. 22 that a $6.8\text{-}\mu\text{H}$ inductor is placed in parallel with the 26-MHz crystal. This inductor forms a parallel-resonant circuit with the holder capacitance of the crystal. The CR24/U has a holder capacitance of 7

pF. At 26 MHz, approximately $6\text{ }\mu\text{H}$ resonates 7 pF, so the closest standard inductor ($6.8\text{ }\mu\text{H}$) in the Miller 9330 series was used.

Balancing the holder capacitance in this way acts as a simple form of mode filter, assuring that the crystal oscillates as marked. I found this simple method to be adequate even with seventh-mode overtone crystals in the 170-MHz range. The method can be applied to nearly any crystal oscillator where the crystal is to be operated in the series mode. However, the addition of a series-blocking capacitor may be necessary (since the crystal may have blocked dc in the original circuit).

fet crystal oscillators

These devices, on the other hand, are quite similar to tubes because the gate *voltage* controls drain current in the common-source connection. So almost any vacuum-tube crystal oscillator finds its direct equivalent in an fet circuit. Two

fig. 21. Pierce oscillator using a bipolar transistor — not recommended (see text). Blocking capacitor denoted by an asterisk.

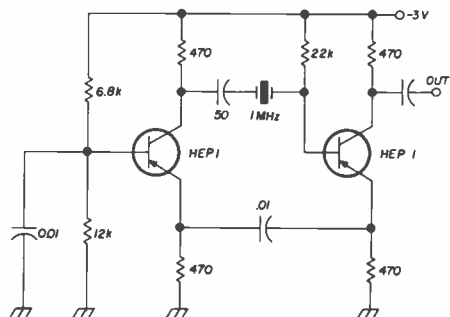
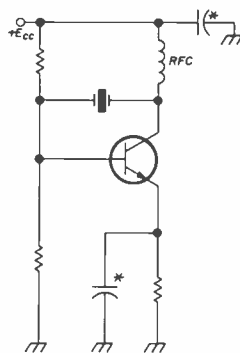


fig. 20. Multivibrator crystal oscillator. The HEP1s can be replaced with 2N1204s.

are illustrated in fig. 23. Notice that both use junction fets. This was done because the diode junction between gate and source conducts (like the grid and cathode of a tube) when forward-biased and allows "gate leak" action. If an insulated-gate field-effect transistor (igfet) is used, a separate diode is usually added across gate and source, as in fig. 24.

Fig. 24 is an ultra-simple Pierce oscillator (a version of the Colpitts), which finds its counterpart vacuum-tube circuit

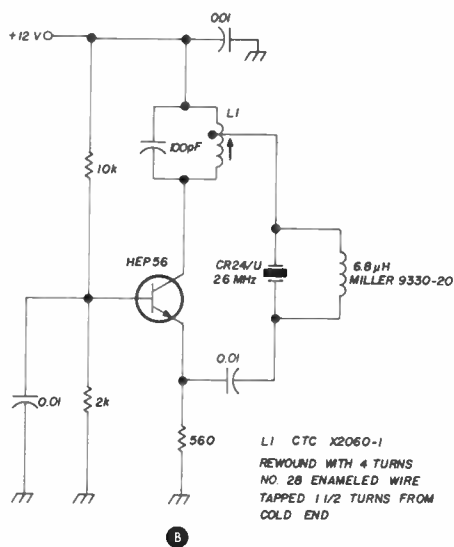
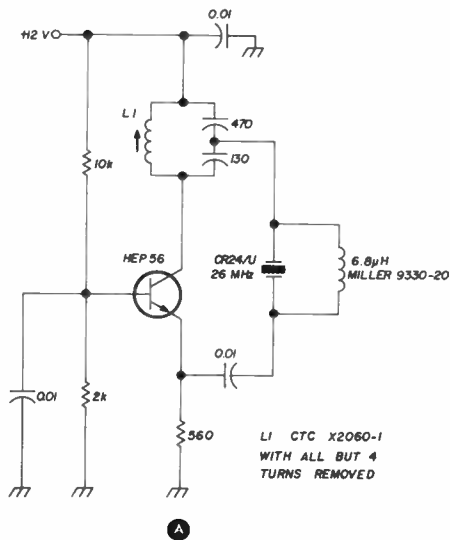
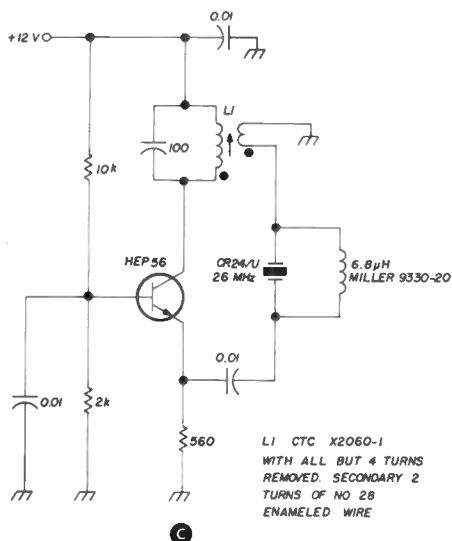


fig. 22. Series-mode crystal oscillators using a bipolar transistor—Colpitts (A), Hartley (B) and tickler-feedback (C).

in fig. 14A. It's useful for checking the general activity of fundamental-mode crystals in the 2- to 20-MHz range. Since fundamental-mode crystals aren't usually cut above 20 MHz, if a higher frequency rock is plugged in (say a 26-MHz, 3rd-mode) it won't oscillate as marked. One indication of oscillation in this circuit is a decrease in drain current, since the nonoscillating circuit draws I_{DSS} until gate-leak action is established.

No circuits using p-channel fets are shown (either junction or insulated-gate types) because rf types haven't become available at the low prices of the n-channel units. The germanium TIXM12 was one of the exceptions (a p-channel jfet, good to over 100 MHz for \$1.07), but it has been discontinued like the TIXM05 we all loved. If and when p-channel fets for rf become available again, the circuits of fig. 23 and 24 can be used by reversing the supply-voltage polarity (and reversing the diode in the circuit of fig. 24).

There are, of course, a number of ways in which npn and pnp bipolar transistors can be combined with n-channel and p-channel jfets and igfets to provide combination circuits. An example is



shown in fig. 25, a Butler oscillator using an npn bipolar transistor and an n-channel jfet.

integrated circuits

In recent years, one of the largest areas of growth in the semiconductor industry has been in integrated circuits. These can be used as crystal oscillators in a number of configurations. Digital ICs, which are the least expensive units, can be made to function as crystal oscillators of the multivibrator-type. Fig. 26 through 29 show the RTL, DTL, TTL, and ECL

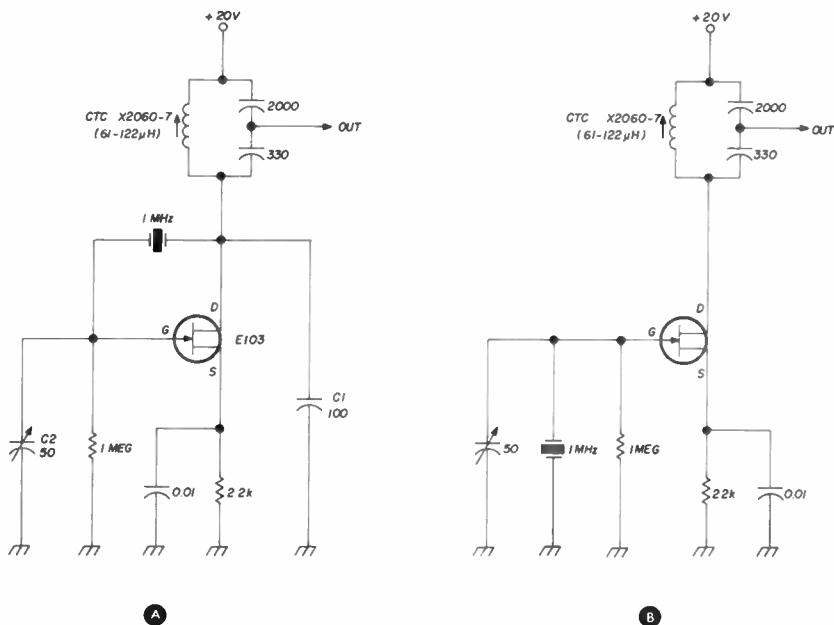


fig. 23. Crystal oscillators using n-channel junction fet's. In (A) crystal performs as an inductor between drain and gate of the fet, with capacitive divider formed by C1 and C2. Circuit (B) is the Miller oscillator, which is useful with crystal having one terminal connected to the can (as in DC-9 octal style).

families as crystal oscillators. Resistor-transistor logic (RTL) is the least expensive family and has been widely used in amateur systems because of its low cost. It is relatively slow, however, and the circuit of fig. 26 can't be expected to work reliably at frequencies higher than a

few megahertz. By using a pair of the higher-power μ L900 buffers, oscillation can be obtained up to about 8 MHz. The circuit of fig. 27 apparently operates at higher frequencies because of lower inherent resistances.

Although RTL is the least expensive, the prices of diode-transistor logic (DTL) and transistor-transistor logic (TTL) have been steadily decreasing, and they're now feasible for ham construction. A cost comparison of several quad-dual gates shows the price per gate to be within reason (table 4).

Fig. 28 shows how DTL and TTL units can be used as a crystal oscillator. Note that the same pin configuration and voltage are used for the two ICs. Also,

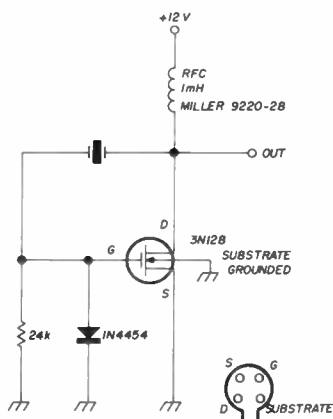


fig. 24. Pierce circuit using n-channel Jfet; useful with fundamental-mode crystals from 2 to 20 MHz.

table 4. Cost comparison of popular integrated circuits.

Logic	Company	Part Number	Cost	Cost Per Gate
RTL	Motorola	MC717P	\$1.08	\$0.27
DTL	Motorola	MC846P	1.65	.41 $\frac{1}{4}$
TTL	TI	SN7400N	2.25	.56 $\frac{1}{4}$
ECL	Motorola	MC1010P	1.80	.45

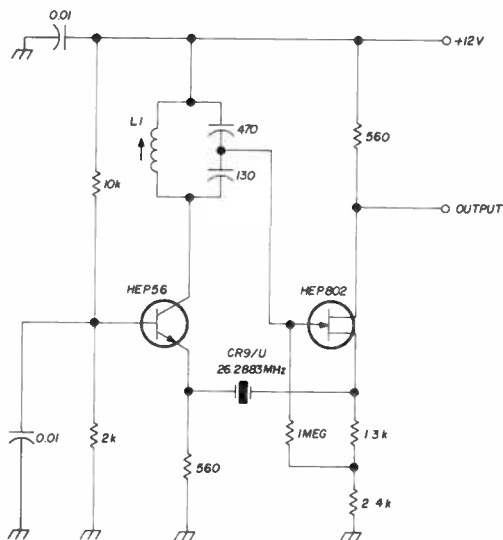


fig. 25. Butler oscillator using npn transistor and n-channel fet. L1 is a CTC X2060-1 slug-tuned coil with all but 4 turns removed.

since only two of the four gates are used, the others can be used as isolation stages or as another crystal oscillator.

One word of caution about use of TTLs however: the pins shown are *only* for the SN7400N. If you try the military version (SN7400 or SN5400), pin connections are quite different!

The emitter-coupled logic (ECL) family is the only one where the internal transistors aren't switched into and out of saturation. This feature makes ECL inherently fast, allowing the MECL I series of Motorola to operate to 30 MHz, and the MECL II series to approach 100 MHz as crystal oscillators. Fig. 29 shows how a MECL I gate can be used as a simple 1-MHz oscillator.

A separate bias driver (MC354G or HEP554) is required with the MECL I series, but it's built in on the MECL II

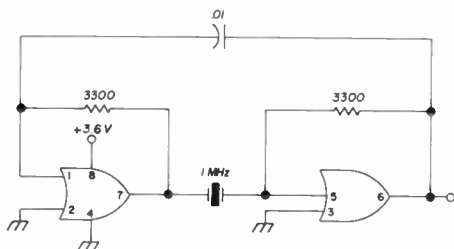


fig. 26. Crystal oscillator using resistor-transistor logic (RTL) integrated circuit.

series. The cost per gate of MECL II, using the MC1010P (quad 2-input gate, at \$1.80) is \$0.45. This compares closely with DTL and TTL units.

the operational amplifier

In linear ICs there is almost limitless variety, but one of the main building blocks is the direct-coupled differential amplifier. Two forms of this monolithic IC have become more or less standard in the semiconductor industry: the operational amplifier (like the Fairchild μ A709 and its descendants) and the rf/i-f amplifier (the Fairchild μ A703 and similar units). Both can be used as crystal oscillators; the operational amplifier is limited to the lower frequencies.

An operational amplifier used as an oscillator brings out many of the basic

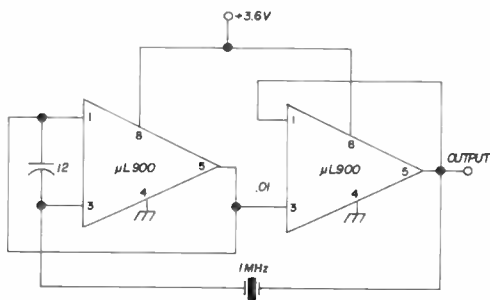


fig. 27. Using RTL buffers in a crystal oscillator circuit.

fundamentals of oscillator design. Because the operational amplifier is such a nearly ideal device, it affords ease of feedback design. Inverting and noninverting inputs are provided on most op-amps, both positive feedback (used to cause oscillation) and negative feedback (used to reduce gain and stabilize output) can be selectively used to produce exactly the

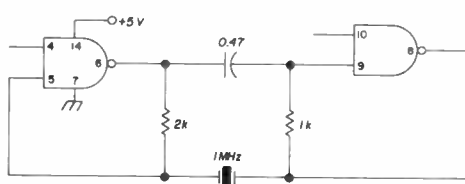


fig. 28. Transistor-transistor logic (TTL) or diode-transistor logic (DTL) gates in a crystal oscillator. The IC is a TI SN7400N or Motorola MC846P.

60-pF padder, is used for setting the frequency range. With these tuned-circuit constants the vfo tunes from 6.93 to 7.24 MHz. K1BBU reports that this circuit exhibits excellent stability and keying characteristics.

The Seiler oscillator circuit shown in **fig. 33** was built by W1DTY to see if the stability of a completely non-temperature-compensated Seiler oscillator was stable enough for 455-kHz ssb generation. Total drift in this circuit, as measured with a frequency counter, was 40 Hz, including warm-up drift. With a 33-pF N750 capacitor in parallel with the inductor, drift was less than 10 Hz from turn on. A 455-kHz crystal oscillator was turned on at the same time for comparison purposes; total drift of this circuit (shown in **fig. 34**) was 4 Hz.

Inductor L1 in the 455-kHz Seiler oscillator consists of 190 turns number 26 enamelled on an Amidon T68-2 form. When the 33-pF temperature compensating capacitor was added to the tank circuit, the 300-pF mica was reduced to 270 pF.

The buffer circuit in **fig. 33** provides excellent isolation for the oscillator and is capable of putting nearly 50 milliwatts of rf into a 50-ohm load.

The simple crystal oscillator circuit shown in **fig. 34** is an excellent circuit for crystals between about 70 kHz and 2 MHz. It presents a 32-pF load for the crystal. The voltage at the collector of the transistor should be approximately one-half the supply voltage and can be ad-

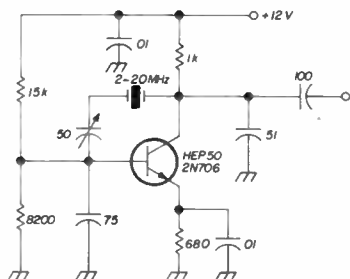
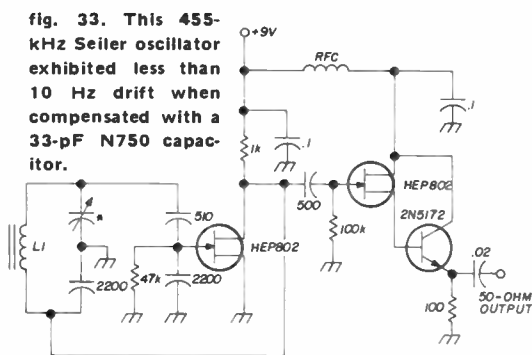


fig. 35. This modified Pierce oscillator is useful for crystals from 2 to 20 MHz. Output should not be loaded with less than 600 ohms.

ⁿ 45 pF PADDER IN PARALLEL WITH 300 pF MICA

justed by changing the value of the 100k base-bias resistor.

A simple crystal oscillator circuit for the frequency range between 2 and 20 MHz is shown in **fig. 35**. This oscillator is a modified Pierce circuit that provides good output and high stability. For proper operation the output of this circuit should work into 600 ohms or more. If the output load is less than 600 ohms,

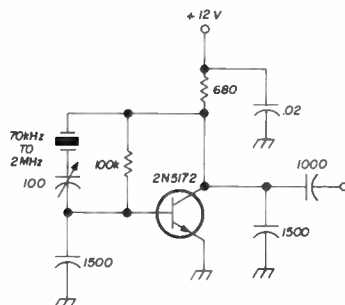
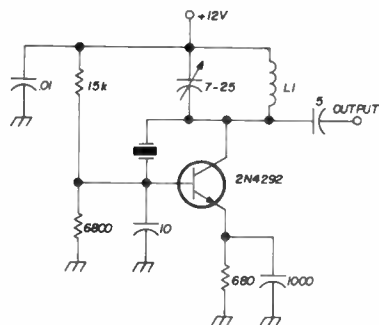


fig. 34. Simple crystal oscillator circuit for crystals from 70 kHz to 2 MHz.

the fet-bipolar buffer circuit shown in **fig. 33** may be used.

overtone oscillator

The simple transistor overtone oscillator shown in **fig. 36** is designed for overtone crystals in the frequency range from 20 MHz to 100 MHz. The frequency of operation is determined by the tuned circuit. Capacitor C1 is a 25-pF trimmer with a minimum capacitance of 7 pF or less. Inductor L1 is chosen for the desired frequency as shown in the table in **fig. 36**.



frequency (MHz)

L1

20-35	2.4 μ H (Miller 4606)
35-60	.68 μ H (Miller 4590)
60-100	.22 μ H (Miller 4584)

fig. 36. Simple overtone oscillator covers the range from 20 MHz to 100 MHz.

Most npn transistors with an f_T of 200 MHz or more will work in this circuit. In operation, the tuned circuit is adjusted to the overtone frequency of the crystal.

Jim Fisk, W1DTY

overtone oscillator

Here is a reliable overtone oscillator using a field-effect transistor that may be

used with crystals from 8 to 60 MHz (see fig. 37). Inductor L2 and the 22-pF capacitor in the source are tuned to approximately 60% of the crystal frequency. The drain tank (C1 and L1) are tuned to the third, fifth or seventh overtone of the crystal. Although the schematic shows a link-coupled output, a small capacitor connected to the drain may be used to couple rf out.

George Tillotsen, W5UQS

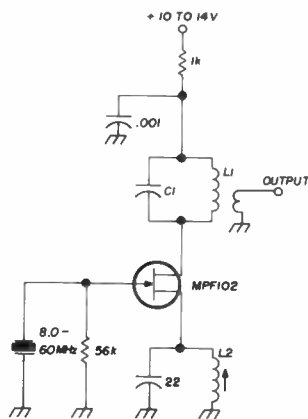


fig. 37. Overtone oscillator circuit.

power supplies

solid-state power supplies

The rectifier-filter section of the basic solid-state power supply is ordinarily one of the five familiar types shown in fig. 1. Of these, I don't generally use any but the full-wave, bridge or conventional doubler. I favor these because they have much better regulation and ripple reduction. Ripple reduction is better because the ripple frequency is twice that of the line frequency. Compared to a half-wave circuit, the chokes and capacitors in a full-wave rectifier effectively have twice the inductive reactance and one-half the capacitive reactance respectively.

Another convincing argument in favor of the full-wave and bridge circuits is that either choke input or capacitive input may be used. Thus, for any given transformer, two dc output voltages are available in the bridge connection, and two more are available in the full-wave circuit.

I use the half-wave types (half-wave and cascade doubler, fig. 1) only when it's necessary to ground one side of the transformer secondary. Such a requirement usually occurs when adding on to existing equipment.

series voltage regulators

The simplest voltage regulator that uses gain is the emitter-follower type, fig. 2. The transistor current gain, h_{fe} , allows this series regulator to regulate much more current than the zener by itself. Another important advantage is that the circuit also functions as a capacitance multiplier. The base voltage is filtered by

R1-C1 in addition to any filtering ahead of the regulator. Since it is an emitter-follower, the emitter "follows" the base, so the output is as well filtered as the base.

Where did the ripple go? It was soaked up by the collector-to-base potential, which varies at the ripple-frequency rate.

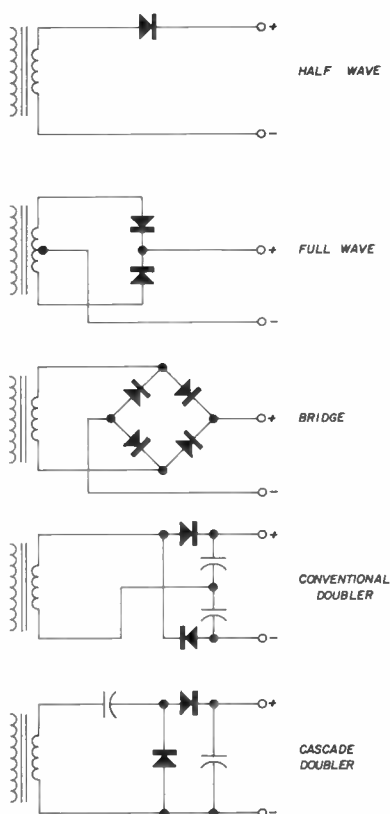


fig. 1. Commonly used rectifier circuits. Full-wave versions provide best regulation and ripple attenuation.

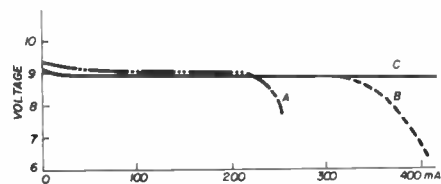
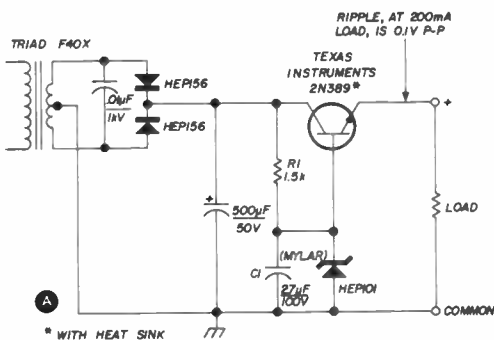


fig. 3. Series regulators with "capacitance multiplication." The Darlington Pair is shown in C; higher zener impedance causes ripple voltage to be the same as in B.

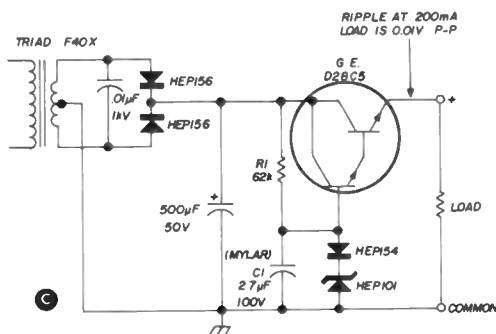
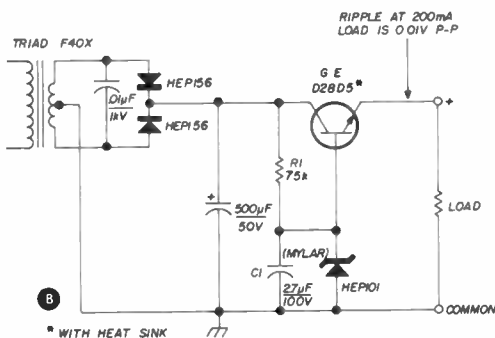
Note that enough voltage must be across the collector-base junction so that negative swings of the ripple voltage don't go below the zener-regulated base voltage.

transistor current gain

The transistor current gain is important in this type of regulator as well as in the others to be discussed. The higher the current gain, the less base current necessary for any given regulated output current. The higher h_{fe} is, the larger R1 can be. This results in better RC filtering (by capacitance multiplication). Fig. 3 illustrates this with three similar emitter-follower regulators, their respective peak-to-peak ripple output, and output voltage as a function of current. All have the same rectifier-filter sections and load resistances.

the darlington pair

Fig. 3C uses a trick called the Darlington pair: two transistors are connected as one to obtain the product of their current gains as the combined h_{fe} . This may seem unfair while we're making comparisons,



but *this* Darlington pair is in one package and is used just like any other transistor. Of course, separate transistors can also be used similarly, but the common-chip types have a thermal-tracking advantage. Note that there are two emitter-base diode voltage drops between the zener and the output, so a higher-voltage zener must be used.

It's also possible to use another form of two-transistor combination in the emitter-follower regulator. Fig. 5 shows such an arrangement; note that complementary transistors are used. Unfortunately, no single-package npn-pnp transistor is yet available for this circuit, so

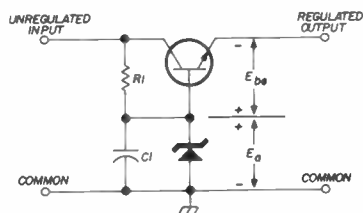


fig. 2. Emitter-follower series regulator. Transistor current gain gives more current regulation than the zener alone.

separate transistors were used. The advantage of this "compound emitter follower" over the Darlington Pair is that it has only *one* base-emitter diode voltage drop between control base and output.

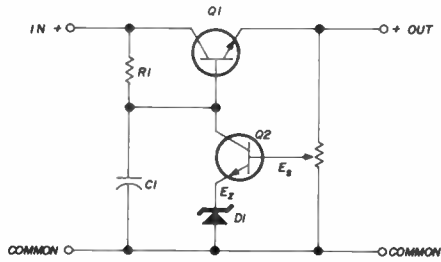


fig. 4. Reference amplifier regulator, allowing adjustable output voltage. R1 is collector load resistance for Q2, a dc amplifier.

adjustable output voltage

The emitter-follower regulator, with its simplicity, is a useful regulating circuit, but it has a troublesome characteristic. The output voltage is fixed at $E_z - E_{be}$ (the zener breakdown voltage less the transistor base-to-emitter diode forward voltage drop).

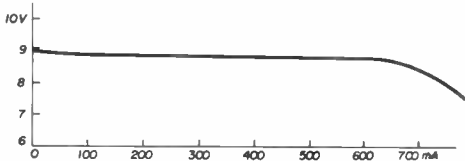
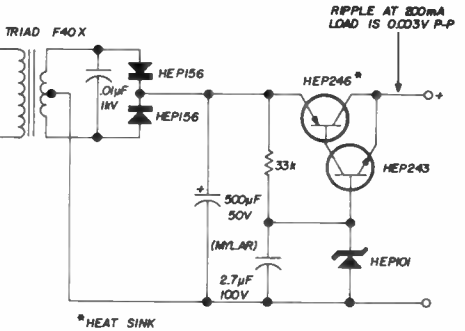


fig. 5. The "compound emitter-follower" regulator. Compared to the Darlington Pair, it has only one base-emitter diode voltage drop between base and output.

By making the circuit a bit more complex, output voltage can be made adjustable. Fig. 4 shows the more sophisticated circuit. Now, R becomes a collector load resistance for Q2, a dc amplifier. Q2 amplifies the difference between E_z , the zener voltage and E_s , a fixed fraction of the regulated output voltage

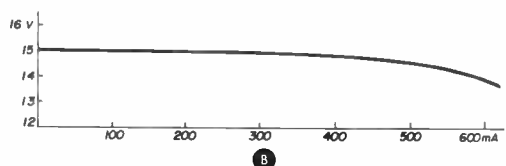
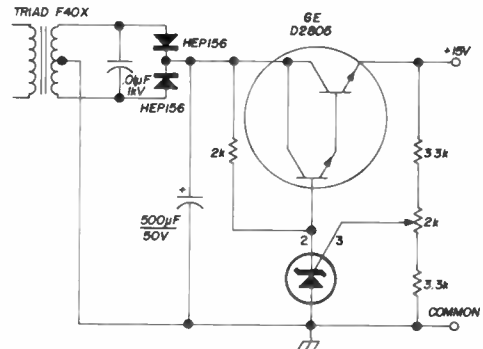
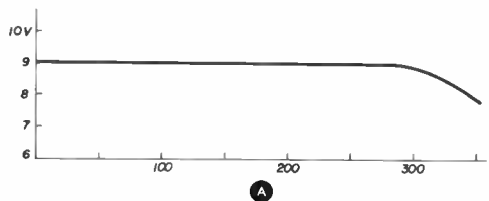
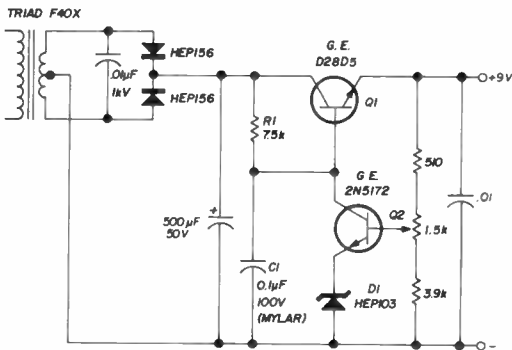


fig. 6. Practical circuits using reference amplifier principle. The new G. E. "programmable zener," a small-scale IC, is shown in B.

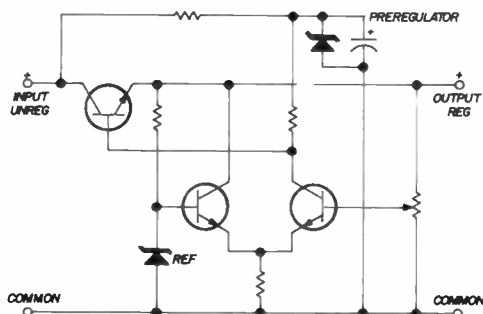


fig. 7. Differential dc amplifier regulator. Zener reference runs at constant low current, improving reference stability.

(error voltage). This form of voltage regulator is widely used because it is simple and adjustable. Because of this wide use, several companies have offered integrated packages containing Q2 and D1. An actual circuit is shown in fig. 6A for a "handle on reality." Note that in such circuits C1 is not made large for capacitance multiplication, but the added gain of the circuit is relied upon for ripple attenuation.

the programmable zener regulator

An interesting new device from General Electric is a monolithically con-

structed regulator called a "programmable zener." A circuit using it, fig. 6B, is similar to that of fig. 6A. Note its new circuit symbol. (I assume that the zener in this IC is an emitter-base breakdown diode.) The circuit using the programmable zener regulator shows good regulation to well over one-half ampere of load current.

differential dc amplifier regulator

By adding one or more transistors to our regulator, even better regulation can be obtained. Fig. 7 shows how a differential dc amplifier allows the zener reference to run at a constant low current, improving reference stability.

The improvement offered by this differential amplifier is especially noticeable if its main function is to compensate for variations in the input voltage. A typical case would be where the input is from a 12-volt line in an auto electrical system. Fig. 8 shows a practical regulated supply using the differential amplifier. The differential amplifier IC is one of the most practical monolithic circuits.

fet regulator

Still another interesting development on the power supply scene is the use of

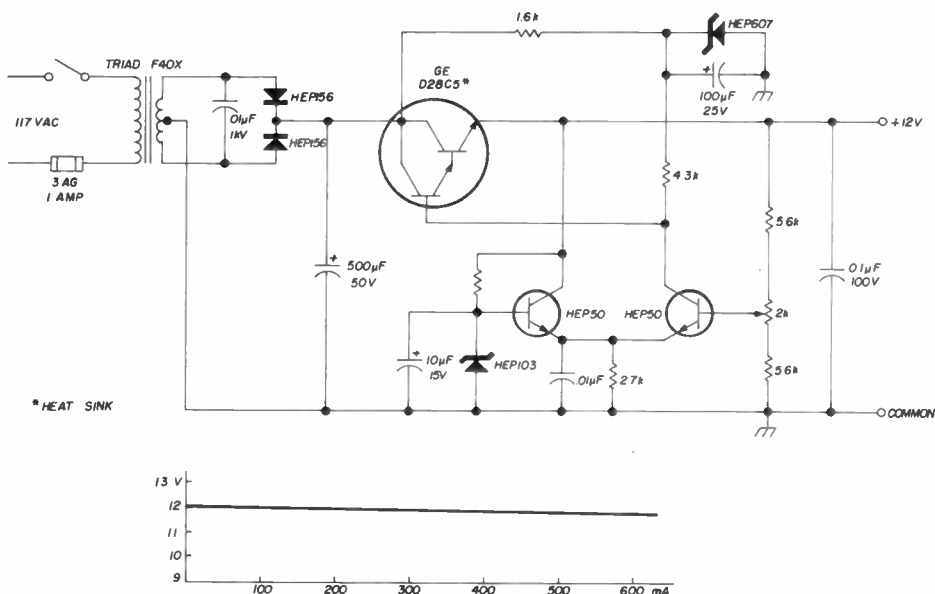


fig. 8. Practical application of the dc differential amplifier.

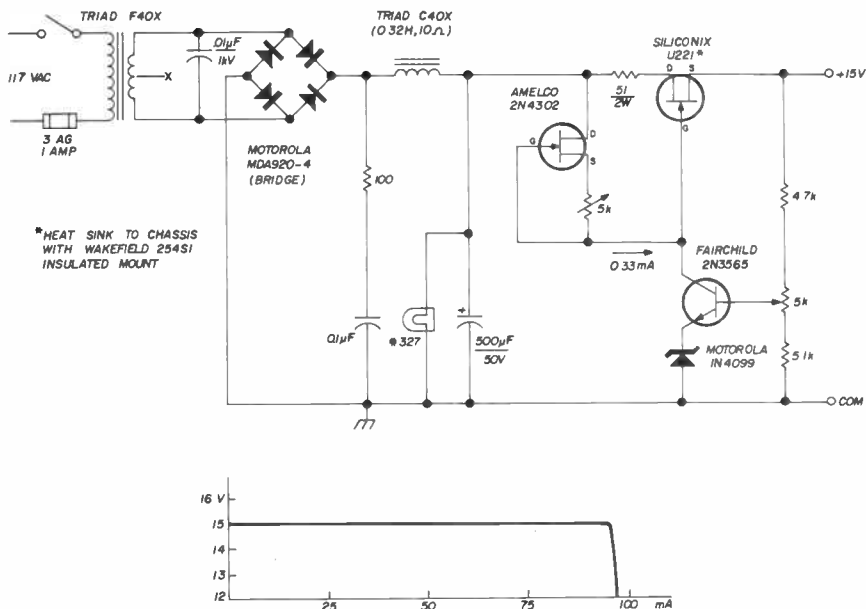


fig. 9. Series-regulated supply using power fets. The second fet replaces the usual resistor load as a constant-current source allowing higher voltage gain.

power fets as series regulators. This has only recently become economically feasible, because several semiconductor manufacturers are now making power fets, and the prices are beginning to come down.

There are several advantages to using a fet as a series regulator. The first is that the regulator has inherent current limiting, since the fet current is limited to I_{dss} . The second advantage is that "thermal runaway," as experienced with bipolar transistors, is not possible. The third reason is that because of the high impedance of the gate, essentially *no* gate current is required to control the fet current. This makes the fet gate even easier for the dc amplifier to control, and higher dc gain may be used.

The circuit shown in fig. 9 uses a series power fet, a bipolar dc amplifier, and a second fet as a constant-current source. The constant-current source replaces the usual resistor load of the dc amplifier. By using the constant-current source as a load resistor, higher voltage gain can be obtained in the dc amplifier. The 5k adjustable source resistor allows the constant-current flow to be set to the value

giving best temperature stability. This is usually around 0.33 mA. Because rather low current is used in the dc amplifier, a

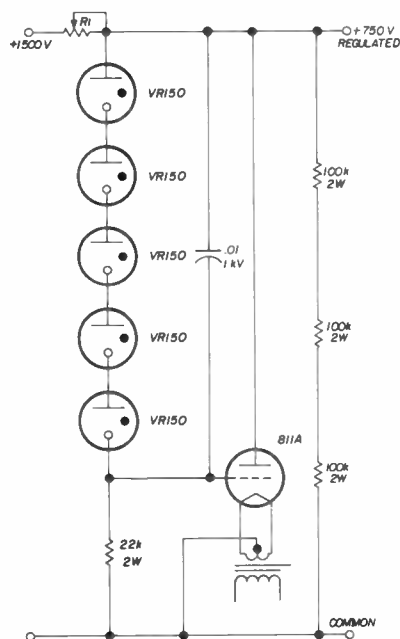


fig. 10. Typical tube-type regulator. The 811A bias, controlled by the VR tubes, changes the voltage across R1 thus regulating the output voltage.

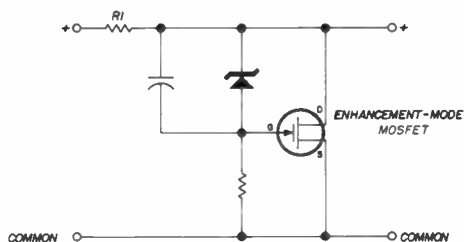


fig. 11. Semiconductor equivalent of the vacuum-tube regulator. (N-channel enhancement-mode fets are not yet commonly available.)

bipolar transistor displaying good h_{fe} at low current (the Fairchild 2N3565) was used. Also, the zener was chosen to have a low impedance at 0.33 mA.

Since the supply of fig. 9 uses choke input, a minimum load is required, which is supplied by the number 327 pilot lamp.

There are a couple of disadvantages to using an fet series regulator, but these are not intrinsic faults. The obvious problem is cost; the less obvious one is loose specification of I_{dss} . The Siliconix U221 and U222 have I_{dss} specifications of about 2 to 1, which are considerably tighter than those of the competition power fets. As power fet technology advances, we can certainly expect to see prices drop and perhaps more predictable I_{dss} from any given off-shelf device.

the shunt regulator

Before discussing the replacement of the discrete differential amplifier with an IC, there's one more circuit we should

look at: the shunt regulator. The vacuum-tube shunt regulator has been in use by hams for many years. Fig. 10 shows a typical HV type. The operation is as follows:

If the +750 V output drops, the drop is directly coupled to the grid of the 811A by the VR tubes, dropping the bias. This causes the 811A to draw less current, thereby decreasing the voltage drop across R1, and raising the +750 V output.

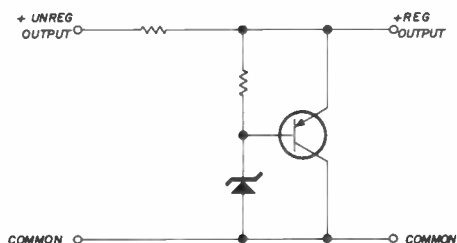


fig. 12. Shunt regulator using bipolar transistor.

A similar shunt regulator can be made using semiconductors; zeners replace VR tubes, and an enhancement-mode fet replaces the zero-bias triode. Such a direct equivalent circuit is shown in fig. 11. It would probably not be built today, because n-channel enhancement-mode fets are not yet commonly available.

A more conventional approach to the shunt regulator is that of fig. 12, using a bipolar transistor. An actual circuit is shown in fig. 13. Another more complex

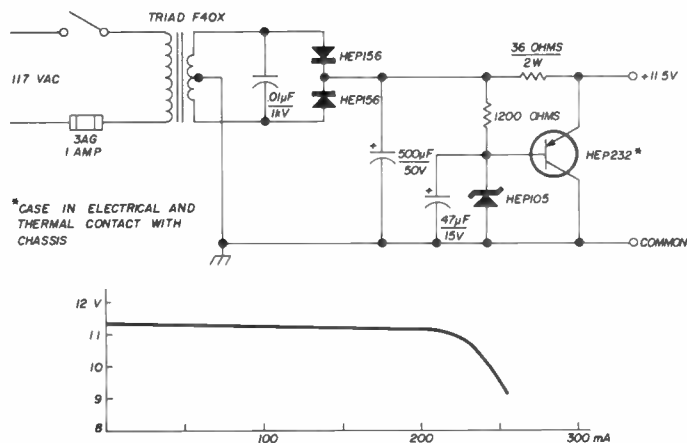


fig. 13. Practical shunt regulator supply using the bipolar transistor.

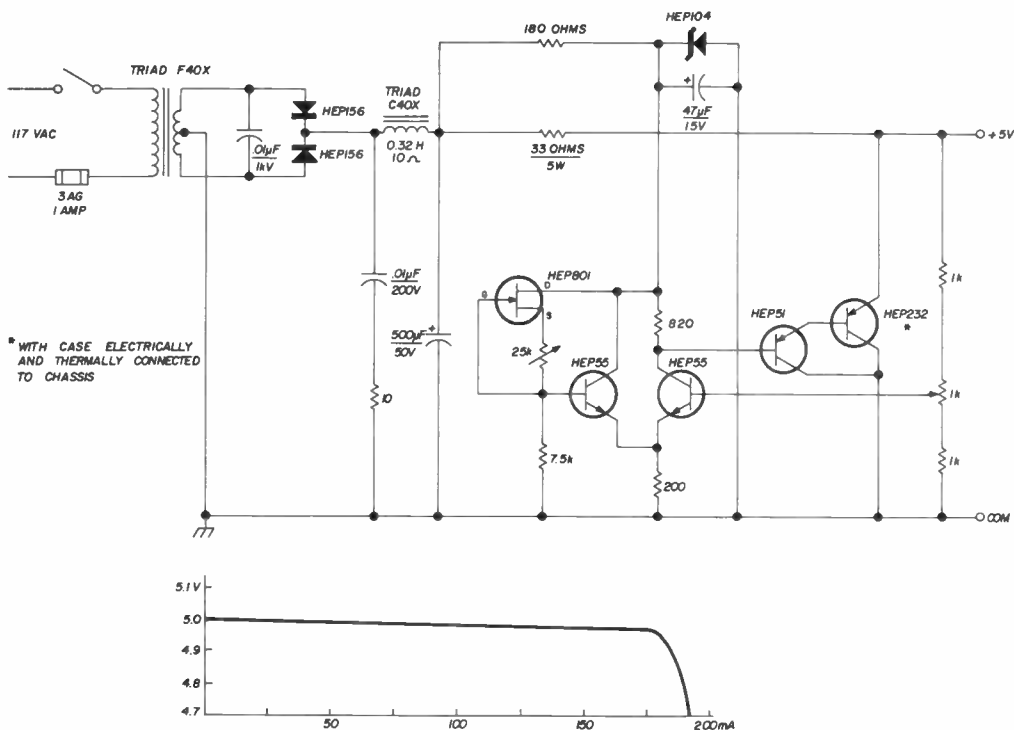


fig. 14. Shunt regulator using a differential amplifier circuit. This was used to power a dozen ICs.

shunt regulator is shown in fig. 14, using a differential amplifier. Note that both of these circuits (fig. 13 and 14) have a convenient feature: the collector (case on most power transistors) is grounded.

The shunt regulator is often overlooked because of its inefficiency when unloaded. However, in many systems the load is nearly constant, and the shunt regulator is used mainly to provide a low power-supply source impedance. The

regulator in fig. 14 was designed to power a system consisting of about a dozen DTL integrated circuits, which had a nearly constant drain. Note also that the shunt regulator is ideal for the choke-input rectifier filter, because it provides a constant load.

ics as voltage regulators

We have seen two beginning steps toward integration of circuits. These are

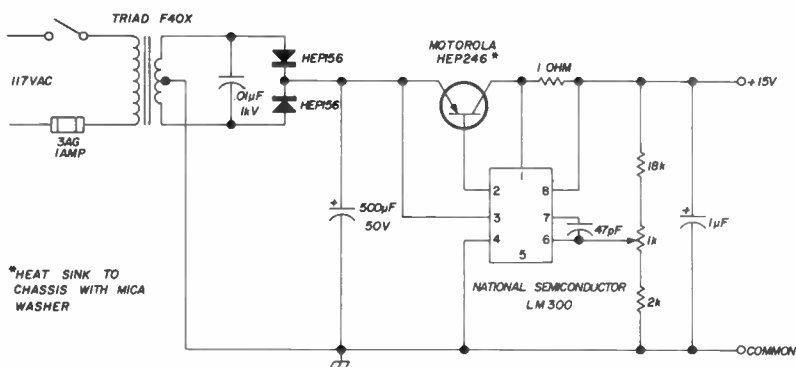


fig. 15. A regulated +15 V supply using an LM300 IC.

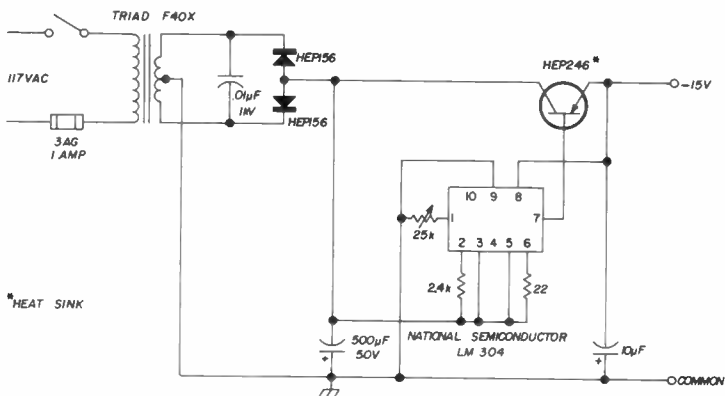


fig. 16. A -15 V supply. This IC is a negative-voltage regulator.

the "ref amp" and the "Darlington pair" as represented by General Electric's RA1 and D28C5. Each of these units contains two semiconductor devices. Modern IC technology has expanded greatly on the number of devices per package. It's now usual for ICs to have dozens, or even hundreds, of devices on one chip.

Among these complex ICs are a few

designed specifically for voltage regulators. National Semiconductor makes the LM300, LM304 and LM305 voltage regulators; Fairchild makes the $\mu A723$ regulator; Motorola makes the MC1460G and MC1460R; and the Continental Devices Corporation makes the CMC513-4. Other companies produce IC voltage regulators, but their prices are more than \$10.00.

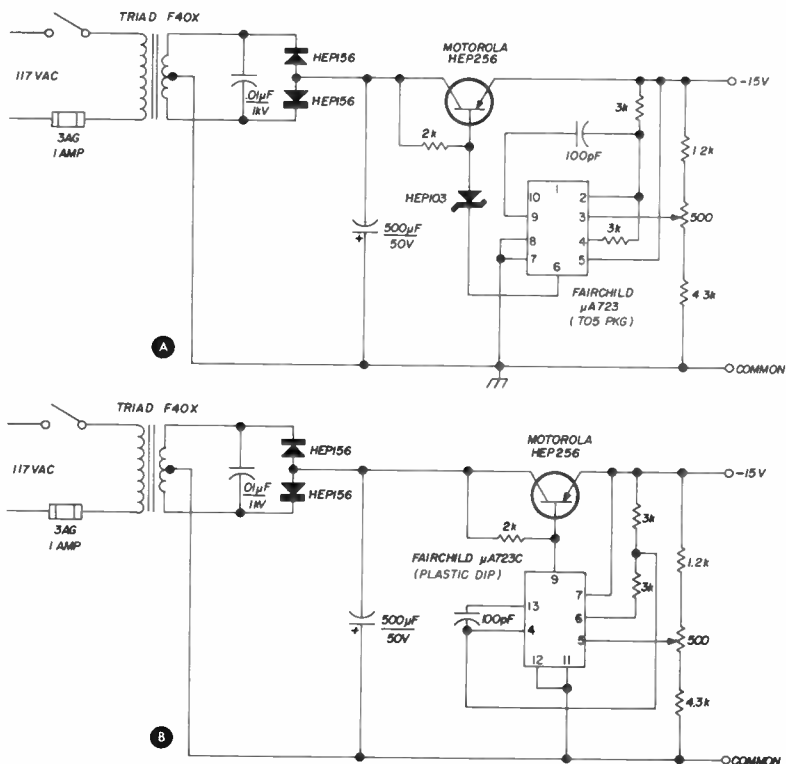


fig. 17. Another application of an IC in a regulated supply. Note that no external zener is needed in B.

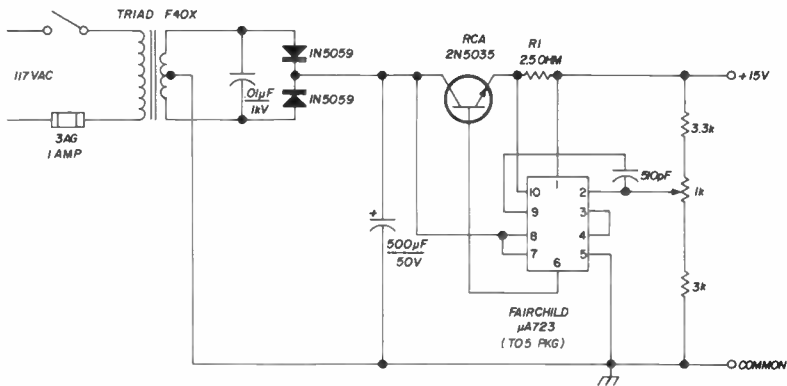


fig. 18. This IC may be used either as a positive-or negative-voltage regulator; here it's shown in a +15 V regulated power supply. R1 determines current limiting. The value shown is for 200 mA.

The LM300 was perhaps the first to hit the market and has been described in a previous article. Its use in a +15 V supply is shown in fig. 15.

Since the original LM300 was first marketed, National Semiconductor has added the LM304 and the LM305. The LM305 is simply an improved LM300 (at a somewhat higher price); it can be plugged into the circuit of fig. 15 to provide improved regulation. The LM304 is a *negative* voltage regulator; its use in a 15 V supply is shown in fig. 16.

Fairchild has entered the integrated regulator field with the $\mu A723$. The $\mu A723$ may be used either as a positive or negative regulator as shown in fig. 17 and 18.

The least expensive IC regulator covered here is a member of a family of five 15-volt regulators, each with the prefix CMC514. The CMC514-4 is the least sophisticated; it is also the only one in the ceramic-epoxy TO5 package (and

therefore the least expensive). The CMC514-4 has only three leads, so it looks for all the world like a 2N3638 — it isn't! Fig. 19 shows it in a simple regulated supply; note that it has only an input, an output, and a ground lead. The

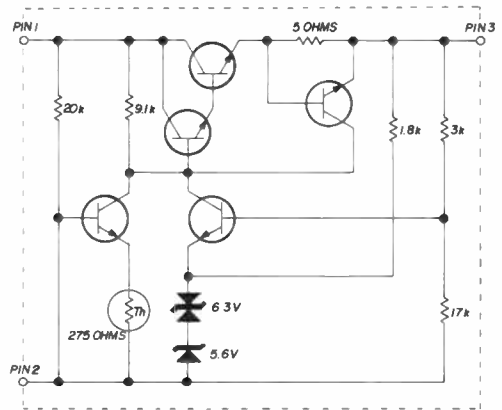


fig. 20. Internal circuit of the CMC 514-4 showing internal thermistor.

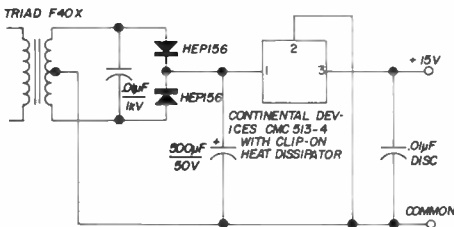


fig. 19. A +15 V regulated supply using the least-expensive IC. It comes in a ceramic-epoxy TO5 package.

CMC514-4 has a built-in thermistor, fig. 20. If too much current is drawn, the chip overheats and the thermistor *turns the regulator off!* When chip temperature drops, the regulator begins regulating again.

Another entry into the regulator IC field is the Motorola MC1460. It comes in a ten-lead TO-5 package (the MC 1460G) and in a small-diamond package (the MC1460R). The R model will, of course,

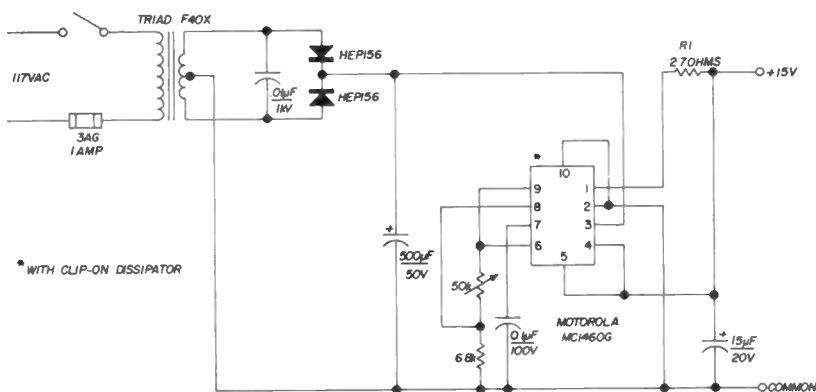


fig. 21. Motorola's MC1460G IC regulator in a +15 V supply has a current-limiting resistor (R1).

handle more current, because it's easier to heat sink. In fig. 21, an MC1460G is used in a +15 V regulated supply.

The MC1460 has a shut-down port (pin 2). In fig. 21, pin 2 is grounded, and so does nothing, but it can be used to turn off the regulator. If pin 2 is made high by a few volts, the regulator shuts down. This high state is low enough so that RTL, DTL or TTL logic levels can be easily used for control. This opens a myriad of possibilities: squelch control of the regulator to conserve power in the output stages of a receiver, shut-down upon demand by a temperature sensor, and so on.

The MC1460 also has a built-in provision for current limiting. The 2.7-ohm resistor in fig. 21 determines at what current the supply will limit. By making this resistor smaller, the regulator will limit at a larger current. A smaller resistor would probably not be used with the MC1460G, but could be used with the MC1460R.

other ics as regulators

We've seen how a number of special ICs can be used (for their intended purpose) as regulators. It's also possible to use others, not sold as regulators, for similar applications. Of course, nearly any

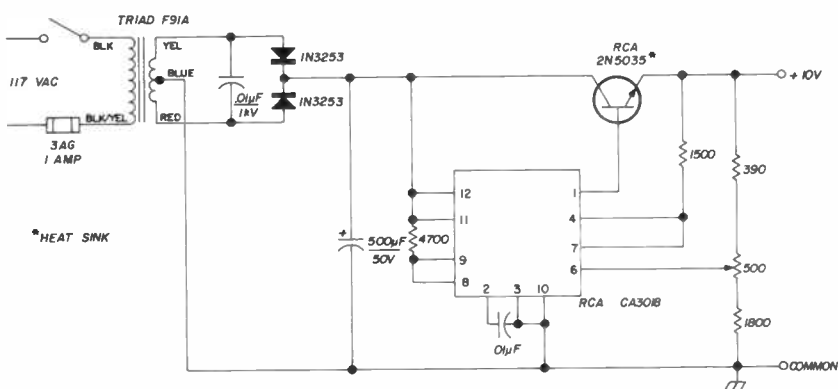
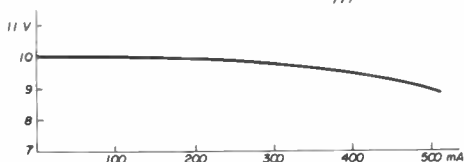


fig. 22. Another inexpensive IC in a voltage-regulator circuit. An emitter-base diode in reverse breakdown provides the reference; an external series pass transistor must be added.



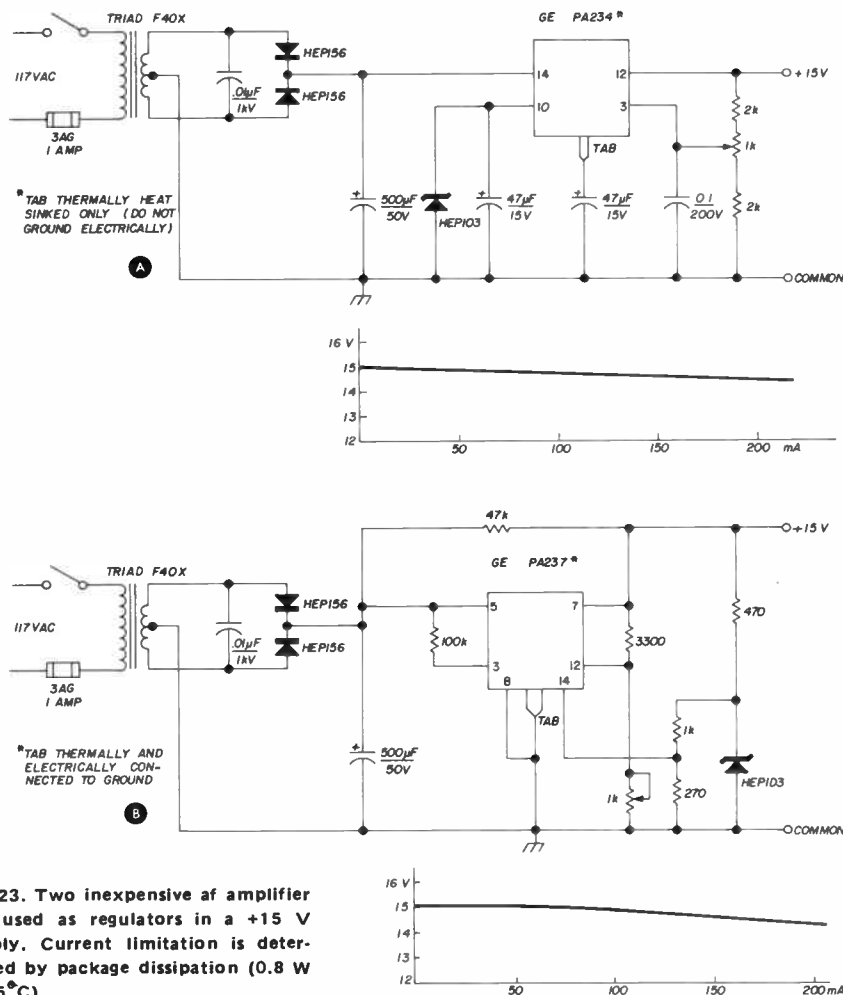


fig. 23. Two inexpensive af amplifier ICs used as regulators in a +15 V supply. Current limitation is determined by package dissipation (0.8 W at 25°C).

operational amplifier can be used as the differential comparator in a regulator, providing that a reference, a series-pass transistor, and a fair number of other external parts are used. This makes a good regulator, but the complexity defeats one of the main reasons for using an IC — simplicity.

Two inexpensive audio amplifiers can be used as regulators. The General Electric PA 234 and PA237 are 1 - and 2-watt audio output amplifiers. Their use in regulated supplies is shown in fig. 23. Their current limitation is set by the package dissipation: 0.8 watt at 25° C.

The RCA CA3018 is another inexpensive IC that can be used as a regulator. In this case, an internal reference is provided

(by an emitter-base diode in reverse breakdown), but an external series pass transistor must be provided. This is shown in fig. 22.

As a final note to the use of ICs as regulators, it is strongly advised that care be used in layout. Since monolithic integrated circuit technology is based on the silicon epitaxial process, ICs almost always have active devices capable of oscillation above 100 MHz. Indiscriminate use of long wire leads, capacitor bypassing, and similar practices will usually make a power supply that can be heard on the local fm set — if the parasitic doesn't eat up the IC first.

Hank Olson, W6GXN

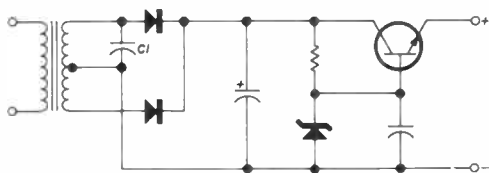
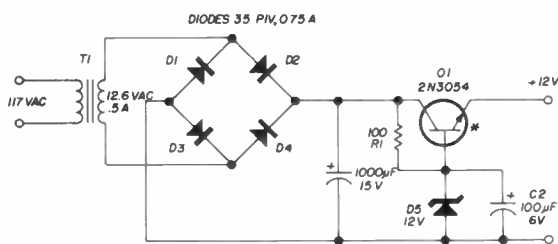


fig. 24. Capacitor C1 reduces power-supply hum.

power-supply hum

The power supply described by Hank Olson (see fig. 3A) was built for a solid-state receiver. The receiver was very quiet until the antenna was connected, then I found a wild example of "tunable hum." A 0.1- μ F disc-ceramic capacitor was connected as shown in fig. 24, and the trouble was cured completely. In my case it was convenient to connect the capacitor across one-half of the secondary, but it is equally effective across the entire secondary winding.

Bill Wildenhein, W8YFB



*MOUNTED ON HEAT SINK

fig. 25. A +12 V regulated power supply using a bridge rectifier and emitter-follower regulator.

regulated 12-volt supply

Most transistor circuits use 12 volts as their power source, and if you like to experiment with these devices (who doesn't these days?), here's a simple supply based on the emitter-follower principle. It will deliver 12 volts at either positive or negative polarity, at a load current of 0.5 ampere. Regulation and ripple attenuation will satisfy most requirements. Its schematic is shown in fig. 25. It has ten components, which works out to about a dollar per component, and you can make the printed circuit board yourself.

The layout of the PC board is the *positive* of the board. If you have a yen for photography, you can make a negative from which the board can be made. Another method is to trace the positive, turn the tracing paper over, and there's the mirror image. This is the actual diagram of the copper side of the printed board. Lay it out with Brady tape or acid resist, then etch away.

For smaller load currents, a smaller transformer can be substituted. Be sure the transformer output voltage is close to 12 volts, because any difference between transformer output voltage and 12 volts will be dissipated across the transistor.

Mike Goldstein, VE3GFN

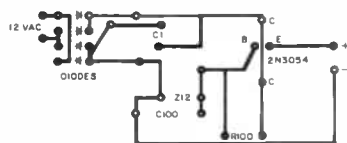


fig. 26. Printed-circuit board for the 12 V regulated power supply shown in fig. 24.

line transient protection

If you're troubled with line transients blowing out solid-state components, fig. 27 shows a simple protective circuit that will tame down the voltage spikes on the line. In this circuit, the two capacitors are charged to the normal repetitive peak reverse voltage; the diodes shunt short-duration transients to the RC filters. The circuit constants shown are for 117-Vac lines. For 220-volt lines, double the voltage ratings of the diodes and capacitors and change the resistors to 200k ohms.

Jim Fisk, W1DTY

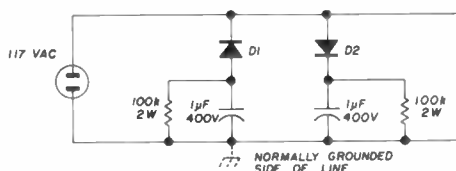


fig. 27. Simple circuit for suppressing voltage transients on the ac line. Diodes D1 and D2 are 1000-PIV silicon diodes.

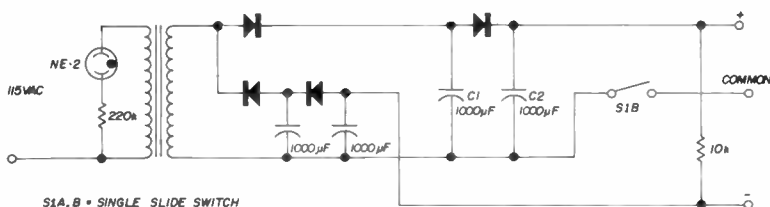


fig. 28. Improved power-supply circuit.

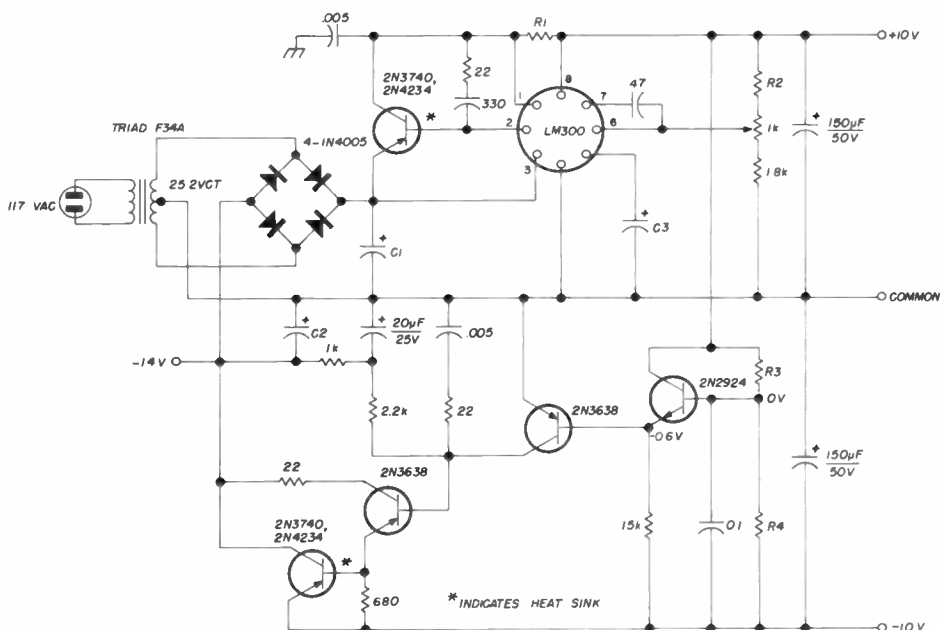
improved power supply

Miniaturized power supplies for small-signal solid-state devices often present special problems, including heat reduction and adequate filtering within a limited space.

If an indicator light is necessary, an incandescent lamp across a secondary winding may be detrimental to the available rectified power output, since the filament often dissipates a sizeable percentage of a small transformer's output. A 1/25-watt NE-2 neon indicator across

the transformer primary will reduce heat as well as providing additional power for the load.

When a filter reactor or resistor is used in a conventional filter, both negative and positive excursions of the ripple voltage at the first capacitor are transferred to a degree to the second filter capacitor, and hence, to the load. Filtering may be improved by replacing the filter reactor or resistor with a garden variety power diode, D1, as shown in fig. 28. In this application D1 performs a gating function, allowing only the positive ripple



C1,C2 value depends on current required. 2000 μ F, 25 V for 425 mA; 2700 μ F, 25 V for 600 mA; 5500 μ F, 25 V for 900 mA (Sprague type 36D or 39D, or Mallory type CG)

C3 zener noise filter; over 10 μ F at more than 3 working volts

R1 sets current limit; 0.22 ohm for 600 mA; 1 ohm for 200 mA

fig. 29. Slaved power supply provides both positive and negative sources with one voltage-regulator IC. Output voltage depends on value of R2: ± 10 V, 600 mA, R2 = 10k; ± 12 V, 500 mA, R2 = 13k; ± 15 V, 250 mA, R2 = 17k.

components to reach C2, resulting in smoother filtering. If a heat-producing series resistor is replaced with the diode, cooler supply operation will result, although ripple will increase.

Also shown in fig. 28 is a simple method for supplying both + and - voltages from one transformer. This arrangement is useful for powering operational amplifiers, cascode circuits and digital devices requiring both polarities.

Gene Brizendine, W4ATE

dual-voltage power supply

Here is a regulated plus-and-minus power supply that uses a National Semiconductor LM-300 integrated-circuit voltage regulator. By changing the values of R2 as shown in fig. 29, the output voltages are ± 10 , ± 12 or ± 15 volts. The transformer charges the input capacitors to nearly 25 volts at light loads, so there is no problem in obtaining 15 volts output. The positive supply is regulated by the LM-300, while the negative supply is "slaved" to the positive one. The slaved supply will follow the positive supply from 4 to 15 volts, but maximum output current is less at higher output voltages.

With +10 volts output at 600 mA, the output resistance is 0.07 ohms; ripple is less than 2-mV rms. The output resistance of the slaved negative supply at -10 volts at 600 mA is 0.01 ohms with less than 2 mV ripple. Line voltage regulation is about 1500:1 with either supply. Current

limiting in the positive supply is controlled by the series-connected 0.22-ohm resistor; a 1-ohm resistor here will current limit the supply to 200 mA.

The input capacitors specified on the schematic are best buys if you're interested in small size and low price. John Meshna, 21 Allerton Street, Lynn, Mass. 01904, has some big ones for a dollar that are very good, but a new Sprague type 36D 5500 μF is inexpensive and about the size of a 2E26.

Hank Cross, W100P

a low-voltage supply

It doesn't take a big power supply with meters, fancy panels and a big transformer for most of the work around the shack. The relatively simple unit shown here is compact, inexpensive, and handles many of my needs, including the broad category of a second supply.

On the premise that one of two voltages will operate most solid-state amateur converters, receivers and transmitters, a two-voltage supply was devised. Either 9 or 15 volts is available and regulated at any current up to 1 ampere. The series pass transistor is a 2N5295 — a low-cost RCA plastic unit, but a TO-66 type, such as the RCA 40310 or 2N3054 is suitable if you can't find a 2N5295 (see fig. 30).

Any small chassis may be used, but my unit was mounted in a 4 x 2¼-inch Mini-box. The power transistor is heat-sunk with insulating washers and thermal compound to the minibox. Wiring is not

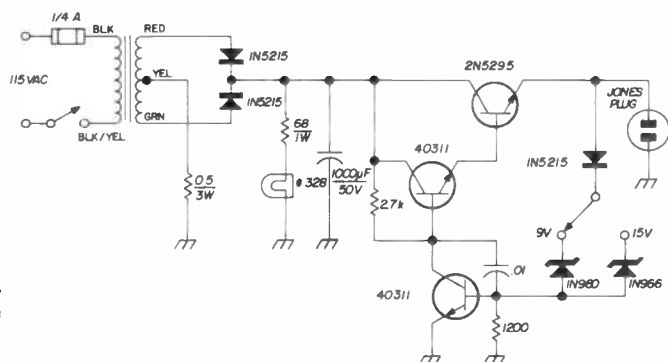


fig. 30. WB2EGZ's low-voltage power supply. The transformer is a Triad 91X or equivalent.